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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

NASA Parts Application Handbook

1. This handbook is approved for use by all elements of the National Aeronautics and Space Administration and is available for use by all departments and agencies of the Department of Defense.
2. Beneficial comments (recommendations, additions, deletions) and any pertinent data which may be of use in improving this document should be addressed to: Manager, NASA Parts Project Office, Goddard Space Flight Center, Greenbelt, Maryland 20771.
3. For user convenience this handbook was structured so that it could be separated into five volumes.

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FOREWORD

This handbook provides a technological baseline for parts used throughout NASA programs. The information included will improve the utilization of the NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List (MIL-STD-975) and provide technical information to improve the selection of parts and their application, and failure analysis on all NASA projects. This handbook consists of five volumes and includes information on all parts presently included in MIL-STD-975.

This handbook (Revision B) succeeds the initial release. Revision A was not released. The content in Revision B has been extensively changed from that in the initial release.

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1. INTRODUCTION

1.1 General.

1.1.1 Application handbook. The NASA Parts Application Handbook (MIL-STD-978) has been prepared to provide a source of technical information for NASA centers and NASA contractors and to maximize standard part usage.

This handbook summarizes current technical knowledge over a broad spectrum of high reliability electrical and electronic component parts. The handbook will not only assist in resolving frequent problems involving component parts but will help avoid such problems by encouraging more knowledgeable part selection and application.

This handbook is an integral part of the NASA Standard Parts Program with MIL-STD-975, the NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List. This handbook should not be used to select specific parts since it may, for information purposes, describe technologies which aren't listed in MIL-STD-975. Specific parts should be selected from those shown in MIL-STD-975.

1.1.2 Objectives. The extensive information in this handbook and MIL-STD-975 should make the following possible:

- a. Improved product reliability and quality
- b. Increased user knowledge for the selection and application of component parts
- c. Improved understanding of component trade-offs
- d. Improved understanding of part design and construction for use when conducting destructive physical analyses or failure analyses
- e. Reduced product cost through increased standardization
- f. Simplified parts procurement system
- g. Simplified logistics and planning
- h. Smaller parts inventory
- i. Uniform incoming inspection routines
- j. Improved understanding and use of the NASA Standard Parts Program.

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1. INTRODUCTION

1.1.3 Handbook organization. This handbook is divided into five volumes. Each volume details specific components as follows:

Volume 1	Introduction Capacitors Resistors and Thermistors
Volume 2	Diodes Transistors Microwave Devices
Volume 3	Microcircuits
Volume 4	Crystals Filters Transformers and Inductors Delay Lines Motors
Volume 5	Connectors, Power Connectors, Radio Frequency Protective Devices Switches Relays Wire and Cable

1.1.4 Special features. This handbook discusses a full range of electrical, electronic, and electromechanical component parts. It provides extensive detailed technical information for each component part. The following list shows some of the subjects covered:

Cost factors	Screening techniques
Conversion factors	Standard parts
Definitions	Environmental considerations
Construction details	Selection criteria
Operating characteristics	Circuit application
Derating	Failure rates
Failure mechanisms	Radiation effects

The handbook is organized so that new part types and additional topics can be easily added. Consistent formats are used to ensure that specific types of information are located in the same place within each section.

The standard format used for each general section (e.g., Capacitor, general) is:

- a. Introduction
- b. Definitions, abbreviations, conversion factors

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- c. NASA standard parts
- d. General device characteristics
- e. General parameter information
- f. General guides and charts
- g. Reliability considerations.

The standard format used for each subsection (e.g., Capacitors, ceramic) is:

- a. Introduction
- b. Usual applications
- c. Physical construction
- d. Military designation
- e. Electrical characteristics
- f. Environmental considerations
- g. Reliability considerations.

1.1.5 Limitations. This handbook was generated to supplement MIL-STD-975 and should not be used for individual part selection. The text often cites individual parts for explanation purposes; in such cases, these parts should not be selected unless they are listed in MIL-STD-975. Some technologies described in this handbook are not included as standard parts in MIL-STD-975. They are included here solely for information.

1.2 NASA Standard Parts Program.

1.2.1 Standard parts program. The NASA Standard Parts Program provides for the selection of standard parts (MIL-STD-975, NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List), defines the guidelines for their use (MIL-HDBK-978 NASA Parts Application Handbook), and establishes policies and direction from the NASA Parts Project Office.

1.2.2 MIL-STD-975. MIL-STD-975 is the standard that is the foundation of the NASA Standard Parts Program. It establishes a list of standard electrical, electronic, and electromechanical parts for use in the selection, procurement, and application for flight and mission-essential ground support equipment. MIL-STD-975 serves the following purposes:

- a. To provide the designer with a list of acceptable parts and the specifications for procuring them

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1. INTRODUCTION

- b. To reduce the quantity of part numbers used in space flight missions and mission-critical ground support applications in order to obtain the benefits of standardization.

Two levels of quality are used in this standard. Grade 2 parts are high quality government-specification-controlled parts for use in noncritical flight and nonmission-essential ground support applications. Grade 1 parts are higher quality government-specification-controlled parts intended for critical flight and mission-essential ground support applications. Parts included in this standard must have application need, technological maturity, and test or usage histories. Such requirements contribute to improved quality and reliability at lower cost with fewer delivery problems.

In addition, MIL-STD-975 includes derating criteria for the different part types. Derating is the reduction of electrical, thermal, and mechanical stresses applied to a part to decrease the degradation rate and prolong the expected life of the part. Derating increases the margin of safety between the operating stress level and the actual failure level for the part and provides added protection from system anomalies unforeseen by the designer. MIL-STD-975 Appendix A contains specific derating conditions.

1.3 Cost.

1.3.1 Cost implication of nonstandard parts. In part selection for a given application, the design engineer considers the suitability of the part for the application. This includes electrical and mechanical characteristics, environmental capability, reliability, availability, purchase cost, and other evident factors. However, various intangibles, particularly in the area of cost, are frequently overlooked or afforded only cursory attention.

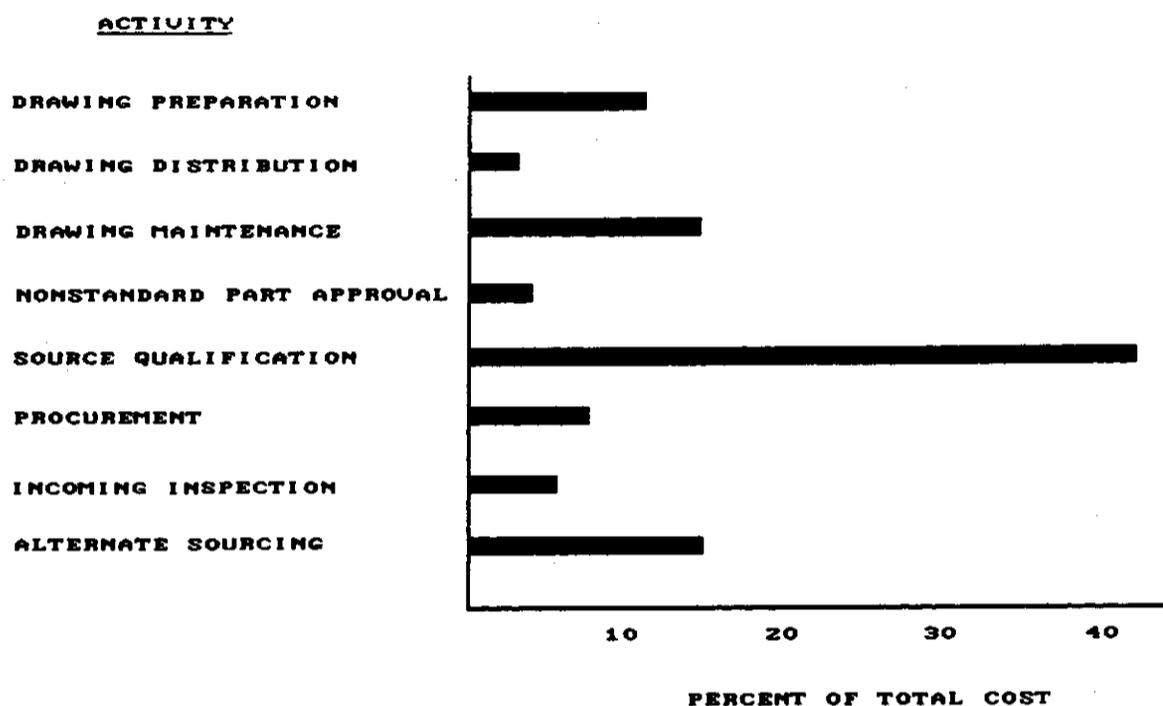
1.3.2 Typical basic costs. Experience has shown that typical costs involved in the specification of a new nonstandard part can range from very low for simple devices to as high as \$50,000 for complex integrated circuits. This includes only the basic costs of introducing a new part into inventory. The contribution of activities involved in the total basic cost is shown in Figure 1.

The relative contribution of each of these activities will vary among the various part types. For example, drawing preparation may be low for resistors but may be 20 times as high for complex integrated circuits. Qualification costs for an initial source can be \$50,000 or more depending on the complexity of the device. Qualification of additional sources, if required, will add considerably to these costs.

1.3.3 Additional costs. Additional considerations for which costs are difficult to estimate, but are still very significant, include:

- a. Stocking costs including handling, storage space, storage facilities, and inventory control

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FIGURE 1. Typical new part activity costs.

- b. Problems entailed by having only a single source which is typical with special or nonstandard parts
- c. Increased cost due to small procurement quantities; this cost is estimated to average an additional 40 percent over the purchase cost of larger quantity standard parts
- d. Problems of schedule slippage, expediting, and decreased vendor response on problems with special or nonstandard parts
- e. Additional failure analysis activity entailed by new, immature, or unproven parts
- f. Cost of equipment repair and replacement of additional component failures entailed by use of unproven parts
- g. Costs of establishing inspection procedures and providing inspection equipment for different part types; inspection costs are also increased because of the larger number of smaller lots of material
- h. Cost of writing and developing programs for inspection of devices on automatic test equipment
- i. Logistic support for maintaining supplies of the new part for field maintenance.

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1.4 Reliability.

1.4.1 Reliability prediction. Reliability is a consideration at all levels of electronics, from materials to operating systems, because materials make up parts, parts compose assemblies, and assemblies are combined in systems of ever increasing complexity and sophistication. Reliability engineering is concerned with the time degradation of materials, physical and electronic measurements, equipment design, processes and system analysis, and synthesis.

The primary information source for reliability prediction is MIL-HDBK-217, "Reliability Prediction of Electronic Equipment." When performing an actual reliability prediction, this handbook should be consulted. It includes formulas and procedures for predicting failure rates of equipment and parts. MIL-HDBK-217 emphasizes the effect of factors such as part count, part quality, environmental stresses, derating, etc. on reliability and provides the basic failure rates for different generic parts. Failure rate data has been compiled from experience and is included as the most complete source of this information. MIL-HDBK-217 includes two methods of reliability prediction, part count analysis and part stress analysis.

1.4.2 Part count analysis. This prediction method is applicable during bid proposal and early design phases. The factors impacting the reliability prediction are part technology, complexity, part count, quality levels, packaging, and application environment. This method provides a basic indication of the system potential to meet reliability goals.

1.4.3 Part stress analysis. This method is applicable when most of the design is completed and a detailed parts list including part stresses is available. It can be used for reliability trade-offs versus part selection.

The quality of the part and the application environment have a direct effect on the part failure rate. The quality levels identified in MIL-STD-975 for standard parts should be used in calculations for part reliability. The environment typically will be space flight or benign ground (for mission-essential ground support equipment).

Other factors which impact reliability predictions are power and current ratings, voltage stress, operating frequency, temperature, matching balance with networks, construction, etc. Microcircuits are treated separately with prediction models for six major classes: digital, linear, microprocessors, memories, hybrids, and converters.

1.4.4 Limitations of reliability predictions. Reliability prediction has at least two practical limitations:

- a. The ability to accumulate data of known validity for new applications
- b. The complexity of the prediction techniques.

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Gathering data to provide statistically valid reliability figures requires effort and diligence. Casual data gathering accumulates data so slowly that a valid level of data may never be reached. When a number of participants gathers data, different methods and conditions are used which prevents exact coordination and correlation of the results. Part reliability data from field use of equipment is difficult to examine due to the lack of suitable data being acquired. The derivation of failure rates is empirically difficult and obtaining valid confidence values is practically precluded due to the lack of correlation.

The failure rates and their associated adjustment factors presented in MIL-HDBK-217 are based upon evaluation and analysis of the best available data at the time of issue of that handbook.

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INTRODUCTION

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2.1 CAPACITORS, GENERAL

2. CAPACITORS

2.1 General.

2.1.1 Introduction. The following sections are intended to help the design engineer select the proper capacitor to fill a particular need. In order to select the proper capacitor, the designer requires not only a description of the device and its specification limits, but also some insight as to its advantages and disadvantages for a given application, peculiarities of construction, mechanical or environmental limitations, reliability, and failure modes or mechanisms.

2.1.1.1 Applicable military specifications. The applicable military specifications are given in table 1 and in the appropriate subsection.

2.1.2 General definitions. This paragraph defines common terms used in the rating and design application of capacitors.

Aging sensitivity. Aging sensitivity is the reduction of the useful life of a device resulting from deterioration mechanisms such as oxidation and wear.

Ambient temperature. The average or mean temperature of the medium (air, gas, liquid, etc.) surrounding a device.

Anode. The positive electrode of a capacitor.

Capacitance. The property of a capacitor which permits the storage of electrical energy when a given voltage is applied. Capacitance is measured in farads, microfarads, or picofarads.

Capacitance tolerance. The maximum deviation (expressed in percent) from the specified nominal value at standard (or stated) environmental conditions.

Capacitive reactance. The resistance to the flow of an alternating or pulsating current by the capacitance, measured in ohms.

Capacitor. An electronic component consisting of two conducting surfaces separated by an insulating (dielectric) material. A capacitor stores electrical energy, blocks the flow of direct current and permits the flow of alternating or pulsating current to a degree dependent on the capacitance and the frequency.

Capacitor, liquid-filled. A capacitor in which a liquid impregnant occupies substantially all of the case volume not required by the capacitor element and its connections. (Space may be allowed for the expansion of the liquid with temperature variations.)

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2.1 CAPACITORS, GENERAL

Capacitor, liquid-impregnated. A capacitor in which a liquid impregnant is predominantly contained within the foil and paper winding, but does not occupy all of the case volume.

Capacitor, temperature-compensating. A capacitor whose capacitance varies with temperature in a known and predictable manner.

Cathode. The negative electrode of a capacitor.

Derating. Derating is the intentional reduction of the stress-vs-strength ratio in an application of the item, for the purpose of extending its operating life.

Dielectric. The insulating material (air, paper, mica, oil, etc) between the plates of a capacitor.

Dielectric absorption. The property of an imperfect dielectric whereby all electrical charges within the body of the material caused by an electric field are not returned to the field.

Dielectric constant. The property of a dielectric material that determines how much electrostatic energy can be stored per unit volume when a unit voltage is applied. (The ratio of the capacitance of a capacitor filled with a given dielectric to that of the same capacitor with a vacuum dielectric.)

Dielectric strength. The maximum voltage that a dielectric material can withstand without rupturing. (The dielectric strength will depend on the thickness of the material and the test method and conditions).

Dissipation Factor (DF). The ratio of resistance to reactance, measured in percent.

Electrolyte. A current-conducting solution (liquid or solid) between two electrodes or plates of a capacitor.

End-of-life design limit. The end-of-life design limit for devices is the expected variation in the electrical parameters of devices for which allowance must be made in circuit design. The parameter variations are expressed as a percentage change from the specified minimum and maximum values.

Equivalent series resistance (ESR). All internal series resistances concentrated or "lumped" at one point in the circuit and treated as one resistance.

Flashpoint of impregnant. The temperature to which the impregnant (liquid or solid) must be heated in order to give off sufficient vapor to form a flammable mixture.

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2.1 CAPACITORS, GENERAL

Impedance (Z). The total resistance to the flow of an alternating or pulsating current, measured in ohms. (Impedance is the vector sum of the resistance and the capacitive reactance; i.e., the complex ratio of voltage to current.)

Impregnant. A substance, usually liquid, used to saturate the paper dielectric and to replace the air between its fibers. (Impregnation increases the dielectric strength and the dielectric constant of the capacitor.)

Insulation resistance (IR). The direct current resistance between two conductors separated by an insulating material. Capacitors are commonly subjected to two insulation resistance tests. One test determines the insulation resistance from terminal to terminal; the other test determines the insulation resistance from one or more terminals to the exterior case or insulation sleeve.

Leakage, dc (DCL). A stray direct current of relatively small value which flows through the capacitor when voltage is impressed across it.

Power factor (PF). The ratio of resistance to impedance, measured in percent.

Quality factor (Q). The ratio of reactance to resistance.

Radio interference. Undesired conducted or radiated electrical disturbances, including transients, which may interfere with the operation of electrical or electronic equipment.

Ripple voltage (or current). The ac component of a unidirectional voltage or current (the ac component is small in comparison with the dc component).

Stability. The ability of a part to resist changes in characteristic values and/or coefficients.

Surge voltage/current. Transient variation in the voltage/current at a point in the circuit. A voltage of large magnitude and short duration caused by a discontinuity in the circuit.

Temperature Coefficient (TC). The change in capacitance per degree change in temperature. It may be positive, negative, or zero and is usually expressed in parts per million per degree centigrade (ppm/°C).

2.1.3 NASA standard parts. See General Section 1.1 for a complete description of the NASA Standard Parts Program. In addition to this handbook, the principal elements of this program include MIL-STD-975(NASA), a standard parts list for NASA equipment.

2.1.4 General device characteristics. The relative size and cost characteristics of the most popular capacitor types are described in Table I. Principal applications are described in Table II.

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TABLE I. Relative size and cost

Dielectric	Applicable Specification	Relative Size For Equiva- $\frac{1}{\text{tent CV Rating}}$	Relative Cost For Equiva- $\frac{1}{\text{tent CV Rating}}$
Ceramic			
Fixed, general purpose } Temperature compensating	MIL-C-123 MIL-C-39014	Small Small	High Very low
Fixed, chip	MIL-C-20 MIL-C-55681	Small Small	Very low Low
Glass	MIL-C-23269	Large	Medium
Mica	MIL-C-39001	Large	Medium low
Paper and plastic			
Metallized plastic	MIL-C-83421	Small	Medium
Metallized paper	MIL-C-39022	Small	Medium
Tantalum electrolytic			
Nonsolid	MIL-C-39006	Very small	High
Solid	MIL-C-39003	Very small	Medium
Solid chip	MIL-C-55365	Very small	Medium

1/ "C" = capacitance, "V" = voltage

TABLE II. Principal applications

Military Specification	Application												
	Established reliability	Capacitor type	Blocking	Buffering	Bypassing	Coupling	Filtering	Tuning	Temperature compensating	Trimming	Motor starting	Timing	Noise suppression
MIL-C-20	X	Ceramic			X	X	X	X	X				
MIL-C-123		Ceramic			X	X	X						
MIL-C-10950	X	Mica			X	X		X		X			
MIL-C-23269	X	Glass	X		X	X		X					
MIL-C-39001	X	Mica	X	X	X	X	X	X				X	
MIL-C-39003	X	Solid Tantalum	X		X	X	X						X
MIL-C-39006	X	Wet Tantalum	X		X	X	X				X		
MIL-C-39014	X	Ceramic			X	X	X						
MIL-C-39022	X	Met. Plastic	X		X		X						
MIL-C-55365	X	Solid Tantalum, Chip			X	X	X						
MIL-C-55681	X	Ceramic, Chip			X	X	X						
MIL-C-83421	X	Met. Plastic	X	X	X	X	X				X		

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2.1 CAPACITORS, GENERAL

2.1.5 General parameter information.

2.1.5.1 Selection. Various factors must be considered when selecting a capacitor type for a particular application. These factors are discussed below, and in the subsection dealing with specific capacitor types.

a. Electrical.

- Capacitance
- Tolerance
- Voltage rating
- AC current-carrying capacity
- Insulation resistance or leakage
- Dissipation factor or equivalent series resistance
- Effects of frequency
- Capacitance change with temperature
- Voltage coefficient
- Dielectric absorption.

b. Mechanical.

- Size
- Terminal configuration
- Mounting.

c. Environmental.

- Operating temperature range
- Moisture resistance
- Shock and vibration
- Altitude
- Radiation.

d. Reliability.

- Derating
- Failure rate
- Failure modes
- Stability
- Operating life.

e. Economic.

- Part cost
- Cost of justifying nonstandard parts
 - Cost of samples
 - Cost of testing
 - Cost of negotiations with customer.

2.1 CAPACITORS, GENERAL

2.1.5.1.1 Important selection factors. The most important factors are discussed below.

Temperature. Temperature can affect capacitance by causing variations in dielectric constant or conductor area and spacing. Temperature can also affect leakage current (through changes in specific resistance), breakdown voltage, current rating, and oil, gas, or electrolyte leakage through seals.

Humidity. Humidity can affect leakage current, breakdown voltage, power factor, or quality factor.

Barometric pressure. Barometric pressure can affect breakdown voltage and oil, gas, or electrolyte leakage through seals.

Applied voltage. Applied voltage can affect leakage current, amount of heating, dielectric breakdown, frequency, corona, and insulation.

Vibration. Vibration can affect capacitance and integrity of the elements, terminals, or case.

Current. Current can affect internal temperature and operational life.

Life. Operating life is affected by all environmental and circuit conditions.

Stability. Stability is affected by all environmental and circuit conditions.

2.1.5.2 Capacitors types and their limitations.

2.1.5.2.1 Ceramic capacitors. There are two major types, NPO and BX. The NPO (negative-positive-zero) has a temperature coefficient that is effectively zero, whereas the BX type may have a capacitance change of +15 percent to -25 percent over temperature range of -55 °C to +125 °C with applied rated dc voltage. In general, the NPO has better characteristics but is larger (because of the low dielectric constant) and more expensive.

Ceramic chip capacitors are brittle and sensitive to thermal shock. Precautions must be taken during mounting to avoid ceramic cracking. The substrate material should have a thermal expansion coefficient that closely matches that of the capacitors. This will help to avoid mechanical stresses that may result from changes in temperature.

The order of this listing does not necessarily imply an order of preference of an individual group.

2.1.5.2.2 Plastic film capacitors. These capacitors offer extremely tight tolerances, very low leakage currents (high insulation resistance), and minimal capacitance changes with temperature (low temperature coefficient). They are especially suited for ac applications since their extremely low dissipation factor limits the I^2R heating loss.

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Plastic film capacitors are limited by their relatively large size and weight and narrow range of available capacitances. Due to the nature of the dielectric, under certain conditions these capacitors may generate voltage transients. The voltage transients occur as a result of clearing action of pin holes in the plastic film, and electrochemical effects which cause spurious, random conduction.

2.1.5.2.3 Tantalum capacitors. The solid tantalum capacitors (both leaded and chips) offer better stability over life and lower capacitance-temperature characteristics than other electrolytic capacitors. The tantalum chips are the smallest of the electrolytic styles.

Solid tantalum capacitors have limited surge-current handling capabilities and high leakage current. In addition, tantalum chips have a relatively narrow range of capacitance values and voltage ratings.

Wet tantalum capacitors are substantially smaller than comparably rated capacitors, except for chips. The wet tantalum devices also have higher surge current ratings, higher ripple current ratings, and lower leakage current than other types of electrolytic capacitors.

Wet tantalum capacitors have limited ability to withstand reverse voltages, and they cost more.

Nonsolid tantalum (tantalum foil) capacitors offer the widest range and highest capacitance values and voltage ratings and also the best ripple current handling capability. The nonpolarized tantalum foil capacitors are the only electrolytic capacitors which can operate continuously on unbiased ac voltages. However, the nonsolid tantalum capacitors cannot be used at frequencies greater than 10 kHz.

Nonsolid tantalum capacitors are limited by their very wide capacitance tolerance and their large size compared to other electrolytic capacitors.

2.1.5.3 Electrical considerations. Specific electrical characteristics may require consideration from a design engineer's standpoint.

Capacitance and Tolerance. The required capacitance value usually limits the capacitor selection. Low capacitance values (up to about 10,000 pF) are provided by glass, mica, and ceramic types. The medium capacitance range (approximately 0.005 to 22.0 μ F) includes paper, plastic, and some ceramic types. The high capacitance range (1 μ F and up) usually is provided by the electrolytic capacitors. These are very general classifications with considerable overlap within the types.

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Within broad limits, the availability of various capacitance tolerances is directly proportional to the absolute capacitance value. Glass and mica types are readily available in tolerances down to $\pm 1\%$ or less; tubular paper or plastic capacitors are normally available in the $\pm 0.5\%$ to $\pm 20\%$ range, while electrolytic capacitors are usually supplied in the range of $\pm 5\%$ and up, depending on the type.

Figure 1 indicates the approximate ranges of the commonly used capacitor types.

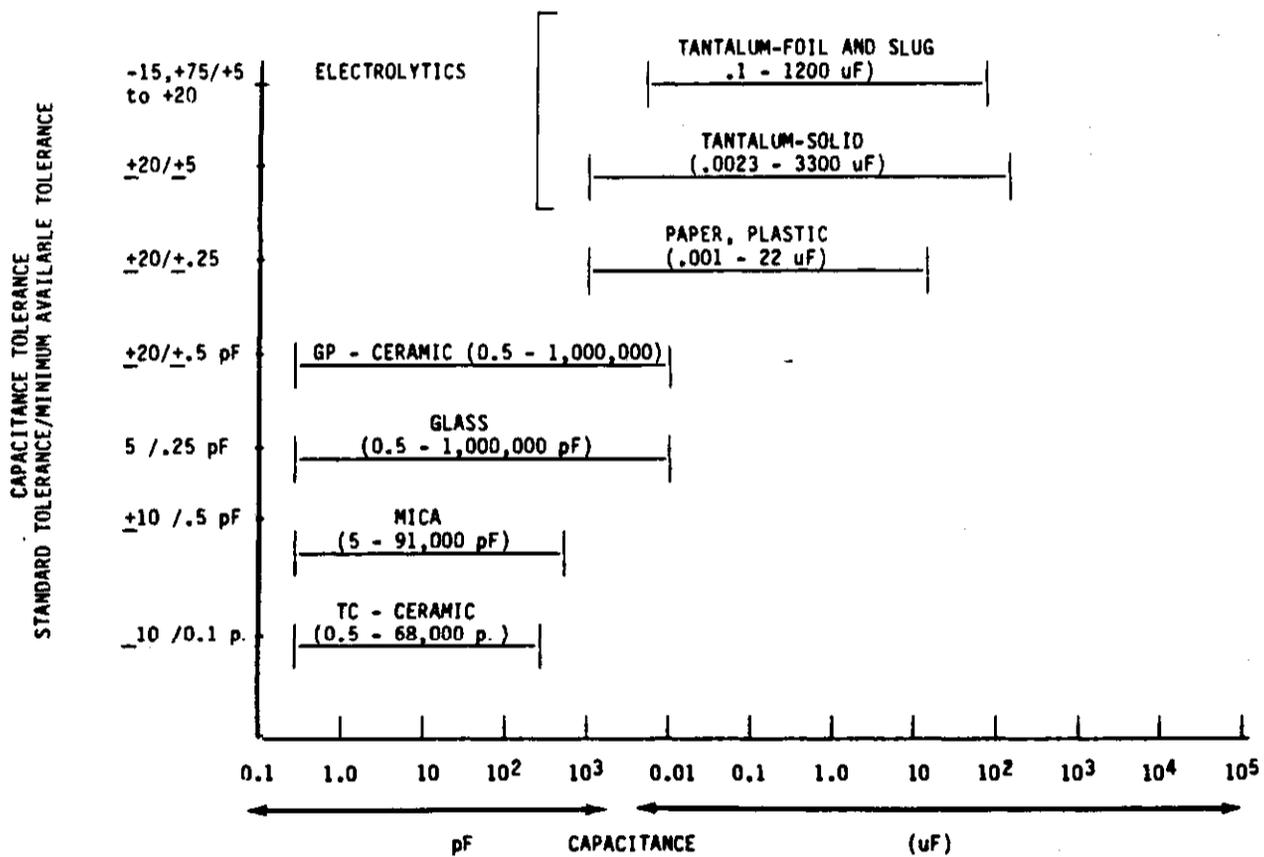


FIGURE 1. Capacitance and tolerance vs dielectric.

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Voltage rating. Figure 2 depicts typical ranges of dc voltage ratings available for different dielectrics. It is important that the voltage rating of the capacitor selected be sufficiently high to allow for reliability derating, and for voltage surges or transients which may occur in the application.

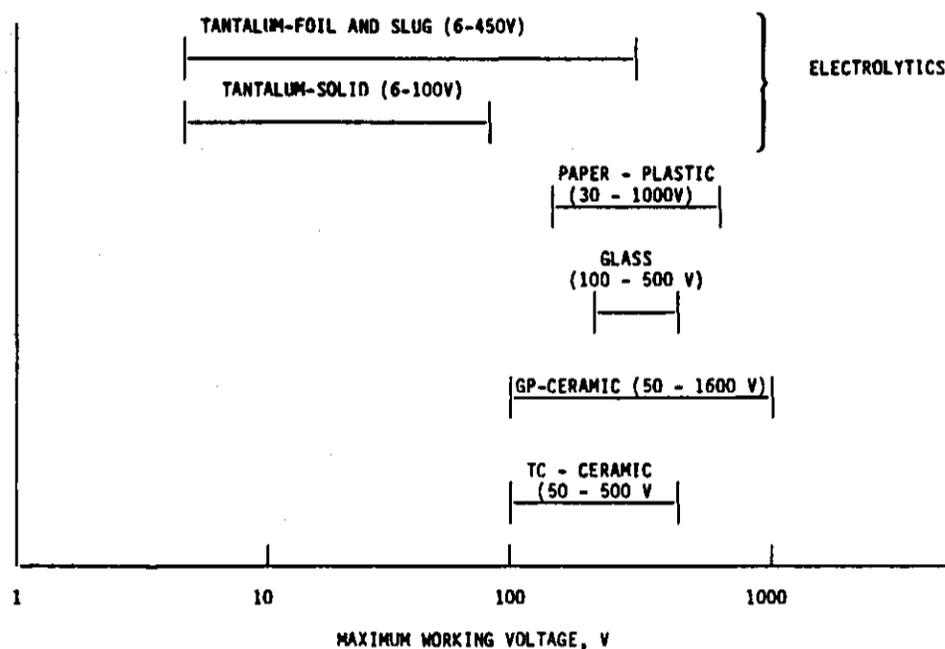


FIGURE 2. Maximum working voltage vs dielectric.

High voltage capacitors (1000 volts and up) must be selected with special care. Corona effects must be considered. In addition to generating spurious electrical signals which may impair equipment performance, corona breakdown results in deterioration of the capacitor dielectric, and can cause eventual capacitor breakdown. Corona results from voids in the dielectric/conductor layers. It is believed to cause dielectric deterioration by generating localized hot spots (depending on the type of dielectric and the corona level generated). Complete dielectric breakdown may occur in a few seconds or after several thousand hours of operation. Corona is likely to occur under ac or pulse conditions.

The first step in selecting a capacitor for a given ac application is to determine the voltage/current wave shape, and the ambient temperature requirements. It is especially important to know the peak voltage, the peak-to-peak voltage, and rms current, and the peak current. In the case of a pure sine wave, the information is generally straightforward. In the case of nonsinusoidal wave shapes, it is sometimes necessary to take an oscilloscope photograph. A scope trace of the capacitor wave shape is very helpful and in some cases absolutely essential to determine the required information.

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AC rating. Operation of capacitors under ac conditions involves three important considerations: the dc voltage rating of the device, the internal heat rise due to I^2R losses, and the corona start level.

Unless the capacitor is rated specifically for ac operation, the ac limitations of the device should be investigated. The peak value of the applied ac voltage must not exceed the dc rating of the device. The temperature rise due to internal heat losses must not exceed the maximum temperature rating of the device. The current-carrying capability of different capacitor types varies widely. The general rule of thumb used by many manufacturers requires that case temperature rise be limited to 10°C .

Corona can be generated at fairly low ac voltage levels. As an example, tests on unimpregnated Mylar capacitors indicate a corona start level of 250 volts peak.

Insulation resistance. Insulation resistance (IR) is expressed in megohms or megohm-microfarads for capacitors with conventional dielectrics, and in terms of leakage current (usually microamperes) for electrolytic capacitors. The effects of this parameter may be significant in timing and coupling applications, or where the capacitor is used as a voltage divider. Leakage current increases with temperature. Figure 3 shows typical values for various dielectric materials.

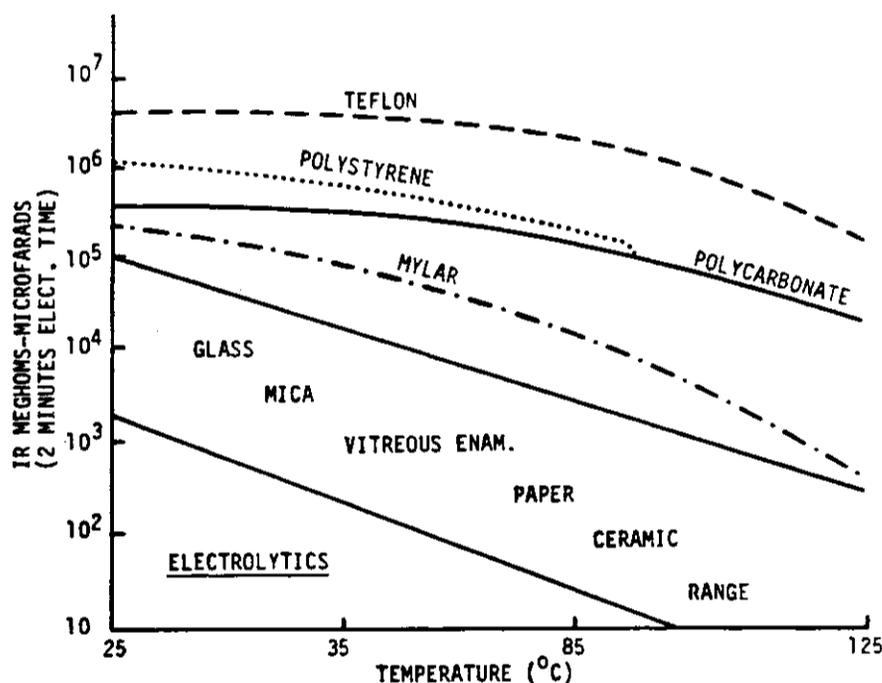


FIGURE 3. Typical values of insulation resistance vs temperature.

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Dissipation factor (DF) or equivalent series resistance (ESR). The dissipation factor is a function of capacitance, ESR, and frequency. Unless otherwise specified, DF is measured at the following frequencies:

- 1 MHz for $C < 100 \text{ pF}$
- 1 kHz for $100 \text{ pF} < C$
- 120 Hz for electrolytic capacitors

DF may vary widely with temperature and to a great extent for ceramic and electrolytic capacitors.

Frequency effects. Most basic capacitor parameter formulas include a frequency term. Capacitor characteristics are to some extent affected by frequency. All capacitors have some inductance associated with their conductors and therefore will resonate at some frequency.

Figure 4 illustrates simplified equivalent circuit of a capacitor wherein all distributed parameters are shown as "lumped" values.

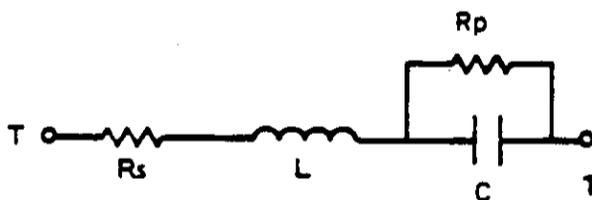


FIGURE 4. Equivalent circuit of a capacitor.

Under dc or low frequency conditions, R_s and L are negligible compared to the C and R_p combination. As the frequency increases, particularly to the megahertz range, both R_s and L increase. R_s increases due to a "skin effect" (where the current tends to travel only through the outer surface metal of a conductor under high frequency conditions). This appears as an increase in the resistance of the conductor. L increases due to the action of the ac current flowing in the leads, electrodes, and terminals, thus generating a magnetic field around them proportional to the frequency.

Since the dielectric constant of most materials will vary with frequency, capacitance is also affected by frequency.

Further examination of the impedance equation shows that as the frequency increases, X_C tends to decrease while X_L increases in value. This means that

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the $(X_C - X_L)^2$ term decreases until at some frequency, the term $(X_C - X_L)^2$ will equal zero and disappear. Then $Z = R_S$ and the capacitor will resonate. This is the point where the capacitor appears as a pure resistor in the circuit.

It also follows that if a capacitor is operated at a frequency higher than its resonant frequency, it will no longer function as a capacitor in the circuit, but will appear as an inductor.

As there are so many variables affected by frequency, no attempt will be made here to present comparative values. As a guide for general frequency applications for different types of capacitors, Figure 5 can be used for an initial approximation. Specific computations or measurements should be implemented to finalize any particular application. Figure 5 reflects frequency ranges for the most efficient application based on normal design values and criteria. Both the upper and lower limits of the frequency range can be extended by special design and construction techniques, as shown by the dashed areas.

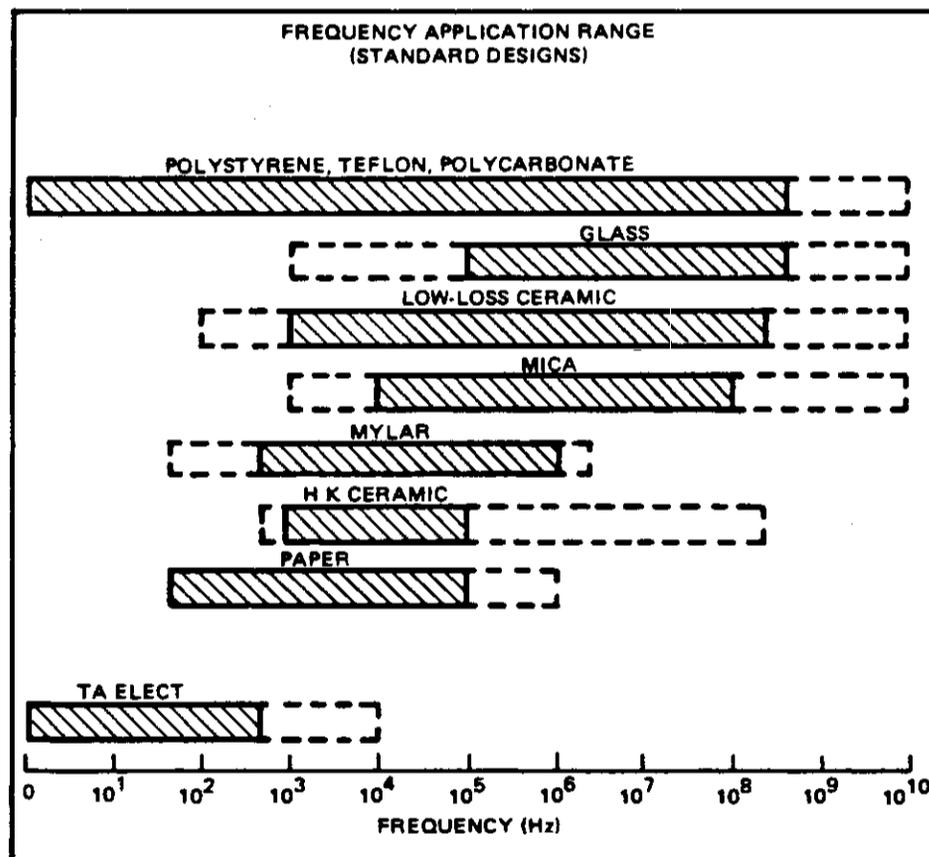


FIGURE 5. Frequency application range (standard designs).

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Temperature effects. Temperature variations (which affect the dielectric constant of capacitor material) result in capacitance changes (ΔC) that can vary from minor to major.

Figures 6 and 7 compare typical temperature coefficient curves of commonly used dielectric types. It should be noted that the curves shown in Figure 7 are for nonimpregnated capacitors. Plastic film capacitors cannot be impregnated and the impregnant (or filler) serves mainly to replace part of the air film. The resultant dielectric constant will vary slightly, thus altering the temperature coefficient curve proportionately.

Voltage coefficient. Capacitance variation with applied voltage is insignificant except with Class II (general purpose) ceramics. High K dielectrics are ferroelectric in nature, and their molecular orientation varies with dielectric stress. See the ceramic subsection for further details.

Dielectric absorption. This phenomenon is due to the tendency of the dielectric to retain electrons it has stored when the capacitor is discharged. When the shorting mechanism is removed, the electrons that remained in the dielectric will eventually accumulate on an electrode and cause a "recovery voltage" gradient to appear across the capacitor terminals. This recovery voltage, divided by the charging voltage and expressed as a percent figure, is called the "percent dielectric absorption."

The magnitude of this percent dielectric absorption figure will vary considerably for different dielectric materials and their impregnants. It is important to note that the measured value of dielectric absorption is a function of the amplitude of the charging voltage, the charging time, the discharge time, the time after discharge that measurements are made, and the temperature.

This tendency of the dielectric to retain its electrons is primarily due to the polarization that takes place at the dielectric dipoles whenever the capacitor is energized. These electrons, in effect, become "bound" or trapped in the dielectric during the discharge period. When the shorting mechanism is removed, these electrons become free again and move to the electrode surface. This results in a potential difference between the electrodes (the "recovery voltage").

A second factor in the magnitude of recovery voltage values is the random movement of "free" electrons in the dielectric. These free electrons take a finite time to move from the dielectric to the electrode, and therefore contribute to the recovery voltage. The magnitude of their contribution is closely related to the discharge time duration.

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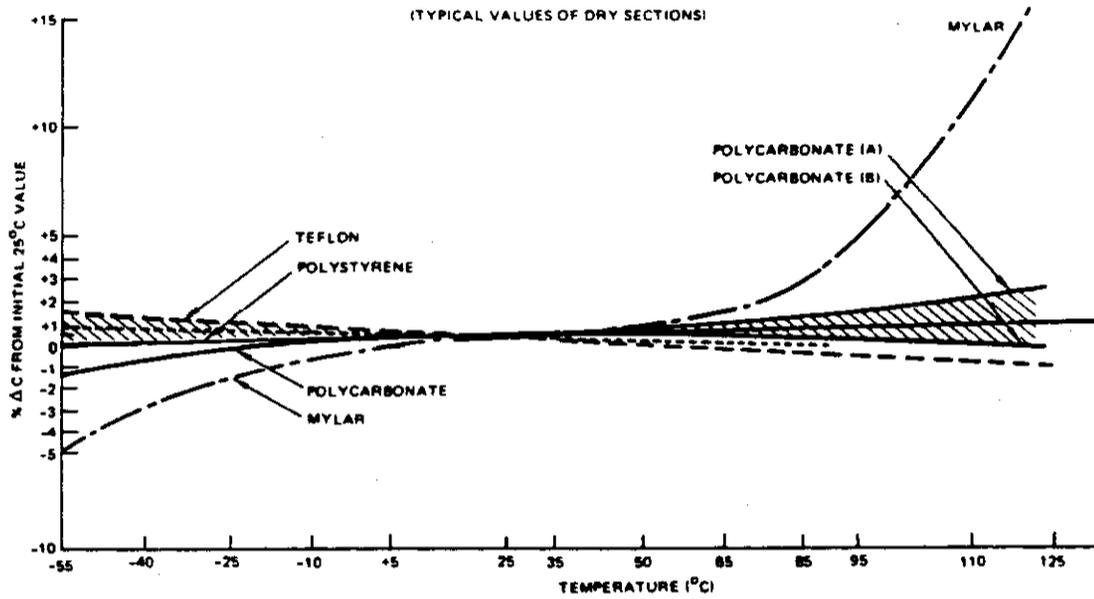


FIGURE 6. Film dielectric capacitance vs temperature.

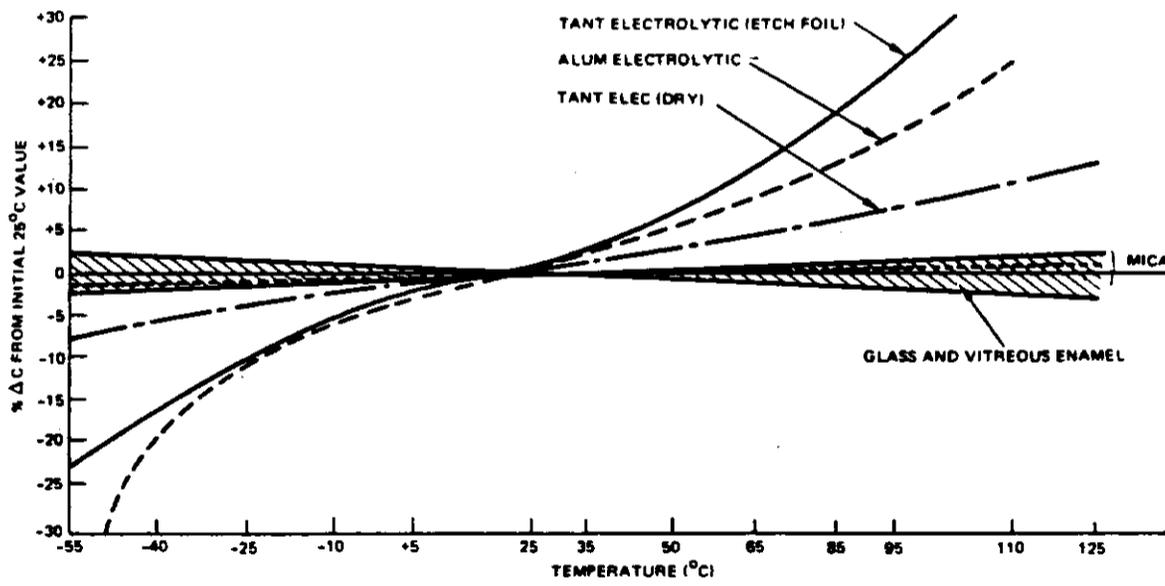


FIGURE 7. Capacitance vs temperature (typical).

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Table III shows approximate percent value of dielectric absorption (DA) for some typical dielectrics. These values are for a given set of conditions. A change in any of the conditions will cause a variation in the percent dielectric absorption.

TABLE III. Dielectric absorption

Conditions: Charging voltage: 200 Vdc Charging time: 1 minute Discharge time: 2 seconds Time after discharge: 1 minute Temperature: 25°C	
<u>Dielectric</u>	<u>% DA</u>
Air	0
*Polystyrene	0.02
*Teflon	0.02
*Polycarbonate	0.08
*Mylar	0.20
Mica (ruby)	0.70
Paper (oil-impregnated)	2.0

*The addition of an oil impregnant will cause the percent DA figure to become essentially that of the impregnant (approximately 2.0 for most oils).

Dielectric absorption is a critical factor in circuitry that is highly dependent upon the speed of response or time delays in the charge and discharge cycles of a pulse circuit.

2.1.5.4 Mechanical considerations. Specific capacitor design characteristics may require special consideration from a product design standpoint.

Mounting by leads. While specifications require that components weighing more than one-half ounce may not be mounted only by their leads, it does not follow that lead-only mounting is satisfactory for all components weighing less than one-half ounce. Nearly all capacitor specifications require rigid mounting of the body during vibration tests.

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Encapsulation. Potting with a hard epoxy can cause capacitor failure by shorting the dielectric as a result of differential pressures exerted after hardening. Solid tantalum capacitors are subject to failure by shorting when potted, and require a buffer coating to prevent internal shorting.

2.1.5.5 Environmental considerations. The behavior and service life of all capacitors are highly dependent upon the environments to which they are exposed. The following is a summary of the environmental factors that are most critical in their effect on capacitors. The design engineer should take into account individual environmental factors as well as combinations of these factors.

Ambient temperature. The temperature of the immediate space surrounding the capacitor is of critical importance since this is one of the factors that determines the temperature at which the dielectric operates.

Service life. The service life of a capacitor will decrease with increased temperature. Another factor affecting service life is dielectric degradation resulting from chemical activity with time.

Capacitance. Capacitance will vary with temperature depending on the dielectric and construction. Both the dielectric constant of the material and the spacing between the electrodes may be affected. These effects may reinforce or cancel each other.

Insulation resistance. The insulation resistance decreases with increased temperature due to increased electron mobility.

Dissipation factor. The dissipation factor is a complex function of temperature and may vary up or down with increased temperature depending on the dielectric material.

Dielectric strength. The dielectric strength (breakdown voltage stress level) decreases with increased temperature. As temperature increases, the chemical activity increases; this will cause a change in the physical or electrical properties of the dielectric.

Sealing. For sealed capacitors, increased temperature results in increased internal pressure that can rupture the seal and result in impregnant leakage and moisture susceptibility.

Humidity (moisture). Moisture absorption by a capacitor can cause parametric changes, reduced service life, and in some cases, early life failures if moisture penetration is sufficient. The most noticeable effect is a decrease in insulation resistance.

The ability of various nonhermetically sealed capacitor types to withstand moist environments is of considerable interest to component or design engineers faced with miniaturization requirements. Generally the nonhermetic unit is con-

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siderably smaller than an equivalent hermetically sealed unit. Most military capacitor specifications require exposure to the moisture tests listed below. Details of test conditions and post-test limits are called out in the individual capacitor specifications.

- a. Immersion cycling. This is a test in which the capacitors are immersed for two or more cycles in fresh or salt water for a period of 15 to 60 minutes per cycle.
- b. Moisture resistance. This test uses a combination of temperature cycling and humidity exposure for 10 cycles, each cycle lasting 24 hours. Subzero temperature exposure and vibration are also included in the cycling phase.

In any nonhermetic unit such as the plastic case or the plastic wrap-epoxy end-filled type of capacitor, moisture can penetrate through the epoxy and plastic casing. The amount of moisture that penetrates will depend on the time, the integrity of the bonding junction between the epoxy and the lead, the density of the plastic material, and the thickness of the epoxy/plastic material.

A distinction should be made between paper and plastic film dielectric designs. Nonhermetic capacitor designs that use paper for all or part of the dielectric are much more vulnerable to moisture than designs using plastic film as a dielectric. Once moisture vapor has penetrated into the dielectric, the paper will absorb the moisture, trapping it and eventually destroying the insulating properties of the paper. For this reason, military equipment specifications disallow the use of paper or paper-plastic dielectrics in other than hermetically sealed metallic cases. In the case of the film dielectric, there is practically no absorption of the vapor by the film and it will either be cycled back out of the capacitor or remain in the air space next to the film surface. While the vapor is in the capacitor, it causes a degradation in the insulation resistance properties of the capacitor. Usually, however, the insulation resistance value of the film dielectric capacitor with moisture vapor present is still superior to that of the paper dielectric capacitor with absorbed moisture.

Vibration, shock, and acceleration. A capacitor can be mechanically destroyed or damaged or may malfunction if it is not designed and manufactured to withstand vibration, shock, or acceleration conditions present in the application. Movement of the internal assembly inside the container can cause capacitance changes, dielectric or insulation failures due to physical movement of the electrode foils or internal roll connections, and fatigue failures of the terminal connections. In addition, external terminals, the case, and mounting brackets, are subject to mechanical stress distortion. Some ceramic capacitors also exhibit a piezoelectric effect which may be a problem in critical circuitry.

Barometric pressure. The altitude at which hermetically sealed capacitors are to operate will affect the voltage rating of the capacitor terminals. As barometric pressure decreases, the ability of the terminal to withstand voltage arc-over also decreases.

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For liquid-impregnated capacitors, the differential between the internal and external pressure becomes greater when the external pressure is reduced. This puts added stress on the seams and terminal seals and can result in rupture of the hermetic seal and impregnant leakage.

Capacitance can be affected by internal dimensional changes due to pressure differentials, and internal arc-overs can result when a partial vacuum condition exists in the capacitor.

Heat transfer by convection is decreased as altitude is increased. This condition must be evaluated in cases where the application results in heat generation within the capacitor.

2.1.6 General guides and charts.

2.1.6.1 Capacitor formulas.

Capacitance. Capacitance is a measure of the quantity of electrical charge per unit of voltage differential that can be stored between electrodes (Figure 8).

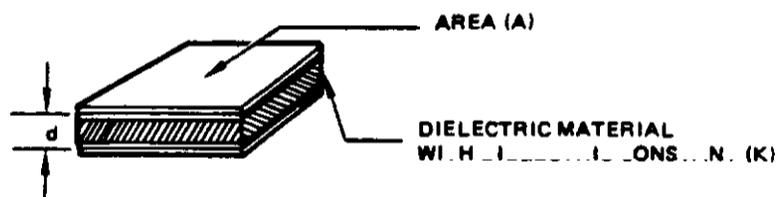


FIGURE 8. Capacitor formulas.

$$C = 0.224K \frac{(N-1) A \times 10^{-12}}{d} \text{ farads}$$

where

A = area of one side of one plate in square inches
 N = number of plates
 d = distance between plates in inches
 K = dielectric constant

$$C \approx KA/d$$

Table IV is a chart of various dielectric materials and their dielectric constants. The approximate (\approx) sign is used rather than an equal sign because these dielectric constants vary somewhat with purity, temperature, frequency, voltage, treatment during manufacture, and various other factors.

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TABLE IV. Dielectric constants at +25°C

Dielectric Material	(K) Dielectric Constant
Vacuum	1.0
Air	1.0001
Teflon	≈ 2.0
Polystyrene	≈ 2.5
Polycarbonate	≈ 2.7
Mylar	≈ 3.0
Polyethylene	≈ 3.3
Kraft paper	≈ 2.0 to 6.0
Mica	≈ 6.8
Aluminum oxide	≈ 7.0
Tantalum oxide	≈ 11.0
Ceramics	≈ 35.0 to 6,000 +

Dissipation factor, power factor, and Q. Each of these terms can be used to express how far a capacitor deviates from being a pure circuit element (see Figure 9).

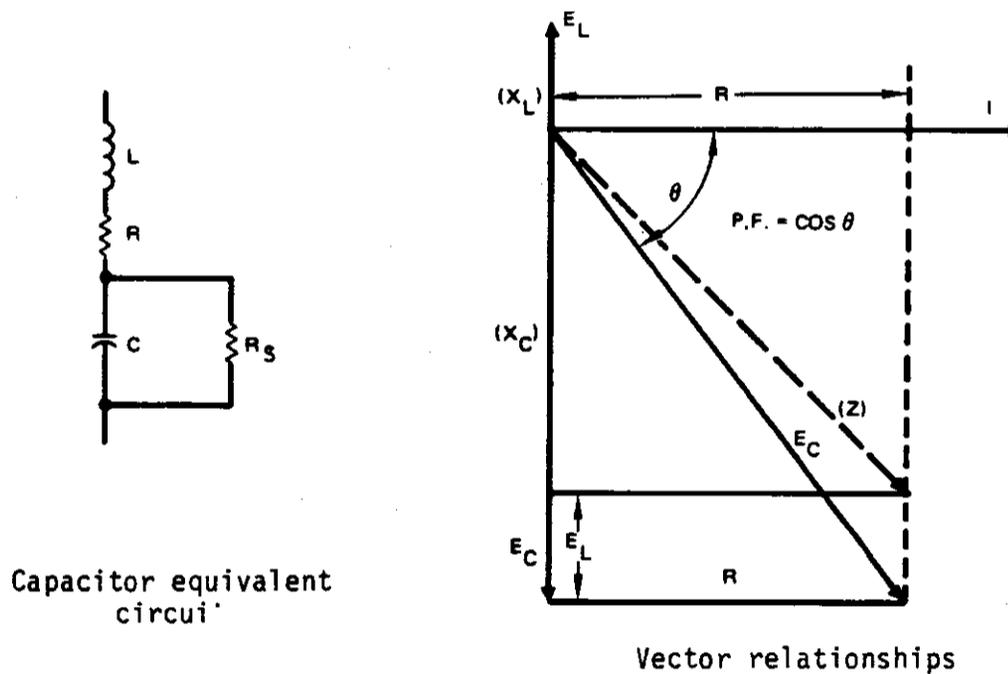


FIGURE 9. Capacitor equivalent and vector relationship.

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where R = effective series resistance in ohms.

$X_L = 2\pi f L = \omega L$ = inductive reactance in ohms.

$X_C = \frac{1}{2\pi f C} = \frac{1}{\omega C}$ = capacitive reactance in ohms

$Z = \sqrt{R^2 + (X_C - X_L)^2}$ = impedance in ohms.

θ = phase angle between current and voltage.

R_S = shunt leakage resistance (negligible for these calculations).

f = frequency in Hertz.

$\omega = 2\pi f$ = frequency in radians per second.

Ohm's Law gives the following: $E_Z = I_Z Z$; $E_C = I_C X_C$; $E_R = I_R R$

and power equations give: Total volt-amperes = $I_Z E_Z = I_Z^2 Z = I^2 Z$

Reactive volt-amperes = $I_C E_C = I_C^2 X_C = I^2 X_C$

Resistive VA (watts) = $I_R E_R = I_R^2 R = I^2 R$

therefore $PF = \frac{P(\text{in}) - P(\text{out})}{P(\text{in})} = \frac{I^2 R}{I^2 Z} = \frac{R}{Z} = \cos \theta$

and DF is defined as the ratio of resistance to reactance,

$$DF = \frac{R}{X_C} = \cot(\theta)$$

Note that for "good" capacitors

R and X_L are $\ll X_C$ such that

$Z^2 = R^2 + (X_C - X_L)^2 \cong X_C^2$ and since $Z \cong X_C$

$$PF = \frac{R}{Z} \cong \frac{R}{X_C} \cong DF$$

Q (figure of merit) is defined as the ratio of reactance to resistance:

$$Q = \frac{X_C}{R} = \frac{1}{DF} \cong \frac{1}{PF}$$

FIGURE 9. (Continued).

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2.1.7 General reliability considerations.

2.1.7.1 Established reliability parts. A large percentage of military type capacitors are available as "established reliability" (ER) parts. Capacitors procured to these specifications have been subjected to special process controls and lot acceptance testing, along with 100% screening and extended life test. This includes an operating voltage conditioning for level "S" parts or better. Level "S" corresponds to a guaranteed failure rate of no more than 0.001%/1000 hrs at 60% or better confidence level under maximum rated operating conditions. The actual failure rate under normal use conditions will be considerably less.

2.1.7.2 Use of accelerated testing techniques on capacitors (Weibull). There are several reasons for using accelerated testing in establishing useful reasons for operating life characteristics (beyond 10,000 hours) and verifying product life characteristics. The Weibull technique shows that by accelerating the stress ratio, it is possible to infer factors such as length of life, optimal sampling techniques, access failure rate level, effectiveness of manufacturing processes and quality control. It is also believed to establish more realistic failure rate computations, which represent critical information for design engineers. Level "B" corresponds to 0.1%/1000 hrs, "C" to 0.01%/1000 hrs, "D" to 0.001%/1000 hrs, and "E" to 0.0001%/1000 hrs. A Weibull sample can be taken at incoming inspection and the actual "real" failure rate can be compared to that claimed by a vendor.

2.1.7.3 Capacitor failure modes. Capacitors usually fail in one of the following modes:

- a. open
- b. short
- c. intermittent
- d. low insulation resistance
- e. capacitance drift
- f. high leakage current (for electrolytic capacitors).

Probable failure modes vary with the type of capacitor. Consult the subsection on individual types for more detailed discussion of failure modes.

2.1.7.4 Failure mechanisms. The classic capacitor failure mechanism is dielectric breakdown. Assuming operation at or beneath maximum rated conditions, most dielectric materials gradually deteriorate with time and temperature to the point of eventual failure. This presumes the early elimination by inspection and screening techniques of infant mortality due to manufacturing defects or mistreatment.

In actual practice the failure mechanisms depend on the construction and type of dielectric and other materials used. More detailed discussions of failure mechanisms are listed under the applicable subsections.

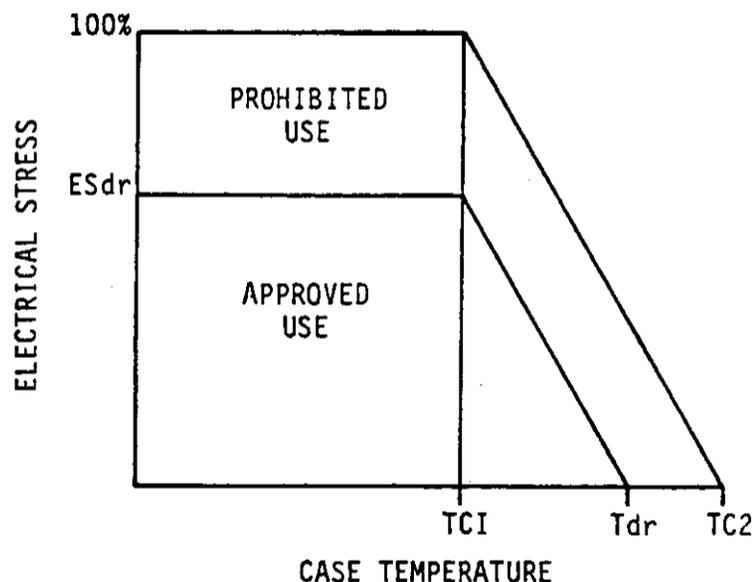
2.1 CAPACITORS, GENERAL

2.1.7.5 Derating. With the exception of failures due to "random" occurrences resulting from manufacturing defects or overstress, capacitor failure rate is a function of time, temperature, and voltage. Operational life can be significantly lengthened by voltage derating and by limiting the operating ambient temperature.

The extent to which electrical stress (e.g., voltage, current, power) is derated depends upon temperature. The general interrelationship between electrical stress and temperature is shown in Figure 10. The approved operating conditions lie within the area below the derated limitation line (ES_{dr}). Operation at conditions between the derated limitation line and the maximum specification curves results in lower reliability (see Handbook 217). Operation in this reduced reliability area requires specific approval.

The derated voltage is the sum of the peak ac voltage and the dc polarizing voltage.

Numerical values are applied to the curves for each part type, based on a percentage of the device manufacturer's maximum rated values. The applicable derating curve or derating percentages are specified in MIL-STD-975.



where

- T_{C1} = case temperature above which electrical stress must be reduced
- T_{C2} = maximum allowable case temperature
- T_{dr} = maximum case temperature for derated operation
- ES_{dr} = maximum electrical stress (e.g., voltage, power, current) for derated operation
- 100% = maximum rated value per the detail specification.

FIGURE 10. Stress-temperature derating scheme.

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2.1 CAPACITORS, GENERAL

End-of-life limits. Circuits shall be designed such that required functional performance is maintained within the identified end-of-life design limits. The end-of-life limit may be assumed to be a 10-year period when the parts are in the approved application region (see Figure 10). The end-of-life values given in the detailed requirements section for each part type are percentage changes from initial values.

Voltage acceleration factor. In essence, if a capacitor is operated for a certain time period at some voltage stress level 1, the voltage acceleration factor can be used to equate this time of operation at voltage (E_1) to an equivalent time at some other voltage (E_2). The formula for this "Voltage Power Law" is shown below.

Voltage Power Law
(constant temperature)

$(L_1/L_2) = (E_2/E_1)^n$ where: E_1 = voltage at Condition 1
 L_1 = life at E_1
 E_2 = voltage at Condition 2
 L_2 = life at E_2
 n = proper exponent for the dielectric material and voltage stress area under consideration.

The expression $(E_2/E_1)^n$ is the acceleration factor, and its accuracy depends upon the proper determination of the exponent n . This value of n will vary for different dielectric materials. It is also affected by design, processing, and test conditions. With all other considerations being equal, operating life or reliability will follow this voltage power law quite closely.

The main problem is to determine the proper value of n for specific dielectrics over specific voltage stress values. Values of n have been empirically determined for various capacitor types, and that these values generally range from about 2 to 6, depending upon the type of unit and the range of stress level.

Typical values which show the variation of the exponent with the stress ratio are listed below:

$n = 5$ for application of 140% to 100% of rated voltage

$n = 3$ for application of 100% to 50% of rated voltage

$n = 2$ for application of 50% to 25% of rated voltage.

Thus, operation at 50% of rated voltage improves the failure rate by a factor of the voltage ratio when compared with operation at maximum rated voltage ($n = 2$ and 3, respectively).

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Temperature acceleration factor. The temperature acceleration factor is somewhat similar in concept to the voltage factor. It is based on a chemical activity rule that states, "For every 10°C increase in temperature, capacitor life expectancy will be cut in half."

This statement, if accepted literally, would mean that the life expectancy of the capacitor would double for a 10°C reduction in temperature, whether from 125°C to 115°C or 35° to 25°C. Again, this is a variable exponent depending on the stress area concerned.

Temperature Rule
(constant voltage)

$$(L_1/L_2) = 2^m \text{ where } \begin{array}{l} L_1 = \text{life at } T_1 \\ L_2 = \text{life at } T_2 \\ m = (T_2 - T_1)/n \\ T_1 = \text{temp at condition 1} \\ T_2 = \text{temp at condition 2} \\ n = \text{°C rule applicable for the temperature stress} \\ \text{area under consideration} \end{array}$$

Note: The expression on the right of the equation is the acceleration factor for the temperature, where n varies according to the dielectric and the temperature stress areas concerned.

2.1.7.6 Capacitor failure rate model. Various types of capacitors require different failure rate models that vary to some degree. To perform an actual reliability prediction, MIL-HDBK-217 should be used.

2.1.7.7 Radiation effects. The principal radiation effects in insulating materials (both organic and inorganic) are related to the ionization dose produced by the particle or photon. The major changes in the macroscopic properties (thermal, mechanical, electrical, and optical) of insulating materials resulting from ionizing radiation are: increase of ionic conductivity and dielectric loss, resulting in decrease of the dielectric Q; changes in dimensions; modification of tensile strength, yield point, and plastic and elastic properties; gas evolution; small changes in the dielectric constant and dielectric strength; and increased optical absorption. Usually the predominant radiation degradation of inorganic materials results from increased conductivity of the material, whereas for organic materials, mechanical changes are usually the major effects. Mechanical changes in organic material occur because of modification of the organic polymer structure caused by radical interaction and formation.

Table V summarizes the radiation resistance properties of various dielectric materials used in making capacitors. It is based on the physical changes taking place and should be used as a guide only. Data based on electrical degradation levels may or may not support the conclusions shown.

2.1 CAPACITORS, GENERAL

TABLE V. Dielectric radiation resistance chart

Material	Absorbed Energy Level For Approx. 25% Degradation
Ceramics	Highly resistant (no levels given)
Glass	Highly resistant (no levels given)
Mica	Highly resistant (no levels given)
Polystyrene	Over 1×10^9 rads
Polycarbonate	Approx. 2×10^8 rads
Mylar	Approx. 1×10^8 rads
Polyethylene	9.3×10^7 rads
Cellulose acetate	1.9×10^7 rads
Teflon	1×10^5 rads
Electrolytic*	Very susceptible (no values given)
Oil-filled capacitors	Very susceptible (no values given)

*Not including solid slug tantalum

The neutron-radiation sensitivity of various types of capacitors is shown in Figure 11. Analysis of this figure shows that permanent change in capacitance value, dissipation factor, and leakage current is not considered severe at a fission neutron-fluence spectrum less than about 10^{14} n/cm². For most capacitor applications this limit is about 10^{17} n/cm², with the exception of paper and paper-plastic capacitors.

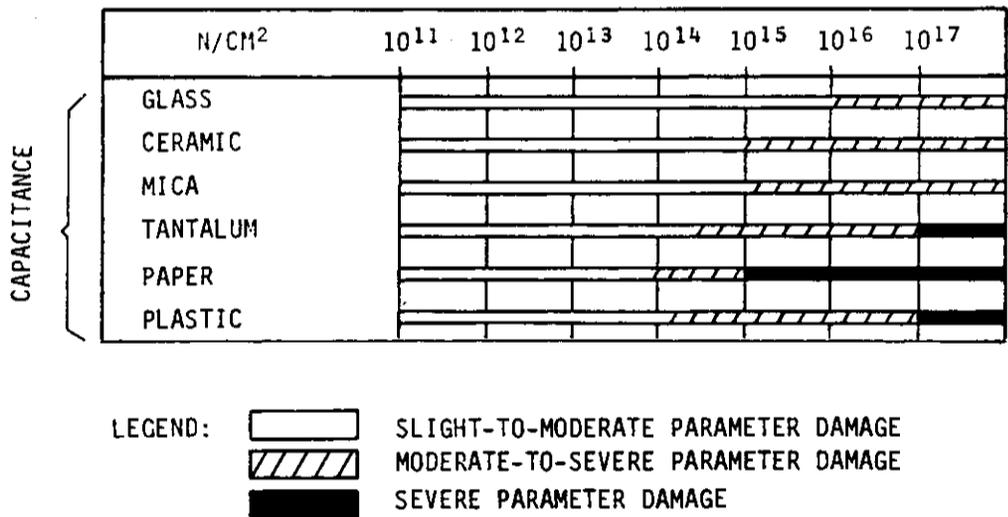


FIGURE 11. Neutron-radiation sensitivity of various types of components.

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The principal cause of radiation-induced capacitance changes is dimensional change in the interelectrode spacing due to gas evolution and swelling. This change is more pronounced in organic-dielectric capacitor construction. Gamma heating and changes in dielectric constants of capacitor dielectrics have been rare and can be considered a second-order effect especially for inorganic materials. Normally, capacitors using organic materials such as polystyrene, polyethylene terephthalate (Mylar), and polyethylene are less satisfactory in a radiation environment by a factor of about 10 than those using inorganic dielectrics. However, it should be noted that even for these types, experimental data indicate no significant permanent changes occurred for exposures up to about $10^{12}/\text{cm}^2$. On the other hand, usage of tantalum and aluminum electrolytic capacitors indicates that both types show capability of surviving extended radiation exposure, with the tantalum being more radiation-resistant. Both decreases and increases in capacitance value have been observed. For example, changes from -10 to +25 percent for tantalum have been observed for exposure up to 10^{17} n/cm^2 . These changes were observed during radiation exposure, with some recovery in the electrical characteristics noted within several days after the end of the radiation exposure. This recovery in some cases was rather slow, and in many instances complete recovery was never attained.

The use of wet-electrolyte capacitors is not normally permitted in high-reliability equipment. Permanent changes in electrical characteristics of these capacitors begin at about $5 \times 10^{13} \text{ n/cm}^2$. The principal mechanism is gas evolution caused by the interaction of the ionizing radiation with the electrolyte, which tends to rupture the capacitor.

Radiation-hardening techniques. In selecting capacitors for a radiation-environment application, a survey of available component part radiation data should be performed to determine whether radiation data exist on that particular part or a similar part. If no data exist, a radiation analysis should be made of the materials that make up the capacitor to try to reduce the number of candidate parts. A radiation exposure is then performed on a few samples of the remaining candidate parts to reduce their population to one or two. These remaining candidate parts then receive extensive investigation in terms of radiation characterization. However, after a capacitor becomes qualified to a nuclear environment, there is always the problem of maintaining this nuclear qualification in any future procurement (i.e., lot-to-lot radiation-quality assurance). In reality, manufacturing processes do change and the vendor does not always inform the procuring agency. Depending on the criticality of the application, some type of screening for usage in radiation environment may have to be performed. This can vary from confirming that there was no change in manufacturing techniques to electrical screening to lot-to-lot sample radiation or to 100 percent radiation screening.

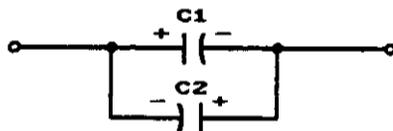
Hardening for fast neutrons. For neutron fluence less than about 10^{14} n/cm^2 (generally the maximum level most semiconductor devices can survive), changes in capacitance value, dissipation factor, and leakage resistance are considered to be minimum if at all detectable. For some capacitors, a fast-neutron fluence

2.1 CAPACITORS, GENERAL

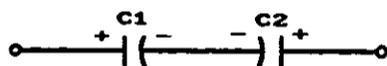
of about 10^{17} n/cm² is necessary before the radiation damage becomes severe. Glass, ceramic, mica, and tantalum are quite radiation-resistant, and they are preferred in that order. Normally, the radiation deterioration of organic dielectric materials is more severe than that of inorganic.

Ceramic, glass or tantalum dielectric material improves the capacitor radiation resistance over paper-dielectric capacitors. In order to minimize radiation effects, the capacitor voltage change, dc voltage and capacitor working voltage should be kept low. Where possible, Zener diodes should be substituted for capacitors in voltage-blocking applications. The utilization of transistor constant-current generators to supply emitter bias rather than emitter bypass capacitors is recommended.

In low-voltage tantalum-capacitor applications (i.e., when the maximum applied voltage is much less than the contact potential), it is possible to cancel out a large portion of ionizing-radiation-induced voltage in the external circuit by using two tantalum capacitors in series or parallel. These two techniques are shown in Figure 12. In the parallel configuration the net radiation-induced charge is the difference between the radiation-induced charges of each capacitor. In the series or back-to-back configuration, the external circuit is subjected to only the difference between the radiation-induced voltage of each capacitor, since the induced voltages are of opposite polarity. However, it should be noted that in either configuration the cancellation effectiveness depends on how well the capacitors are matched and how equally they are irradiated. The cancellation will not apply when the applied voltage is large compared with the contact potential.



A. Parallel configuration.



B. Series or back-to-back configurations.

FIGURE 12. Electrical configuration of sintered-anode solid-electrolyte tantalum capacitor for ionizing-radiation hardening.

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2.1 CAPACITORS, GENERAL

Tantalum capacitors are frequently used in the back-to-back configuration where large capacitor values in a nonpolar type of application are required.

Although long-term exposure to neutron or gamma radiation can eventually cause permanent degradation of component part parameters, the design engineer should also be aware of the effects of short pulses of high-level gamma (or x-ray) radiation.

The voltage across a biased or unbiased tantalum capacitor tends to approach a value of approximately -1.1 V when the capacitor is exposed to a pulse of high-level gamma radiation. The actual voltage change for a given capacitor is a function of both gamma level and initial charge, and varies from unit to unit for a given set of gamma and initial-charge conditions. These facts are not changed by connecting capacitors back to back, but the effects experienced by the circuit are changed. The greatest benefit from back-to-back operation is obtained for the unbiased state, when there is a net charge of zero across the capacitor pair. As noted previously, this does not necessarily imply zero voltage across each capacitor, but rather that their voltages are equal and opposite. Whenever this condition exists, the net result of a gamma burst is a voltage across the pair equal to the difference in the change in voltage across each capacitor. This net resulting voltage may be of either polarity or may even be initially of one polarity and then change to the opposite polarity.

The preceding theoretical considerations, substantiated by considerable data, yield the following guidelines for analysis and design of circuits using back-to-back tantalum pairs:

- a. Because of cancellation effects, induced voltage across a pair of zero-biased capacitors is roughly an order of magnitude less in amplitude than for a single tantalum capacitor, but the decay time is of the same order of magnitude.
- b. Inasmuch as the induced voltage represents the difference between two induced voltages, it may be of either polarity.
- c. The voltage induced in large capacitors is no greater than that induced in small ones. In circuits involving operational amplifiers, therefore, it usually is advisable to use large capacitors and small resistors instead of the opposite to obtain a given time constant, since this tends to minimize the effect of operational-amplifier (op-amp) offset current. This consideration can become quite important in circuits exposed to neutron radiation, since high neutron dosage causes op-amp offset currents to increase rather drastically. For example, the zero-bias 3-sigma response for a group of back-to-back 22- μ F, 20 V tantalum capacitors from the same manufacturer was only 2.7 mV/krad. The time constant of the circuit was the same in both cases.

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- d. Back-to-back tantalum capacitors with a net charge (bias) across the pair will generally lose charge at a rate of approximately 1.0 to 1.5 percent per krad (25°C). Of course, for very small biases (a few millivolts), the voltage loss may be masked by the random voltage which is normally induced on zero-bias capacitors. If it is imperative that the initial charge be retained in spite of high gamma dosage, tantalum capacitors should not be used. Tests on a limited number of high-k ceramic capacitors indicate that these devices yield about the same level of induced charge as a pair of back-to-back tantalum capacitors, while losing an order of magnitude less initial charge. There is a considerable penalty in both size and cost when using ceramic capacitors in lieu of tantalum capacitors; solid-tantalum capacitors are also generally more stable with age and temperature.
- e. Where the gamma radiation to which the capacitor pair is exposed is of short duration (a few microseconds or less), the response is almost purely gamma-dose instead of gamma-dose-rate-dependent. That is, the response to a pulse of 1×10^9 rads/s for 10 μ s would have essentially the same effect as a 100-ns pulse of 1×10^{11} rads/s amplitude.
- f. Virgin (previously nonirradiated) parts usually respond more than previously irradiated parts. This seems to be true of ceramic as well as tantalum capacitors and occurs even when several hours or even days elapse between exposures, although it is much more pronounced in the case where the second burst follows the first by only a few seconds.

2.2 CAPACITORS, CERAMIC

2.2 Ceramic.

2.2.1. Introduction. Ceramic capacitors can be defined as capacitive devices in which the dielectric material is a high-temperature, sintered, inorganic compound. These dielectric materials generally contain mixtures of complex titanate compounds such as barium titanate, calcium titanate, strontium titanate and lead niobate.

Ceramic capacitors are available in a variety of physical forms, ranging from discs to monolithic multilayered types. Tubular types, feed-through styles and variations of these are also manufactured. Ceramic dielectric capacitors are used more than any other single dielectric family, because of their low cost, wide range of characteristics, good volumetric efficiency, and excellent high-frequency capabilities. However, not all desirable characteristics are available in any given style.

2.2.1.1 Classes. The characteristics are determined by the dielectric class. There are two classes, Class II and Class I dielectrics.

2.2.1.2 General purpose (Class II dielectrics). This term embraces a broad family of capacitors manufactured with high K dielectric formulations. These types are characterized by relatively large changes in capacitance with temperature, voltage, frequency, and time. Their main advantage is high volumetric efficiency. They include type BX.

2.2.1.3 Temperature compensating (Class I dielectrics). As the term implies, capacitance varies in a fairly precise and predictable manner over the operating range. These capacitors are made with low K ceramic dielectric. They include NPO (Nominal Zero Temperature Coefficient) type, as well as both positive and negative temperature coefficient styles. They are stable with time and voltage.

2.2.2 Usual applications.

2.2.2.1 General purpose types (Class II). These devices are not intended for precision applications, but are suitable for use as bypass, filter, and noncritical coupling elements in circuits where appreciable changes in capacitance can be tolerated. Consideration must be given to changes in dielectric constant caused by temperature, electric field intensity, applied frequency, and shelf aging. The piezoelectric effect of barium titanate may also be a limiting factor in low-level circuitry. See Table VI.

2.2.2.2. Temperature compensating types (Class I). These can be used wherever high capacitance stability with temperature is required, or where a specific characteristic is required to compensate for temperature variations of other circuit components. These styles are stable with time, temperature, voltage and frequency, although variations with frequency extremes may require consideration.

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TABLE VI. Usual applications

Circuit Application	Ceramic	
	NPO	BX
Blocking	No	Yes
Coupling	No	Yes
Bypassing	No	Yes
Frequency discrimination	Yes	Yes
Transient voltage	No	Yes
ARC suppression	No	Yes
Timing	Yes	No

2.2.2.3 Application notes. Consideration must be given to the following application requirements:

- a. Capacitance values above 0.33 μF are not recommended for use in critical applications because the devices are more susceptible to delaminations due to the thinness of the dielectric material.
- b. For low voltage applications it is recommended that the rated voltage be a minimum of 100 volts dc.
- c. If potting in a hard material is required for some styles of capacitors, then a resilient materials shall be applied to the capacitor as a buffer.
- d. For space flight use, wax impregnates or other volatile materials must not be applied to the capacitor.

2.2.3 Physical construction. Class I dielectrics use calcium titanate or titanium dioxide to which other titanates may be added in varying proportions to obtain the desired characteristics. The temperature coefficients in ppm/ $^{\circ}\text{C}$ of these materials are typically NPO, N750, N2200, and so on. The primary consideration here is the essentially non-ferroelectric nature of the material. This is the factor responsible for the temperature, voltage and time stability characteristics of the finished capacitor.

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The Class II dielectrics are based on barium titanate. Modifiers may be added in the form of bismuth stannate, niobium or tantalum pentoxides, or various combinations of other materials. These dielectric materials are ferroelectric and exhibit those characteristics for which the so-called "high-K" general application ceramic capacitors are known. These include high capacitance, nonlinear temperature coefficients, ac and dc voltage sensitivity, and capacitance hysteresis. The principal advantage offered by these materials is a high dielectric constant, which ranges from 200 to as high as 16,000.

Table VII shows the typical range of dielectric constants and the variations in electrical characteristics that may be expected with different formulations. Since stability normally varies inversely with dielectric constant, a device with a K15 dielectric may have a 0.1 percent capacitance change with temperature over a given range, whereas a device with a K10,000 dielectric may have a 90 percent change over the same range and a device with a K1800 dielectric may have a 15 percent change.

TABLE VII. Parameter variation with material change

Parameter	Dielectric Constant Range
Dissipation factor	0.01% to 4%
Capacitance change with temperature	0.1% to 90%
Capacitance drift	0.05% to 15%
Voltage coefficient	Negligible to 60%
Aging	Negligible to 15%
Frequency effects	Negligible to 50%

Four styles of constructions are used in ceramic capacitors. They are the disc style, the feed-through or standoff style, the monolithic style and the tubular style.

2.2.3.1 Disc style. This style consists of a metallized electrode on each face of a flat ceramic disc. Silver bonded with a glass frit is the most common electrode material, although other metals may be used.

2.2 CAPACITORS, CERAMIC

The disc devices are basically commercial types, but have been adapted to military specification requirements to a limited extent. They are not recommended for normal NASA applications.

2.2.3.2 Feed-through or standoff style. Feed-through styles are made with both tubular and discoidal ceramic elements. Figure 13 illustrates the difference between electrode configurations.

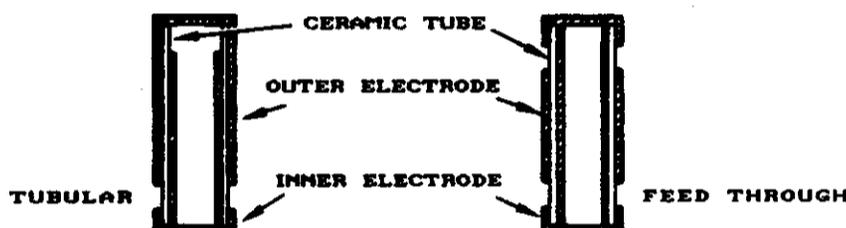


FIGURE 13. Electrode configuration of tubular and feed-through capacitors.

The discoidal version is made with a donut type disc. One electrode connects to the inner hole to permit attachment to the terminal. The other electrode connects to the outer edge of the disc for attachment to the flange or housing. This style is more expensive and is designed with frequency characteristics for UHF application.

2.2.3.3 Monolithic style. This style offers significant size reductions as well as much improved environmental capability.

Monolithic ceramic chip capacitors are brittle and sensitive to thermal shock. Precautions must be taken during mounting to avoid ceramic cracking. To avoid mechanical stresses that may result from temperature changes, the substrate material should have a thermal expansion coefficient that closely matches the thermal expansion coefficient of the capacitor.

Although process details vary among manufacturers, a typical process is as follows: green ceramic sheet is screened with electrode patterns, stacked so that alternating electrodes terminate on opposite ends, and pressed. This laminated material may consist of 30 or more sheets and contain up to 100 individual capacitors. The laminated sheet is then diced to separate each capacitor.

The individual chips are fired at temperatures upwards of 2000 °F to fuse the assembly into a monolithic block. Where the electrodes are exposed on the ends a silver paste combined with glass frit is painted on and fired. This connects all the alternate plates together and provides a pad to which the leads may be

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soldered. The chips are now functional capacitors which are essentially impervious to most environmental conditions (Figure 14). Final operations consist of lead attachment and encapsulation (Figure 15).

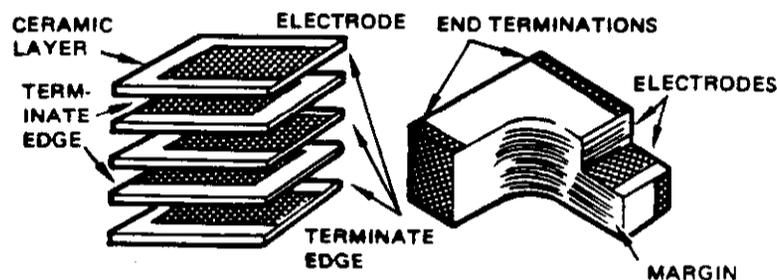


FIGURE 14. Typical construction of a monolithic capacitor chip.

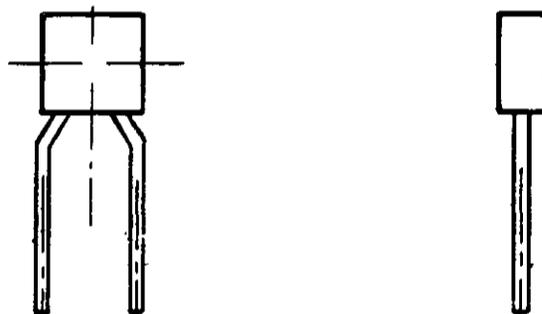


FIGURE 15. Outline drawing of a radial leaded ceramic capacitor (CKR05 Style).

2.2.3.4 Tubular style. The tubular configuration is still used (particularly in the temperature compensating devices), but is rapidly becoming obsolete for military use. The tubular capacitor is fabricated from a tube of ceramic with one electrode painted on the inner surface and terminated at one end of the tube, and the other electrode painted on the outer surface and terminated at the opposite end. Silver bonded with glass frit is the most common electrode material, although others are also used. Its capacitance is limited to low values because of the configuration, and cannot approach the volumetric efficiency of monolithic styles.

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The term "tubular" used here applies specifically to the type of construction described above, and not to the general category of axial-led styles in a cylindrical configuration such as the CKR12 and similar styles. These have monolithic capacitor elements and only the external configuration is tubular in shape.

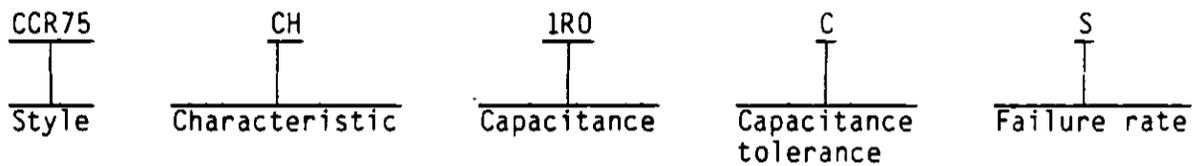
2.2.4 Military designation.

2.2.4.1 Applicable military specifications. Ceramic capacitors are covered by the following specifications:

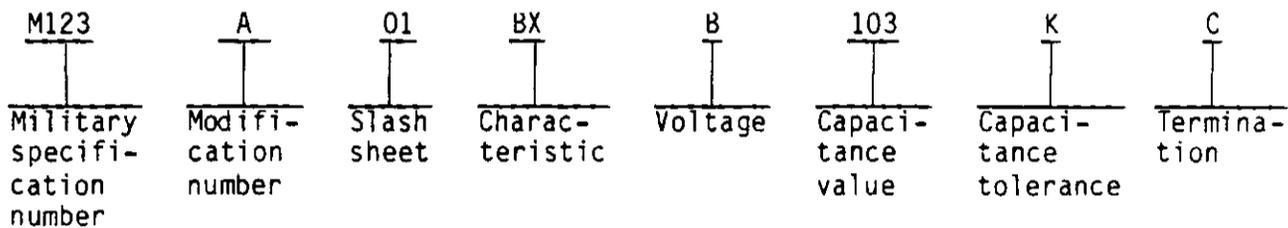
MIL-C-20	Capacitors, Fixed, Ceramic Dielectric (Temperature Compensating) Covers CC and CCR Styles.
MIL-C-123	Capacitors, Fixed, Ceramic Dielectric, (Temperature Stable-BP and General Purpose - BX) High Reliability - covers CKS Style. This is designed for space programs and is an upgrade of MIL-C-55681 and MIL-C-39014. However, it does not cover all values and styles called out in MIL-C-55681.
MIL-C-39014	Capacitors, Fixed, Ceramic Dielectric (General Purpose) - Established Reliability - covers CKR styles.
MIL-C-55681	Capacitors, Fixed, Ceramic Dielectric, Established Reliability - covers CDR Style.

2.2.4.2 Part designation. Parts are currently selected by specifying part numbers detailed in the appropriate Military Specification slash sheet. For example, the part numbers listed herein describe failure rate level, voltage rating, capacitive value, tolerance, characteristic, style, and other features.

MIL-C-20



MIL-C-123



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MIL-C-39014/05-2237

The numbers following the slash describe failure rate, level, voltage rating, capacitance value, and tolerance.

MIL-C-55681

CDR01	BX	100	A	K	S	S
Style	Rated temperature and voltage-temperature limits	Capacitance	Rated voltage	Capacitance tolerance	Termination finish	Failure rate level

Many styles are available in the specifications listed, each with separate capacitance ranges and voltage ratings. For selection, consult MIL-STD-975 (NASA).

2.2.4.3 Space application. MIL-C-123 was developed for space applications in order to meet the severe environmental conditions encountered in space. MIL-C-123 is a replacement for MIL-C-55681 and MIL-C-39014. MIL-C-123 is designed to include stringent testing requirements such as:

- One quality level
- Pretermination Destructive Physical Analysis (DPA)
- Pre-encapsulation terminal strength
- Humidity, steady state low voltage criteria
- Tighter moisture resistance requirements
- DPA inspection (finished product)
- 100% inspection for thermal shock, voltage conditioning dielectric withstanding voltage and other electrical parameters.

2.2.5 Electrical characteristics. Because of the wide difference in characteristics, capacitance ranges, and applications of Class I and Class II dielectrics, refer to MIL-STD-975 (NASA) and the individual military specification for the requirements. See Table VIII.

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TABLE VIII. Electrical characteristics of ceramic capacitors

Characteristic	Ceramic Capacitor	
	NPO	BX
Capacitance range (pf)	10 to 68,000	12.0 to 1,000,000
Tolerances	0.1 pf to $\pm 10\%$	± 0.5 pf to $\pm 20\%$
Voltage rating (Vdc)	50	50 to 1600
Aging (percent/decade)	None	1 to 2
Curie point ($^{\circ}\text{C}$)	Outside the range of interest	115
Temperature coefficient	30 ppm/ $^{\circ}\text{C}$	+15 to -25 percent
Insulation resistance (Megohm- μf)	1,000 minimum	1,000 minimum
Operating temperature ($^{\circ}\text{C}$)	-55 to +125	-55 to +125

NOTE:

Deaging: The aging process can normally be reversed by exposure to 125 $^{\circ}\text{C}$. Complete deaging will occur at 150 $^{\circ}\text{C}$. Both the capacitance and dissipation factor will revert to the former 10-hour levels and then the aging process will resume.

2.2.5.1 Derating. The failure rate of ceramic capacitors under operating conditions is a function of time, temperature, and voltage. Refer to MIL-STD-975 (NASA) for specific derating conditions.

2.2.5.2 End-of-life design limits. For general purpose ceramics, the capacitance change is ± 30 percent and insulation resistance change is -50 percent. Capacitance change for temperature compensated ceramics is ± 0.5 percent or 0.45 pF (whichever is greater), and the insulation resistance change is -50 percent.

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2.2.5.3 Design precautions. BX and BR dielectric ceramic capacitors are rated for dc use and may be used with ac if the sum of the dc voltage and ac peak-to-peak voltage does not exceed the dc rating, and the heat generated does not cause a temperature rise in excess of 35°C. Power can be calculated with the following formula: $\text{Watts} = E^2 (2\pi fc)$ (Dissipation Factor). Nominal ratings are 1/8 watt.

Dielectric absorption occurs in all solid-dielectric capacitors and is a significant parameter in pulse and other circuits where rapid charge-and-discharge characteristics are important. In such applications, time delays caused by dielectric absorption may be detrimental to circuit performance. The ratio of the recovery voltage of a capacitor to the initial-charge voltage, expressed as a percentage, is called the percent dielectric absorption. The percent dielectric absorption for a BX dielectric is 2.5 and for BR dielectric is 4.5, and for an NPO is 0.75.

BX and BR dielectric ceramic capacitors exhibit a piezoelectric response when subjected to vibration. Values up to several millivolts can be expected, therefore circuits with a high gain following a capacitor may be sensitive to this effect.

Capacitors, fixed, ceramic, (CCR, CKR, and CDR). Fixed ceramic capacitors (i.e., CCR, CKR, and CDR styles) should meet the destructive physical analysis criteria defined in Table IX, Group A Inspection, of MIL-C-123, and the humidity, steady-state, low voltage criteria (+85°C/ 85% RH) defined in Table X, Group B Inspection of MIL-C-123 (100% in order to minimize the risk of low IR failures).

2.2.5.4 Measurement conditions. The capacitance of Class II dielectrics is affected by variations in voltage and frequency, particularly as the dielectric becomes thinner as in the case of monolithic styles. Therefore, measurement voltage must be specified; e.g. MIL-C-123 specifies 1.0 Vrms @ 1,000 Hz as a measurement potential. This is used as a standard for all measurements on parts having a 50 Vdc rating or greater.

Class I dielectric capacitance is largely independent of voltage and frequency within the ratings of the device. However, standard measurement conditions similar to those of Class II dielectrics are normally specified.

2.2.5.5 Dissipation factor or "Q". For Class II dielectrics, the dissipation factor will seldomly affect circuit operation except in specific applications requiring the high Q of the Class I dielectrics. Class II dissipation factors range between 1.0 and 4.0 percent, although military grades are normally no higher than 2.5 percent. The primary purpose for checking this parameter is that for a given dielectric, variation in dissipation factor may provide the manufacturer with information on formulation and firing variations. An exceptionally high dissipation factor indicates poor termination (cold solder joint, lifted electrode, separated lead, etc.).

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Often differences in dissipation factor as slight as 1.5 percent vs 2.0 percent are exploited as sales gimmicks. Figure 16 shows the effects of temperature-vs-dissipation factor. The dissipation factor at -55°C increases by 5 to 10 percent depending on formulation. In this case, the dissipation factor of the bismuth-free body, although slightly higher at room temperature, is significantly lower at -55°C .

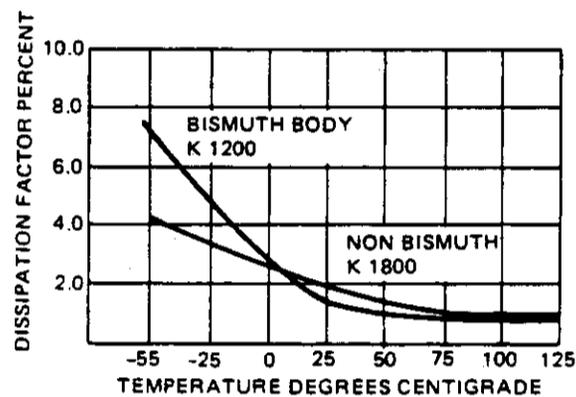


FIGURE 16. Dissipation factor vs temperature.

The dissipation factor is affected by measurement voltage and frequency; since this measurement is normally made at the same time as capacitance, the same comments apply.

The Class I dielectric dissipation factor for these devices is less than 0.1%. This parameter is sometimes expressed as Q , which for capacitors is the reciprocal of the dissipation factor expressed as a pure number. A dissipation factor of 0.1% corresponds to a Q of 1000.

2.2.5.6 DC voltage coefficient. Class II ceramic capacitors have a voltage coefficient that normally becomes greater as dielectric constant increases. This effect is dependent on the formulation as well as dielectric thickness. Figure 17 illustrates the effect of dc voltage on K1800 material. Note that this effect is not linear; although there is a change of 10 percent at rated voltage, the change at 50 percent of rated value is only about three percent, while below 30 percent of rated voltage the change is either negligible or slightly positive.

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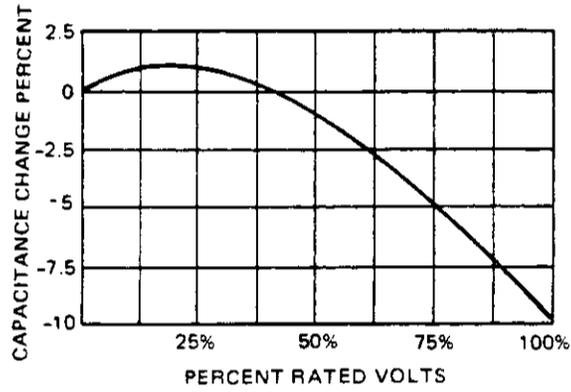


FIGURE 17. Capacitance change vs dc voltage.

The curves in Figure 18 show the combined effect of voltage and temperature. Voltage coefficient is greater at low temperatures than at high temperatures and therefore -55°C is usually the point of greatest change for a given characteristic.

Class I dielectric curves are not shown for voltage coefficient, since these dielectrics are not particularly sensitive to voltage. Depending on the formulation, the voltage coefficient varies from negligible to a maximum of about 2%.

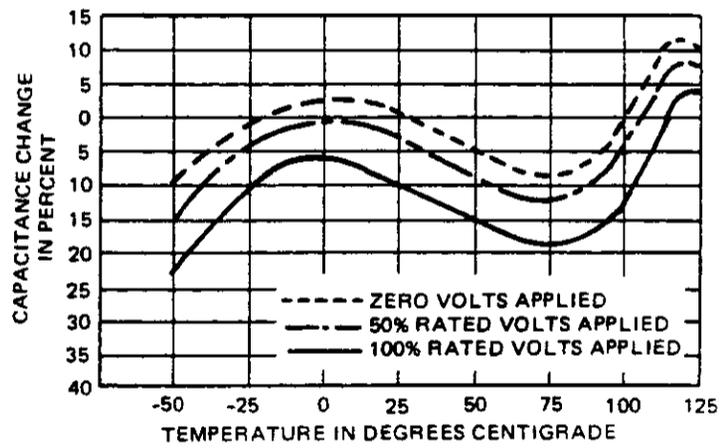


FIGURE 18. Typical combined effect of voltage and temperature.

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2.2.5.7 AC voltage effects. For Class II dielectrics, the effects of ac voltage on capacitance and dissipation factor are also dependent on material and volts per mil stress. It is this latter effect that makes measurement voltage more critical as the dielectric decreases in thickness. A capacitor with a 1 mil dielectric thickness is much more sensitive to 2.0 Vrms than a 20-mil-thick disc.

Figure 19 shows the effects of ac voltage on the capacitance of typical capacitors. Note that, an increase in ac voltage causes an increase in capacitance, opposite to the effects of dc voltage.

Dissipation factor also increases with ac voltage. For example, dissipation factor measured at 1.0 Vrms for K1800 dielectric is about 3% and increases to about 8% at a measurement voltage of 3.0 Vrms, 1 kHz.

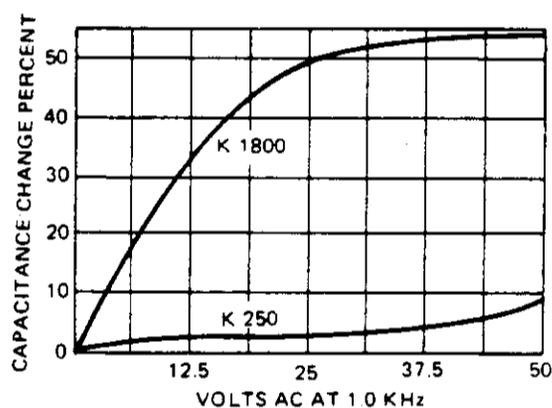
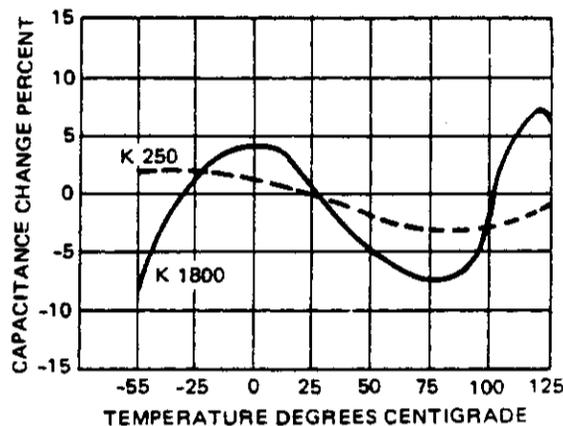


FIGURE 19. Capacitance vs ac voltage.

Class I dielectrics maintain their capacitance and dissipation factor characteristics when used within their voltage ratings.

2.2.5.8 Temperature characteristics. Figure 20 shows typical temperature characteristics of Class II ceramic dielectrics. It can be seen that the temperature stability of the capacitor decreases with increasing dielectric constant.

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FIGURE 20. Capacitance change vs temperature.

Higher dielectric constant materials are available, but with decreasing stability and a narrower operating range of -55°C to $+85^{\circ}\text{C}$. For example a K5600 dielectric would give 4 times the capacitance of K1800 dielectric at 25°C , but could lose 60 percent of its capacity over the temperature range. A K8000 dielectric would give 6 times the capacitance but lose 80 percent of its capacity. These dielectrics are effective only in controlled ambient conditions where full use may be made of the increased capacitance at 25°C without concern for the falloff at temperature extremes.

The most distinguishing feature of the Class I ceramics is their relatively linear change in capacitance with temperature. This change will vary depending on formulation from P100 (positive 100 ppm/ $^{\circ}\text{C}$) to N5600 (negative 5600 ppm/ $^{\circ}\text{C}$). The Class I dielectrics are identified by their nominal temperature coefficient (TC). It is important to remember that this TC is determined from a two point measurement of capacitance at 25°C and at 85°C . These materials have a more negative TC as the temperature approaches -55°C . This is covered by specific TC limits over the temperature range. Additionally, there is a tolerance on the nominal TC which varies from ± 30 ppm for NPO to $\pm 1,000$ ppm for N5600. Effectively, an NPO ± 30 ppm will have a temperature coefficient falling between a $+30$ ppm and a -30 ppm limit. Figure 21 shows the temperature characteristic for a N330 ± 60 ppm envelope (note that it is more negative on the cold side). Also important is the fact that the tolerance is necessarily greater for low capacitance values (less than 10 pF), because stray capacitance becomes a factor and the temperature coefficient of the stray capacitance has an effect.

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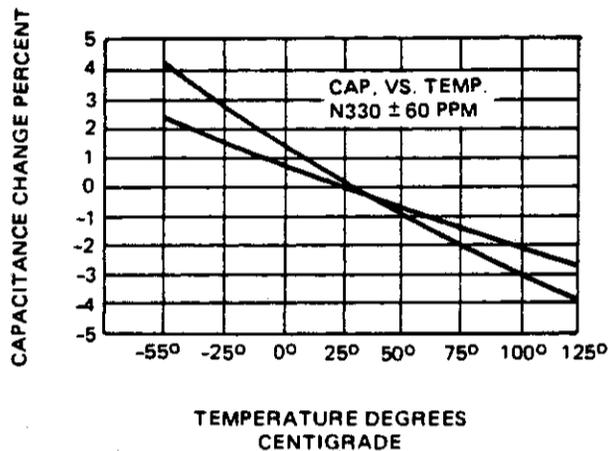


FIGURE 21. Temperature vs capacitance.

2.2.5.9 Effects of frequency. Figures 22 and 23 show comparative effects of frequency for three typical formulations. Actual plots vary, depending on capacitance value, configuration, and lead length. However, these illustrations provide general information in selecting the capacitor type for a particular application.

It can be seen that Class I (NPO) dielectrics are least affected by frequency, and that the Q of the NPO dielectric is several times better than that of the high-K material.

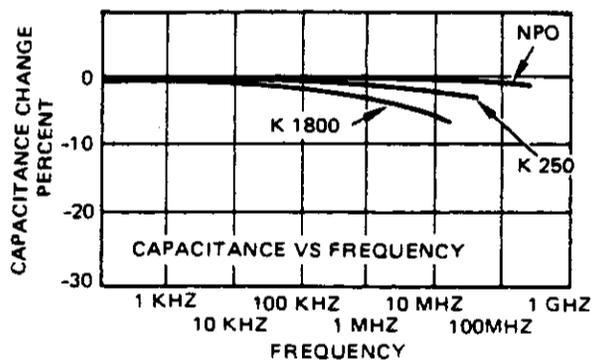


FIGURE 22. Capacitance vs frequency.

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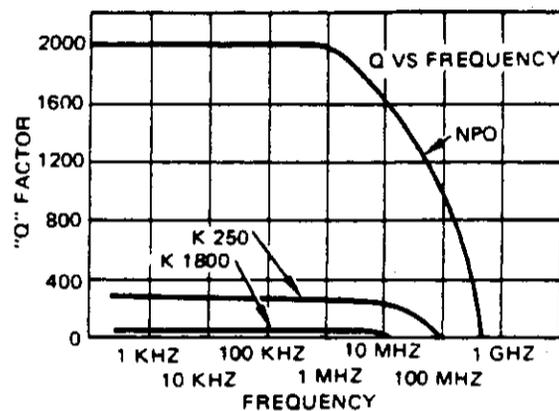


FIGURE 23. "Q" vs frequency.

2.2.5.10 Aging. Class II ceramic capacitors exhibit a characteristic referred to as aging; i.e., a decrease in capacitance with time. Class I dielectrics are not subject to this phenomenon and are quite stable in storage and operation.

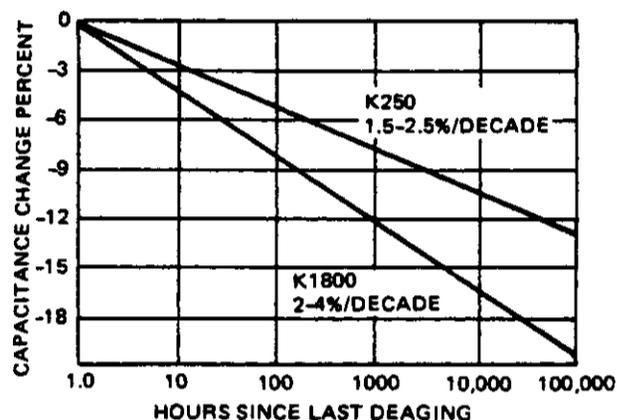
During storage at room temperature there is a decrease in the dielectric constant due to crystalline structural changes in barium titanate. The magnitude of these changes increases as the dielectric constant increases.

These changes are exponential with time, and aging is normally expressed as percent capacitance change per logarithmic unit of time (such as 2 percent per decade). Time zero in such computations relates to the last exposure to a temperature in excess of 125°C, and time is expressed in hours. Figure 24 shows typical aging curves based on an aging rate of 2.5 percent for K250 dielectric and 4.0 percent for K1800 dielectric material.

Due to its exponential rate, aging on a linear time base is relatively low after 1,000 hours. Manufacturers normally allow for aging when measuring capacitance, so that purchase tolerances will tend to run on the high side of nominal. Longer shelf life can be guaranteed by preaging and sorting to tighter guard bands.

Dissipation factor is also affected by aging, with a gradual decrease occurring during storage. Since this change is favorable and not of great magnitude, it is usually unimportant.

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FIGURE 24. Aging curves, Class II dielectrics.

Application of a high voltage, (e.g., during a dielectric strength test) will cause a decrease in capacitance. Provision is made for this effect in the applicable specification, which permits a waiting period for capacitance and power factor to age into tolerance after such exposures.

2.2.5.11 Deaging. The aging process can be reversed by exposure to 125°C with complete deaging usually occurring at 150°C. Both capacitance and dissipation factor will revert to the former 1.0-hour levels and the aging process will resume.

2.2.5.12 Life. Both Class I and Class II ceramic capacitors will normally operate for several thousand hours when properly derated. However, Class II ceramic capacitors will generally demonstrate a decrease in capacitance of 10-15% in a standard thousand-hour life test at maximum conditions. Although this is apparently the result of different phenomena within the crystalline structure, it is similar in effect to shelf aging. Because of the ferroelectric nature of the Class II formulations, these capacitors usually exhibit less change in capacitance when tested at 125°C, as compared to being tested at 85°C, because some deaging takes place at the higher ambient.

Class I ceramic capacitors typically exhibit less than one percent change in a similar life test.

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2.2.6 Environmental considerations. Barium titanate has a piezoelectric effect and voltage transients may be produced by mechanical stresses such as vibration and shock. This effect is greater on high-K bodies. It occurs after the dielectric has been polarized by the application of voltage at high temperature. It is for this reason that voltage conditioned capacitors exhibit a greater piezoelectric effect.

Voltages in the order of 150 microvolts have been measured under some vibration test conditions. Where this effect may be deleterious to satisfactory circuit operation, consideration should be given to using a Class I dielectric or isolating the capacitor from mechanical stress.

2.2.7 Reliability considerations.

2.2.7.1 Failure modes. Ceramic capacitors usually fail because of an open circuit, short circuit, or because of parameter degradation, (low insulation resistance or excessive change in capacitance with temperature).

The most common failure modes are the catastrophic failures, i.e., opens and shorts. Significant parameter degradation failures are relatively rare and in many applications are unnoticed unless the changes are extreme.

2.2.7.2 Failure mechanisms.

- a. Shorts. Operation under voltage stress can result in shorts due to puncture through the ceramic. This failure mode is common to all styles of ceramic capacitors and results from contaminants in the dielectric, hairline cracks, thin spots in the dielectric, or voids. These failure mechanisms can be precipitated before installation in the end equipment by proper screening techniques, particularly voltage conditioning.
- b. Opens. Opens are the major problem with monolithic ceramic capacitors, particularly the smaller axial-leaded configurations (0.1" diameter and less). These devices employ nailhead or paddle-shaped leads soldered to the end terminations and depend mainly on the epoxy encapsulation for mechanical integrity. Unfortunately, the encapsulation process, which is usually accomplished by transfer molding at high pressures, will fill any voids in the joint between the lead and the ceramic. The epoxy will then tend to open the connection by lifting the lead away from the chip. This failure mechanism is insidious and often occurs only at specific temperatures, or may result in intermittent opens at a given temperature.
- c. Low insulation resistance. This may occur as the result of surface contamination during manufacture or because of moisture penetration through a defective epoxy encapsulation. Monolithic ceramic capacitors, when properly cured and fired, are impervious to moisture as

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far as the chip is concerned, but the shunt path formed by contaminants or surface moisture may result in circuit failures. Low insulation resistance failures may also occur in low-voltage, high-impedance circuit applications. Applying rated voltage will temporarily clear the part. Failures of this nature should be verified by measuring insulation resistance at 1.5 Vdc.

- d. Processing problems. Contaminants in the dielectric, hairline cracks, thin spots, voids, and delaminations are some of the processing related problems that may cause failure of monolithic multilayered ceramic capacitors. These defects result in time, temperature, and environmental related failure mechanisms which have been a serious cause of ceramic capacitor failure both at the part and equipment level.

The problem is more acute with thinner dielectrics, i.e. higher value capacitors, where voids can reduce effective dielectric thickness to dangerously low levels, and hairline cracks and delaminations can easily propagate under stress. When exposed to humid conditions ceramic capacitance with dielectric defects can demonstrate failures even at low applied voltages (low voltage failure mode). This mechanism results in low impedance between the plates leading to a short circuit.

Adequate tests and screening can be applied to detect these dielectric problems. Destructive physical analysis is sometimes used to detect voiding and delamination. MIL-C-123 employs both screening and lot acceptance tests to control such defects for high reliability applications.

2.2.7.3 Screening. Temperature cycling, operating voltage conditioning and x-ray are usually performed to screen out manufacturing defects and potential early life failures.

2.2.7.4 Reliability derating. If manufacturing anomalies and early life failures have been eliminated by screening techniques, the failure rate of all ceramic capacitors under operating conditions becomes a function of time, temperature, and applied voltage. Operational life can be significantly lengthened by voltage derating and by limiting the maximum operating temperature.

Assuming an empirically established failure rate for a given style of ceramic capacitor, the actual failure rate is approximately proportional to the 3rd power of the ratio of the applied voltage to the test voltage. Thus an average failure rate of 0.20%/1000 hours at maximum rated voltage can be reduced to about 0.025%/1000 hours (an improvement by a factor of 8) by operation at 50% of rated voltage.

The derating guideline for ceramic capacitors are specified in MIL-STD-975 (NASA).

2.2.7.5 Failure rate determination. Because of the wide variety of styles and voltage ratings available in ceramic capacitors, and rapid changes in technology, the latest issue of MIL-HBK-217 should be consulted for quantitative failure rate determination.

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2.3 CAPACITORS, MICA AND GLASS

2.3 Mica and glass.

2.3.1 Introduction.

2.3.1.1 Mica. Mica is one of a very few natural materials directly adaptable for use as a capacitor dielectric. Its physical and electrical properties, and its rare tendency to nearly perfect cleavage make it probably the best natural capacitor dielectric known. It is inherently stable, both dimensionally and electrically, so that mica capacitors exhibit excellent temperature coefficient characteristics and very low aging with operation.

The property of perfect cleavage enables blocks of mica to be split into sheets as thin as 0.0001 inch. The surfaces of the split sheets are parallel along the planes of natural crystalline structure.

Most capacitors are made from muscovite mica, which is the best of several natural varieties. It has a dielectric constant between 6.5 and 8.5, and is capable of operation at temperatures up to 200°C.

2.3.1.2 Glass. Glass dielectric capacitors were developed mainly as a substitute for mica. Their electrical characteristics are very similar to those of mica capacitors. They have excellent long-term stability, a low temperature coefficient, and a history of good reliability.

2.3.2 Usual applications. Both glass and mica capacitors are designed for use in circuits requiring relatively low capacitance values, high Q, and good stability with respect to temperature, frequency, and aging. They may be used for high-frequency coupling and bypassing, or as fixed elements in tuned circuits. Their inherent characteristics of high insulation resistance, low power factor, low inductance, and excellent stability make them particularly well suited for high frequency applications.

Although glass and mica styles are essentially interchangeable from an electrical standpoint, first consideration should be given to the mica units because of the difference in cost.

Glass dielectric types cost 2 to 10 times as much as an equivalent mica type, depending on capacitance value, tolerance, and quantity.

2.3.3 Physical construction. The commonly used mica styles have either a molded case or dipped construction, such as the CMR04, CMR05, and CMR06. In either case, the capacitor consists of a stack of mica sheets onto which a thin layer of silver is screened and fired. Thin slips of conducting foil are inserted at alternate ends to provide a conducting path to the silvered plates, folded over at the top of the stack, and clinched together with a clamp-lead assembly. The axial leaded styles are then molded in a polyester plastic case. The more popular radial leaded styles are dipped several times

2.3 CAPACITORS, MICA AND GLASS

in an electrical grade phenolic resin, followed by a final vacuum impregnation with epoxy resin. This results in a physically rugged assembly with high moisture resistance and excellent electrical properties (see Figure 25).

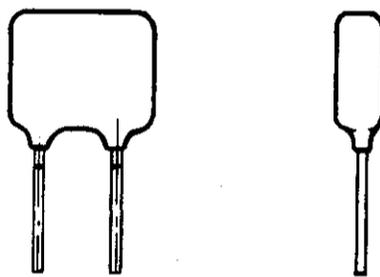


FIGURE 25. Outline drawing of a radial leaded mica capacitor (CMR05 style).

Glass capacitors are made by stacking alternate layers of aluminum foil and glass ribbon until the desired capacitance is obtained, then fusing the assembly into a monolithic block as shown in Figure 26. The same glass composition is used for both the case and the dielectric, insuring that the electrical properties of the capacitor are entirely those of the dielectric material. Leads are welded to the electrodes, and a glass-to-metal seal is formed at the entrance to the case. There is some question as to whether a true hermetic seal is formed, but the capacitors are highly resistant to environmental moisture.

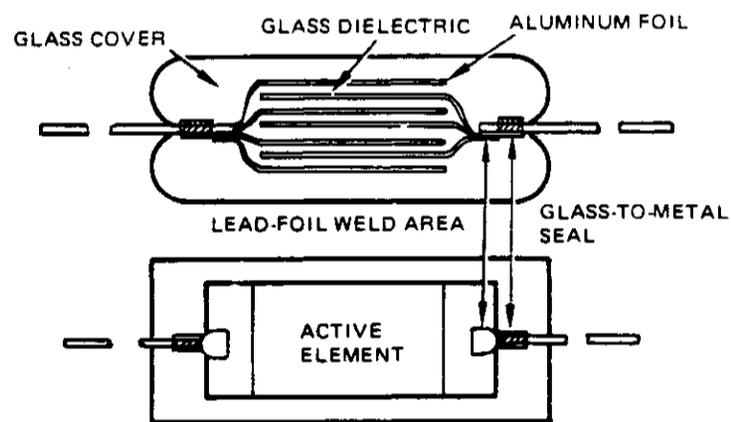


FIGURE 26. Typical construction of the glass-dielectric capacitor.

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2.3.3.1 Feedthrough and stand-off configurations. Mica capacitors are also made in "button" style for RF bypassing and feedthrough applications. These are available in the CB series of styles in accordance with MIL-C-10950. The hermetically sealed styles such as the CB60 series are preferred over the non-hermetic resin-sealed styles such as the CB11 series. They are high-quality units intended for use at frequencies up to 500 MHz.

These capacitors are composed of a stack of silvered-mica sheets connected in parallel. This assembly is encased in a metal shell with a high potential terminal connected through the center of the stack. The other terminal is formed by the metal shell connected at all points around the outer edge of the electrodes. This design permits the current to fan out in a 360-degree pattern from the outer terminal, providing the shortest RF current path between the center terminal and chassis. The internal construction results in minimum external inductance associated with the capacitor. The units are then welded and hermetically glass sealed.

2.3.4 Military designation.

2.3.4.1 Applicable military documents. Mica and glass capacitors are covered by the following military specifications:

MIL-C-10950	Capacitors, Fixed, Mica Dielectric, Button Style CB, (covers feedthrough and stand - off styles, hermetically sealed and resin sealed.)
MIL-C-23269	Capacitors, Fixed, Glass Dielectric, Established Reliability, (covers glass dielectric styles CRY).
MIL-C-39001	Capacitors, Fixed, Mica Dielectric, Established Reliability, General Specification For (covers radial lead dipped styles CMR).

2.3.4.2 Part designation. The type designation for mica capacitors per MIL-C-39001 is shown below. Although similar for other types of mica and glass capacitors, the applicable specification should be consulted for other types.

CMR05	C	100	D	0	D	S
Style	Character- istic	Capacitance	Capacitance tolerance	Temperature range	Rated voltage	Failure rate level

2.3.5 Electrical characteristics.

2.3.5.1 Derating. The failure rate of mica and glass capacitors under operating conditions is a function of time, temperature, and voltage. Refer to MIL-STD-975 (NASA) for specific derating conditions.

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2.3 CAPACITORS, MICA AND GLASS

2.3.5.2 End-of-life design limits. Capacitance change for mica capacitors is ± 0.5 percent and for glass dielectric it is $+0.5$ percent or 0.5 pF (whichever is greater). The insulation Resistance change for mica is -30 percent and for glass it is $500,000$ megohms at $+25^{\circ}\text{C}$ and $10,000$ megohms at $+125^{\circ}\text{C}$.

2.3.5.3 Voltage rating. Glass and mica dielectric capacitors are available as standard parts in the following voltage ranges:

Dipped mica style: 50 to 500 V	Button mica style: 500 V
Molded mica style: 300 to 2500 V	Glass style: 100, 300, and 500 V

These are generally operable over the temperature range of -55°C to $+125^{\circ}\text{C}$ at full rated voltage.

2.3.5.4 Capacitance and tolerance. Glass capacitors are available in capacitance values from 0.5 to about $10,000$ pF in the commonly used styles, and up to $150,000$ pF in the larger transmitting types. Tolerances down to $\pm 1\%$ or 0.25 pF, whichever is larger, are standard.

The same general range is covered by the dipped and molded mica styles of MIL-C-5 and MIL-C-39001, except that minimum available value is 1.0 pF, and the maximum value is somewhat higher. Dipped micas are made up to about $50,000$ pF, but are not widely used in the higher values because of their size. The most popular range for both glass and mica capacitors is 1000 pF and less.

Button mica styles are available up to 2400 pF, in tolerances of ± 1 , ± 2 , 5 , or 10% .

2.3.5.5 Dissipation factor or "Q." Capacitor losses, expressed as DF (dissipation factor) or Q , which is equal to $1/\text{DF}$, are a function of the measurement frequency. Maximum dissipation factors for mica and glass capacitors procured to MIL-C-23269 and MIL-C-39001 are shown in Figure 27. DF is measured at 1 MHz for values equal to 1000 pF or less, and at 1 KHz for values above 1000 pF.

Glass capacitors demonstrate somewhat lower losses, particularly in the small capacitance values. This is because tighter controls can be maintained on the composition of the glass and less allowance must to be made for material variations. Also, the internal terminations are welded, rather than clinched, allowing for greater uniformity.

A simpler evaluated comparison for many applications is shown in Figure 28, which depicts comparative Q at 1 MHz. In the midranges of capacitance, both glass and mica are comparable, but glass is superior at both the high and low ends of the capacitance range. Further comparisons are discussed in paragraph 2.3.5.7 on effects of frequency.

2.3.5.6 AC voltage ratings. Glass capacitors of the CYR series may be operated with an impressed ac voltage provided that the peak value of the applied voltage

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does not exceed the maximum rated dc voltage. This rating applies for power line frequencies and through the audio frequency range. For operation at high frequencies, the power dissipated in the capacitor becomes the limiting factor.

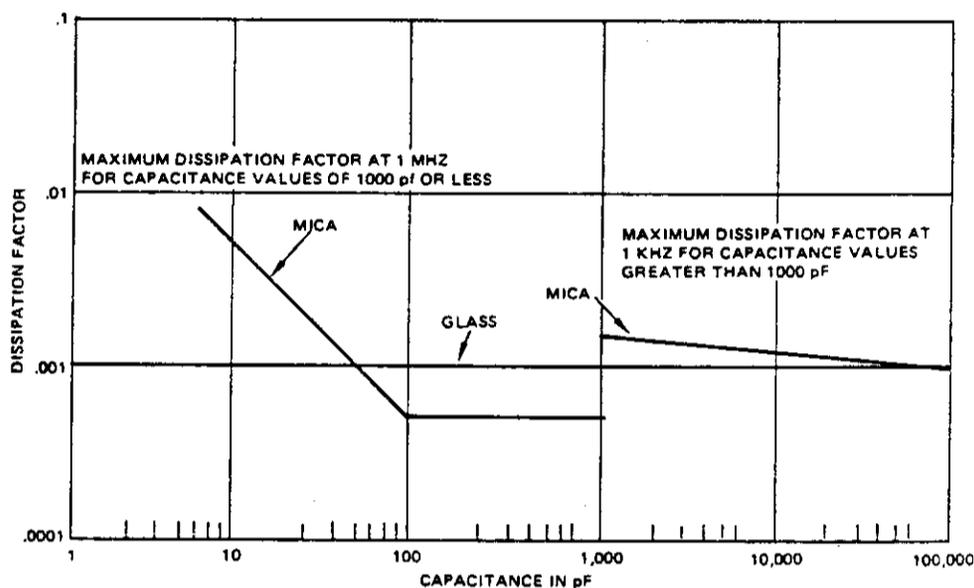


FIGURE 27. Dissipation factor vs capacitance.

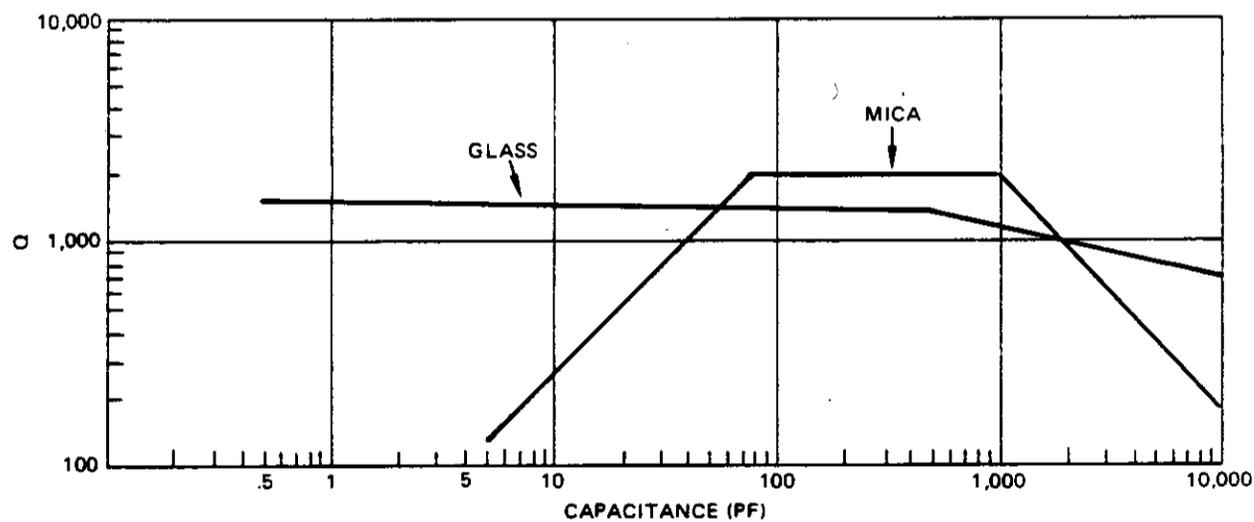


FIGURE 28. Q vs capacitance at 1 MHz.

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Mica capacitors can be operated under ac conditions, but the generally higher dissipation factor, as well as the possibility of corona initiation at relatively low ac voltages must be considered. Each potential ac application of mica capacitors should be reviewed for part capability before the design is fixed. Mica capacitors are more susceptible to corona problems at high ac voltages because of the presence of microscopic voids which may contain air.

2.3.5.7 Effects of frequency. Since these types of capacitors are often used in circuits where the designer is concerned with high frequency performance, the following should be considered: capacitance variation with frequency, self-resonant frequency, Q vs frequency, and RF current capability.

Capacitance vs frequency. Mica capacitors remain quite stable with frequency from low audio through radio frequencies. Capacitance variation in this range is measured in hundredths of a percent. In the VHF range (30 MHz and up) the apparent capacitance tends to gradually increase and may show variations as high as 10% above the 1 MHz measurement as frequency is raised still further. This variation of capacitance with frequency becomes more pronounced as the capacitance value increases.

Glass capacitors are also stable with frequency, as shown in Figure 29. In this case, apparent capacitance tends to gradually decrease with increasing frequency.

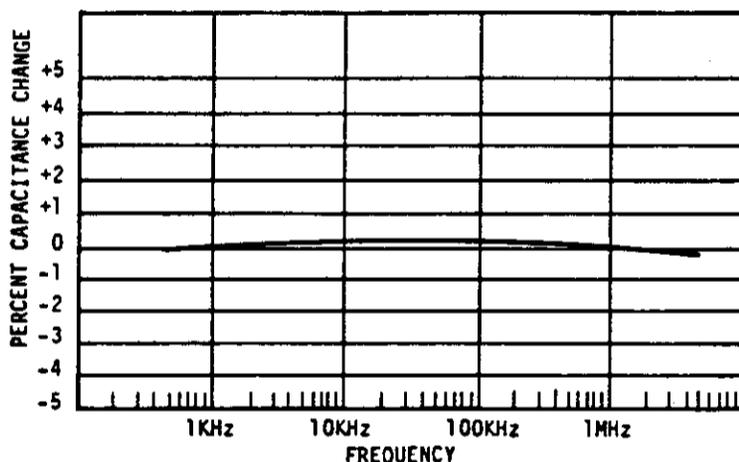


FIGURE 29. Typical capacitance change vs frequency for glass dielectric capacitors.

Self-resonant frequency (SRF). An upper limit of usual operating frequency range is established by the self-resonant frequency of a capacitor. For both mica and glass styles the length of the leads is a significant factor in the SRF. For a given case size, the inductance of the leads and internal terminations is approximately constant, so that the resonant frequency is approximately inversely proportional to the square root of the capacitance.

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Figures 30, 31, and 32 show typical curves for glass and dipped mica styles.

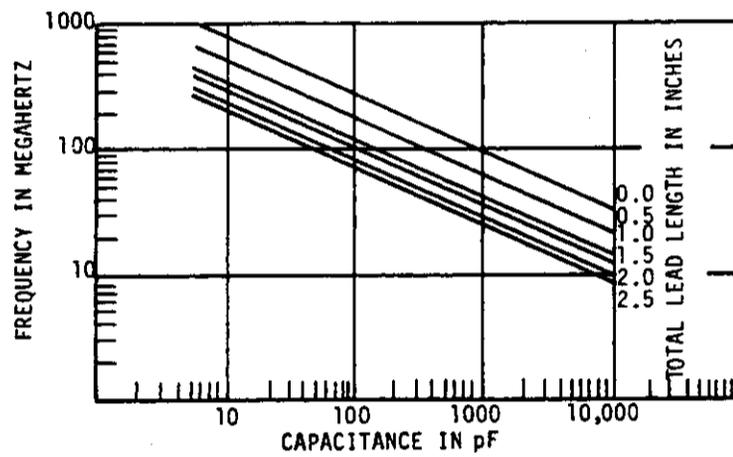


FIGURE 30. Typical curves of self resonant frequency vs capacitance for types CYR10, CYR15, CYR20, and CYR30 glass dielectric capacitors.

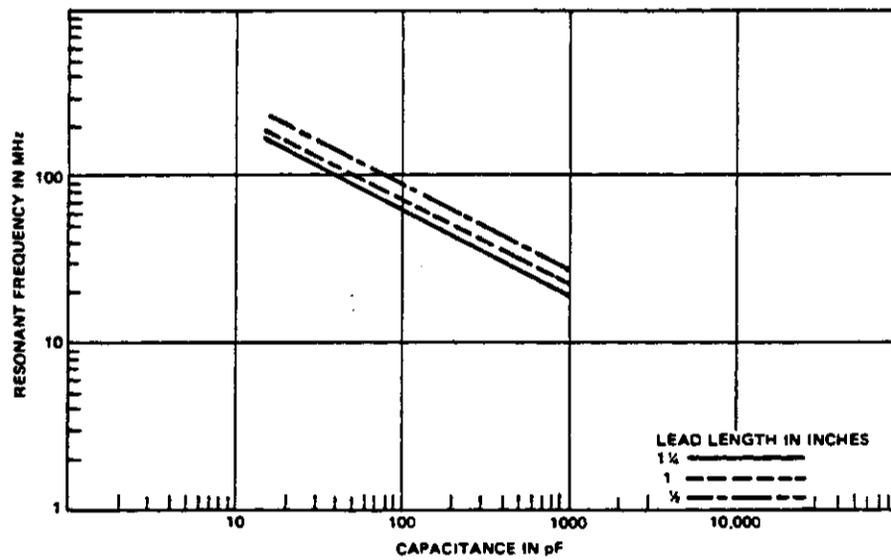


FIGURE 31. Resonant frequency of a typical dipped mica capacitor (type CMR05).

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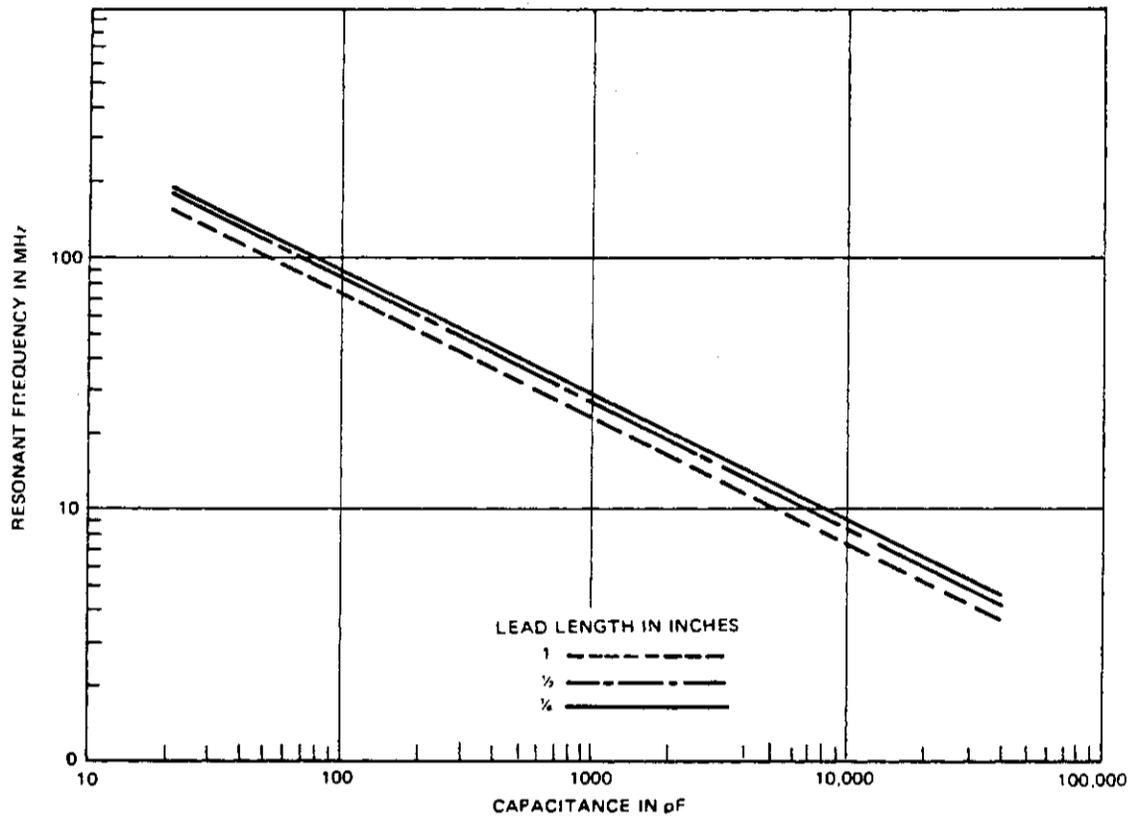


FIGURE 32. Resonant frequency of a typical dipped mica capacitor (type CMR07).

Q vs frequency. Figures 33 and 34 show typical curves of Q and DF as a function of frequency. Glass capacitors are superior to mica in this respect, particularly at frequencies in the range of 1 MHz and above.

RF current capacity. Except for large transmitter type capacitors, little information is available concerning RF current-carrying capability of glass and mica capacitors. Such information is often required, since these units are used in high frequency applications. However, the practically infinite variety of voltage, frequency, and case size combinations makes compilation of comprehensive data a formidable task. The following information can be used as a guideline for such applications (Figure 35).

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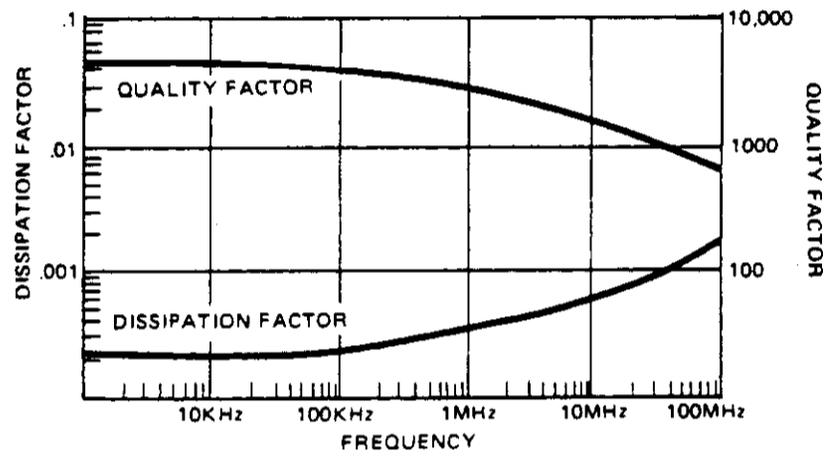


FIGURE 33. Q and DF vs frequency for typical glass capacitor.

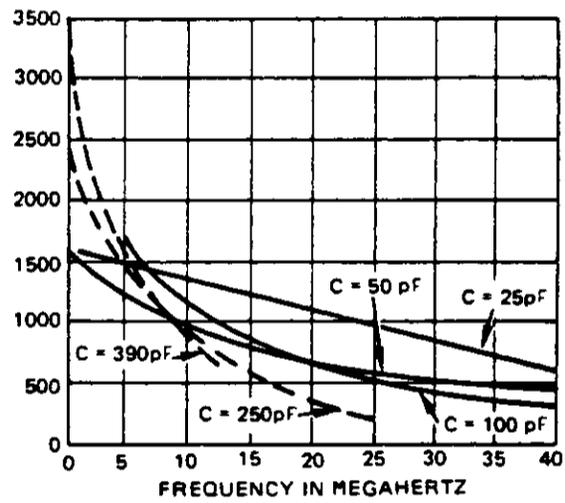


FIGURE 34. Q vs frequency for typical dipped mica capacitor (type CMR05, 500 VDC).

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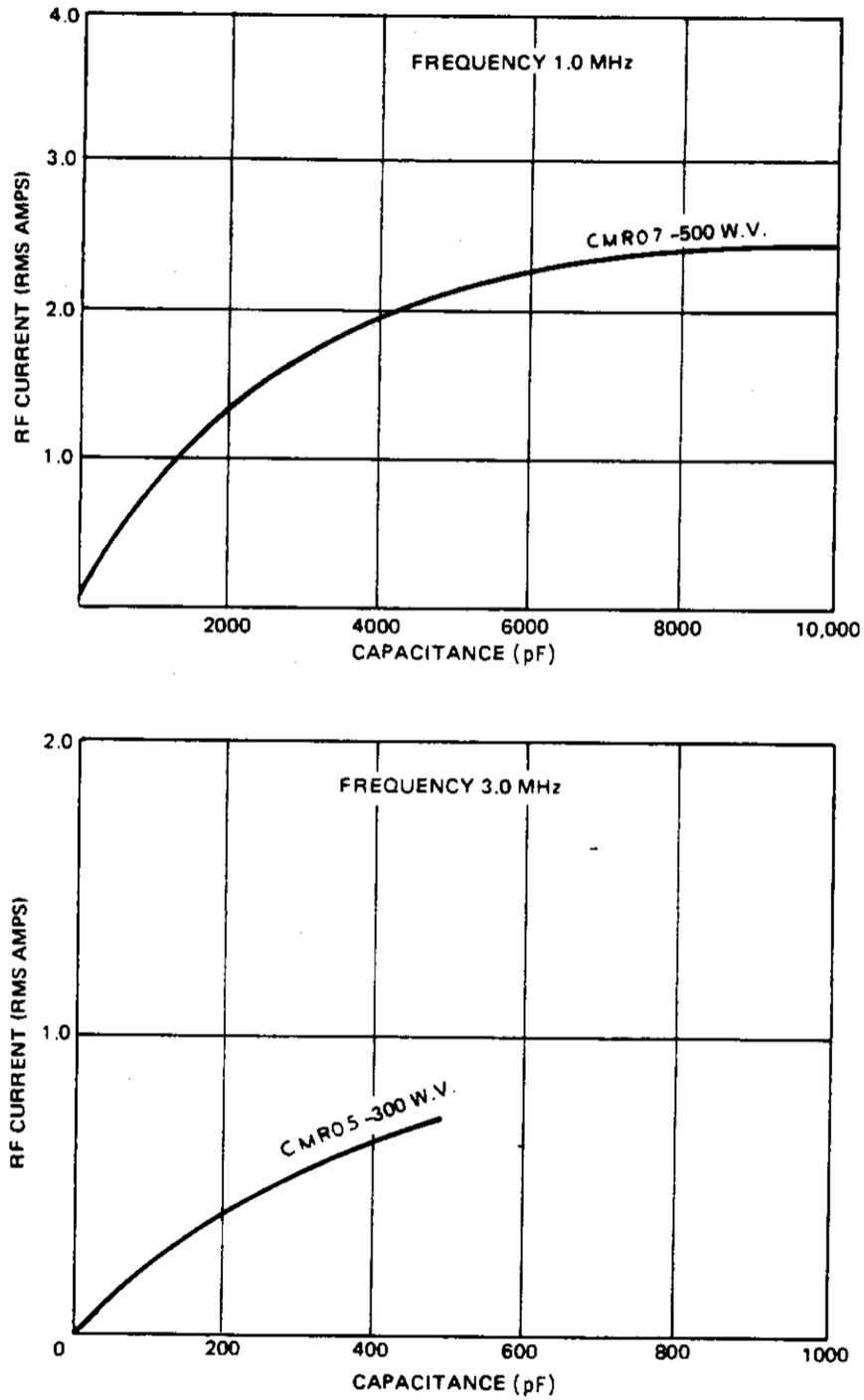


FIGURE 35. Radio frequency current ratings for dipped mica capacitors.

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Glass capacitors (CYR10, CYR15, CYR20). The following maximum voltampere ratings are acceptable limits for these capacitors:

CYR10 - 400 VA
CYR15 - 600 VA
CYR20 - 1100 VA

These ratings are based on a case temperature rise of 45°C which limits the operating ambient temperature to 70°C. For operation at higher ambients, the dissipation must be reduced to limit the maximum case temperature to 125°C. In any case, the peak value of the ac voltage plus dc bias must not exceed the dc rated voltage of the capacitor.

Dipped mica capacitors. The curves shown in Figure 35 represent typical manufacturers' recommendations for safe limits of RF current for dipped mica units of various case sizes. These curves are based on a case temperature rise of 15°C. This lower temperature rise limit is imposed because of the greater tendency of mica units to develop hot spots due to corona breakdown or dielectric imperfections, as compared to the more nearly homogeneous glass devices. Again, the peak amplitude of the combined ac and dc applied voltages must not exceed the dc voltage ratings, and the temperature rise plus ambient temperature must not exceed the maximum operating temperature rating of the capacitor.

2.3.5.8 Effects of temperature. As stated previously, one of the significant advantages of glass and mica capacitors is their stability with temperature.

Glass has a fairly uniform positive temperature coefficient of capacitance. As measured at 100 kHz, the TC is equal to 140 ±25 PPM/°C. This translates into a capacitance change (from the 25°C value) of approximately +1.5% at 125°C, and -1.0% at -55°C.

2.3.6 Environmental considerations. Both glass and mica devices, as procured to the appropriate specification, are adequate for most environments, including exposure to extreme humidity and high levels of shock and vibration.

For severe vibration environments, the cases of the radial-leaded dipped mica styles in particular must be adequately anchored to prevent failure due to lead fatigue.

2.3.7 Reliability considerations. Both glass and mica dielectric capacitors when used well within their voltage and temperature ratings, will operate reliably for several thousand hours. However since construction methods vary with different vendors, physical differences should be examined and analyzed for use in the intended application.

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2.3 CAPACITORS, MICA AND GLASS

2.3.7.1 Failure modes and mechanisms. The most common failure mode for these capacitors is eventual short circuit due to dielectric breakdown. Failure rate varies directly with capacitance value for a given case size, simply because of the increase in the number of plates as capacitance increases. The failure rate of micas would be expected to be higher than glass, because of the greater opportunity for flaws in the natural dielectric, but historical data does not indicate this. The development of the dipped mica style represented an improvement over the molded types because of the elimination of the high heat and pressures of molding. These stresses tend to weaken or fatigue the mica films sufficiently to make the effect noticeable in long-term life tests.

A broadly experienced failure mode exhibited by mica capacitors is caused by poor mechanical clamping of the lead end clips to the capacitor foil stack. This mechanical connection, if not processed properly, is prone to develop electrical intermittants over temperature excursions. Because of this, mica capacitors are not listed in MIL-STD-975. Use of these should be avoided for critical applications. If it is absolutely necessary to use these capacitors, a 100% monitored temperature cycling test should be performed to detect and remove any intermittant devices.

2.3.7.2 Screening. Early life failures are best screened out by a conditioning period of 50 hours or more under over-voltage stress. Glass capacitors are typically conditioned at 300% of rated voltage at 25°C, and micas at 200% of rated voltage at 150°C. These conditions reflect variations in manufacturers' standard approaches to screening, rather than any basic difference in part susceptibility to failure. Temperature cycling is also often specified prior to conditioning, in order to assist in precipitation of failure of parts with mechanical weakness or poor internal connections.

To screen out early life failures, the following screening tests are usually performed: temperature cycling, operating burn-in or high voltage stabilization, and X-ray.

2.3.7.3 Failure rate. Figures 36 and 37 depict typical curves of capacitor failure rate as a function of voltage and temperature stress. These curves represent an accumulation of data including field experience and controlled laboratory tests. They do not represent screened or established reliability parts. For quantitative reliability predictions, current failure rate data should be consulted. These curves, however, can be used to provide an order-of-magnitude estimate as to probable performance. For further information refer to MIL-HDBK-217.

2.3 CAPACITORS, MICA AND GLASS

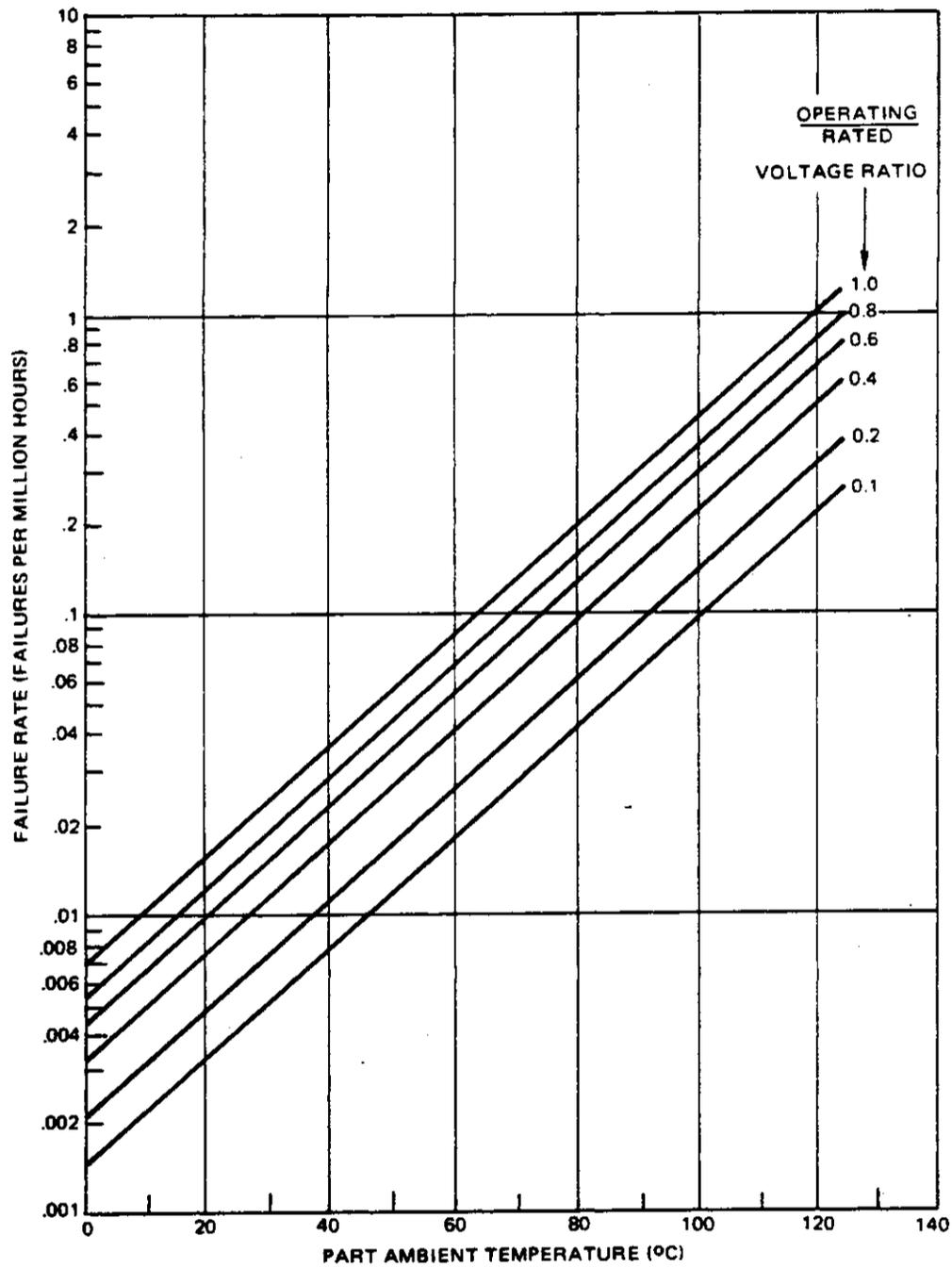


FIGURE 36. Failure rates (in failures per 10⁶ hours) for MIL-C-11272, glass and porcelain capacitors.

2.3 CAPACITORS, MICA AND GLASS

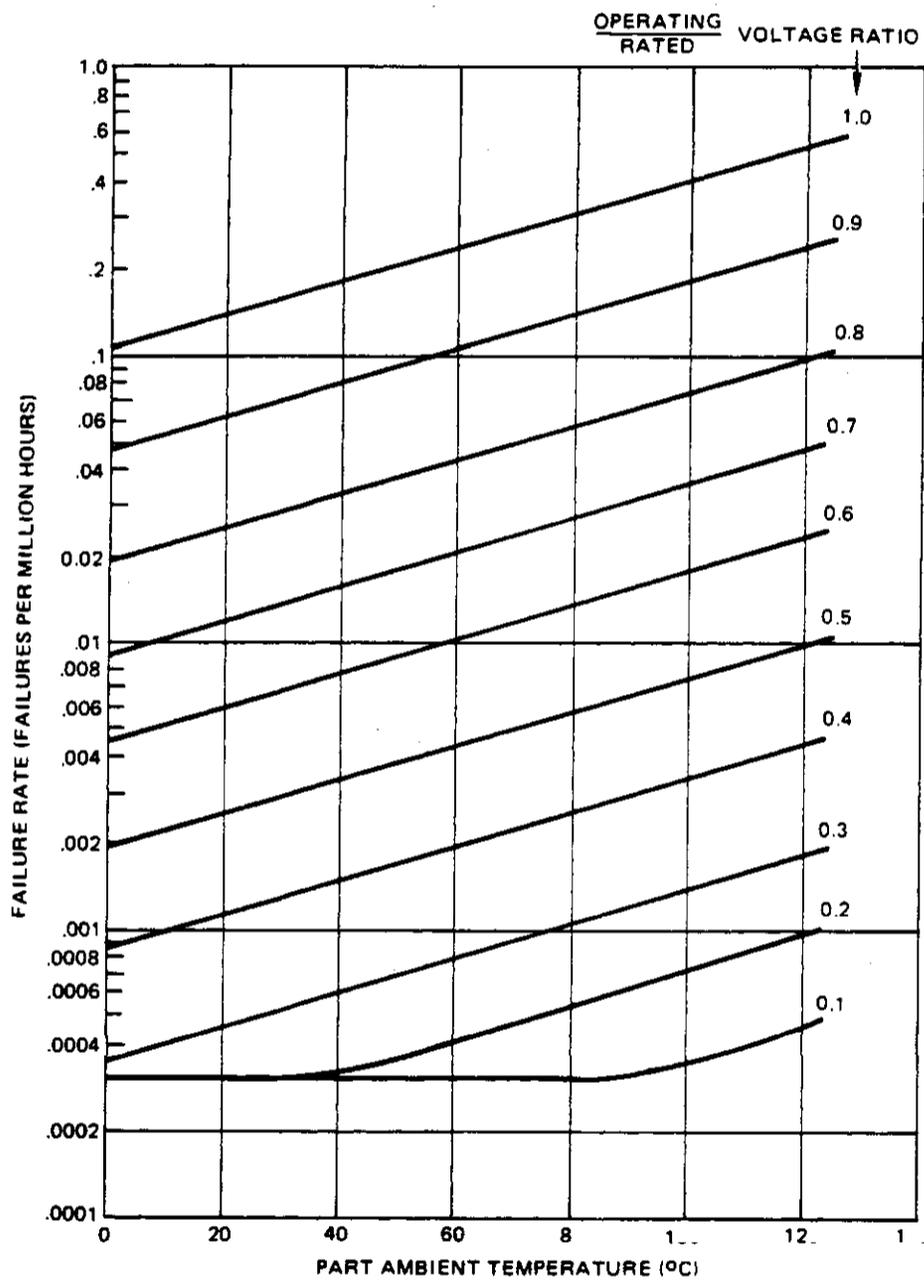


FIGURE 37. Failure rates (in failures per 10⁶ hours) for MIL-C-5, dipped mica capacitors, temperature range 0-125°C, characteristic 0.

2.4 CAPACITORS, PAPER AND PLASTIC

2.4 Paper and plastic.

2.4.1 Introduction. Paper, plastic, and paper-plastic dielectric capacitors serve the broad middle ground of capacitor requirements. This group includes a wide variety of dielectric systems, styles, voltage ratings and temperature characteristics useful in applications once restricted mainly to wound paper-foil devices. Although the remarks in this section apply mainly to the common tubular configuration, these capacitors are available in a great variety of combinations of case styles and electrical ratings for particular applications.

Some of the more commonly used types are: paper-foil, metallized paper, Mylar-foil, metallized Mylar, metallized paper-Mylar, polystyrene-foil, Teflon-foil, polycarbonate-foil, and metallized polycarbonate.

2.4.2 Usual applications. These capacitors are used in most types of applications, including coupling, bypassing, filtering, timing, noise suppression, and power factor correction. They are useful over frequency ranges up to 10 MHz or more, depending on capacitance value and type of construction. They have high insulation resistance, fairly good stability, and can operate at ambient temperatures up to 125°C and above. Certain types of plastic dielectric such as polycarbonate, polystyrene, and Teflon also have excellent temperature coefficient characteristics.

Any ac-rated capacitor can be used in an equivalent dc circuit. However, the converse is not true because of dielectric heating, corona, and resistance heating (R_S).

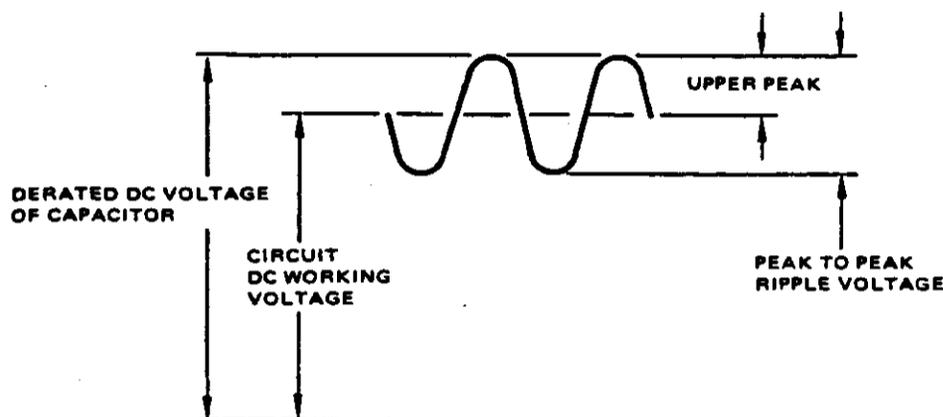
For high frequency ac applications, the dissipation factor (DF) of each capacitor should be measured at the frequency of intended use; capacitors with DF excessively higher than the lot average (even though within specifications) should not be used. The capacitor chosen for a lot acceptance life test should be one with the highest DF ratings and tested under accelerated actual-use stress conditions.

The combined dc and ac voltages should not exceed the recommended derated dc voltage of the capacitor (Figure 38):

Capacitors may be operated at higher frequencies and at reduced rms voltage such that maximum ac current ratings are not exceeded.

Film capacitors should not be used in circuits with less than 500 microjoules of energy available for clearing and should not be used in circuits that would be degraded by voltage transients during clearing. Because the film is thin, it contains pin holes. When the dielectric strength at the hole is not sufficient to withstand the voltage stress, a momentary short develops (10 - 10,000 ohms). High peak currents at the fault site cause a clearing action when the metal vaporizes from around the hole and the short clears.

2.4 CAPACITORS, PAPER AND PLASTIC

FIGURE 38. AC/DC rated voltages.

All film capacitors (metallized film, single or dual wrap, and extended foil) can function intermittently when operated under certain conditions. Electrochemical effects or contamination within the capacitor enclosure can cause spurious, random conduction when the capacitor is operated during temperature changes and where total circuit energy is less than 500 microjoules. The random resistance for film capacitors at 125°C may vary from 1 to 10,000 megohms for capacitance values below 1.0 microfarad.

2.4.3 Physical construction. Both paper and plastic dielectric capacitors are made in conventional wound foil form or as metallized dielectric units. Combinations of two dielectrics are also used such as paper and Mylar or metallized paper/metallized Mylar to obtain some advantages from each dielectric material.

2.4.3.1 Wound foil construction. The capacitor is produced by winding two metal foils separated by two or more sheets of dielectric into a compact roll. The foil is usually high purity aluminum. After the roll is wound, paper units are vacuum dried and impregnated with resin or a synthetic compound. Plastic dielectric types do not require impregnation for moisture protection, though impregnation is sometimes used to reduce corona and to improve voltage breakdown characteristics.

End-connection contacts to the metal foil electrodes are either the tab type or the extended foil type, as shown in Figures 39 and 40. In the tab type, short metallic strips are inserted into the roll during the winding process, and the terminations are soldered or welded to the tabs before final encasement. This type of construction is not preferred, since self-inductance is higher, dissipation factor may be poorer, and the possibility exists of uneven mechanical stresses on the dielectric material at the points of tab insertion.

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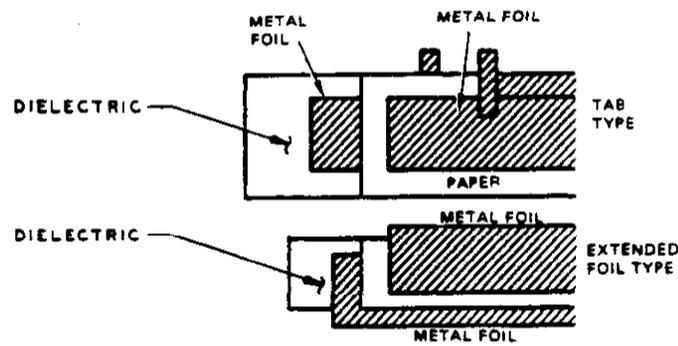


FIGURE 39. End-connection contacts to foils.

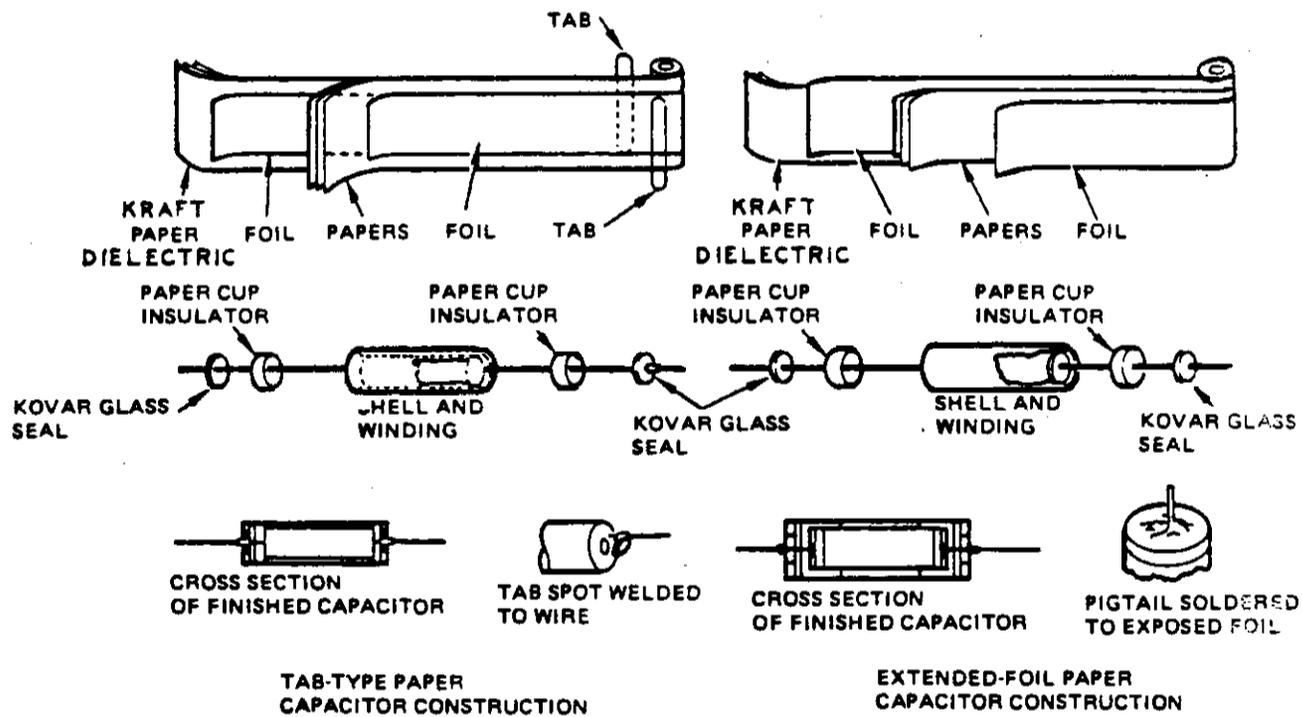


FIGURE 40. Construction of a kraft paper dielectric capacitor.

The extended foil construction is used on most high quality types. The conducting foils overlap the opposite edges of the dielectric, and the entire exposed edge of each foil is soldered to the lead termination.

2.4 CAPACITORS, PAPER AND PLASTIC

After winding and termination, the roll is inserted into its case. For most styles the case is metallic with glass-to-metal end seals forming hermetically sealed unit, as shown in Figure 41. For some dielectric materials, notably Mylar, nonhermetically sealed styles have proven adequate for most environments. Here an additional wrap of several turns of Mylar film is wound around the terminated roll to provide an outer wrap. This additional wrap extends beyond the edges of the foil. Epoxy resin is then poured into the cups thus formed at either end. This provides mechanical rigidity, insulation, and moisture protection for the terminations.

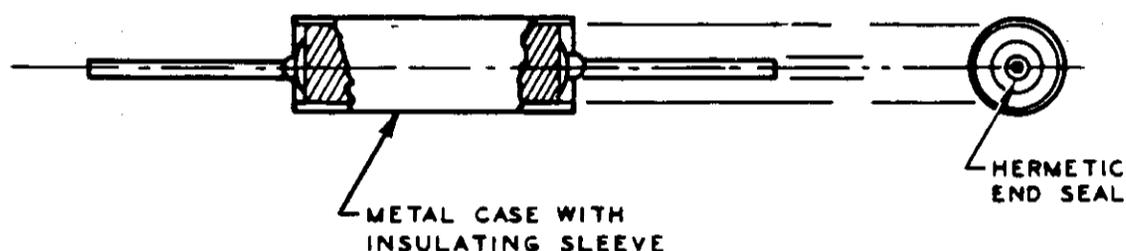


FIGURE 41. Typical hermetically sealed metallized capacitor (CRH style).

2.4.3.2 Metallized film construction. Metallized film construction significantly reduces overall capacitor size. In this type of construction, the aluminum foil conductors are replaced by a thin film of metal which is evaporated directly onto the dielectric film. The metallized strips are then wound in the same manner as the foil types.

Another basic difference in construction is the method of termination. Since the exposed edges of the metallizing are quite thin (about 25 microns thick) it is impossible to solder directly to the ends of the foils. Therefore the ends of the winding are coated with a molten metallic spray. The wire leads are then soldered to this coating.

Another advantage of the metallized film types is that the capacitors are self-healing. If breakdown occurs, a tiny area of the thin film surrounding the breakdown point burns away, leaving the capacitor operable, but with a slightly reduced capacitance. In the conventional paper-foil type (where the foil is thicker), a permanent condition can occur on breakdown, causing a large area of the paper surrounding the breakdown to be carbonized, resulting in a permanent short-circuit.

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2.4 CAPACITORS, PAPER AND PLASTIC

2.4.4 Military designation.

2.4.4.1 Applicable military specification. Following are the specifications for paper and plastic dielectric capacitors:

MIL-C-83421	Capacitors, Fixed Supermetallized Plastic Film Dielectric (DC, AC, or DC and AC) Hermetically Sealed in Metal Cases, Established Reliability, General Specifications for. This specification covers CRH series capacitors.
MIL-C-39022	Capacitors, Fixed, Metallized, Paper - Plastic Film, or Plastic Film Dielectric, Direct and Alternating Current, (hermetically sealed in metal cases), Established Reliability. This specification covers CHR Series tubular and rectangular metallized styles for general use.

2.4.4.2 Part designation. Parts are selected by specifying part numbers detailed in the appropriate specification slash sheet. An example of such a part number is M83421/01-1090 which describes failure rate level, voltage rating, capacitance value and tolerance. For further information refer to MIL-STD-975 (NASA).

2.4.5 Electrical characteristics.

2.4.5.1 Derating. The failure rate of paper and plastic capacitors under operating conditions is a function of time, temperature, and voltage. Refer to MIL-STD-975 (NASA) for specific derating conditions.

2.4.5.2 End-of-life design limits. Capacitance change for paper and plastic dielectric capacitors is rated at ± 2 percent and the Insulation Resistance change is -30 percent.

2.4.5.3 Capacitance and voltage ratings. Paper and plastic capacitors are wound in a wide range of sizes and dielectric thicknesses. They are available in capacitance values from about 1000 pF to several microfarads, and in voltage ratings from 30 volts through several thousand volts. The very high voltage-capacitance ratings are usually restricted to oil-filled paper or paper-Mylar, but the range from about 0.001 to 1 μ F in voltage ratings from 100 to 600 Vdc offers the designer a wide selection of dielectrics and configurations.

2.4.5.4 Capacitance tolerance. Paper, Mylar, and paper-Mylar devices in either wound foil or metallized construction are normally specified to tolerances of ± 5 , ± 10 or ± 20 %. While it is possible to wind them to closer initial tolerances, capacitance variations with temperature and long term drift under operational conditions make it impractical to do so.

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Polycarbonate, polystyrene, and Teflon dielectric units, in contrast, have good temperature coefficient characteristics, and are stable to within $\pm 2\%$ over their respective operating temperature ranges. These types are available in tolerances of 0.25 percent and greater.

2.4.5.5 Dissipation factor. Paper and film dielectric capacitors typically have dissipation factors (DF) of less than 1% at 25°C. The DF can vary widely on a relative basis as measured at temperature extremes, but typically remains at 2% or less for most dielectrics. Losses in paper capacitors are largely dependent on the impregnant used (see Figure 42).

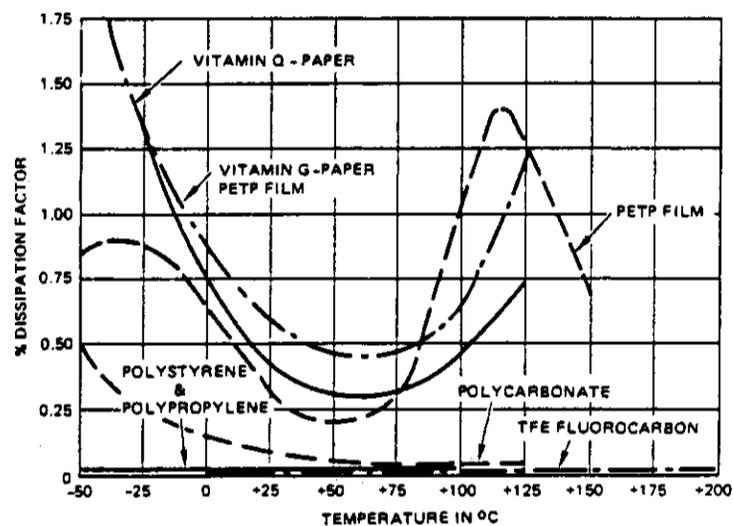


FIGURE 42. Dielectric loss comparison of various dielectric materials.

For most applications, the DF characteristic of the capacitor is not important unless ac voltages are involved. In this case the heat rise induced in the capacitor as a result of ac current is a direct function of the effective DF at the frequency of operation. This effective DF may vary significantly from that measured under standard test conditions, and may result in internal heat rise sufficient to cause early failure due to accelerated dielectric degradation.

2.4.5.6 Insulation resistance. As with dissipation factor, insulation resistance (effective shunt resistance) is usually so high as to be of little concern to the designer. In some cases, however, such as long-time constant integrating networks, holding capacitors, or capacitive voltage dividing networks, the shunt leakage path may be important to circuit operation, particularly when IR degradation with temperature is considered. Under such circumstances the final choice of dielectric system may be determined by its insulation resistance characteristics.

2.4 CAPACITORS, PAPER AND PLASTIC

Figure 43 shows comparative values of insulation resistance as a function of temperature for several commonly used dielectrics. Note that insulation resistance is specified in terms of megohm-microfarads, i.e., the product of the insulation resistance in megohms and the capacitance in microfarads.

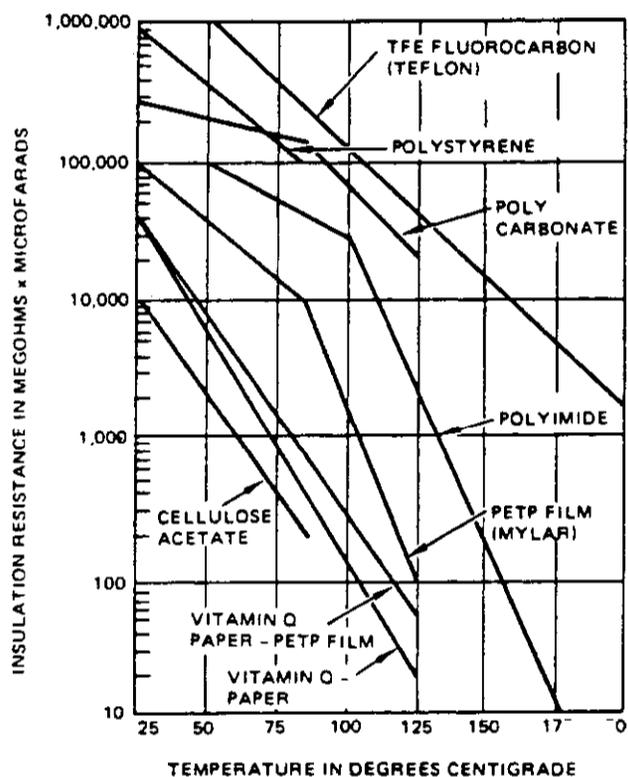


FIGURE 43. Insulation resistance vs temperature for various dielectrics.

Since insulation resistance is inversely proportional to capacitance value for a given style, this product forms a convenient device for comparison among various dielectric systems. The curves show typical values for wound foil capacitors; corresponding values for metallized dielectrics would be decreased by a factor of about 2 to 5.

2.4.5.7 AC operation. Two main factors to be considered in ac applications of paper and plastic capacitors are corona and internal heat rise. Corona is not strictly an ac phenomenon but must be considered because of the relatively low voltages at which ac corona is initiated.

Corona offset voltage is the ac rms voltage at which corona once initiated is extinguished as the voltage is reduced. It should always be used as the criterion for establishing safe operating levels. The corona onset voltage is defined as the point at which corona begins to occur as the voltage is increased from zero. The offset voltage is the lower of the two values.

2.4 CAPACITORS, PAPER AND PLASTIC

A typical curve of empirically determined corona offset voltage as a function of dielectric thickness (dc voltage rating) is shown in Figure 44. These curves apply to unimpregnated Mylar capacitors. While the curves show the results of tests conducted at 25°C, other data indicates that corona offset voltage does not decrease significantly at temperatures up to 125°C. These curves represent upper limits for any impregnated plastic dielectric capacitor, since the presence of air voids is the problem in all cases.

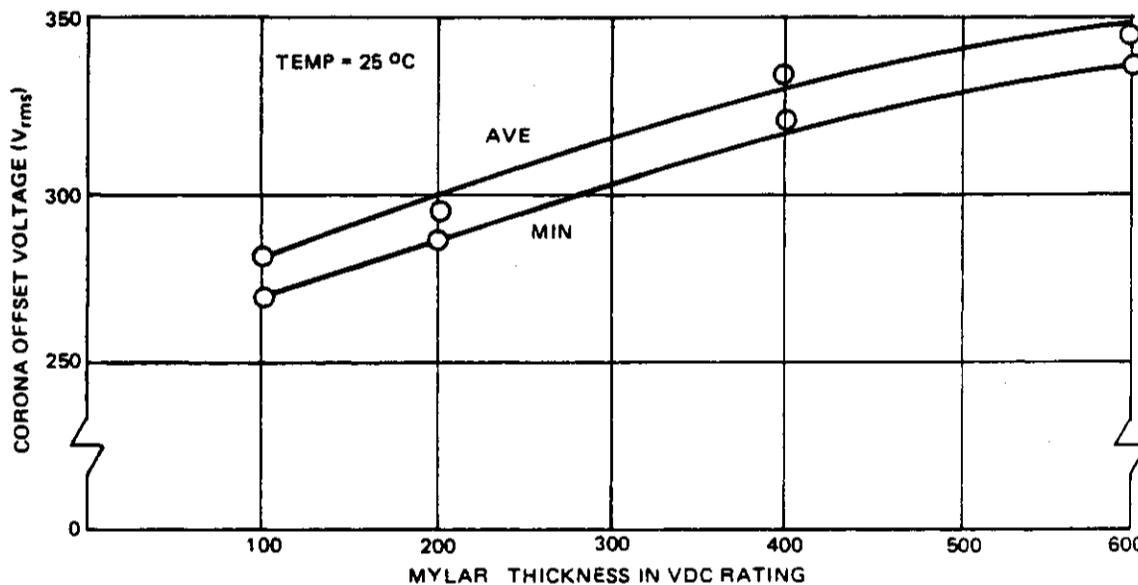


FIGURE 44. Corona offset voltage vs Vdc rating.

The final and often principal limiting factor in selecting a capacitor for ac application is internal heat loss, represented by I^2R . The "R" in this expression is composed of the following series elements:

- a. The resistance of the metals used for the leads, electrodes, solder, and metal spray. This resistance is primarily controlled in initial design stages by choice of materials, sizes, etc.
- b. The inherent equivalent series resistance of the dielectric material. This resistance is also controlled by initial choice of material.
- c. The resistance of the oxides resulting from the interface connections between the various elements comprising these connections. The main controls on the resistance of these oxides are manufacturing processes and workmanship.

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If other factors have been optimized to provide minimum series resistance, the choice of dielectric material becomes the limiting item in minimizing heat generation. Figure 42 shows that some dielectrics have significantly low dissipation factors, particularly polystyrene, Teflon and polycarbonate. Although these differences compared with paper or Mylar are insignificant in most applications, they can be of considerable importance under ac conditions.

Polystyrene and Teflon have other limitations as dielectrics, e.g. limited operating temperature for polystyrene and large physical size for Teflon. Polycarbonate is an excellent choice for ac operation. In addition to its low dissipation factor, polycarbonate also has an excellent temperature coefficient, is capable of operation up to 125°C, and is readily available in metallized form.

Figure 45 shows some comparative ac ratings for typical metallized polycarbonate, metallized paper-Mylar, and impregnated paper styles.

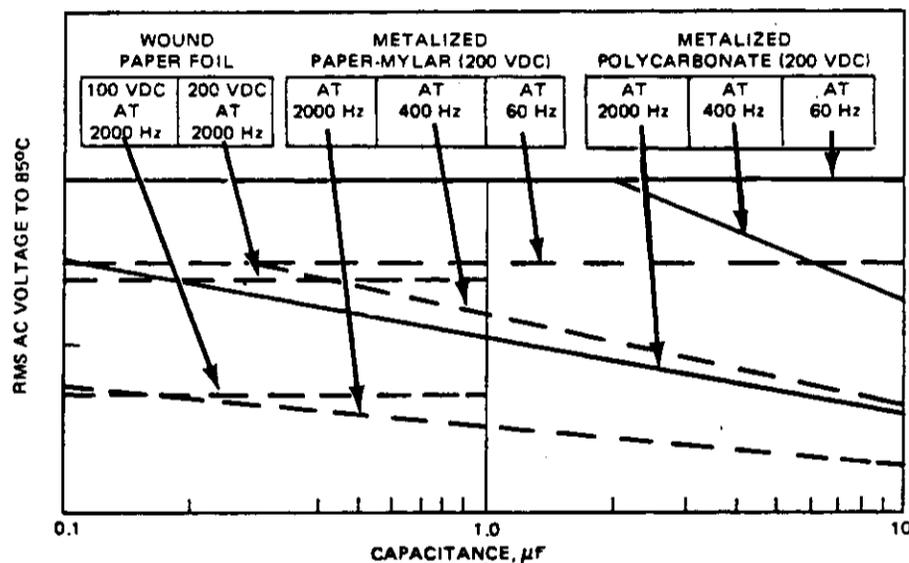


FIGURE 45. Dielectric loss characteristics.

2.4.5.8 Effects of frequency. Dielectric losses tend to increase at high frequencies. The characteristics of high quality paper and plastic dielectric are such that dielectric losses can be considered constant over the usable frequency range. The upper usable frequency is then limited by the self-resonant frequency of the device.

As a general rule, the self-resonant frequency of the paper-plastic family can be considered a function of capacitance value and lead length. Figure 46 can be used as a guide to the typical self-resonant frequency of all tubular styles, regardless of dielectric material, voltage rating, or case size.

2.4 CAPACITORS, PAPER AND PLASTIC

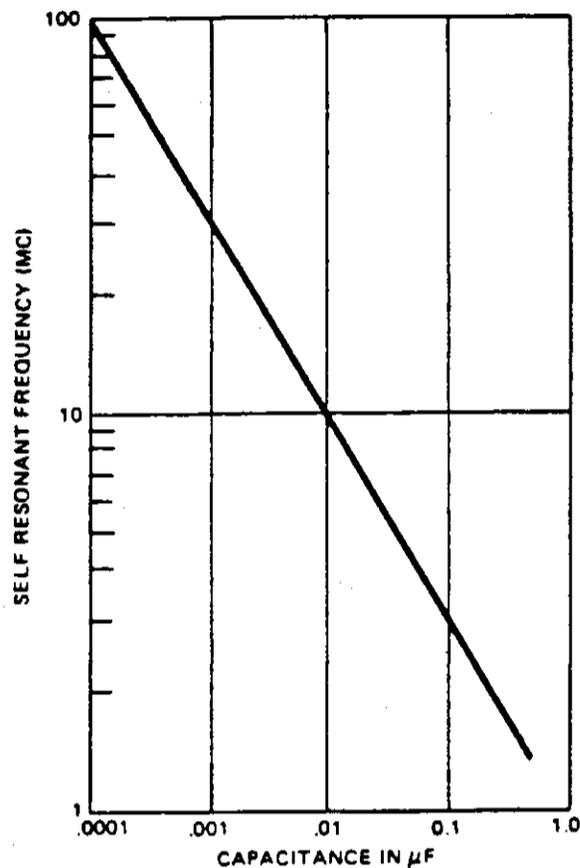


FIGURE 46. Typical self-resonant frequency as a function of capacitance for tubular capacitors with 1/4-inch leads.

2.4.5.9 Effects of temperature. The capacitance variation with temperature is often a prime consideration in selection of dielectric type. Figure 47 illustrates typical temperature characteristics of the dielectric types.

2.4.5.10 Dielectric absorption. Dielectric absorption of unimpregnated plastic dielectric capacitors is 0.2% or less.

For oil-impregnated paper styles, the figure is about 2%, which is essentially the dielectric absorption of the oil. For oil-impregnated plastic dielectric styles, the dielectric absorption also rises to 2%.

2.4.6 Environmental considerations. Except for MIL Type CTM capacitors, which are Mylar dielectric units enclosed with an outer Mylar wrap and epoxy end fill, most paper and plastic capacitors are of hermetically sealed metal case construction. As such, they are highly resistant to moisture and other hazardous environments and operate reliably when used well within their ratings.

2.4 CAPACITORS, PAPER AND PLASTIC

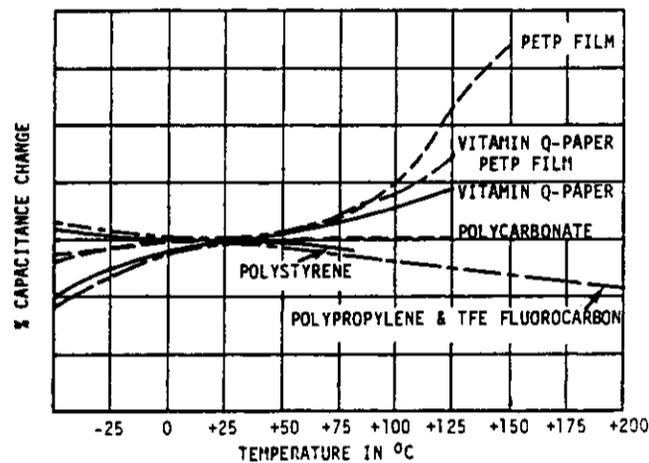


FIGURE 47. Capacitance change characteristics at different temperatures for various dielectric materials.

2.4.6.1 Vibration. Although most paper and plastic capacitors are rated for operation at vibration levels up to 15 G and 2000 MHz, the capacitor specifications require that the devices be rigidly mounted by the body during qualification tests. For severe vibration requirements, particularly in the larger case sizes, supplementary mounting means should be used to prevent part failure due to lead fatigue.

2.4.7 Reliability considerations.

2.4.7.1 Failure modes and mechanisms. Paper and plastic capacitors are subject to two primary failure modes: opens and shorts. This includes intermittent opens and intermittent or high-resistance shorts. In addition, a capacitor may fail in other ways, such as capacitance drift, high dissipation factor, instability with temperature, and low insulation resistance.

Failures can be the result of manufacturing defects, electrical, mechanical, or environmental overstress, or eventual wearout due to dielectric degradation with operation.

Open capacitors usually occur as a result of manufacturing defects, but occasionally result from overstress in application. Failure analysis indicates that about 40% of capacitor failures are "opens", and the large majority of this population is made up of paper or plastic types. Most of the open type failures are the result of a poorly consummated solder or weld joint between the end or the roll and the wire lead.

Such latent failures will often endure several hours of operation with no indication of failure, but will eventually manifest themselves as open, intermit-

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2.4 CAPACITORS, PAPER AND PLASTIC

tent, or high-resistance joints (high dissipation factor) under some set of operating conditions, sometimes over a very limited range of ambient temperature. The frequency of occurrence of this problem is greatest with low capacitance values in small-diameter cases, particularly in diameters of less than 0.250 inch.

As noted above, an open condition can sometimes be caused by operating over-stress. For instance, operation of dc rated capacitors at high ac current levels can cause localized heating at one of the internal terminations as a result of I^2R losses at that point. Continued operation results in oxidation of the joint, increased termination resistance and heating, and eventual failure of the connection. This is one reason that care must be exercised in ac applications.

Capacitance drift is not normally of serious consequence as a failure mode since most applications tolerate relatively wide variations in capacitance value without equipment failure. Most specifications allow a maximum change of $\pm 10\%$ for paper and plastic devices when subjected to a life test of 1000 hours at maximum rated voltage and temperature. Derating by the design engineer will enhance the long-term stability of these capacitors.

Temperature instability beyond specified manufacturers' limits rarely occurs, but can be induced by excessive clamping pressures on nonrigid containers, or can result from loose windings.

Insulation resistance failures usually result from moisture entrapped in the winding or case during manufacture, or from operation of nonhermetically sealed or improperly sealed units under prolonged exposure to a humid atmosphere. Particular care must be taken by the manufacturer in the drying and impregnating of paper capacitors, since Kraft paper contains about 13% water in its natural form. Failure to completely remove this water or to thoroughly impregnate the winding will result in high leakage during operation.

Dielectric breakdown and a consequent shorted condition is predominantly the most common failure mode, and nearly always the ultimate reason for failure of a properly designed and manufactured unit.

All the active dielectric material in capacitors is subjected to the full potential to which the capacitor is charged. To achieve small physical size, relatively high electrical stress levels are common. Breakdowns develop after many hours of satisfactory operation. There are numerous causes for these breakdowns, many associated with slowly changing physical or chemical conditions. The ultimate failure is sometimes brought about by abnormal electrical or mechanical stress.

If a conducting particle is embedded among the electrode and dielectric layers, it may cause an immediate short circuit. However, if it is sufficiently small and blunt, it may only create a region of high electrical and mechanical stress.

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This particle will always be under pressure and will tend to slowly push through the adjacent materials. Many dielectrics become softer with increasing temperature, therefore increasing the internal pressure that produces the particle penetration. Thus, higher temperatures promote earlier failure.

Contamination during assembly and traces of impurities in basic materials can cause general or localized degradation of the dielectric or electrode. The process is basically chemical and higher temperatures result in greater activity and earlier failure. One common contaminant is water, which promotes hydrolysis. Along with the associated decomposition, this process provides an abundance of ions to initiate a discharge through a weak area.

Many dielectric materials, particularly paper and the plastics, go through a slow aging process wherein they gradually become more brittle and susceptible to cracking. The higher the temperature, the faster the process. Once the capacitor has aged, it becomes particularly susceptible to temperature cycling, which produces excessive stresses in the capacitor body.

Dielectric breakdown may also occur as a result of switching transients or voltage surges induced by malfunction of associated circuitry. Transient exposure is of particular concern. The device may survive several applications of temporary overvoltage without apparent degradation, but repeated overstress will eventually cause premature breakdown.

2.4.7.2 Screening. Proper screening techniques can be highly effective in screening out manufacturing defects and potential early life failures. Screens for paper and plastic capacitive devices usually include temperature cycling, ESR and DF measurement high frequencies, operating voltage conditioning, and seal test.

2.4.7.3 Reliability derating. Capacitor failure rate rises at an increasing rate with applied voltages and temperatures. Failure rate of paper and plastic capacitors is proportional to a power of the ratio of the applied to the rated voltage, where the value of the exponent is usually in the range of 3 to 6, depending on the type of dielectric and the actual portion of the operating range under consideration. The general rule of 50% decrease in operating life with 10°C increase in operating temperature up to rate conditions also applies.

2.4.7.4 Failure rate determination. Basic failure rate can be determined from MIL-HDBK-217 for any combination of voltage stress and ambient temperature within the rating of the device. This failure rate should then be multiplied by the K-factor applicable for the intended use. For flight applications, the failure rate obtained from the curves must be multiplied by 10.

These data are intended only as a guide. They indicate the general area of failure rate to be expected with paper and plastic capacitors in particular areas of application. For quantitative reliability predictions, current failure rate data for the part type as procured to a particular specification must be considered.

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2.5 CAPACITORS, TANTALUM FOIL

2.5 Tantalum foil.

2.5.1 Introduction. Foil type tantalum capacitors are probably the most versatile of all electrolytic capacitors. They are available in both polar and nonpolar construction and in voltage ratings from 3 to 450 volts. They are capable of operation at 125°C with proper derating, and are electrically the most rugged of the three basic tantalum types.

These capacitors are most commonly supplied in nonhermetically sealed styles, with elastomer seals at either end of the tubular metal case. They are also available in hermetically sealed cases except in the smaller case sizes. The elastomer sealed style is both more economical and more readily available.

Their prime disadvantages when compared with other tantalum types are relatively large size, fairly large change in capacitance with temperature, and high equivalent series resistance especially at cold temperatures. In addition, they are not suitable for timing or precision circuits due to their very wide design tolerances. Etched foil types can have as much as ten times the capacitance per unit area as plain foil types. However, the plain foil types can withstand 30 percent higher ripple current, have better capacitance temperature characteristics, and a low power factor (lower dissipation factor). Plain foil types are as reliable as etched foil types. The life and capacitance-temperature (stability at temperature extremes) characteristics of these devices are excellent.

2.5.2 Usual applications.

2.5.2.1 Polarized styles. The polarized foil types are used where low-frequency pulsating dc components are to be bypassed or filtered out and for other uses in electronic equipment where large capacitance values are required and comparatively wide capacitance tolerances can be tolerated. When used for low-frequency coupling in transistor circuits allowance should be made for the leakage current.

This leakage current could cause excessive base, emitter, or collector currents. These polarized capacitors should be used only in dc circuits with polarity properly observed. If ac components are present, the sum of the peak ac voltage plus the applied dc voltage must not exceed the dc voltage rating. Even though those units rated at 6 volts and above can withstand a maximum of 3 volts in the reverse direction, it is recommended that they not be used in circuits where this reversal is repetitious.

2.5.2.2 Nonpolarized styles. The nonpolarized types are primarily suitable for ac applications or where dc voltage reversals occur. Examples of these uses are in tuned low-frequency circuits, phasing of low voltage ac motors, computer circuits where reversal of dc voltage occurs, and servo systems.

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2.5.3 Physical construction. These capacitors consist essentially of two thin tantalum foil sheets, approximately 0.5 to 1.0 mils thick, with a tantalum wire lead spot-welded to each foil. The anode foil is electrochemically treated to form tantalum pentoxide (Ta_2O_5) on the surface of the foil. This extremely thin oxide is the dielectric material of the capacitor. The cathode foil is left unanodized. See Figure 49.

The two foils are then wound into a cylindrical configuration with two or three sheets of Kraft paper as spacers. These spacers serve a dual purpose. They prevent the possibility of short-circuits between the two foils as a result of rough surfaces or jagged edges on the foils and when later impregnated with electrolyte, they help to maintain intimate and uniform contact of the electrolyte with all surfaces of the anodized foil.

The rolls are then taped to prevent unwinding and inserted into a metallic case. They are impregnated with a suitable electrolyte (e.g. ethylene glycol) and sealed. The tantalum leads are brought out through the end seals. A solderable lead, usually nickel, is butt-welded to the tantalum wire. See Figure 48.

For nonpolar units, the construction is the same as outlined above except that the surfaces of both foils are formed with an oxide dielectric film.

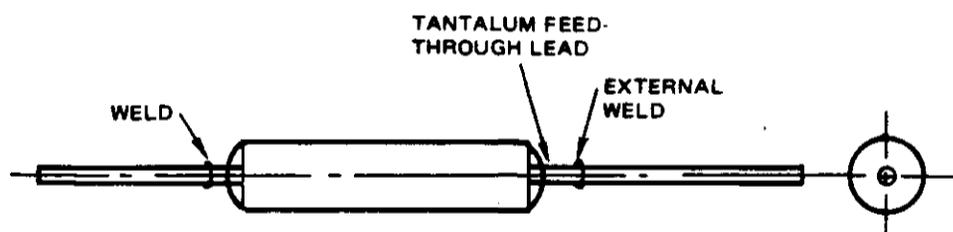


FIGURE 48. Outline drawing of a tantalum capacitor (CLR style).

2.5.3.1 Etching. Both polar and nonpolar units are available either in plain foil styles or with various degrees of etched foil. By etching the surface of the tantalum foil, it is possible to increase the surface area several-fold and to correspondingly increase the capacitance. This increased capacitance, however, is attained at the expense of higher dissipation factor, poorer capacitance-temperature characteristics, and lower ripple current-carrying ability than the plain foil styles.

2.5 CAPACITORS, TANTALUM FOIL

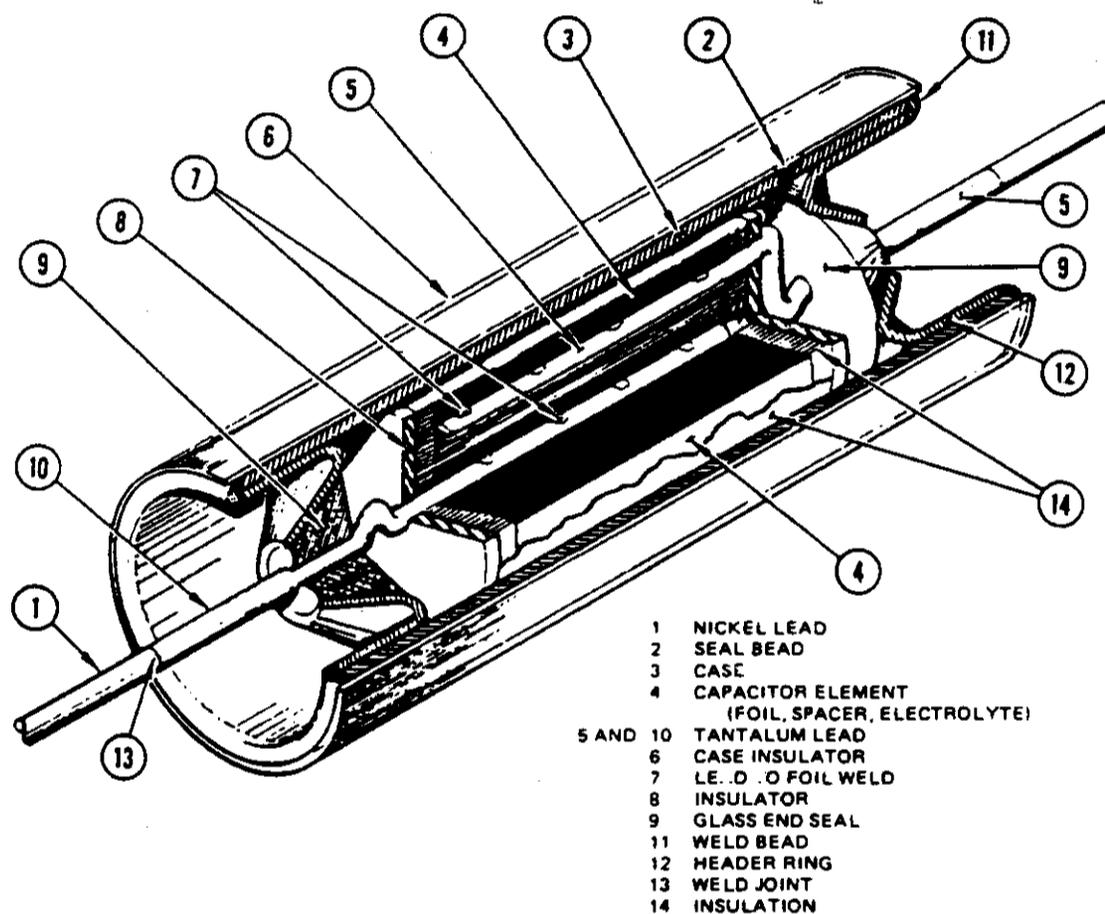


FIGURE 49. Typical construction of a tubular tantalum foil capacitor.

2.5.3.2 Mounting. These capacitors are not intended to be mounted by their leads, particularly in the larger case sizes. Clips or other restraining devices should be used to prevent lead breakage in shock and vibration environments.

In addition, the elastomer end seals on the nonhermetically sealed styles provide little restraint to torsional stress on the body of the capacitor. If the body is twisted after the leads have been soldered into place to examine the

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2.5 CAPACITORS, TANTALUM FOIL

marking, it is quite possible that the foils will be torn away from the leads at the internal welds. This results in an open or intermittent capacitor. The smaller case sizes, and particularly the etched foil types, are especially susceptible to such mishandling. For this reason the capacitors should be assembled with the marking properly exposed.

2.5.4 Military designation.

2.5.4.1 Applicable military specification. Tantalum foil capacitors are covered by the following specification:

MIL-C-39006 CLR (established reliability) styles

2.5.4.2 Military type designations. The following is an example of a typical military callout, along with a description of the significance of each of the letters and digits in the designation. This is included only for reference information.

Type designation example

CLR25	B	D	600	U	G	S
Style	Characteristic	Voltage	Capacitance	Capacitance tolerance	Type of seal	Failure rate level

2.5.5 Electrical characteristics.

2.5.5.1 Derating. The failure rate of tantalum foil capacitors under operating conditions is a function of time, temperature, and voltage. Refer to MIL-STD-975 (NASA) for specific derating conditions.

2.5.5.2 End-of-life design limits. When operated under electrical and environmental conditions defined in the applicable military specification, the capacitance can be expected to change by $\pm 15\%$. The leakage current can increase to 130% of the initial value.

2.5.5.3 Voltage ratings. The dc rated voltages for typical military styles are shown in Table IX.

2.5.5.4 Operating temperature range. These capacitors are suitable for operation over a temperature range of -55° to $+85^{\circ}\text{C}$ with full rated voltage applied.

2.5.5.5 Reverse voltage. While it is advisable to operate polarized styles only in the forward direction, these units are capable of withstanding reverse voltages up to a value of 3 volts without damage. Nonpolarized styles may be operated at full rated voltage in either direction.

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TABLE IX. Voltage ratings

Style	Anode	Voltage Range (volts)
CLR25	Etched foil	15 to 150
CLR27	Etched foil	15 to 150
CLR35	Plain foil	15 to 450
CLR37	Plain foil	15 to 375
CLR71	Etched foil	15 to 150
CLR73	Etched foil	15 to 150

2.5.5.6 Ripple voltage. Tantalum foil capacitors are the only tantalum electrolytic capacitors capable of operating continuously on unbiased ac voltages. Nonpolarized styles may be operated continuously on an unbiased ac voltage with a peak value of 150 volts or the dc rating of the capacitor, whichever is less. Polarized styles must be biased to prevent dc voltage reversals.

However, in most applications, voltage is not the limiting factor. Except at relatively low frequencies, ripple is generally limited by the I^2R loss in the capacitor. This loss is high because of the relatively high equivalent series resistance. The allowable ripple current is a function of case size, capacitance value, frequency, and ambient temperature.

When complex ripple wave shapes are involved they should be measured on an oscilloscope or by some other method which will give the peak rating. Tantalum foil capacitors should be limited to operation at ripple frequencies between 60 and 10,000 Hz. Above 10,000 Hz effective capacitance rapidly drops off to the point where these devices act as practically pure resistance at frequencies of only a few hundred kHz.

Figure 50 indicates maximum allowable ripple voltage and current for typical tantalum foil styles. This figure shows allowable rms values at 60 Hz and 25°C for the most popular case sizes, as a function of case size and capacitance value. Note that for etched foil styles, the values must be multiplied by 0.5.

The allowable ripple obtained from Figure 50 must then be multiplied by a frequency correction factor (Figure 51), and a temperature derating factor (Figure 52). Thus, for operation with an 800 Hz ripple frequency at 35°C ambient, the voltage or current value obtained from Figure 50 must be reduced by a factor of approximately 8.

2.5.5.7 DC leakage current. The leakage current of foil tantalum capacitors ranges from less than 1 μA to 100 μA or more depending on electrical rating, the type of construction, and temperature.

2.5 CAPACITORS, TANTALUM FOIL

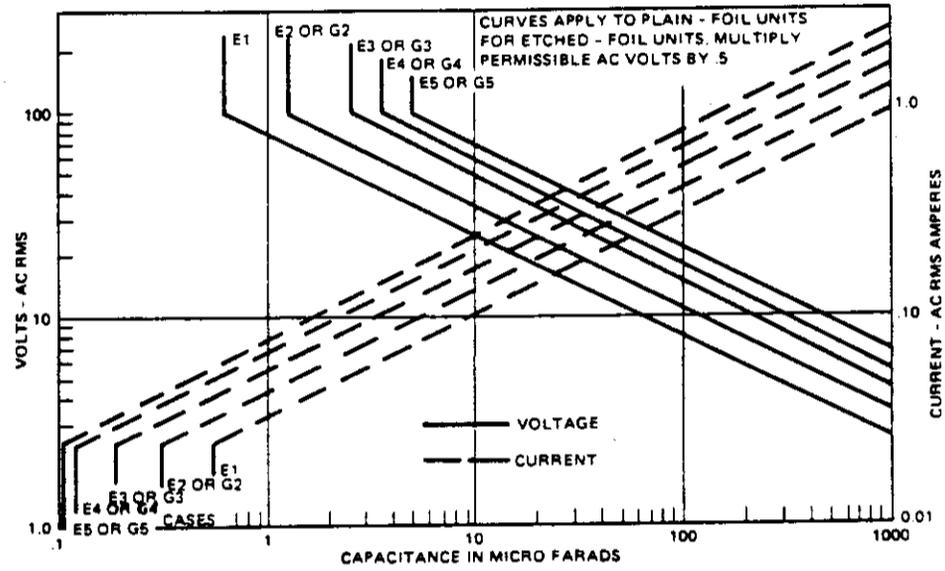


FIGURE 50. Maximum allowable ripple voltage and current for styles CLR25, CLR27, CLR35 & CLR37.

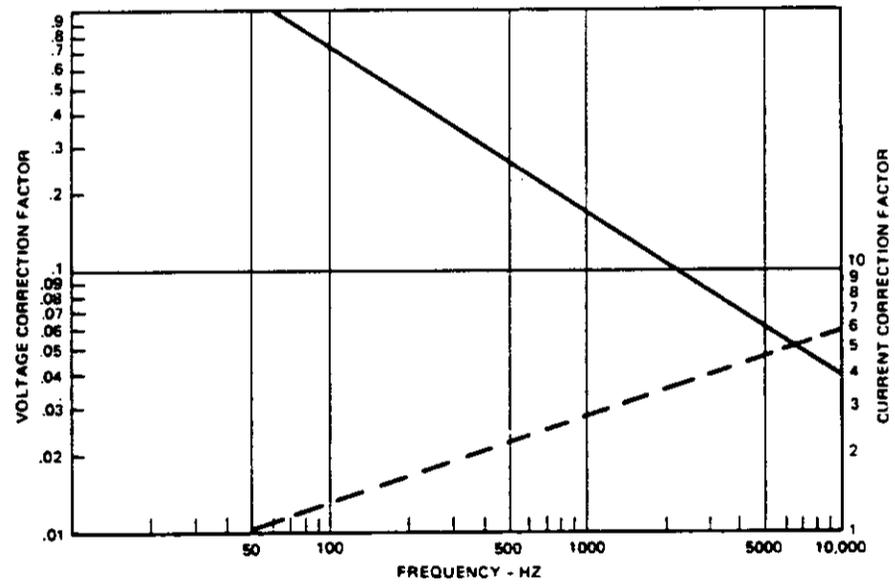


FIGURE 51. Correction factor for maximum allowable ripple current vs frequency for tantalum foil capacitors.

2.5 CAPACITORS, TANTALUM FOIL

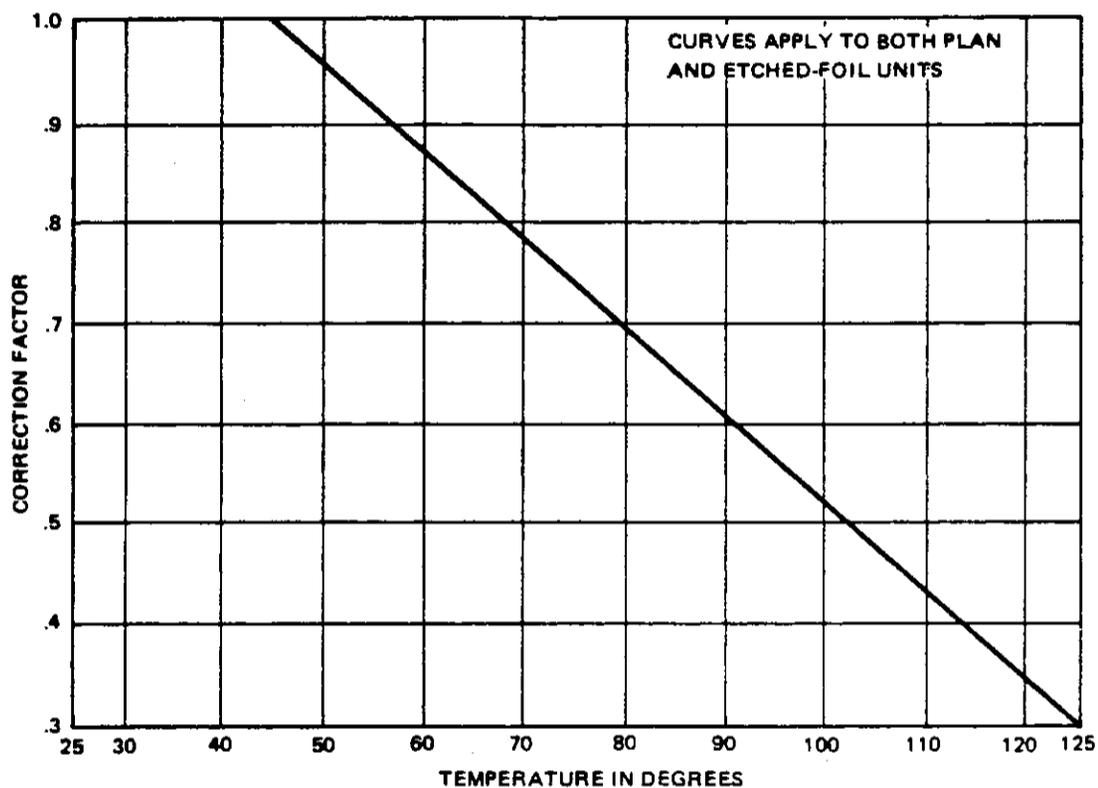


FIGURE 52. Correction factor for maximum allowable ripple voltage vs temperature for tantalum foil capacitors.

Leakage current, as in all tantalum capacitors, is the result of minute faults in the dielectric film. These faults tend to be self-healing under applied voltage, so that leakage current will normally decrease exponentially with life.

Leakage current is roughly proportional to applied voltage up to the maximum dc voltage rating.

Leakage current increases rapidly with temperature as shown in Figure 53. Note that leakage at 125°C is about 30 times the room temperature value.

2.5.5.8 Effects of frequency. Capacitance, dissipation factor or effective series resistance, and impedance will vary with the frequency of the applied voltage.

Capacitance vs frequency. Typical curves are shown in Figure 54 for frequencies up to 5 kHz.

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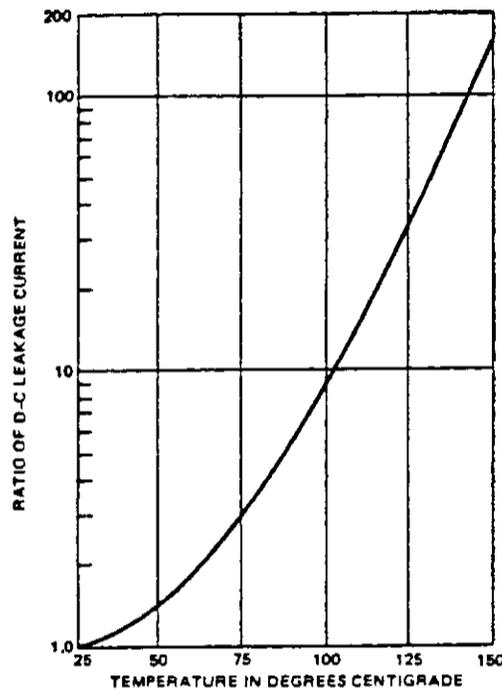


FIGURE 53. Typical curve, ratio of dc leakage current at rated voltage vs temperature for foil tantalum capacitors.

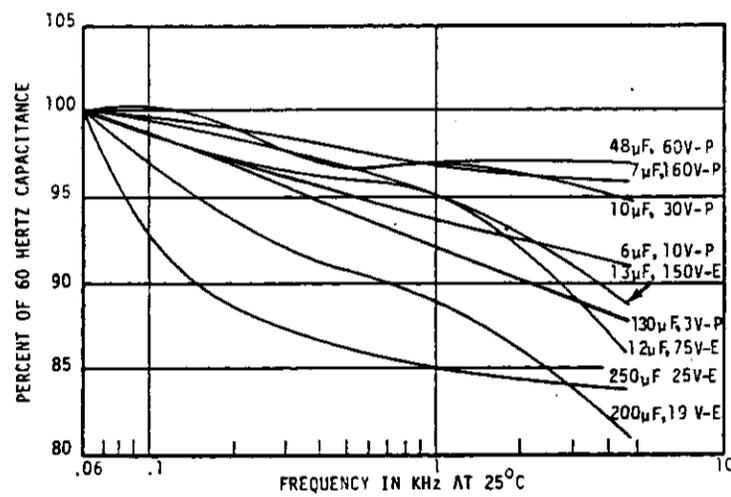


FIGURE 54. Effect of frequency on capacitance of typical foil type tantalum capacitors.

2.5 CAPACITORS, TANTALUM FOIL

Dissipation factor vs frequency. Dissipation factor (ESR/X_C) increases rapidly with frequency for the higher capacitance values, mainly because of the large time decrease in capacitive reactance relative to the fairly small decrease of equivalent series resistance. See Figure 55 for typical curves.

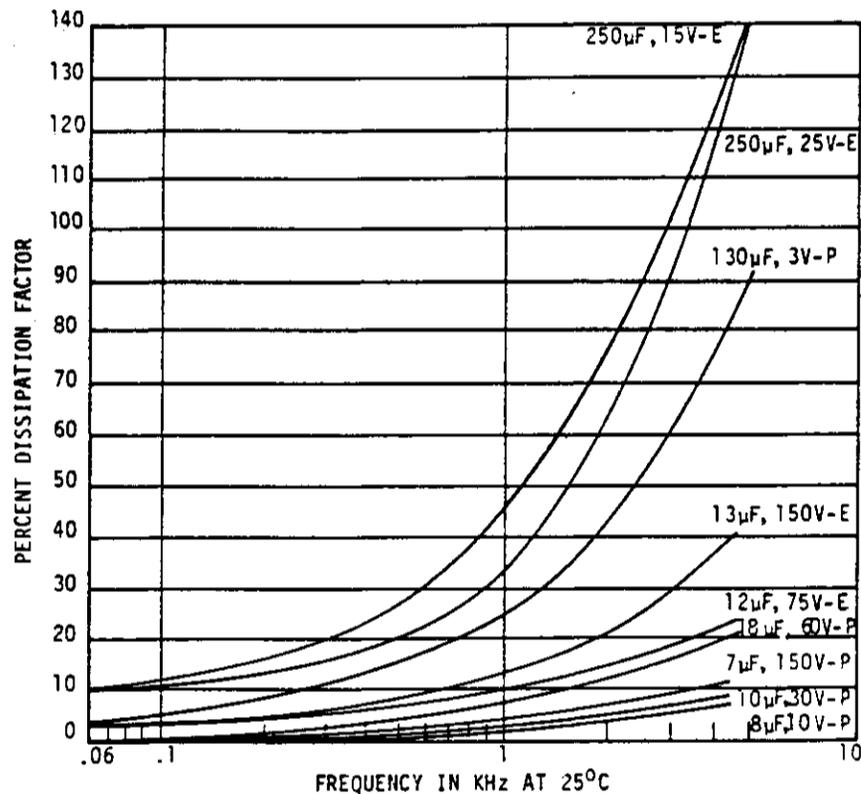


FIGURE 55. Effect of frequency on dissipation factor of foil type tantalum capacitors.

Impedance vs frequency. Figure 56 shows typical curves for various capacitance values over the range from 60 Hz to 100 MHz for foil tantalum capacitors. The curves are similar to those of other tantalum types, with a downward slope at low frequencies, a trough in the range about 10-500 KHz, and an inductive increase in impedance at higher frequencies. Figures 57, 58 and 59 illustrate temperature correction factors to be applied to the impedance value obtained from Figure 56. Since equivalent series resistance of tantalum foils increases significantly at low temperatures, these correction factors must be taken into consideration in design. Note that impedance at -55°C at a given frequency may be several times the 25°C value.

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2.5 CAPACITORS, TANTALUM FOIL

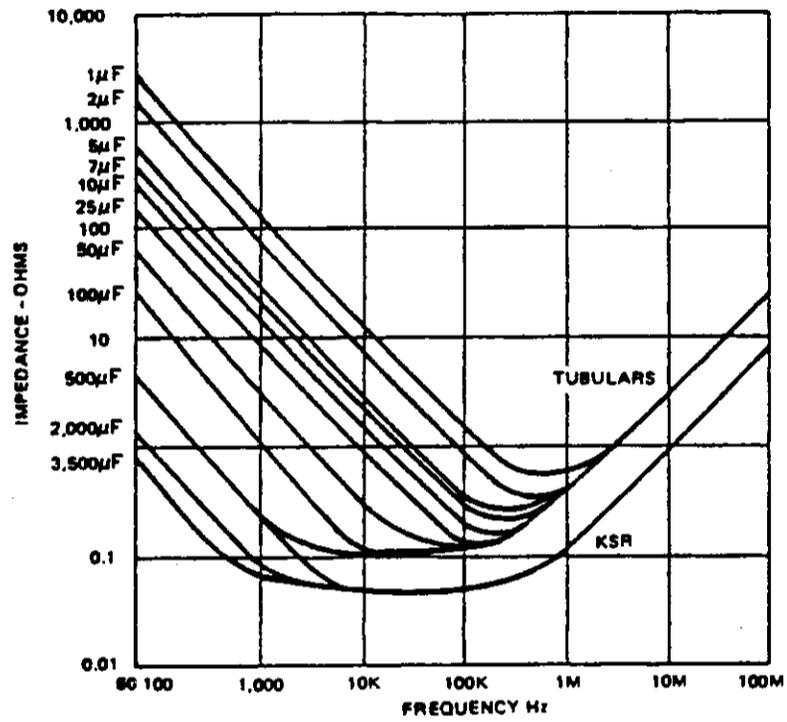


FIGURE 56. Impedance curves for tantalum foil capacitors at 25°C.

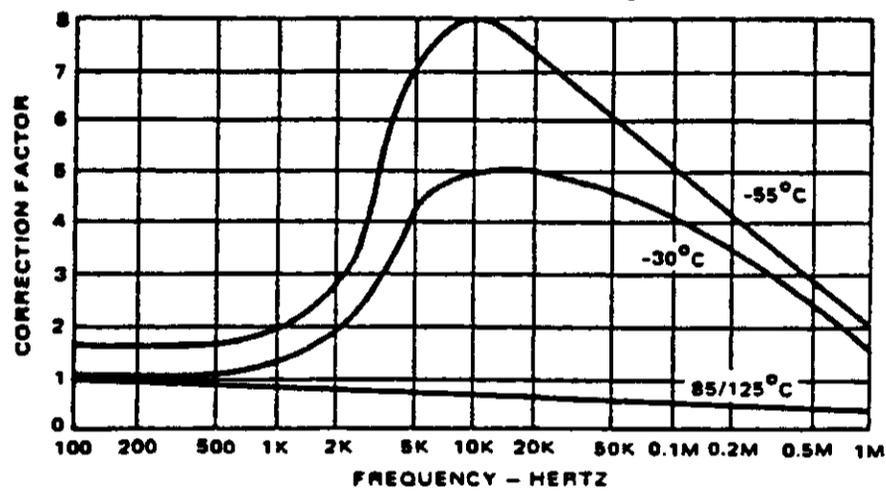


FIGURE 57. Tantalum foil impedance correction factors for capacitance up to 2µF.

2.5 CAPACITORS, TANTALUM FOIL

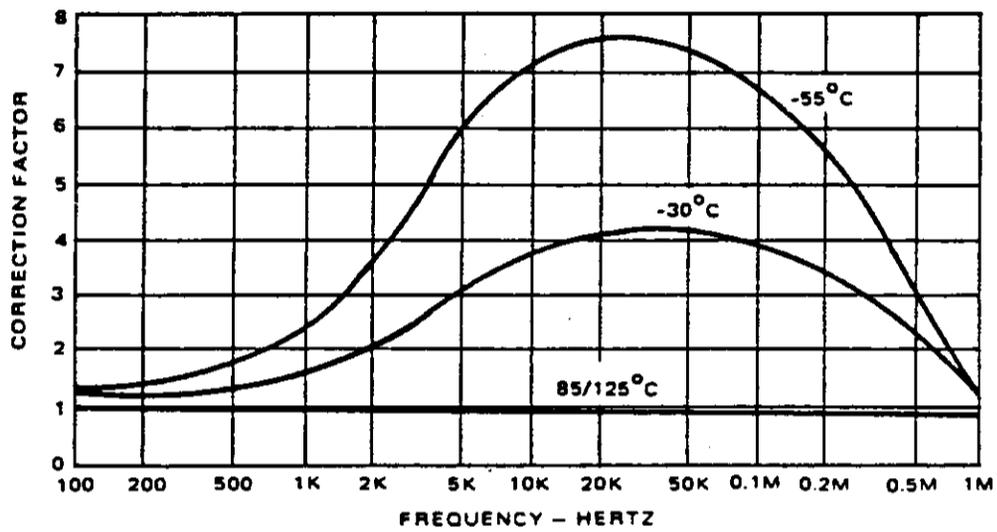


FIGURE 58. Tantalum foil impedance correction factors for capacitance 2-50 μF

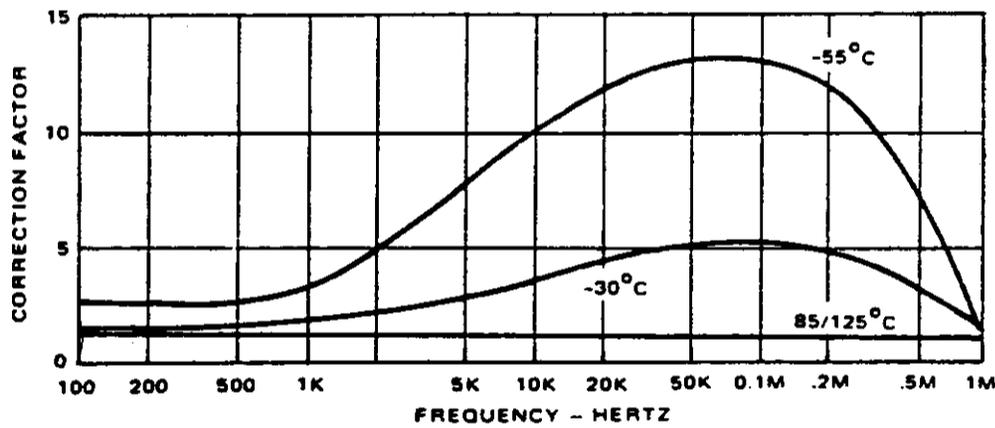


FIGURE 59. Tantalum foil impedance correction factors for capacitance 50 μF and over.

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2.5 CAPACITORS, TANTALUM FOIL

2.5.5.9 Circuit impedance. No special precautions are required with foil tantalum capacitor current limiting. These units will withstand sudden inrush and discharge currents without deleterious effects.

2.5.5.10 Series applications.

Series operation. Whenever tantalum capacitors are connected in series for higher voltage operation, a resistor should be paralleled across each unit. Unless a shunt resistor is used, the dc rated voltage can easily be exceeded on the capacitor in the series network with the lowest dc leakage current. To prevent capacitor destruction, a resistance value not exceeding a certain maximum should be used. This value will depend on capacitance, average dc leakage, and capacitor construction for plain foil types. The resistance across each capacitor should not exceed $5/C$ megohms, where C is in microfarads. For etched foil types use $15/C$ megohms.

Parallel operation. When tantalum foil capacitors are connected in parallel the sum of the peak ripple and the applied dc voltage should not exceed the recommended derated dc voltage rating of the capacitor with the lowest rating. The connecting leads of the capacitors in parallel should be large enough to carry the combined currents without reducing the effective capacitance due to series lead resistance.

2.5.6 Environmental considerations.

2.5.6.1 Stability and life. Tantalum electrolytic capacitors have excellent life and shelf life characteristics. Life, at higher temperatures will show a comparatively lower decrease in capacitance. With rated voltage applied, more than 4,000 hours of life can be expected at +85°C. These devices may be expected to operate at least 1,000 hours at +85°C with less than 10 percent loss of capacitance.

Because the more stable tantalum film is less subject to dissolving by the surrounding electrolyte than the film in an aluminum capacitor, the shelf life of the tantalum unit is much longer, and less reforming is required. After storage for long periods, the reforming current required is low and the time required for reforming is comparatively short. Reforming may be expected to take less than 10 minutes. These properties are affected by the storage temperature to a significant degree, being excellent at temperatures from -55° to +25°C; good at +65° and relatively poor at +85°C.

2.5.6.2 Effects of temperature. The characteristics of these capacitors vary significantly with temperature, particularly with low temperatures.

Capacitance (C). Capacitance change with temperature is positive and depends on capacitance value and voltage rating. Capacitance change ranges from a maximum allowable of -20% to -40% at -55°C to an increase of +10 to +50% at maximum operating temperature.

2.5 CAPACITORS, TANTALUM FOIL

Equivalent series resistance (ESR). ESR increases rapidly at temperatures below 0° as shown in Figure 60 for a typical capacitor. It may also increase at temperatures over 80°C and voltages over 50 Vdc. Figure 61 depicts variation of impedance with temperature and frequency. The increase in impedance at -55°C in the mid frequencies is essentially due to an increase in ESR.

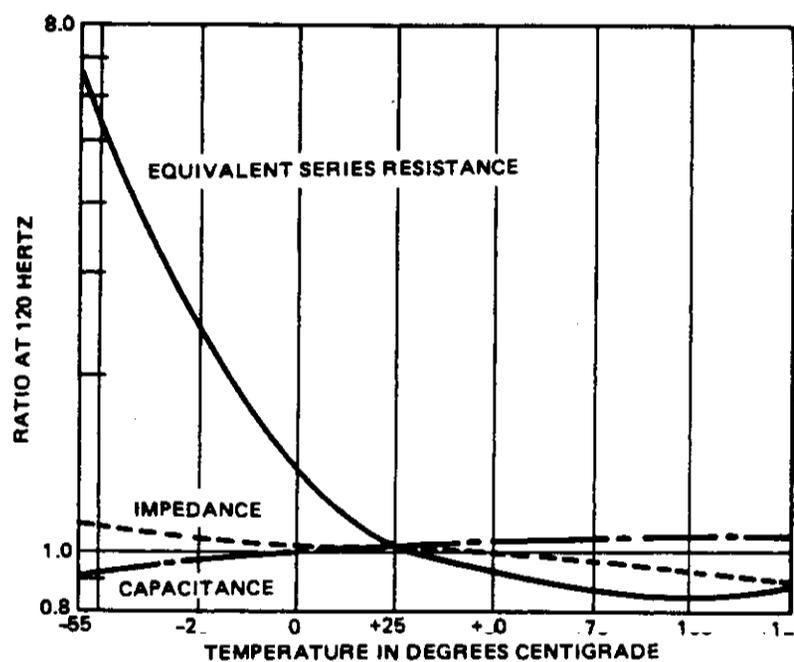


FIGURE 60. Typical curves of impedance, capacitance, and equivalent series resistance with temperature for 26- μ F, 100-volt polarized etched-foil capacitors.

2.5.7 Reliability considerations. Foil tantalum capacitors properly used are the most reliable of the three types of tantalum capacitors. Unless subjected to electrical or mechanical overstress, the normal failure mode is by degradation rather than a complete open or short.

2.5.7.1 Failure modes and mechanisms. Except for manufacturing defects, such as poor welds or improper end seals, a foil tantalum capacitor eventually fails by a decrease in capacitance beyond some acceptable limit. This results from vaporization of the electrolyte and its escape through the end seals, so that the capacitor dries out. Such failures are unlikely for several thousand hours of operation when used within their ratings.

2.5 CAPACITORS, TANTALUM FOIL

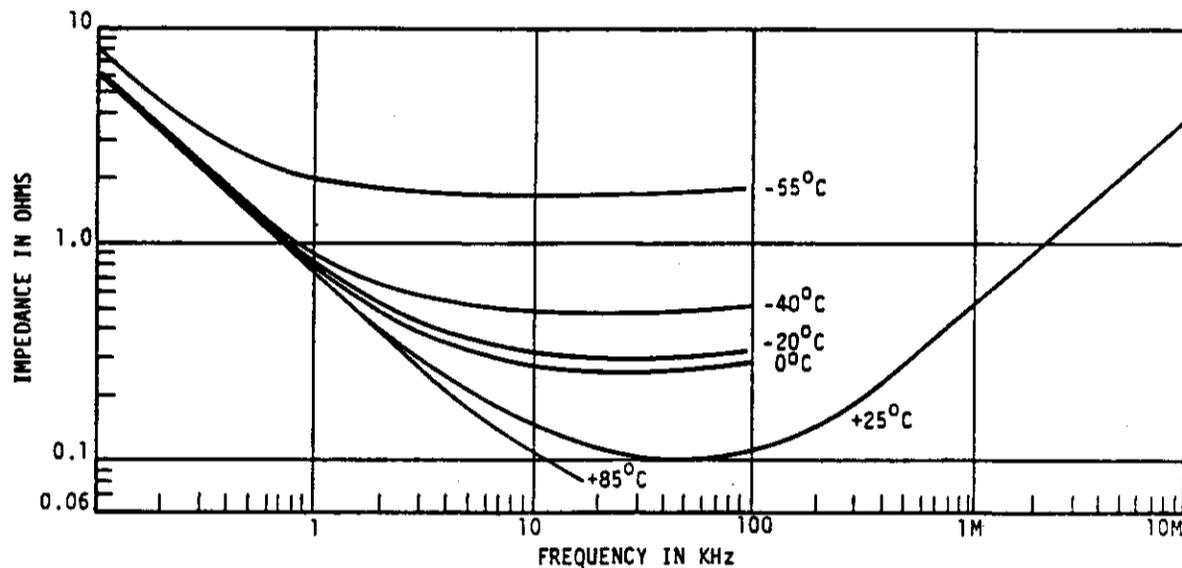


FIGURE 61. Typical curves of impedance with frequency at various temperatures for 200- μ F, 6-volt plain-foil capacitors.

2.5.7.2 Screening. Established reliability specification MIL-C-39006 requires 100% operating voltage conditioning to screen out potential early life failures. Capacitors which exhibit excessive leakage are screened out as reliability hazards. For hermetically sealed types, a 100% seal test is also conducted.

Mounting. Clips or other restraining devices/methods should be used to prevent lead breakage and fatigue in shock and vibration environments.

2.5.7.3 Failure rate level determination. Consult MIL-HDBK-217 for current data on the particular style and quantitative reliability level predictions.

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2.6 CAPACITORS, SOLID TANTALUM

2.6 Solid tantalum.

2.6.1 Introduction. Solid tantalum capacitors are the most widely used electrolytic capacitors for electronics equipment. They have high volumetric efficiency, good stability with time and temperature, and are reliable devices when properly applied.

Because the electrolyte used is solid and dry, these capacitors have a more stable capacitance-temperature characteristic than any of the other electrolytic capacitors. Maximum capacitance variation is less than 10% over the operating temperature range of -55°C to $+125^{\circ}\text{C}$.

Their limitations are relatively high leakage current, possible dielectric punctures, limited voltage range available (6 to 100 volts), and a maximum allowable reverse voltage of 15 percent of the rated dc voltage at $+25^{\circ}\text{C}$ and only 1 percent at $+125^{\circ}\text{C}$.

These capacitors are available in polarized and nonpolarized units under military type designations.

2.6.2 Usual applications. These capacitors are generally used where low-frequency pulsating dc components are to be bypassed or filtered and where large capacitance values are required where space is at a premium, and where there are significant levels of shock and vibration. These capacitors are mainly designed for filter, bypass, coupling, blocking, energy storage, and other low voltage dc applications (such as transistor circuits in missile, computer, and aircraft electronic equipment) where stability, size, weight, and shelf life are important factors.

2.6.3 Physical construction. The anode consists of tantalum powder mixed with an organic binder and pressed into pellet form. The pellets are then sintered in a vacuum oven to decompose and evaporate the binder, yielding a pellet of high porosity and high surface area.

The pellets are anodized in an acidic bath to form the tantalum pentoxide dielectric on all surfaces reached by the electrolyte. The pellets are then impregnated with an aqueous solution of manganous salt, which is pyrolytically decomposed to yield manganese dioxide. The manganese dioxide is the working electrolyte in solid form. A carbon compound is applied over the surface of the manganese dioxide (MnO_2) to allow for application of a silver paint. The completed pellet is then inserted into a pool of molten solder inside the metal case which is then sealed. The solder is displaced up around the sides of the pellet to provide an electrical and mechanical bond to the case. Figure 62 shows the construction of a typical solid tantalum capacitor.

2.6 CAPACITORS, SOLID TANTALUM

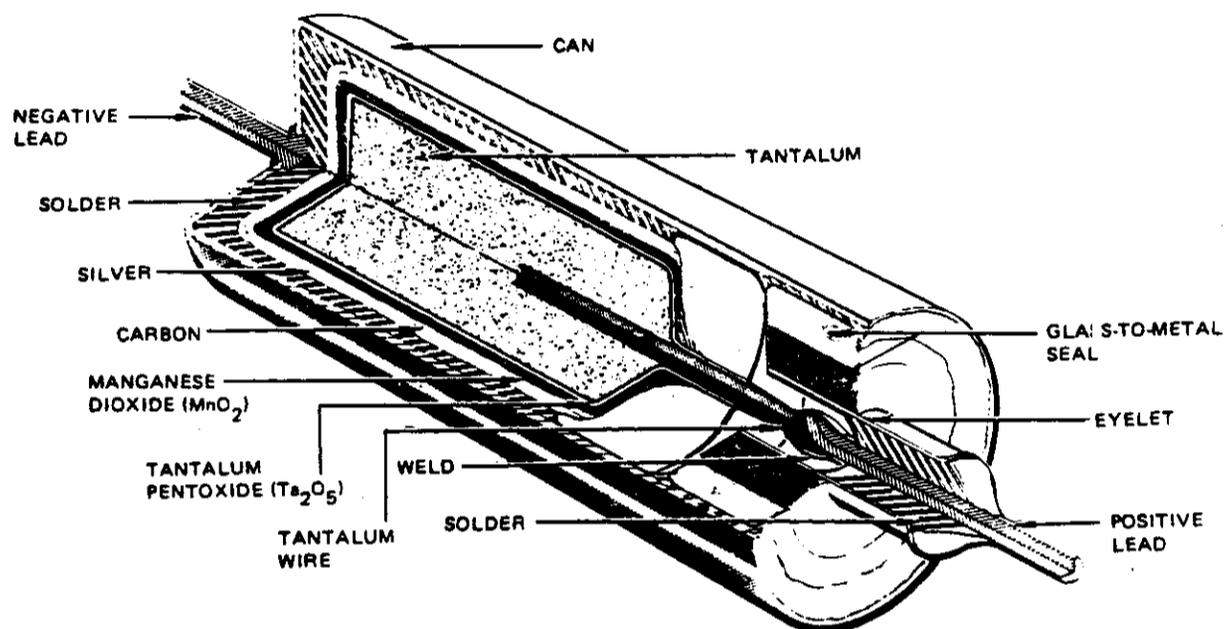


FIGURE 62. Typical construction of a solid tantalum electrolytic capacitor.

2.6.3.1 Mechanical considerations.

Mounting. Precautions must be observed when encapsulating these capacitors in hard epoxy resins, particularly with the smaller case sizes.

Shrinking of the potting material induces high pressures of varying intensity on the surfaces of the parts enclosed. For solid tantalum capacitors, these differential pressures can induce strains (in the case and leads) sufficient to result in fractures of the dielectric film. The conductive MnO_2 coating on the pellet penetrates these minute openings, and a shorted capacitor results. Upon removal of the potting material, the capacitor will often return to normal.

To eliminate this condition, these capacitors must be protected with a buffer coat of resilient material such as silicone rubber. It is important that the positive end seal be coated, as well as the metallic case surfaces.

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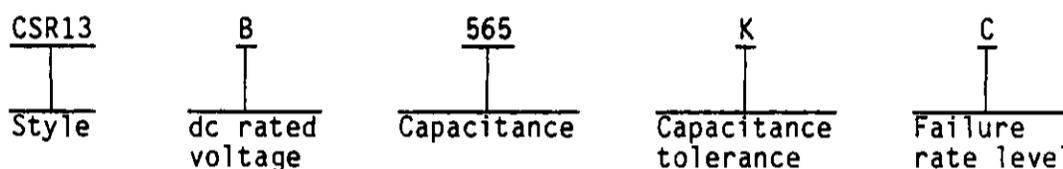
2.6 CAPACITORS, SOLID TANTALUM

2.6.4 Military designation.

2.6.4.1 Applicable military specification. Solid tantalum capacitors are covered by the following specification:

MIL-C-39003 Capacitors, Fixed, Electrolytic (Solid Electrolyte), Tantalum, Established Reliability, General Specification for.

2.6.4.2 Military type designation. The following is an example of a typical designation for an established reliability solid tantalum capacitor. For actual ordering reference, the latest issue of the applicable M39003 slash sheet must be consulted.



2.6.5 Electrical characteristics. These capacitors are available in three basic case configurations under MIL-C-39003. The CSR13 style is procured per specification sheet M39003/01, and the CSR09 style per M39003/02 and the CSR33 per M39033/06.

2.6.5.1 Voltage derating. Of all the electrolytic capacitors, the solid tantalums are the most stable over life and possess the lowest capacitance temperature characteristic. The limitations of the solid tantalum capacitors are relatively high leakage current (DCL), limited voltage range, and relatively low allowable reverse voltage.

When properly derated, these units may be operated over a temperature range of -55°C to +125°C. Refer to MIL-STD-975 (NASA) for specific derating conditions.

The failure rate of solid tantalum capacitors is a function of temperature, voltage, and circuit impedance. With each 10°C rise in operating temperature, the failure rate approximately doubles. The failure rate is also approximately proportional to the cube of the ratio of applied voltage to the rated voltage.

DC leakage current increases when either voltage or temperature are increased; the rate of increase is greater at the higher voltages and temperatures. A point can be reached where dc leakage current will avalanche causing the capacitor to be permanently shorted. For this reason the maximum ratings should never be exceeded and the derating guidelines should be observed.

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By increasing circuit impedance, the leakage current is reduced. In high-impedance circuits momentary breakdowns will self-heal. In low-impedance circuits the self-healing characteristics under momentary breakdown of the dielectric do not exist. The large currents in low-impedance circuits may cause the capacitor to short.

To minimize the incidence of catastrophic shorts due to momentary dielectric breakdowns, and to allow self-healing action to take place, a series resistance of 3 ohms per applied volt should appear in series with the capacitor and its power source, or there should be limitations on the "turn-on" surge current. The charging current available to the capacitor should be one ampere or less. This need not be a discrete resistor at the capacitor terminals, but can include the impedance of the power source and associated circuitry provided that no other large capacitors are directly in parallel with the one to be protected.

2.6.5.2 End-of-life design limits for solid tantalum capacitors. Expected capacitance change is $\pm 10\%$ and the leakage current change is 200% of initial limit.

2.6.5.3 Reverse voltage. These capacitors are capable of withstanding peak voltages in the reverse direction equal to 15 percent of their dc rating at +25°C, 10 percent at +55°C, 5 percent at +85°C, and 1 percent at +125°C.

2.6.5.4 Ripple voltage. These capacitors may be operated with an impressed ac ripple voltage, provided the capacitors do not exceed their heat dissipation limits. Total heat dissipation limits depend on the ambient operating temperature and the operating frequency.

The individual detail specification should be consulted for maximum allowable ripple voltage at the frequency and temperature required. However, Figures 63 and 64 give typical ripple voltage limits for CSR13 styles.

In addition, the sum of the applied dc bias voltage and the peak of the ac ripple must not exceed the allowable rated dc voltage for the applicable temperature. The permissible ripple voltage may also be applied without a dc bias voltage, provided that the negative peak ripple does not exceed the allowable reverse voltage.

For example: Referring to Figure 63, a 10 μF capacitor of any voltage rating may be operated at 1.9 Vrms, 120 Hz, at 25°C; at 125°C the permissible voltage is reduced to 0.75 Vrms. When this same capacitor is subjected to a ripple frequency of 1000 Hz, the permissible ac must be reduced by a factor equal to the ratio of the allowable ripple at 1000 Hz, 25°C, (Figure 64) to the allowable ripple at 120 Hz, 25°C (Figure 63), i.e.,

$$\text{Vrms (1000 Hz, 25°C)} = 0.47 \text{ Vrms (Figure 63)}$$

$$\text{Vrms (1000 Hz, 125°C)} = 0.75 \left(\frac{0.47}{1.9} \right) = 0.19 \text{ Vrms}$$

2.6 CAPACITORS, SOLID TANTALUM

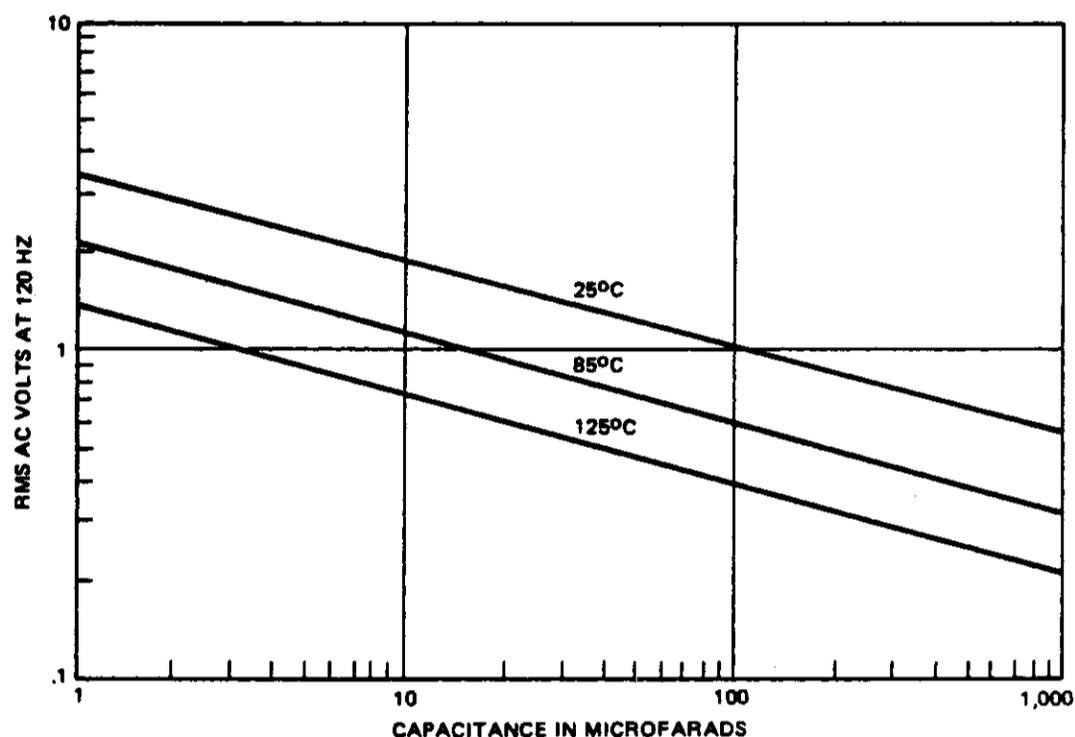


FIGURE 63. Permissible ripple voltage vs capacitance and ambient temperature at 120 Hz (style CSR13).

2.6.5.5 Series networks. When solid tantalum capacitors are connected in series, the maximum voltage across the series should not exceed the recommended derated voltage of the lowest rated capacitor in the group, or else voltage divider resistors should be used so that no capacitor in the group operates at more than its recommended derated voltage.

2.6.5.6 Parallel network. Whenever solid tantalum capacitors are connected in parallel, the sum of the peak ripple and applied dc voltage should not exceed the recommended derated voltage of the capacitor with the lowest rating.

2.6 CAPACITORS, SOLID TANTALUM

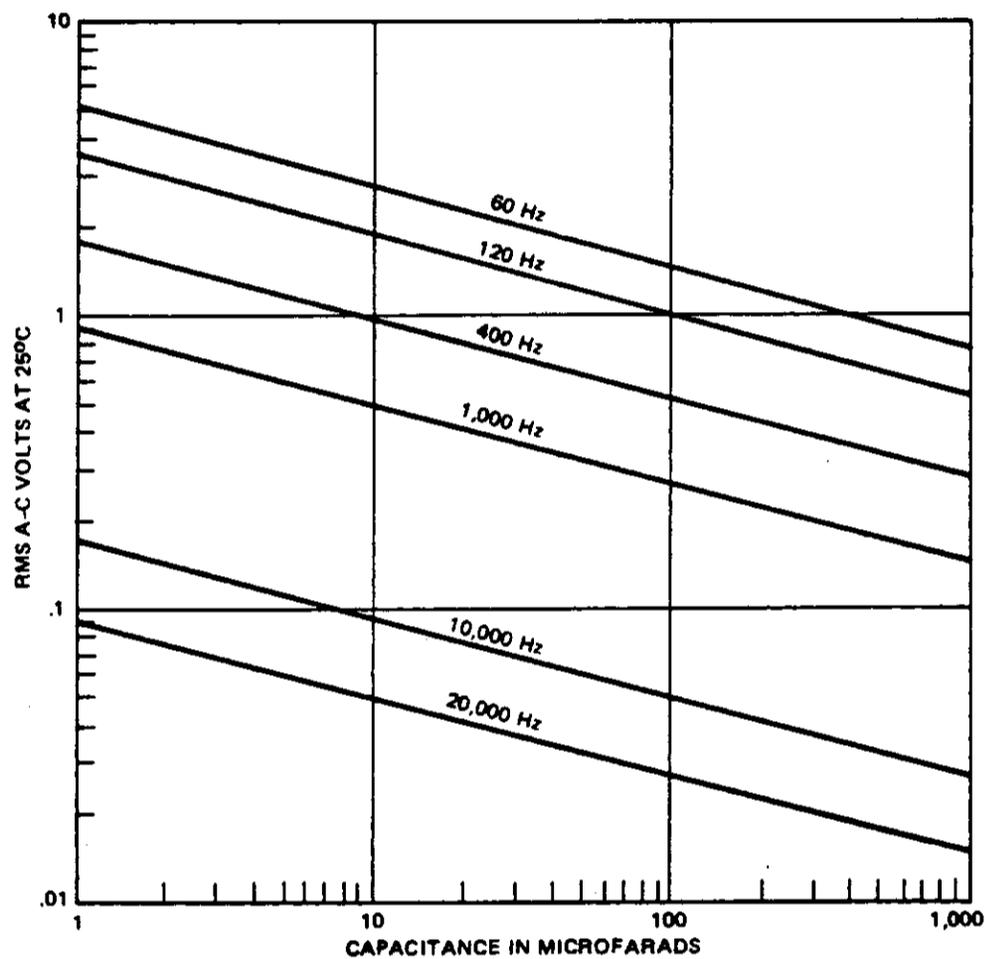


FIGURE 64. Permissible ripple voltage vs capacitance and frequency at 25°C (style CSR13).

To obtain a higher capacitance than that available from a single capacitor, a number of units may be connected in parallel. However, because of the peculiar failure mechanism associated with solid tantalum capacitors, this is not always a practical approach since each capacitor requires a resistance in series. Where there is no series impedance in the parallel legs and a minute breakdown occurs, the parallel capacitors attempt to dump their charge into the low impedance fault. Thus, what might have been a clearing action may become a catastrophic failure.

2.6 CAPACITORS, SOLID TANTALUM

2.6.5.7 Dielectric absorption. Dielectric absorption may be observed by the reappearance of potential across the capacitor after it has been shorted and the short removed. This characteristic is important in RC timing circuits, triggering systems, and phase-shift networks. The curves shown in Figure 65 were established by charging capacitors for 1 hour at rated voltage and then discharging them through a dead short for 1 minute.

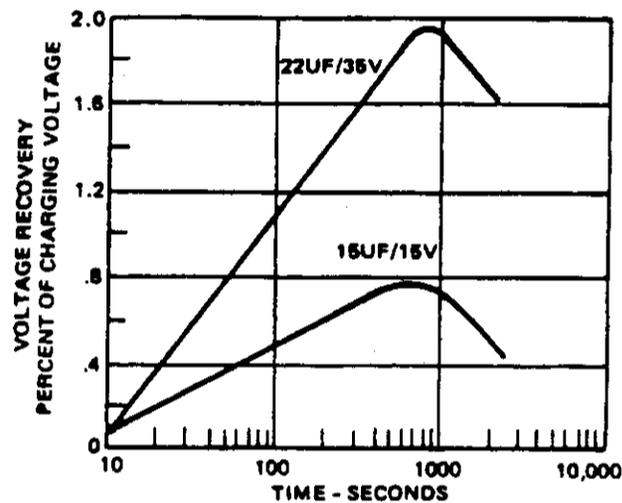
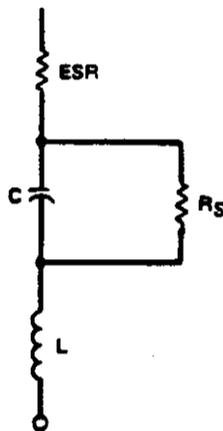


FIGURE 65. Typical dielectric absorption of solid-electrolyte tantalum capacitors at 25°C.

Voltage recovery was measured with a high-impedance electrometer at the given intervals indicated on the curves. Increasing the ambient temperature shifts the curves to the left and decreases the amplitude but does not affect the shape. Shortening charge time, lengthening discharge time, or decreasing charging voltage results in reduction of the peak amplitude of the curve, but has little effect on its shape or relative position.

2.6.5.8 The solid tantalum capacitor as a circuit element. The equivalent circuit of the capacitor may be represented in simplified form as shown in Figure 66.

2.6 CAPACITORS, SOLID TANTALUM

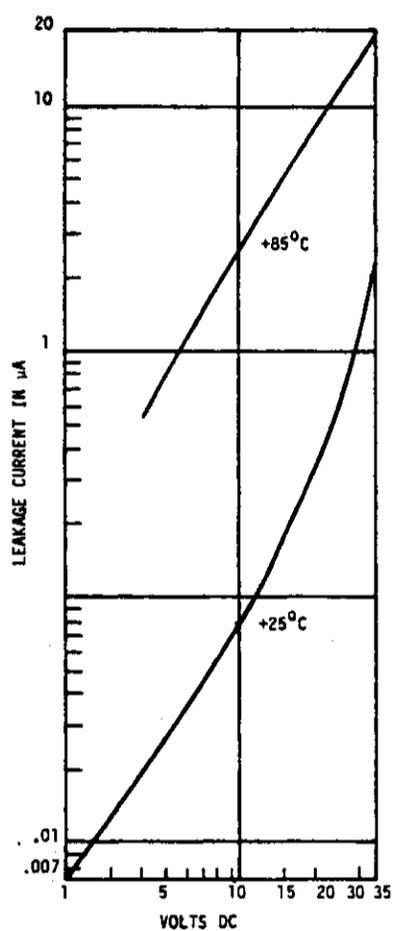
FIGURE 66. Simplified equivalent circuit of a capacitor.

2.6.5.9 DC leakage current. DCL for solid tantalum capacitors at 25°C range from less than 1 μA in low capacitance values to about 20 μA in the larger case sizes and capacitance values. Leakage current generally decreases with lower temperature, but increases by an order of magnitude at high temperature.

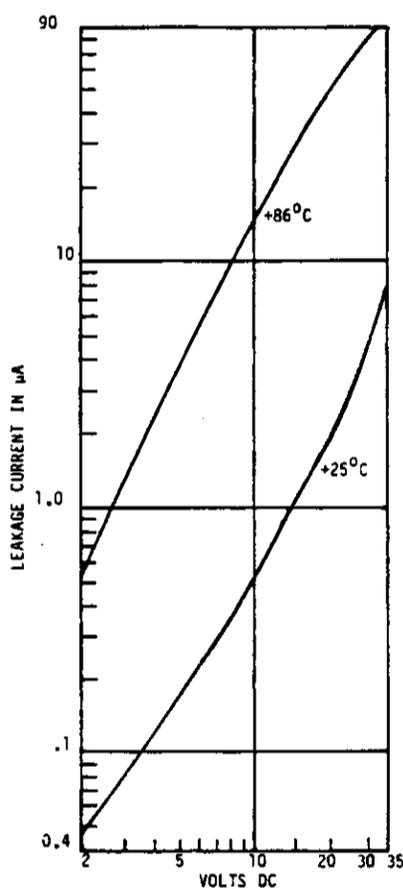
The shunt resistance, R_s , shown in Figure 66 represents the leakage path formed by an accumulation of individual faults in the tantalum pentoxide dielectric. These faults occur primarily because of a finite residue of impurities which remain in the tantalum metal despite the best practical purification techniques. Wherever an impurity is encountered, anodization cannot produce a continuous Ta_2O_5 layer of uniform thickness. The result is a minute hole or thin spot in the dielectric layer that can be filled with the manganese dioxide (MnO_2) solid electrolyte or with air. Compared with Ta_2O_5 , either of these materials will allow relatively heavy conduction under conditions existing within the capacitor.

Simple probability dictates that the larger the capacitive area, the larger the number of impurity sites encountered, and the higher the leakage current. Actual leakage current in a given application is then a function of capacitance value, voltage rating, applied voltage, and temperature. Typical curves of leakage current vs applied voltage are shown in Figure 67 (A, B, and C).

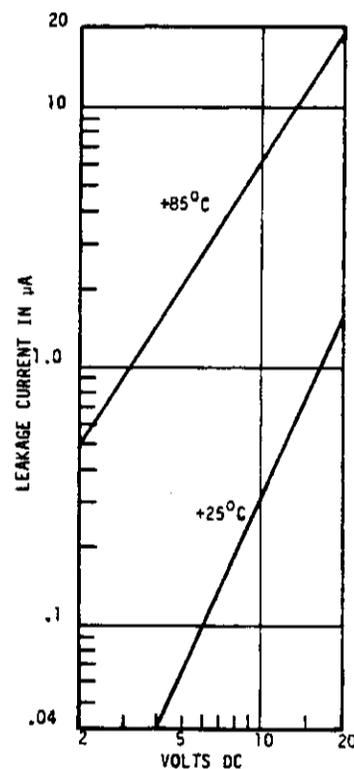
2.6 CAPACITORS, SOLID TANTALUM



A. Typical curves for 6.8 μF, 35 volt capacitor



B. Typical curves for 47 μF, 35 volt capacitor



C. Typical curves for 68 μF 20 volt capacitor

FIGURE 67. DC leakage current vs applied voltage for solid tantalum capacitor.

2.6 CAPACITORS, SOLID TANTALUM

2.6.5.10 Effects of frequency. Capacitance, effective series resistance, dissipation factor, and impedance all vary with the frequency of the applied voltage. Usually the main concern will be the cumulative total effect of the equivalent circuit, i.e., impedance. A brief discussion of each of these parameter variations follows:

Capacitance vs frequency. The rated capacitance value of these capacitors is specified as measured at 120 Hz. The apparent capacity decreases with frequency as shown in Figure 68.

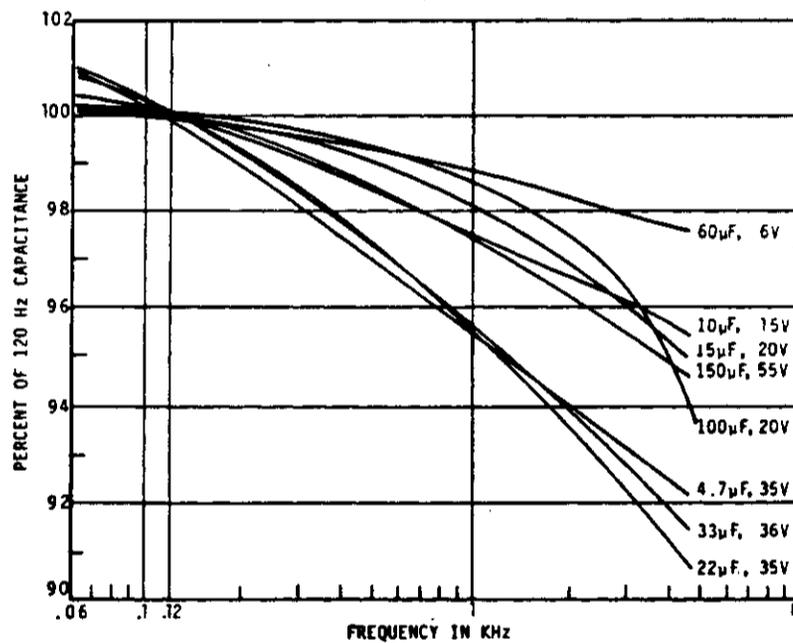


FIGURE 68. Capacitance vs frequency for typical solid tantalum capacitors at 25°C.

Dissipation factor vs frequency. Dissipation factor (ESR/X_C) is directly proportional to frequency, and would theoretically be a straight line plot if R and C were ideal. Actual typical curves are shown in Figure 69.

2.6 CAPACITORS, SOLID TANTALUM

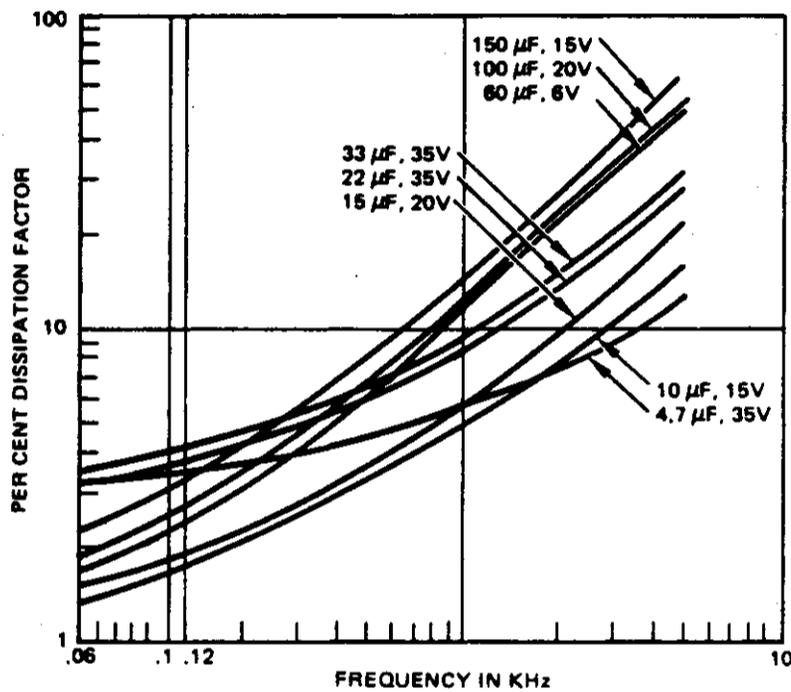


FIGURE 69. Dissipation factor vs frequency for typical solid tantalum capacitors at 25°C.

Impedance vs frequency. The impedance of a solid tantalum capacitor at lower frequencies consists essentially of the combination of capacitive reactance and equivalent series resistance. As shown in Figure 70 thru 73, the initial downward slope of the curves is due primarily to capacitive reactance. The trough of the curves is almost totally resistive, and the self-inductance of the device no longer functions as a capacitor, and impedance increases with increasing frequency.

It can be seen that the troughs of the curves typically bottom out in the 500 kHz to 1 MHz range in the higher capacitance values, and that in the lower capacitance ranges (1 μF and less) the impedance curve tends to bottom out in the 1 to 10 MHz range. For these lower capacitance values, variation of impedance with frequency closely resembles that of paper capacitors.

It should be emphasized that the curves shown are typical. In design applications where impedance over a particular frequency range is critical, special requirements may be in order. Actual impedance values, particularly at high frequencies, may vary significantly from the values shown on the curves.

2.6 CAPACITORS, SOLID TANTALUM

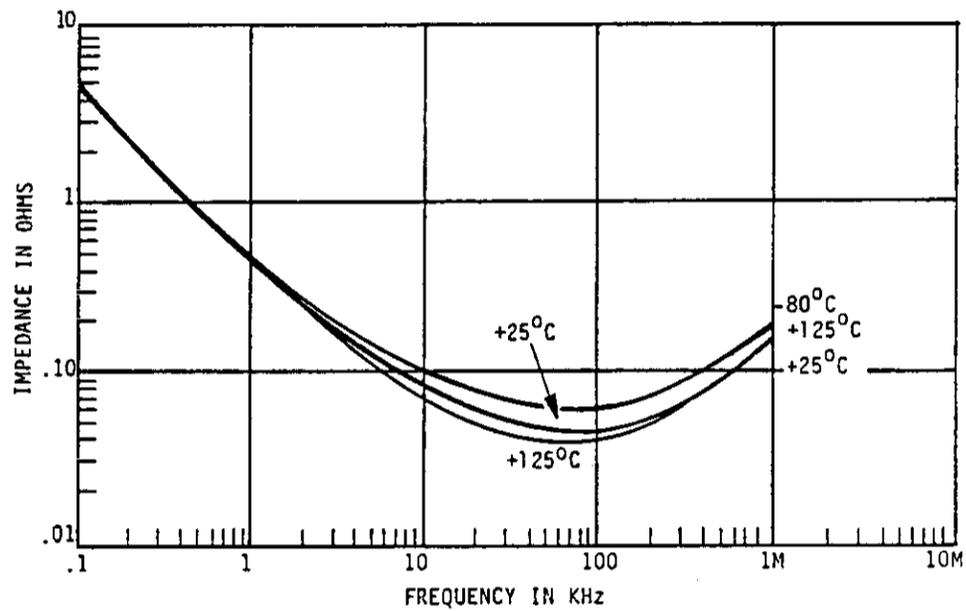


FIGURE 70. Typical curves of impedance vs frequency at various temperatures for solid tantalum 330-μF, 6-volt capacitors.

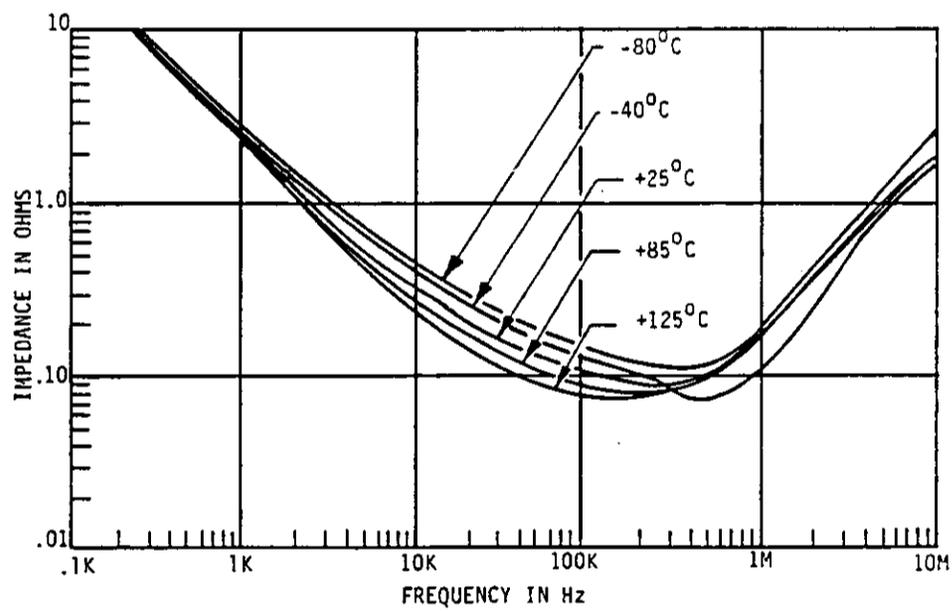


FIGURE 71. Typical curves of impedance vs frequency at various temperatures for solid tantalum 68-μF, 20-volt capacitors.

2.6 CAPACITORS, SOLID TANTALUM

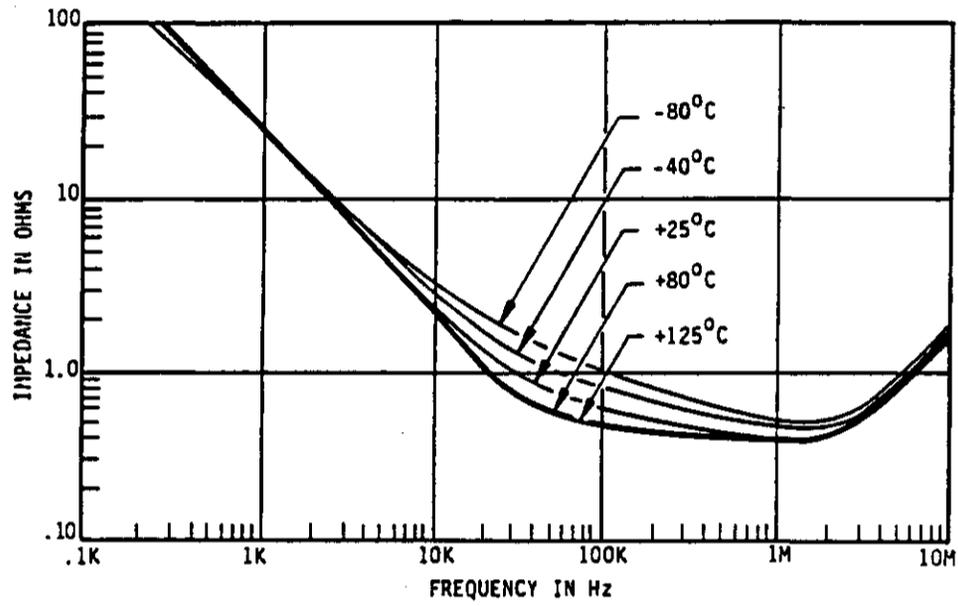


FIGURE 72. Typical curves of impedance vs frequency at various temperatures for solid tantalum 6.8-μF, 35-volt capacitors.

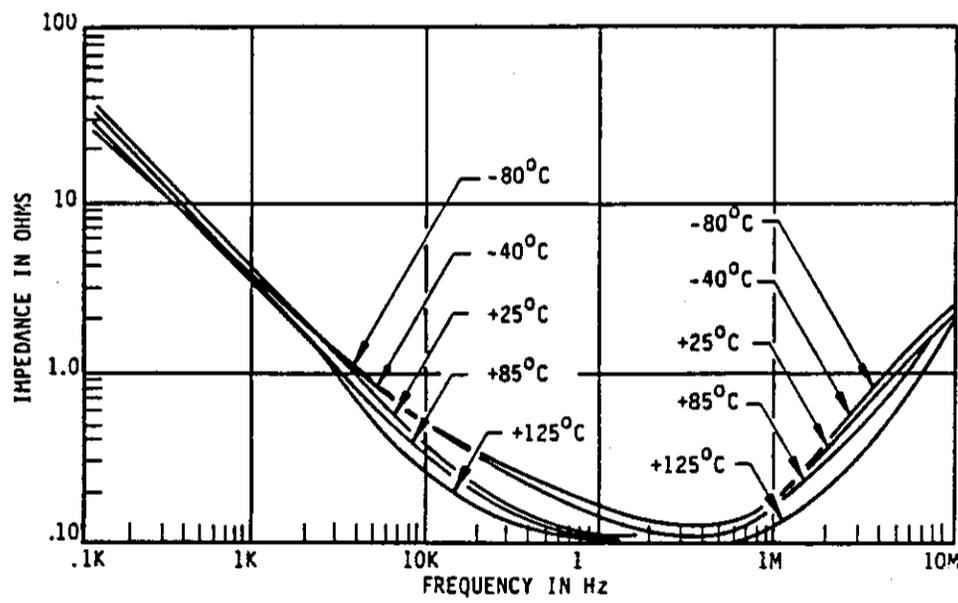


FIGURE 73. Typical curves of impedance vs frequency at various temperatures for solid tantalum 47-μF, 35-volt capacitors.

2.6 CAPACITORS, SOLID TANTALUM

2.6.6 Environmental considerations.

2.6.6.1 Effects of temperature. The capacitance of solid tantalum capacitors is relatively stable with temperature. This is also true of impedance, as shown in Figures 74 through 77.

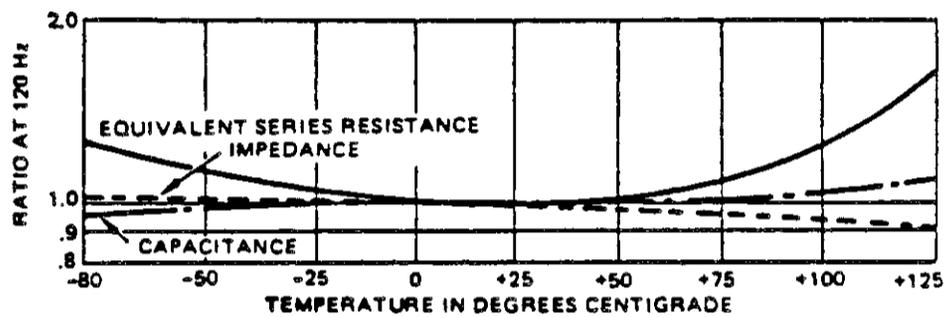


FIGURE 74. Typical curves of impedance, capacitance, and equivalent series resistance vs. temperature for 330- μ F, 6-volt capacitors.

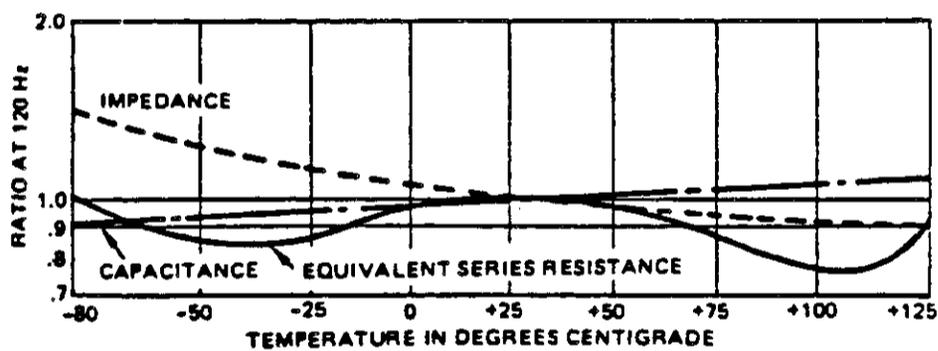


FIGURE 75. Typical curves of impedance, capacitance, and equivalent series resistance vs. temperature for 68- μ F, 20-volt capacitors.

2.6 CAPACITORS, SOLID TANTALUM

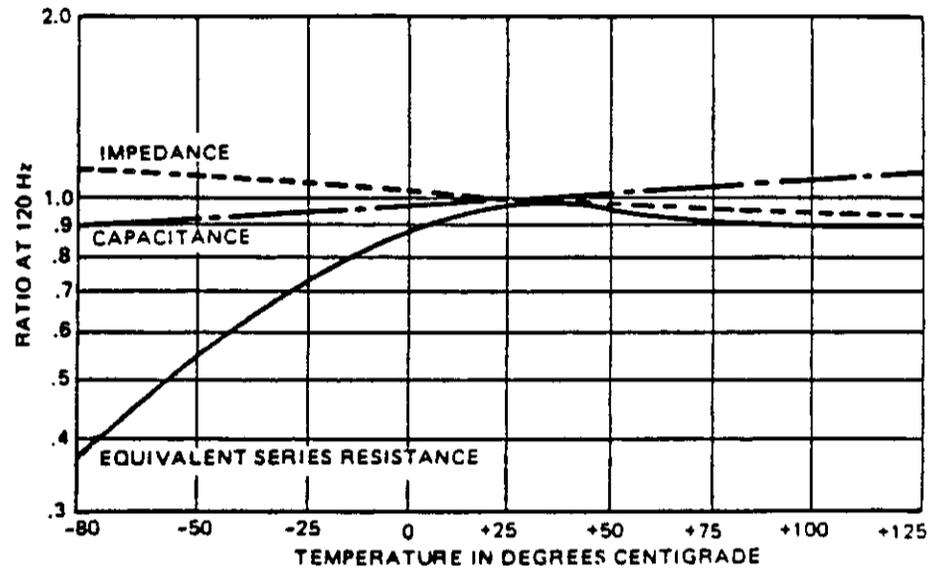


FIGURE 76. Typical curves of impedance, capacitance, and equivalent series resistance vs. temperature for 6.8- μ F, 35-volt capacitors.

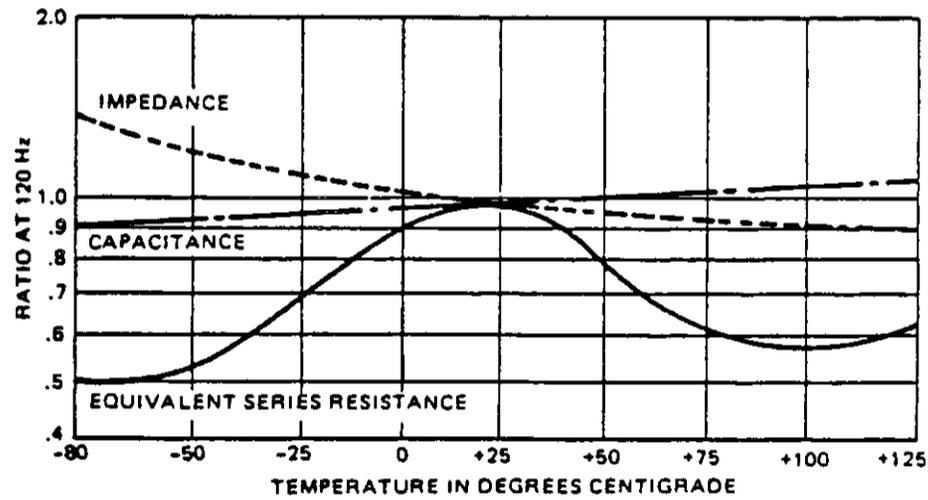


FIGURE 77. Typical curves of impedance, capacitance, and equivalent series resistance vs. temperature for 47- μ F, 35-volt capacitors.

2.6 CAPACITORS, SOLID TANTALUM

2.6.6.2 Operating temperature range. These capacitors are suitable for operation over a temperature range of -55° to $+85^{\circ}\text{C}$ at full rated voltage.

2.6.6.3 Effects of temperature on failure rate. Failure rate approximately doubles with each 10°C rise in operating temperature.

2.6.7 Reliability considerations.

2.6.7.1 Failure modes. A solid tantalum capacitor is subject to two primary failure modes; open or intermittent internal connections and shorted dielectric. Other possible failure modes include change in capacitance or dissipation factor with operation, and an increase in leakage current beyond certain limits.

Leakage current changes, even up to 100%, would not normally be significant in applications of these capacitors. In fact, leakage current normally decreases during initial hours of operation, tending to remain constant or decreasing slightly thereafter. Similarly, changes in capacitance and dissipation factor are small enough to present no real problem in normal applications.

Open or intermittent conditions are usually the result of discrete manufacturing defects, such as loose slugs due to inadequate soldering, poor internal welds, or solder shorts.

The primary failure concern in solid tantalum capacitors then, is the short circuit, which usually constitutes a catastrophic circuit failure as well as a capacitor failure.

2.6.7.2 Failure mechanism. As discussed under leakage current, all solid tantalum capacitors have faults in the dielectric because of a residue of impurities.

Each impurity site in the dielectric produces leakage current when a conductive material is present. Leakage currents would be many times larger than they are, if it were not for a change in MnO_2 structure that takes place opposite the fault sites. As dc voltage is applied to the capacitor, high current densities are produced in the faults. The high current produces heating in the MnO_2 opposite the fault. At elevated temperature, MnO_2 undergoes spontaneous reduction to lower oxides. The lower oxides coincidentally display much higher electrical resistivities, effectively reducing the leakage current that originally produced the heating.

If a relatively small fault site is encountered, the mechanism just described may permanently reduce the leakage current associated with that site to a very low value. Since this process can continue indefinitely in service, the capacitor tends to improve with age.

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2.6 CAPACITORS, SOLID TANTALUM

Failure is believed to occur because relatively large impurities migrate to the anode-dielectric interface during operation. This action creates defects which are too large to be self-healing. The amount of current flow may be sufficient to produce localized heating which destroys a still larger area of dielectric. Since the oxides of manganese have a negative temperature coefficient of resistivity, excess temperature rise in the defect zone can cause loss of control of leakage current. This leads to thermal runaway and catastrophic failure. Since the migration rate is temperature dependent, the failure rate is also temperature dependent.

Because of this inherent thermal runaway problem, solid Tantalum capacitors should not be used in power supply filter applications or other applications where the effective series resistance is less than one ohm per volt.

For Grade 1 applications, additional surge current testing is required as outlined in MIL-STD-975.

2.6.7.3 Screening. To screen out manufacturing defects and potential early life failures, the following screening tests are usually performed: temperature cycling, operating voltage conditioning, vibration, surge current at -55° , $+25^{\circ}$ and $+85^{\circ}\text{C}$ and X-ray.

2.6.7.4 Reliability derating. The failure rate of solid tantalum capacitors under operating conditions is a function of time, temperature, and voltage. Refer to MIL-STD-975 (NASA) for specific derating conditions.

2.6.7.5 Failure rate determination. For actual quantitative prediction purposes, current failure rate data should be consulted for the part as procured to a particular specification, style, and reliability level. For further information, refer to MIL-HDBK-217.

2.7 CAPACITORS, WET SLUG TANTALUM

2.7 Wet slug tantalum.

2.7.1 Introduction. The liquid electrolyte sintered-anode tantalum electrolytic capacitor, commonly known as the "wet slug," was the first type of tantalum capacitor to be developed for large-scale production. The outstanding advantage of the wet-slug type is its high volumetric efficiency. For capacitance values in the microfarad range, it will generally provide the smallest case size available in a given voltage rating. However, it has two outstanding disadvantages. It cannot tolerate reverse voltage of any magnitude for even short periods of time, and its electrolyte, which is a sulfuric acid solution in liquid or gel form, is highly corrosive and can damage neighboring circuitry if it leaks out. For these reasons, these capacitors should be considered nonpreferred styles. First consideration should always be given to solid or foil type tantalum capacitors. Wet-slug capacitors should be used only when the required function can be filled by no other available style. Recently, the CLR 79, MIL-C-39006/22 has been added to MIL-C-39006. This unit is identical to the CLR 65 except that the case is all tantalum rather than silver, it has a three-volt dc reverse capability, and a high permissible ripple current capability than other wet-slug capacitors.

Due to the all-tantalum construction of the part, there is no silver migration. Electrical specifications are included in MIL-STD-975 (NASA).

2.7.2 Usual applications. These capacitors are used mainly in power supply filter circuits. They are available as single-cell units with voltage ratings to about 125 Vdc, and in multiunit series or series-parallel packages with ratings to several hundred volts. They are strictly polar devices and are not available in a nonpolar configuration because of their extreme sensitivity to voltage reversal.

2.7.3 Physical construction. The anode consists of a slug formed by pressing and sintering tantalum powder into a porous cylindrical structure. This slug is then electrochemically treated to form a coating of tantalum pentoxide on all the surfaces of the granules. The anodized film serves as the capacitor dielectric, and the thickness of the film determines the voltage rating. The slug, with a tantalum wire lead extending, is assembled into a drawn case which serves as the cathode. The case is filled, and the slug impregnated with a highly conductive electrolyte, usually a dilute sulfuric acid solution in gel form.

The electrolyte serves as the connection between the dielectric film and the cathode. The case is sealed at the positive end, and a solderable lead is butt-welded to the tantalum lead emanating from the end seal.

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Until recently these capacitors were available only in a nonhermetically sealed style, the end seal consisting of a combination of Teflon and elastomer bushings and O-rings which were compressed and retained by crimping the case (see Figure 78). While this type of seal is normally quite effective over temperature extremes, extended temperature cycling will often cause electrolyte leakage.

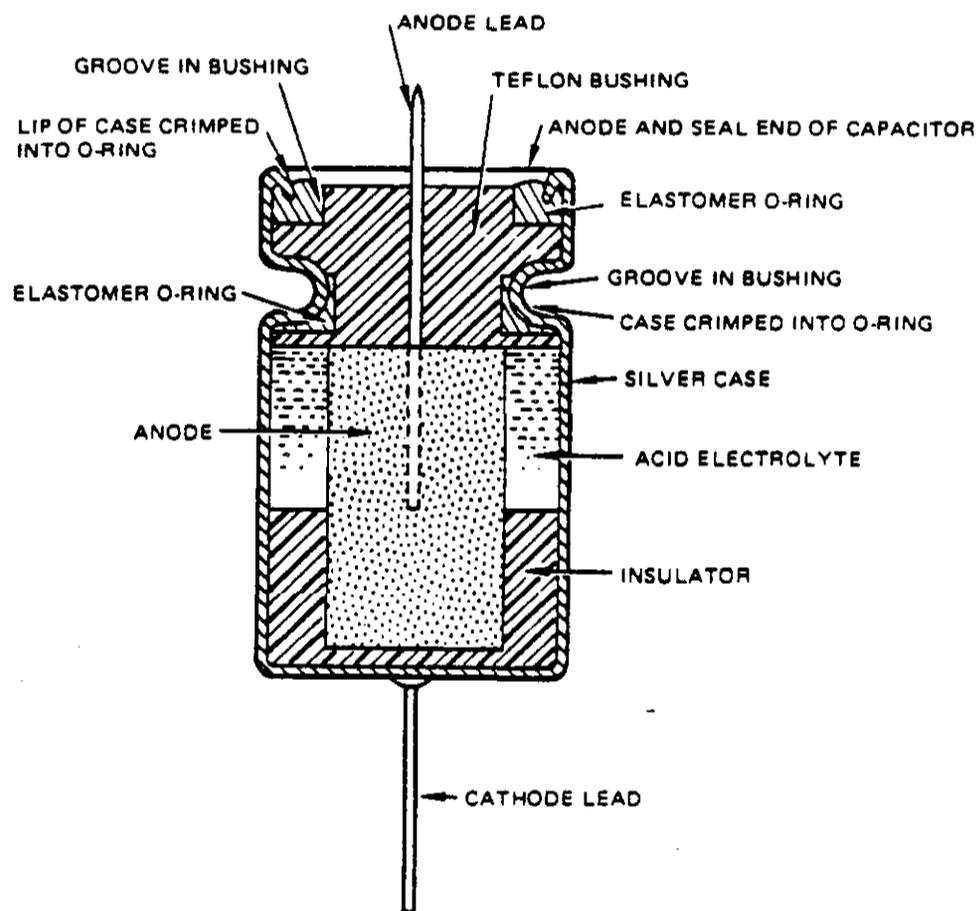


FIGURE 78. Typical construction of an elastomer seal design.

Hermetically sealed styles, such as MIL type CLR65, are also available and are shown in Figure 79. This type of construction eliminates the problem of external leakage but the capacitor is still subject to failure by leakage of electrolyte past the inner seal. Displacement of electrolyte past the Teflon seal into the area of the adjacent glass seal will result in a high leakage path between the positive lead and the case.

2.7 CAPACITORS, WET SLUG TANTALUM

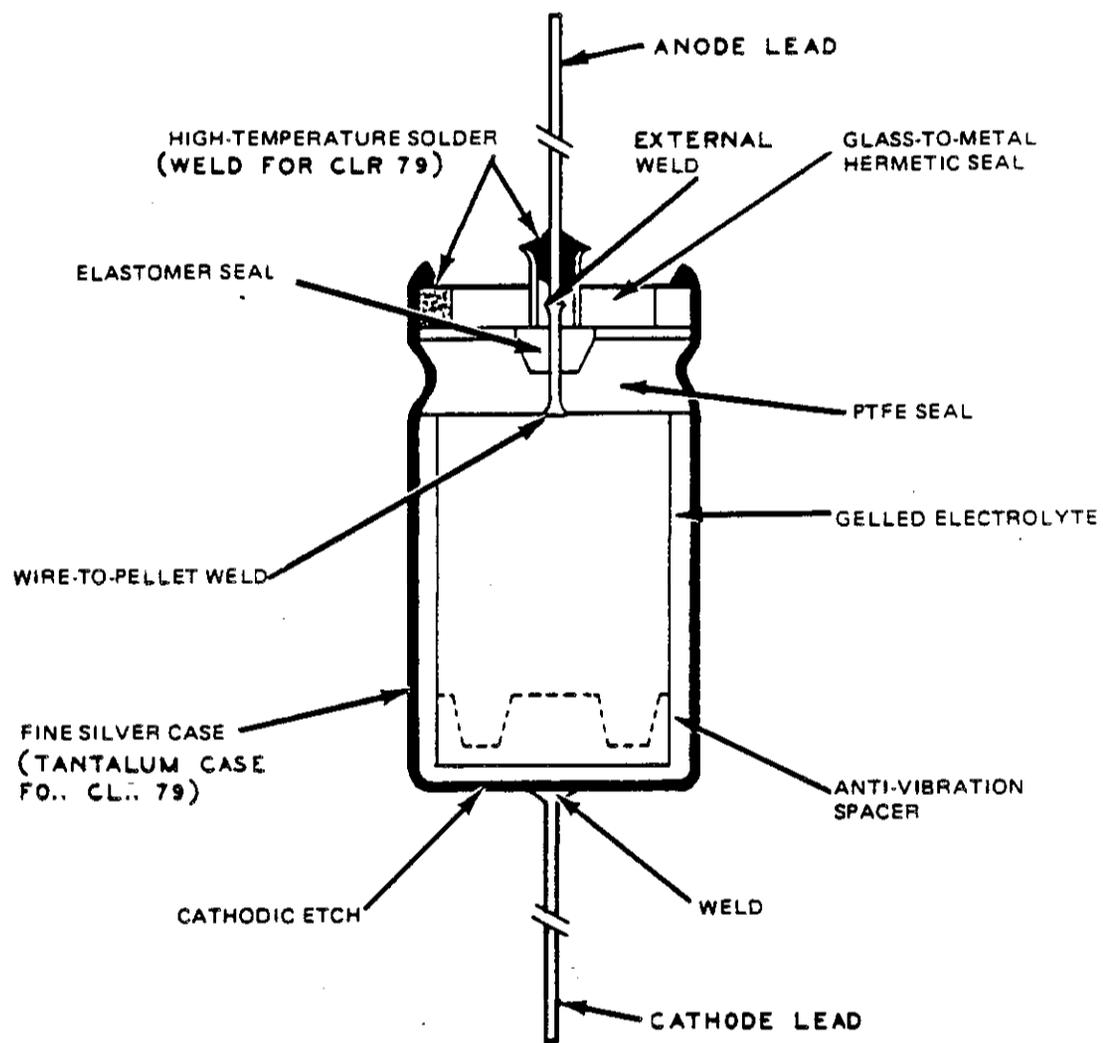


FIGURE 79. Typical construction of a hermetically sealed wet-slug tantalum capacitor.

2.7.4 Military designation. Wet-slug tantalum capacitors are covered by the following specification:

MIL-C-39006

Established Reliability. Styles CLR 65 and CLR 79 covers the hermetically sealed types, Figure 80.

2.7 CAPACITORS, WET SLUG TANTALUM

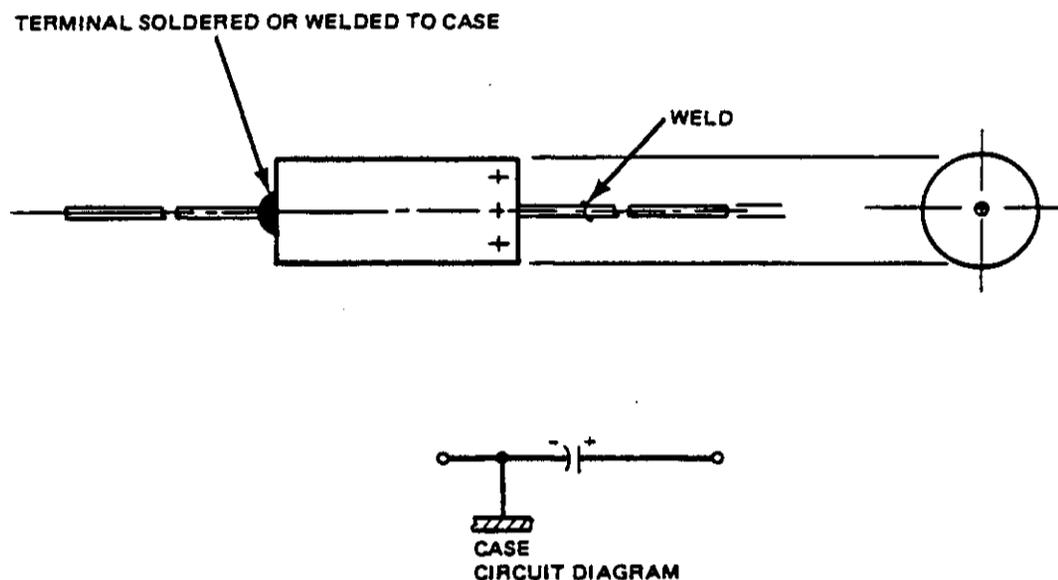


FIGURE 80. Outline drawing of a hermetically sealed capacitor.

2.7.5 Electrical characteristics.

2.7.5.1 Ratings. These capacitors are available as single units in voltage ratings up to about 125 Vdc, and in capacitance values to about 600 μF , depending on the voltage rating. They are also available in packaged units, consisting of several individual capacitors in parallel or series. Parallel arrangements are at ratings to several thousand microfarads at low voltages, and to about 10 μF at 600 Vdc. These ranges are also covered by the preferred solid or foil types, though usually at some increase in size.

2.7.5.1.1 Derating. The failure rate of wet-slug tantalum capacitors under operating conditions is a function of time, temperature and voltage. Refer to MIL-STD-975 (NASA) for specific derating conditions.

2.7.5.2 Reverse voltage. These capacitors, with the exception of the CLR 79 which has a 3 volt reverse capability at 85°C, cannot be operated with any reverse bias. The capacitor will fail in a short time due to silver plating action.

2.7 CAPACITORS, WET SLUG TANTALUM

2.7.5.3 Ripple voltage. The maximum allowable ripple voltage is a function of case size, working voltage and frequency. Figure 81 shows the maximum allowable ripple voltage at 1 kHz at ambient temperatures up to 85°C for CLR 65 case sizes. Multiplying factors for the applied frequency can be obtained from Figure 82.

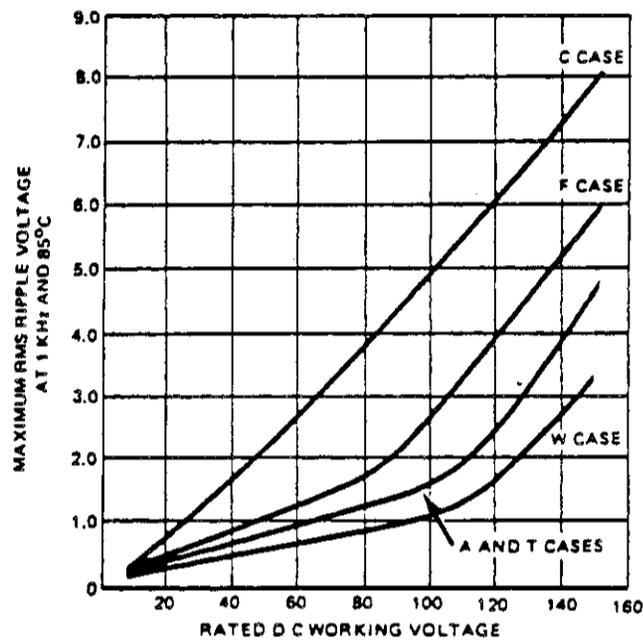


FIGURE 81. Maximum permissible ripple voltage as a function of rated working voltage for typical CLR 65.

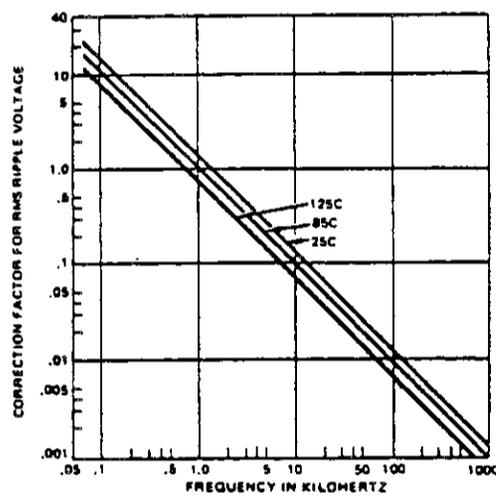


FIGURE 82. Correction factor for RMS ripple voltage as a function of frequency at various temperatures.

2.7 CAPACITORS, WET SLUG TANTALUM

2.7.5.4 Ripple currents. Figures 83 through 86 represent a number of typical temperature rise vs. rms ripple current plots for various capacitance/voltage case sizes (CLR 79).

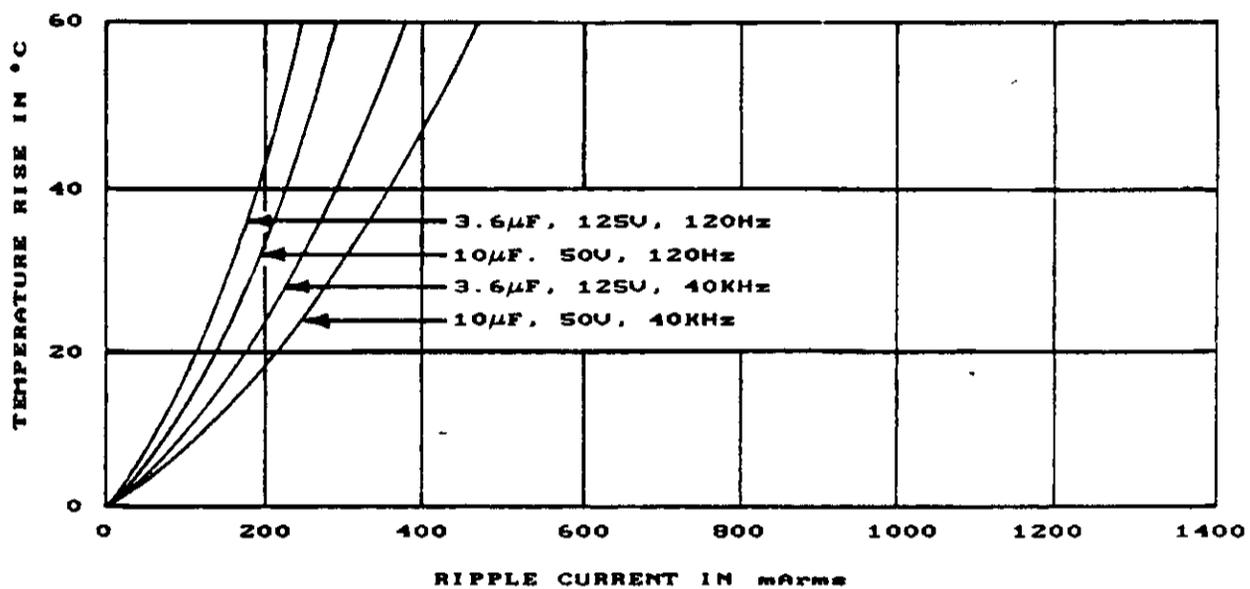


FIGURE 83. Case T1 temperature rise as a function of ripple current.

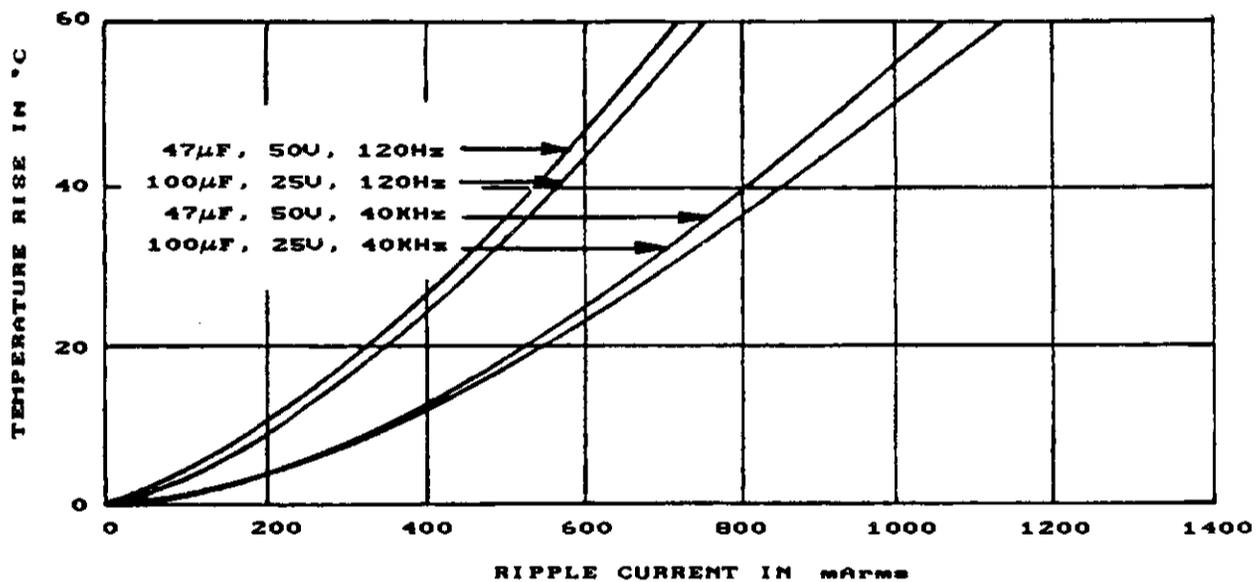


FIGURE 84. Case T2 temperature rise as a function of ripple current.

2.7 CAPACITORS, WET SLUG TANTALUM

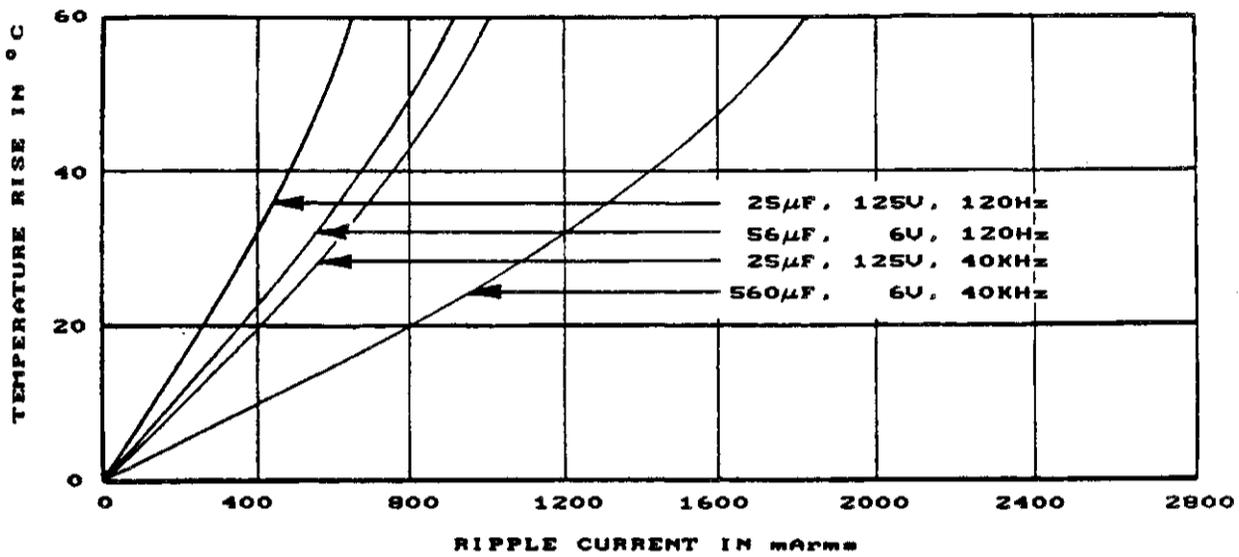


FIGURE 85. Case T3 temperature rise as a function of ripple current.

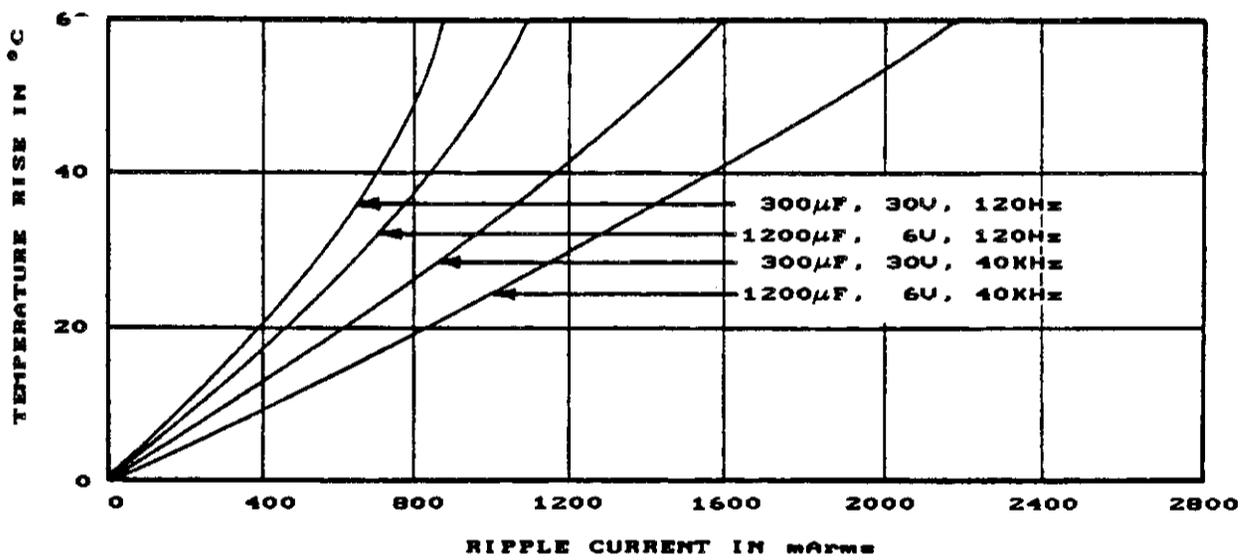


FIGURE 86. Case T4 temperature rise as a function of ripple current.

2.7 CAPACITORS, WET SLUG TANTALUM

2.7.5.5 DC leakage current. Leakage current of wet-slug capacitors is quite low, in the order of a few μA or less at room temperature and rated voltage. Leakage current increases by approximately an order of magnitude at maximum rated temperature.

2.7.5.6 Power factor and equivalent series resistance. The power factor of wet-slug capacitors measured at standard conditions (120 Hz, 25°C) ranges from allowable maximum values of about 2% to 60%, depending on capacitance and voltage rating. This order of magnitude is typical of all types of electrolytic capacitors and is usually of little concern in typical filtering applications. At low temperatures, however, equivalent series resistance increases rapidly. This effect, in combination with a typical correspondingly large decrease in capacitance at low temperatures, may require a capacitor of large nominal value to be selected for proper circuit operation over the temperature range. (Figure 87).

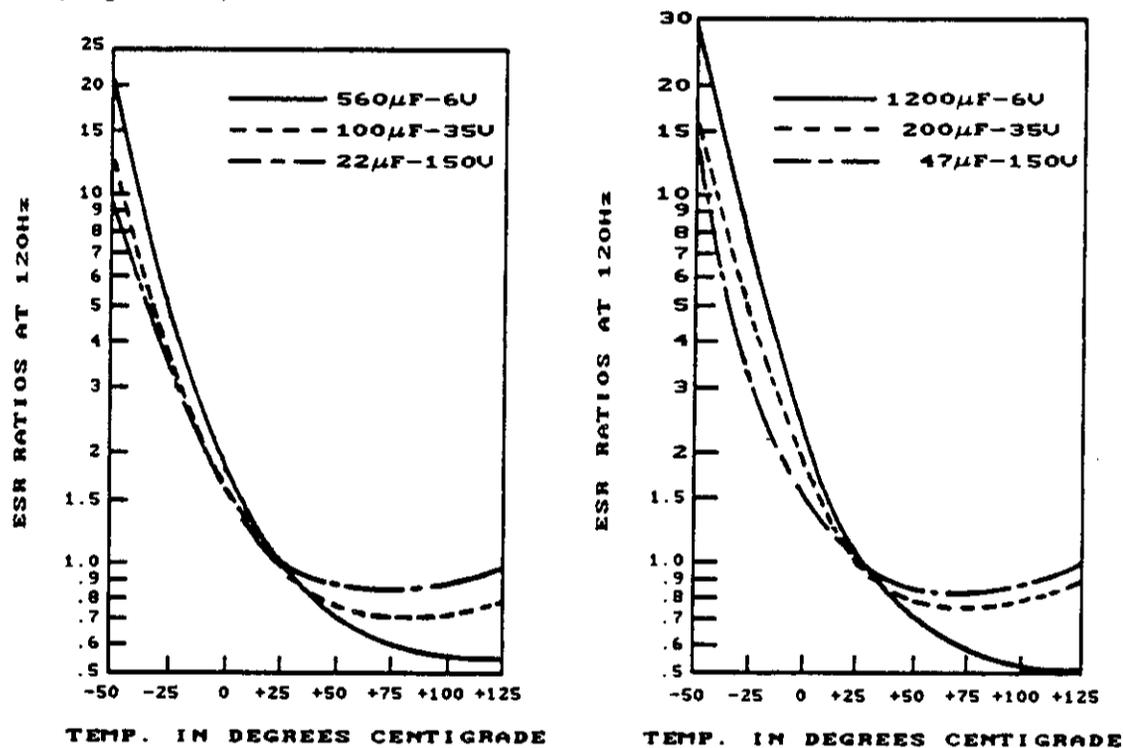


FIGURE 87. Typical curves of equivalent series resistance as a function of temperature (all voltage ratings shown are 85°C ratings).

2.7 CAPACITORS, WET SLUG TANTALUM

Figure 88 is a plot of equivalent series resistance (ESR) vs frequency for various case sizes. When capacitors are to be used in circuits operating between 10 kHz and 100 kHz, impedance measurements at 40 kHz (CLR 79) should be considered as a screening requirement.

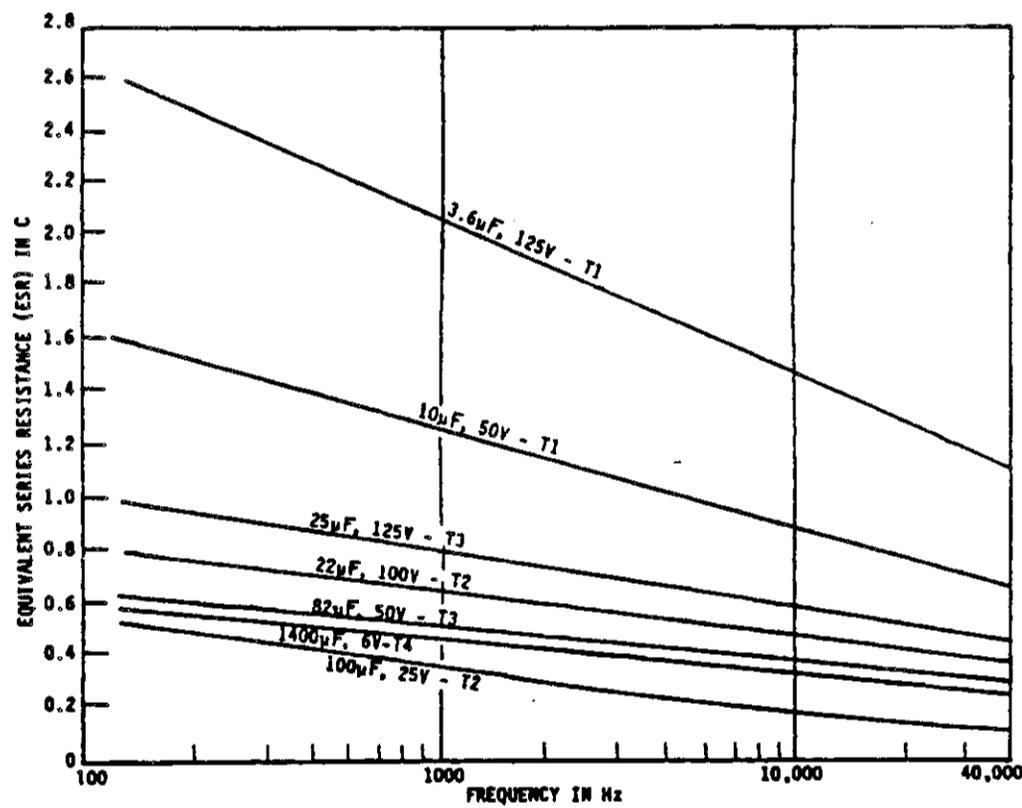


FIGURE 88. Effects of frequency on ESR.

2.7.5.7 Effect of frequency. Figures 89 and 90 show typical curves of impedance vs frequency and temperature for wet-slug capacitors. The effect of capacitance decrease and ESR increase at low temperature is readily apparent.

2.7 CAPACITORS, WET SLUG TANTALUM

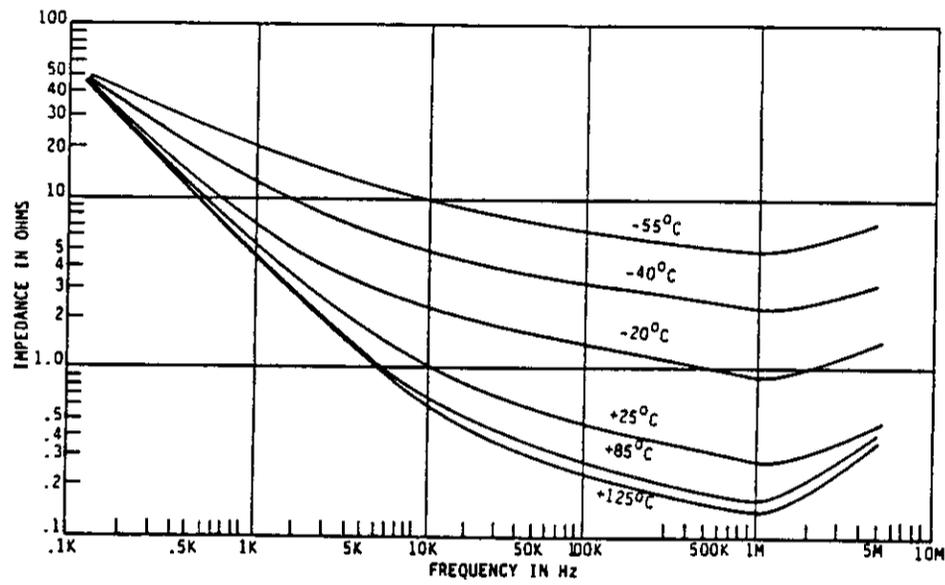


FIGURE 89. Typical curves of impedance with frequency at various temperatures for wet-slug capacitors 25µF at 125°C Vdc (all voltage ratings shown are 85°C ratings).

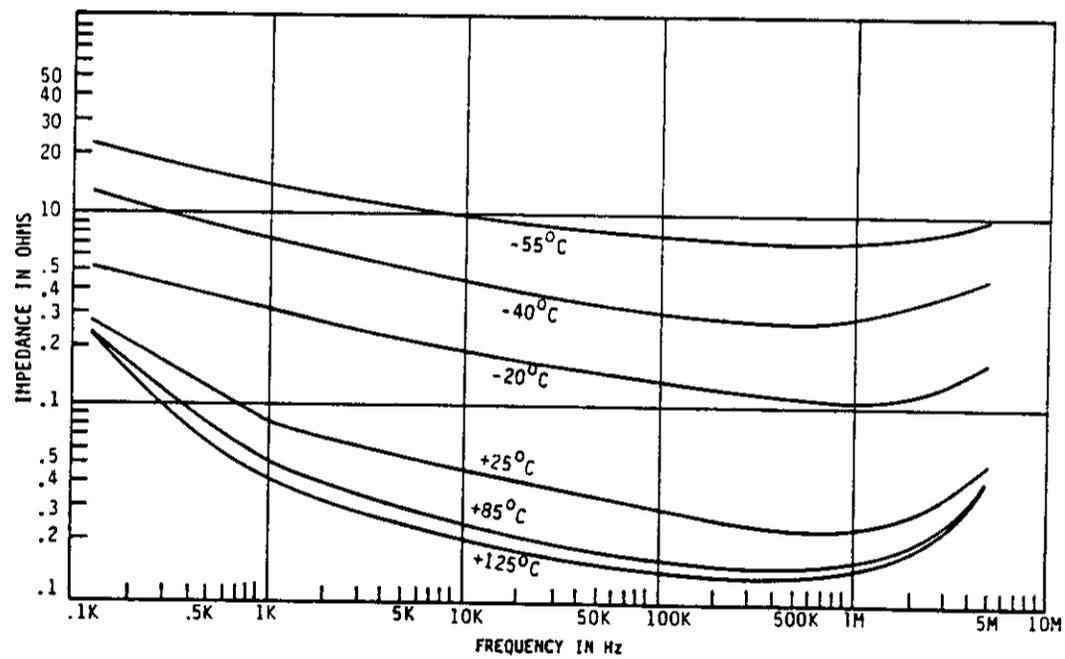


FIGURE 90. Typical curves of impedance with frequency at various temperatures for wet-slug capacitors 560µF at 6 Vdc (all voltage ratings shown are 85°C ratings).

2.7 CAPACITORS, WET SLUG TANTALUM

2.7.5.8 Circuit impedance. There are no particular limitations on wet-slug tantalum capacitors with respect to inrush current. There is some evidence to indicate that repeated discharge into a low impedance load may have a cumulative degrading effect. Sudden discharge of these devices sometimes results in the generation of transient reverse voltages. Since exposure to reverse voltage will cause these capacitors to fail, operation under conditions where such discharges may occur should be avoided.

2.7.6 Environmental considerations.

2.7.6.1 Effects of temperature. At low frequencies, the capacitance of these devices will decrease by about 15 to 65%, depending on rating. Equivalent series resistance also increases rapidly at low temperatures (Figures 87 and 91).

2.7.6.2 Temperature cycling. Repeated temperature cycling will usually result in electrolyte leakage in a high percentage of elastomer-sealed wet-slug capacitors. This will often result in no apparent decrease of capacitance or some other functional degradation of the capacitor itself, but the corrosive electrolyte may damage adjacent circuitry. Depending on the type and quality of the seal and the severity of the temperature excursions, leakage may occur in as few as 5 or 6 cycles, or may not occur until 100 or more cycles.

There is some evidence that electrolyte leakage is aggravated by exposure to low temperatures in test chambers cooled by CO₂.

Sulfuric acid has an affinity for carbon dioxide, and will readily absorb it if negative internal pressure developed at low temperatures allows passage of the gas into the electrolyte. Then, as the temperature is raised, expansion of the absorbed CO₂ generates a positive pressure inside the capacitor case and forces electrolyte out past the seal. Thus, it is possible that some of the failures observed in temperature cycling are artificially accelerated. In any event, the hermetically sealed style should always be specified when wet-slug capacitors are used.

2.7.6.3 Shock and vibration. As with other capacitor types, component specification vibration tests are always conducted with the body of the capacitor securely mounted. For high levels of shock and vibration, these capacitors may require supplementary mounting, particularly in the larger case sizes.

2.7.7 Reliability considerations.

2.7.7.1 Failure modes and mechanisms (see Table X). According to most manufacturers' descriptions, the predominant failure mode of elastomer-sealed wet-slug capacitors is gradual loss of capacitance with operation, and an ultimate open circuit condition. This is the result of the gradual vaporization of the electrolyte past the seal into the surrounding atmosphere. The rate of electrolyte loss is directly affected by the capacitor temperature. With the decrease in capacitance there is concurrent increase in ESR, again because of electrolyte loss.

2.7 CAPACITORS, WET SLUG TANTALUM

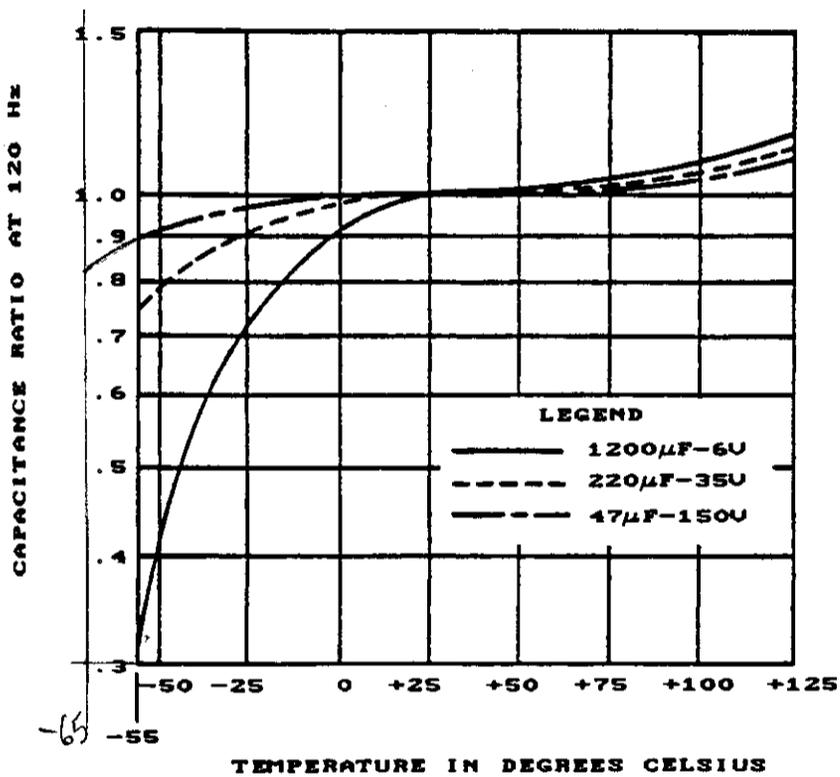
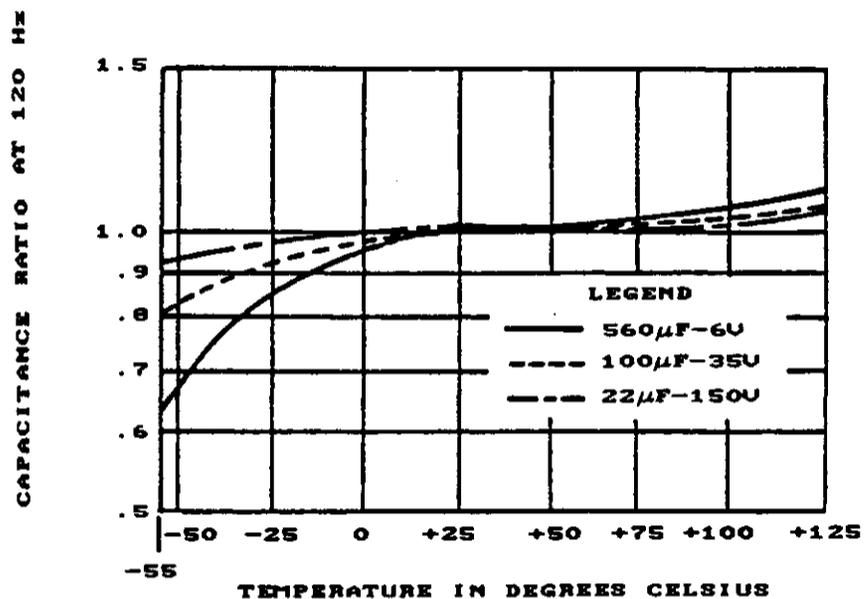


FIGURE 91. Typical curves of capacitance as a function of temperature (all voltage ratings shown are 85°C ratings).

2.7 CAPACITORS, WET SLUG TANTALUM

TABLE X. Failure modes and mechanisms (CLP 65 style)

Failure Mechanism	Description/Cause	Detection Method	Method to Minimize or Eliminate Cause
Electrolyte leakage	Faulty crimping of elastomer seals, riser wire out of round or has scratches and abrasions.	Thymol blue or litmus paper test. Increase in dissipation factor or decrease in capacitance. Temperature cycling, burn-in, seal test, insulation resistance.	Use of hermetically sealed capacitor design; use gelled electrolyte to minimize mobility.
	Leakage past inner seal causing electrolyte to bridge anode lead to cathode.	Increased leakage current or short circuit. Temperature cycling, burn-in.	Use of hermetic capacitors having fully anodized anode riser through the seal and external nickel lead attachment.
Silver deposition on anode	Reverse voltage on capacitor.	High leakage current, lower ESR, short circuit.	Application must ensure voltage is never reversed during use or test. Minimize ripple current. Use of designs utilizing titanium or tantalum cases.
Mechanical defects	Warped slugs, slugs cocked in case, canted seals, bent or improper length risers, etc.	Radiographic inspection.	Improved process control.

For a properly manufactured and screened device that has not been mishandled during test inspection and that has never been subjected to reverse voltage, even of a transient nature, the above failure mode is probably the most likely. However, failure analysis experience indicates that the main problem source for this device is application of reverse voltage.

Since the tantalum pentoxide dielectric is a rectifier, the film conducts in the direction of the reverse voltage and can be damaged or destroyed, depending on the magnitude and duration of the reverse voltage. In addition, reverse current causes electroplating of silver from the case onto and beneath the oxide layer. This increases the current leakage paths, resulting in increased dissipation and internal heat rise, liberation of gases, and catastrophic failure. The formation of hydrogen and oxygen gases at the electrodes creates excessive internal pressure and can result in electrolyte leakage or bursting of the case.

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Although these devices are capable of some self-healing action, voltage reversals introduce an incipient failure mechanism which is largely unpredictable. The final effect of reverse bias depends on the magnitude of the voltage and time of application, since the amount of plating action is a coulombic function. The voltage threshold is very low, and levels of less than 100 mv have caused failure, given enough time. In addition, there appears to be no practical way of testing these capacitors to determine whether they have been subjected to short-time reversals.

Reverse polarity can be applied in several ways, even from the terminals of an ohmmeter during part inspection, board tests, or tests at the subsystem or system level. In addition, reverse polarities have been applied at some unsuspected operational phase, such as during system turn-on or stand-by position, interactions of failures or removal of other parts, power line transients, etc.

For these reasons, these devices must be used with a full knowledge of their limitations.

2.7.7.2 Screening. Temperature cycling and operational voltage conditioning are the most commonly used screening techniques for these capacitors. Temperature cycling is used to screen out units with improper or defective seals, as evidenced by electrolyte leakage. Voltage conditioning helps to detect units with improperly formed or contaminated anodes. Forward dc leakage current on a good unit will normally decrease to some value and remain relatively constant, or continue to decrease at a lower rate with continued operation. Measurements on a potential failure will tend to show an increase in leakage current. These devices should be subjected to vibration screening to eliminate devices that exhibit voltage spikes (5 to 20 volts) observed at 20 g and 80 g axial vibration and 51 g random vibration.

2.7.7.3 Reliability derating. The failure rate of wet-slug tantalum capacitors under operating conditions is a function of time, temperature and voltage. Where wide temperature variations are expected on a continual basis throughout the life of the system, solid or foil tantalum capacitors should be specified. If a wet-slug tantalum capacitor must be used, the hermetically sealed type should be specified.

Since present designs of hermetically sealed depend on an elastomer seal to prevent internal electrolyte leakage of the capacitor itself, reliability derating should be considered from the temperature cycling standpoint. When wet tantalum capacitors are connected in parallel, the sum of the peak ripple and the applied dc voltage should not exceed the recommended derated dc voltage rating of the capacitor with the lowest rating. The connecting leads of the capacitors in parallel should be large enough to carry the combined currents without reducing the effective capacitance due to series lead resistance.

2.7.7.4 Failure rate. For purposes of reliability predictions, MIL-HDBK-217 should be consulted. Since the failure rate of these devices depends so greatly on the conditions to which they have been exposed and transient conditions which they may see, published failure rate data based on controlled conditions of operation must be treated mainly as a base for predicted performance.

2.8 CAPACITORS, VARIABLE

2.8 Variable.

2.8.1 Introduction. Variable capacitors are not included in MIL-STD-975 (NASA); they are included in this handbook to provide a technical understanding of this type of part.

Variable capacitors are small-sized trimmers designed for use where fine tuning adjustments are periodically required during the life of the equipment. The three types most popularly used are the piston tubular trimmer, ceramic dielectric, and air trimmer.

2.8.2 Usual applications. The variable capacitor is normally used for trimming and coupling in such circuits as intermediate frequency, radio frequency, oscillator, phase shifter, and discriminator stages. Because of their low mass, these units are relatively stable against shock and vibration, which tend to cause changes in capacitance.

2.8.3 Physical construction.

2.8.3.1 Piston type, tubular trimmer. These capacitors are constructed of glass or quartz dielectric cylinders and metal tuning pistons. A portion of the cylinder is plated with metal and forms the stator. The metal piston (controlled by a tuning screw) acts as the rotor.

The overlap of the stator and the rotor determines the capacitance. The self contained piston within the dielectric cylinder functions as a low inductance coaxial assembly. The piston type capacitor is further classified as a rotating or nonrotating piston type.

2.8.3.2 Rotating piston. The rotating piston is constructed in such a way as to secure the piston to the tuning screw. As the tuning screw is rotated the piston rotates with it (Figure 92).

2.8.3.3 Nonrotating piston. The nonrotating piston is constructed so that the tuning screw is secured at each end and cannot move up and down as it is turned. The screw is threaded into the piston and the piston moves up and down without rotating (Figure 93).

2.8.3.4 Ceramic dielectric trimmer. This capacitor consists of a single rotor and stator for each section, with each section fabricated of ceramic material impregnated with transformer or silicone oil. Pure silver is fired and burnished on the top of the base of the stator in a semicircular pattern. The rotor (usually of titanium dioxide) has pure silver contact points. The contact surfaces of both the rotor and stator are ground and lapped flat, eliminating air space variations with temperature. (See Figure 94).

2.8 CAPACITORS, VARIABLE

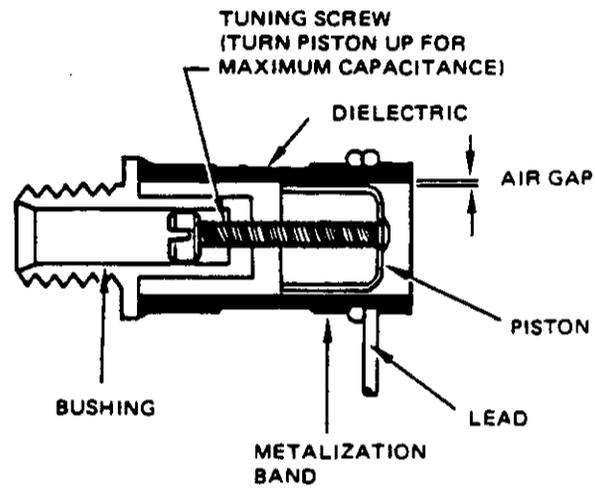


FIGURE 92. Typical construction of a rotating piston style.

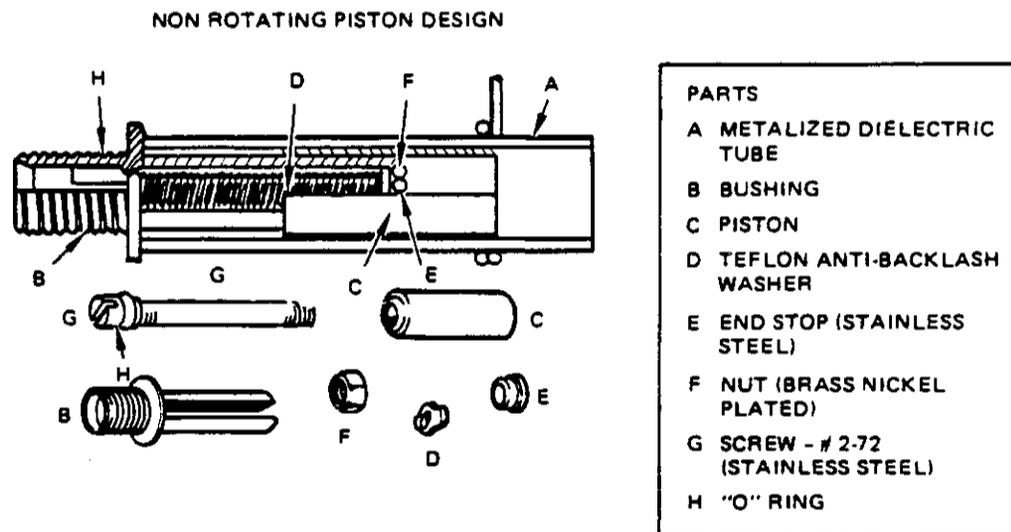


FIGURE 93. Typical construction of a non-rotating piston design.

2.8 CAPACITORS, VARIABLE

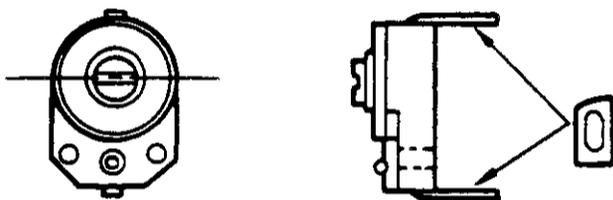


FIGURE 94. Outline drawing of a ceramic dielectric variable capacitor.

2.8.3.5 Air dielectric trimmer. This capacitor consists of multiple rotors and stators, each having a half-moon shape. The overlap of rotors and stators determine the capacitance. The rotors can be rotated continuously and full capacitance change occurs during each 360° rotation. (See Figure 95).

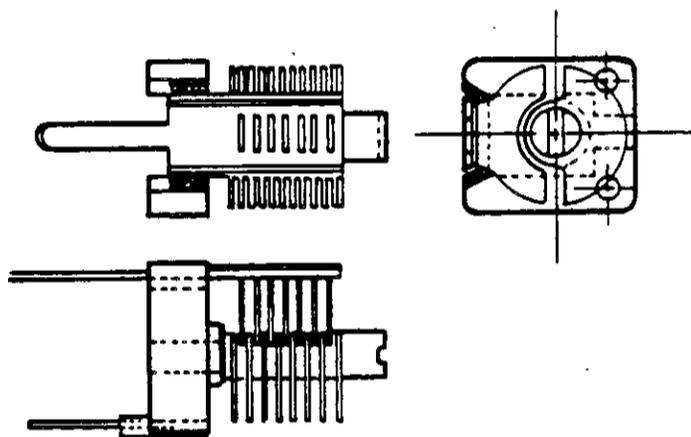


FIGURE 95. Outline drawing of a variable air dielectric trimmer.

2.8.3.6 Mounting. The trimmer capacitors are generally leaded devices and may be mounted close to a metal panel with little increase in capacitance. Care should be exercised to avoid cracking or chipping the ceramic mounting base.

2.8.4 Military designation. Variable capacitors are covered by MIL-C-81 for variable ceramic, MIL-C-92 for variable air and MIL-C-14409 for piston type tubular. The following examples show typical military designations for variable capacitors.

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2.8 CAPACITORS, VARIABLE

2.8.4.1 MIL-C-81 (variable ceramic dielectric).

CV11
|

Style

A
|

Characteristic

070
|

Capacitance

2.8.4.2 MIL-C-92 (variable air dielectric).

CT06
|

Style

A
|

Voltage

004
|

Capacitance

J
|

Rotational Life

2.8.4.3 MIL-C-14409 (variable piston type tubular)

PC51
|

Style

H
|

Characteristic

380
|

Capacitance

2.8.5 Electrical considerations.

2.8.5.1 Voltage ratings. The MIL-C-81, "CV" type capacitors are available to 500 Vdc from -55°C to +85°C with voltage derating to 125°C.

The MIL-C-92, "CT" type capacitors are available in a range of voltages to 700 Vdc from -55°C to +85°C.

The MIL-C-14409, "PC" types capacitors are available in a range of voltages to 1250 Vdc from -55°C to +125°C and with characteristic letter "Q" from -55°C to +150°C.

2.8.5.2 Available capacitance values. Variable ceramic dielectric capacitors are available in a range of capacitance values from approximately 1.5 to 60 pF with a DF of 0.2 percent maximum.

Variable air dielectric capacitors are available in a range of capacitor values from approximately 1.3 to 7.5 pF with a Q greater than 1500 at 1 MHz.

Variable piston type tubular capacitors are available in a range of capacitance values from 0.5 to 120 pF with Q ranging from 10,000 to 250, inversely proportional to the capacitance value.

2.8 CAPACITORS, VARIABLE

2.8.5.3 Q vs frequency. The Q of a trimmer capacitor depends on the dielectric system and the size of the unit. The internal mechanism and plating do not affect it. The RF current flows along the unit to the base, providing a long inductive path which decreases Q and lowers the self-resonant frequency. Typical self-resonance of glass trimmer capacitors varies from 100 MHz to 1000 MHz according to size and construction. A quick comparison of the characteristics of a ceramic trimmer and a glass tubular trimmer shown in Table XI.

Table XI. Comparison of characteristics of ceramic and glass trimmer

Characteristic	Ceramic	Glass
Tuning Resolution	1/2 turn	Multi-turn
Temperature Range	To +85°C	To +125°C
Temperature Coefficient	200 to 1200 ppm/°C	±50 ppm/°C
Capacitance Drift	75% or 0.5 pF	0.01 to 0.1 pF
DCWV	To 500 Vdc	To 1200 Vdc
Q	500 at 1 MHz	1000 at 20 MHz
Rotational Life	100 Turns	Up to 10,000 Turns

2.8.5.4 Derating. The failure rate of variable capacitors under operating conditions is a function of time, temperature and voltage.

2.8.5.5 End-of-life design limits for variable capacitors. Expected capacitance change is ±5 percent from set value and Insulation Resistance change is -30 percent from initial value.

2.8.6 Reliability considerations.

2.8.6.1 Failure modes and mechanisms. The trimmer capacitors have had a reliable history of electrical operation. Most of the problems associated with the parts are of a mechanical nature. Rough handling will cause shorting of the plates of an air trimmer. Cracking of the rotor or stator of the ceramic trimmer during improper soldering or cleaning methods can allow solder flux onto the walls of a piston type, thereby binding the piston or fracturing the screw adjust.

2.8.6.2 Screening. Early life failures are best screened out by a voltage conditioning period of 50 hours or more under voltage stress. Tubular trimmers are typically burned-in at 100% of rated voltage while ceramic trimmers can be burned-in at 200% of rated voltage. Temperature cycling is also specified prior to voltage conditioning to accelerate failure of parts with mechanical weaknesses or poor internal connections.

2.8.6.3 Failure rate determination. For failure rate information refer to MIL-HDBK-217.

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2.8 CAPACITORS, VARIABLE

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3.1 RESISTORS, GENERAL

3. RESISTORS

3.1 General.

3.1.1 Introduction. This section contains information on the various types of resistors intended for use in NASA applications.

The information in this section is designed to help the engineer select the resistors he will specify. As with other types of components, the most important thing a user must decide is which of the numerous types of resistors will be best for use in the equipment he is designing. Proper selection is the first step in building reliable equipment. To properly select the resistors to be used, the user must know as much as possible about the types from which he can choose. He should know their advantages and disadvantages, their behavior under various environmental conditions, their construction, and their effect on circuits and the effect of circuits on them. He should know what makes resistors fail. He should also have an intimate working knowledge of the applicable military specification.

All variable and fixed resistors can be grouped into one of three general basic types. They are composition, film, or wirewound types (see Figure 1). The composition type is made of a mixture of resistive material and a binder which are molded into the proper shape and resistance value. The film type is composed of a resistive film deposited on, or inside of, an insulating cylinder or filament. The wirewound type is made up of resistance wire which is wound on an insulated form. These basic types differ from each other in size, cost, resistance range, power rating, and general characteristics. Some are better than others for particular purposes; no one type has all of the best characteristics. The choice among them depends on the requirements, the environment and other factors which the designer must understand. The designer must realize that the summaries of the following general characteristics and costs are relative, not absolute, and that all the requirements of a particular application must be taken into consideration and compared with the advantages and drawbacks of each of the several types before a final choice is made.

The detailed requirements for standard resistor types are contained in the applicable military specification and applicable subsection of this section.

3.1.2 Applicable military specifications.

<u>Mil Spec</u>	<u>Title</u>
MIL-STD-202	Test Methods for Electronic & Electrical Components Parts
MIL-HDBK-217	Reliability Prediction of Electronic Equipment

3.1 RESISTORS, GENERAL

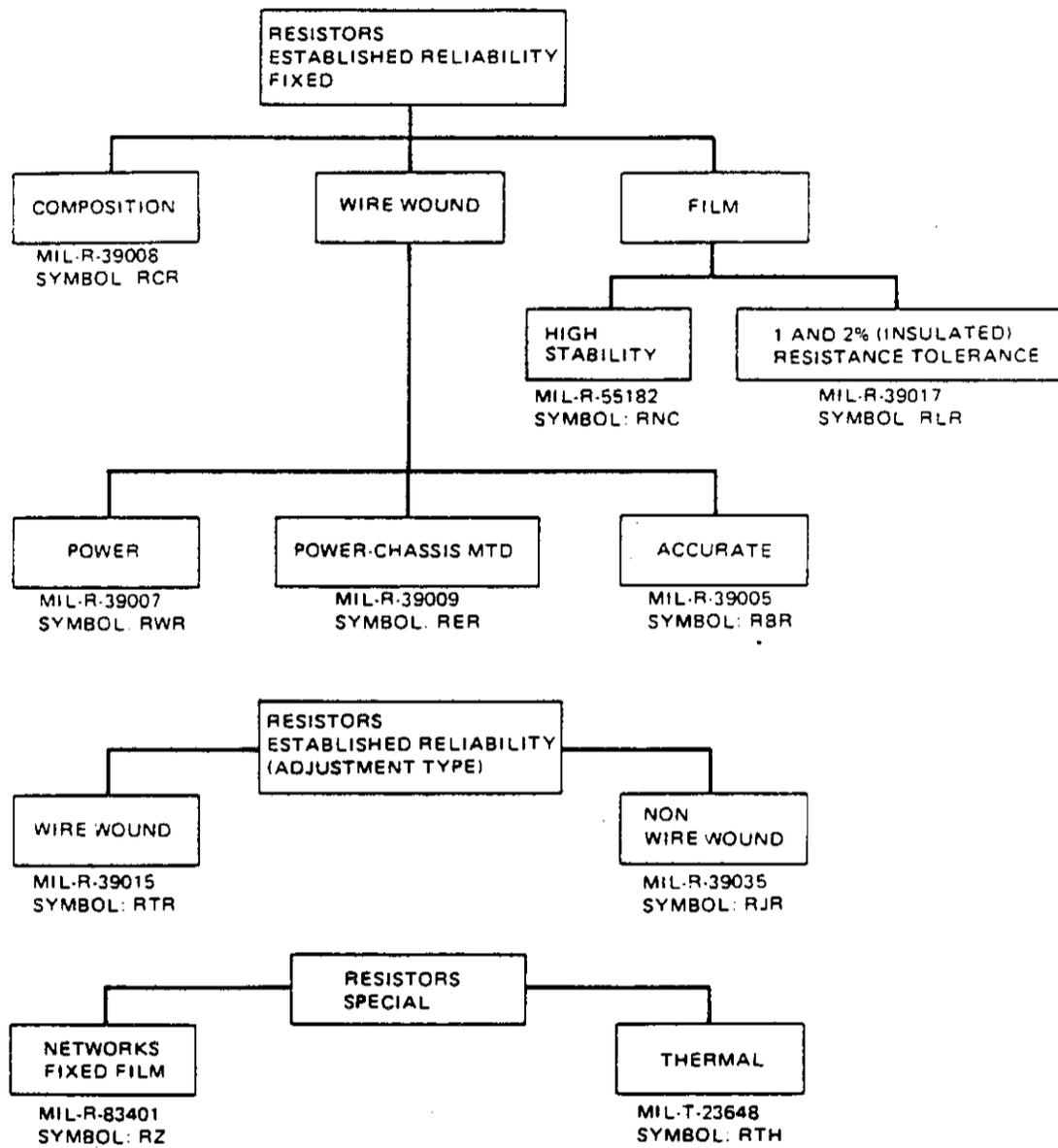


FIGURE 1. Resistor categories.

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MIL-STD-790	Reliability Assurance Program for Electronic Parts specifications
MIL-STD-1276	Leads for Electronic Components Parts
MIL-R-39008	Fixed Composition
MIL-R-55182	Fixed Film (High Stability)
MIL-R-39017	Fixed Film (Insulated)
MIL-R-39005	Fixed Wire-wound (Accurate)
MIL-R-39007	Fixed Wire-wound (Power Type)
MIL-R-39009	Fixed Wire-wound (Power Type, Chassis Mounted)
MIL-R-39015	Variable Wire-wound (Lead Screw Actuated)
MIL-R-39032	Resistors, Package of
MIL-R-39035	Variable Nonwire-wound (Adjustment Type)
MIL-R-83401	Fixed, Film Resistor Networks
MIL-R-23648	Thermistor (Thermally Sensitive, Resistor) Insulated
GSFC-S-311-P18	Thermistor (Thermally Sensitive, Resistor) Insulated, NTC

3.1.3 General definitions. A list of common terms used in rating and design application of resistors follows.

3.1.3.1 Resistors, general.

Ambient operating temperature. The temperature of the air surrounding an object, neglecting small localized variations.

Critical value of resistance. For a given voltage rating and a given power rating, this is the only value of resistance that will dissipate full rated power at rated voltage. This value of resistance is commonly referred to as the critical value of resistance. For values of resistance below the critical value, the maximum voltage is never reached; and for values of resistance above critical value, the power dissipated becomes lower than rated. Figure 2 shows this relationship.

3.1 RESISTORS, GENERAL

Dielectric strength. The breakdown voltage of the dielectric or insulation of the resistor when the voltage is applied between the case and all terminals are tied together. Dielectric strength is usually specified at sea level and at simulated high air pressures.

Hot-spot temperature. The maximum temperature measured on the resistor due to both internal heating and the ambient operating temperature. Maximum hot-spot temperature is predicted on thermal limits of the materials and the design. The hot-spot temperature is also usually established as the top temperature on the derating curve at which the resistor is derated to zero power.

Insulation resistance. The dc resistance measured between all terminals connected together and the case, exterior insulation, or external hardware.

Maximum working voltage. The maximum voltage stress (dc or rms) that may be applied to the resistor is a function of the materials used, the required performance, and the physical dimensions (see Figure 2).

Noise. An unwanted voltage fluctuation generated within the resistor. Total noise of a resistor always includes Johnson noise which is dependent on resistance value and temperature of the resistance element. Depending on type of element and construction, total noise may also include noise caused by current flow and noise caused by cracked bodies and loose end caps or leads. For variable resistors, noise may also be caused by jumping of the contact over turns of wire and by an imperfect electrical path between contact and resistance element.

Standard resistance value. The resistance value tabulated by a decade chart is specified in the applicable military specification. Resistance values not listed in the chart for the appropriate tolerance are considered as nonstandard for that specification.

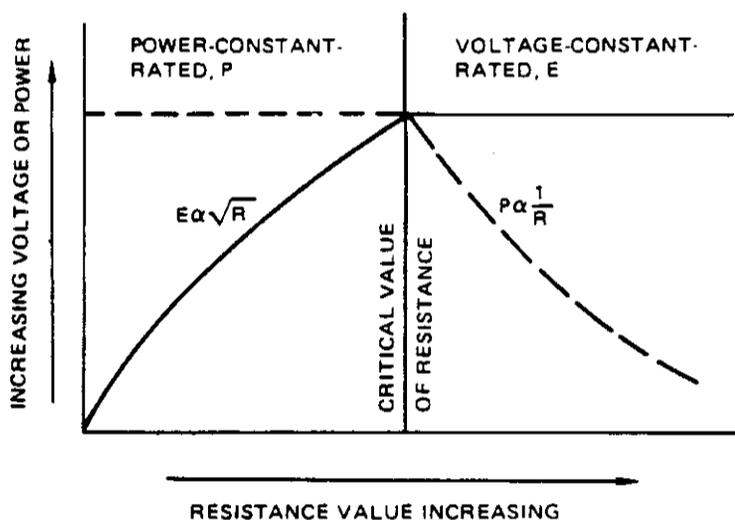


FIGURE 2. Maximum working voltage and critical value of resistance.

3.1 RESISTORS, GENERAL

Resistance tolerance. The permissible deviation of the manufactured resistance value (expressed in percent) from the specified nominal resistance value at standard (or stated) environmental conditions.

Stability. The overall ability of a resistor to maintain its initial resistance value over extended periods of time when subjected to any combination of environmental conditions and electrical stresses.

Temperature coefficient (resistance temperature characteristics). The magnitude of change in resistance due to temperature; it is usually expressed in percent per degree centigrade or parts per million per degree centigrade (ppm/°C). If the changes are linear over the operating temperature range, the parameter is known as "temperature coefficient (TC)"; if nonlinear, the parameter is known as "resistance temperature characteristic."

Voltage coefficient (resistance voltage characteristic). Certain types of resistors exhibit a variation of resistance entirely due to changes in voltage across the resistor. This characteristic is the voltage coefficient; it is generally applicable to resistors of 1,000 ohms and over.

3.1.3.2 Resistors, variable.

End resistance. The resistance measured between the wiper terminal and an end terminal with the wiper element positioned at the corresponding end of its mechanical travel.

Noise. The effective contact resistance introduced in the wiper arm while rotating the wiper across the resistor element. Wirewound potentiometers are normally referred to in terms of equivalent noise resistance (ENR) which is caused primarily by variations in wiper contact during travel along the wire element. Although nonwirewound potentiometers have a continuous resistor element, there is usually a built-in dc offset due to measurable contact resistance between the wiper and the resistor. Their noise is therefore normally referred to in terms of contact-resistance variation (CRV) which is caused primarily by the changes in contact resistance between the wiper and the resistor element during rotation.

Resolution or adjustability. The ability of an operator to predetermine ohmic value, voltage, or current. This is a measure of the sensitivity or degree of accuracy to which a potentiometer may be set. Wirewound potentiometers, having noncontinuous elements, are typically referred to in terms of a theoretical resolution which is the reciprocal of the number of turns of wire. Adjustability is affected by the material and the uniformity of the resistor and the wiper elements, the length of the resistor element, and the design of the adjustment mechanism.

Rotational life. The number of cycles of rotation which can be attained at certain operating conditions while remaining within specified allowable parametric criteria. A cycle comprises the travel of the wiper along the total resistor element in both directions.

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Setting stability. The ability of a potentiometer to maintain its initial setting during mechanical and environmental stresses, it is normally expressed as a percentage change in output voltage with respect to the total applied voltage.

Torque. The mechanical moment of force applied to a potentiometer shaft. Starting torque is the maximum torque required to initiate shaft rotation. Stop torque is the maximum torque which can be applied to the adjustment shaft at a mechanical end-stop position.

Total resistance. The resistance measured between the two end terminals of a potentiometer.

Travel. The clockwise or counterclockwise rotation of the wiper along the resistor element. Mechanical travel is the total rotation of the wiper between end-stop positions. Electrical travel is the rotation between maximum and minimum resistance values. These are not identical owing to discontinuities at the end positions.

3.1.3.3 Resistor network, fixed, film.

Aspect ratio. The geometical relationship of the length and width (L/W) of a rectangular resistor element which is used in the layout of networks to establish "as fired" resistance values (e.g., when L = W, the aspect ratio is 1:1, or 1 "square."

DIP. Dual-in-line-package.

Formulation. A specific mix of thick-film material where the conductor, glass, and other additive ingredients are formulated to provide certain properties such as a specific sheet resistivity or TCR.

Paste/ink. Screenable thick-film material comprised of metals, oxides, and glasses in an organic vehicle which when fired, produces a circuit element such as a resistor or conductor.

Power density. The power dissipation per unit area of a resistor or substrate (in watts per square inch) used to determine the optimum layout design of a network.

Screen. The process of printing a network pattern of thick-film ink or paste onto a substrate by means of a squeegee applied to a photoetched wire-mesh "silk screen" or metal mask.

Sheet resistivity. The nominal resistance per unit area of a thick-film ink or paste which is usually expressed in ohms per square inch (assuming a constant thickness) where the design resistance value of the screened resistor is determined by $R = \rho L/W$.

3.1 RESISTORS, GENERAL

Substrate. The base or carrier for the thick-film network which is usually a ceramic plate.

Thick-film/cermet/metal glaze. Resistor and conductor materials comprised of metals or metal oxides in a glass binding system which can be screened onto a substrate and fired to provide circuit elements or networks.

Tracking. The inherent capability of resistors from the same formulation and screened onto the same substrate to exhibit similar performance characteristics (e.g., drift, TCR).

Trim/abrade/adjust. The process of tailoring a thick-film-resistor element to a specific value or tolerance by the removal of resistor material (by means of sandblasting or laser abrading) which increases the ohmic value.

Voltage gradient/field strength. The linear voltage stress applied across a resistor element (in volts per inch) used to determine the optimum geometry of a high-voltage resistor.

3.1.3.4 Resistors, thermal.

Current-time characteristic. The relationship, at a specified ambient temperature, between the current through a thermistor and time elapsed from the application of a step function of voltage.

Dissipation constant (σ). The ratio, at a specified ambient temperature, of change in power dissipation in a thermistor to the resultant body temperature change.

Maximum operating temperature. The maximum body temperature at which a thermistor will operate for an extended period of time with acceptable stability of its characteristics. This temperature is the combination of external and internal self heating.

Maximum power rating. The maximum power rating of a thermistor is the maximum power which a thermistor will dissipate for an extended period of time with acceptable stability of its characteristics.

Negative temperature coefficient (NTC). An NTC thermistor is one in which the zero-power resistance decreases with an increase in temperature.

Positive temperature coefficient (PTC). A PTC thermistor is one in which the zero-power resistance increases with an increase in temperature.

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Resistance ratio. The ratio of the zero-power resistances of a thermistor measured at two specified reference temperatures.

$$\frac{R_o (T_1)}{R_o (T_2)} = e^{\beta \left(\frac{1}{T_1} - \frac{1}{T_2} \right)}$$

where:

$R_o (T_1)$ is the resistance at absolute temperature T_1 .

$R_o (T_2)$ is the resistance at absolute temperature T_2 .

e is 2.718.

β is a constant which depends on the material used to make the thermistor.

Stability. The ability of a thermistor to retain specified characteristics after being subjected to environmental and/or electrical test conditions.

Standard reference temperature. The thermistor body temperature at which nominal zero-power resistance is specified.

Temperature-wattage characteristic. The relationship, at a specified ambient temperature, between the thermistor temperature and the applied steady-state wattage.

Thermistor. A thermally sensitive resistor whose primary function is to exhibit a change in electrical resistance with a change in body temperature.

Thermal time constant (τ). The time required for a thermistor to change 63.2% of the difference between its initial and final body temperature when subjected to a step function change in temperature under zero-power conditions.

Zero-power resistance (R_o). The resistance value of a thermistor at a specified temperature with zero electrical power dissipation.

Zero-power resistance temperature characteristic. The relationship between the zero-power resistance of a thermistor and its body temperature.

Zero-power temperature coefficient of resistance (α_T). The ratio at a specified temperature, T , of the rate of change of zero-power resistance with temperature to the zero-power resistance.

$$\alpha_T = \frac{1}{R} \left(\frac{dR}{dT} \right)$$

3.1.4 NASA standard parts. See the introduction Section 1.1 for a complete description of the NASA Standard Parts Program. In addition to this handbook, the principal elements of this program include MIL-STD-975.

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- b. Nominal minimum resistance tolerance available is ± 0.1 percent for fixed, film resistors and ± 1.0 percent for the resistor networks.
- c. Maximum practical full-power operating ambient temperature should not exceed 125°C for metal film RNC types and 70°C for RLR resistors. Type RZ resistor networks are continuously derated from 70°C to 125°C .
- d. Operation at rf (above 100 MHz) may produce inductive effects on spiral-cut type fixed, film resistors, and capacitive effects on the resistor networks.
- e. The resistance temperature characteristic is fairly low (± 200 ppm/ $^{\circ}\text{C}$) for thick-film types (RLR) and very low (± 25 ppm/ $^{\circ}\text{C}$) for metal-film types (RNC). The resistance temperature characteristic is fairly low (± 300 ppm/ $^{\circ}\text{C}$, ± 100 ppm/ $^{\circ}\text{C}$ and ± 50 ppm/ $^{\circ}\text{C}$) for resistor networks (RZ).
- f. Electrostatic effects should be considered since resistors can change value when subjected to electrostatic charges. Resistors with a tolerance of 0.1 percent should be packaged in accordance with MIL-R-39032.

3.1.5.3 Fixed, wirewound (accurate) resistors, RBR.

- a. Fixed, wirewound, accurate resistors are physically the largest of all types for a given resistance and power rating. They are very conservatively rated and are available in standard tolerances as low as ± 0.02 percent.
- b. They are used where high cost and large size are not critical and operational climate can be controlled.
- c. Application of voltages in excess of voltage rating may cause insulation breakdown in the thin coating of insulation between element coating.
- d. Operation above 50 kHz may produce inductive effects and intrawinding capacitive effects.
- e. The resistance element is quite stable within specified temperature limits.
- f. Use of good soldering techniques is extremely important since higher contact resistance may cause overall resistance shifts far outside of resistance tolerance on low value units.
- g. The presence of moisture may degrade coating or potting compounds.

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3.1 RESISTORS, GENERAL

MIL-STD-975 is a standard parts list for NASA equipment with section 10 containing a summary of standard resistors.

These parts all use basic military ordering references conforming to the type designation according to the applicable military specification for the respective part as described in the paragraphs of this section.

3.1.5 General device characteristics.

3.1.5.1 Fixed, composition resistors, RCR.

- a. The nominal minimum resistance tolerance available for fixed, composition resistors is ± 5 percent. Combined effects of climate and operation on unsealed types may raise this tolerance to ± 15 percent from the low value (i.e., aging, pressure, temperature, humidity, voltage gradient, etc.).
- b. High-voltage gradients will produce resistance change during operation.
- c. High Johnson noise levels at resistance above 1 megohm preclude use in critical circuits of higher sensitivity.
- d. Radio frequency will produce end-to-end shunted capacitive effects because of short resistor bodies and small internal distances between both ends.
- e. Operation at VHF or higher frequency reduces effective resistance due to losses in the dielectric (the so-called Boella effect).
- f. Exposure to humidity may have two effects on the resistance value. Surface moisture may result in leakage paths which will lower the resistance values or absorption of moisture into the element may increase the resistance. These phenomena are more noticeable in higher values of resistance.

When exposed to a humid atmosphere while dissipating less than 10% of rated power or in shelf storage, nonoperating equipment, and shipping conditions, resistance values may change as much as 15%. Before being considered failures, out of tolerance resistors should be conditioned in a dry oven at $100 \pm 5^\circ\text{C}$ for 48 hours.

Resistors which continue to be out of tolerance after conditioning should be considered failures.

3.1.5.2 Fixed, film resistors, RNC, RLR, and fixed, film networks, RZ.

- a. These resistors are low tolerance, high stability, low environmental changes, low temperature coefficient, space and weight saving, and low noise.

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3.1.5.4 Fixed, wirewound resistors (power type), RER and RWR. This type resistor is generally not supplied in low tolerances and its frequency response is poor. However, its temperature coefficient is low.

3.1.5.5 General characteristics of variable resistors, RJR and RTR.

- a. All types of variable resistors should be derated for operation above their rated ambient temperature.
- b. Wirewound types should not be used in frequency-sensitive rf circuits due to introduction of inductive and capacitive effects.
- c. High humidity conditions may have a deleterious effect on unenclosed types due to the resistance shift in composition types and winding-to-winding shorts in wirewound types.
- d. Nonwirewound elements may wear away after extended use leaving particles of the elements to permeate the mechanism and resulting in warmer operation, high-resistance shorts, etc. Wirewound types are subject to noise because of stepping of the contact from wire-to-wire.
- e. With either wirewound or nonwirewound resistors, good practice indicates the use of enclosed units to keep out as much dust and dirt as possible and to protect the mechanism from mechanical damage. The presence of oil through lubrication may cause dust or wear particles present to concentrate within the unit.
- f. Select a variable resistor with a power rating sufficient to handle the higher current produced when the resistor is reduced, particularly if it is being used in series as a voltage-dropping resistor.
- g. When a variable wirewound linear resistor is being used as a voltage divider, the output voltage through the wiper will not vary linearly if current is being drawn through it. This characteristic is usually called the "loading error." To reduce the loading error, the load resistance should be at least 10 to 100 times as great as the end-to-end potentiometer resistance.
- h. No current should be drawn from the wiper of a nonwirewound resistor.

3.1.6 General parameter information. Resistors must be selected to be compatible with the conditions to which they are exposed. Numerous factors must be considered in this selection process. The most important are noted in the following:

3.1.6.1 Resistance value. These are initially determined by the circuit requirements. Usually these values need to be adjusted to make them closely match the standard resistance values supplied by the manufacturer, or listed in the military specifications. If it is impossible to adjust circuit values to a standard value, parallel or series combination resistors can be used.

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The design engineer should also remember that the resistance value of the resistor that is put into the physical circuit will differ from the value he has on his circuit schematic. This difference will change as time passes. The purchase tolerance of the resistor to be used will allow it to differ from the nominal stated value, depending on the type of resistor specified. Furthermore, the temperature at which the resistor works, the voltage across it, and the environment it encounters will affect the actual value at particular times. For example, the designer should allow for a possible variation of ± 15 percent from the nominal value of a purchased ± 5 percent composition resistor if he expects his circuit to continue to operate satisfactorily over a very long time under moderate ambient conditions. Such a figure is a rule of thumb based on many tests and most resistors will remain much nearer their starting value; but if many are used, chance will ensure that some will go near this limit.

3.1.6.2 Power rating. The minimum required power rating of a resistor is another factor that is initially set by the circuit usage but is markedly affected by the other conditions of use. As mentioned previously, the power rating is based on the hot-spot temperature the resistor will withstand while still meeting its other requirements of resistance variation, accuracy, and life.

Self-generated heat. Self-generated heat in a resistor is calculated as $P = I^2R$. This figure, in any circuit, must be less than the actual power rating of the resistor used. It is practice to calculate this value and to use the next larger power rating available in the standard. This calculation should, however, be considered only as a first approximation of the actual rating to be used.

Rating versus ambient conditions. The power of a resistor is based on a certain temperature rise from an ambient temperature of a certain value. If the ambient temperature is greater than this value, the amount of heat that the resistor can dissipate is correspondingly reduced, and therefore it must be derated because of temperature. All military specifications contain derating curves to be used for the resistors covered.

Rating versus accuracy. Because of the temperature coefficient of resistance that all resistors possess, a resistor which is expected to remain near its measured value under conditions of operation must remain cool. For this reason, all resistors designated as accurate are very much larger physically for a certain power rating than are ordinary "nonaccurate" resistors. In general, any resistor, accurate or not, must be derated to remain very near its original measured value when it is being operated.

Rating versus life. If especially long life is required of a resistor, particularly when "life" means remaining within a certain limit of resistance drift, it is usually necessary to derate the resistor even if ambient conditions are moderate and if accuracy by itself is not important. A good rule to follow when choosing a resistor size for equipment that must operate for many thousands of hours is to derate it to one-half of its nominal power rating. Thus,

3.1 RESISTORS, GENERAL

if the self-generated heat in the resistor is 1/3-watt, do not use a 1/2-watt resistor, but rather a 1-watt resistor. This will automatically keep the resistor cooler, will reduce the long-term drift, and will reduce the effect of the temperature coefficient. For equipment demanding small size but with a relatively short use life, this rule may be impractical. The engineer should adjust his dependence on rules to the circumstances at hand. A "cool" resistor will generally last longer than a "hot" one and it can absorb overloads that might permanently damage a "hot" resistor.

Rating under pulsed conditions and under intermittent loads. When resistors are used in pulse circuits, the actual power dissipated during the pulses can sometimes be much more than the maximum rating of the resistor. For short pulses the actual heating is determined by the duty factor and the peak power dissipated. Before approving a resistor application, the engineer should be sure that: (1) the maximum voltage applied to the resistor during the pulses is never greater than the permissible maximum voltage for the resistor being used, (2) the circuit cannot fail in such a way that continuous excessive power can be drawn through the resistor and cause it to fail, (3) the average power being drawn is well within the agreed-on rating of the resistor, and (4) continuous steep wavefronts applied to the resistor do not cause any unexpected troubles.

3.1.6.3 Derating. With the exception of failures due to "random" occurrences resulting from manufacturing defects or overstress, resistor failure rate is a function of time, temperature, and applied power. Operational life can be significantly lengthened by power derating and by limiting the operating ambient temperature.

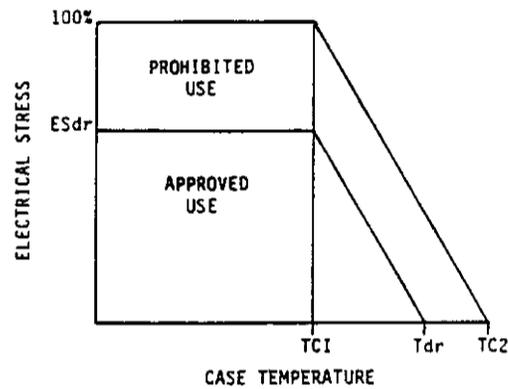
The extent to which electrical stress (e.g., voltage, current, and power) is derated depends upon temperature. The general interrelationship between electrical stress and temperature is shown in Figure 3. The approved operating conditions lie within the area below the derated limitation line (ESdr). Operation at conditions between the derated limitation line and the maximum specification curves results in lower reliability (see MIL-HDBK-217). Operation in this reduced reliability area requires specific approval.

Numerical values are applied to the curves for each part type based on a percentage of the device manufacturer's maximum rated values. The applicable derating curve or derating percentages are specified in MIL-STD-975.

3.1.6.4 Contact resistance variation. The apparent resistance seen between the wiper and the resistance element when the wiper is energized with a specified current and moved over the adjustment travel in either direction at a constant speed. The output variations are measured over a specified frequency bandwidth, exclusive of the effects due to roll-on or roll-off of the terminations and are expressed in ohms or percent of total nominal resistance.

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T_{C1} = case temperature above which electrical stress must be reduced.

T_{C2} = maximum allowable case temperature.

T_{dr} = maximum case temperature for derated operation.

ES_{dr} = maximum electrical stress (e.g., voltage, power, and current) for derated operation.

100% = maximum rated value in accordance with the detail specification.

FIGURE 3. Stress-temperature derating.

3.1.6.5 High frequency. For most resistors the lower the resistance value, the less total impedance it exhibits at high frequency. Resistors are not generally tested for total impedance at frequencies above 120 hertz (Hz). Therefore, this characteristic is not controlled. The dominating conditions for good high frequency resistor performance are geometric considerations and minimum dielectric losses. For the best high frequency performance the ratio of resistor length to the cross-sectional area should be a maximum. Dielectric losses are kept low by proper choice of the resistor base material and when dielectric binders are used, their total mass is kept to a minimum. The following is a discussion of the high-frequency merits of these major resistor types.

- a. Carbon composition. Carbon composition resistors exhibit little change in effective dc resistance up to frequencies of about 100 kHz. Resistance values above 0.3 megohm start to decrease in resistance at approximately 100 kHz. Above frequencies of 1 megahertz all resistance values exhibit decreased resistance.
- b. Wirewound. Wirewound resistors have inductive and capacitive effects and are unsuited for use above 50 kHz, even when specially wound to reduce the inductance and capacitance. Wirewound resistors usually exhibit an increase in resistance with high frequencies because of skin effect.

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- c. Film type. Film-type resistors have the best high frequency performance. The effective dc resistance for most resistance values remains fairly constant up to 100 MHz and decreases at higher frequencies. In general, the higher the resistance value, the greater the effect of frequency.

3.1.6.6 Insulation or coating. All resistors intended for use in reliable electronic equipment must be protected by an insulating coating. Coatings can be a molded phenolic case, epoxy coating, or ceramic or glass sleeves. Wire-wound power resistors use various cement and vitreous enamel coatings to protect the windings and to insulate and provide moisture barriers. Not all of the coatings and insulations applied to commercial resistors are satisfactory for extreme variations in ambient conditions. The various military specifications include tests used to qualify the various manufacturers' products thus providing a greater confidence in the coating. Some insulation coatings may be susceptible to outgassing under vacuum conditions.

3.1.6.7 Effects of ambient conditions. In the establishment of ratings for resistors, the design engineer has implicitly considered the mechanical design of the equipment. This is because the ambient conditions in which the resistor must operate determine the power rating and mechanical construction of the resistor.

Resistor heating. A very important question in the application of resistors is how hot will they get in service. In a piece of equipment the heat in a resistor comes from two sources: (1) self-generated heat and (2) heat that the resistor receives from other heat-producing components in the same neighborhood. The important thing to remember is that under these conditions each resistor will be heated more than I^2R . When much heat is produced, as in stacked wire-wound resistors, the design engineer would do well not to freeze his design until he has measured a typical assembly with power on to see just how hot the resistors get. The same thing is true of the extra heating given the resistors by convection.

This is another way of saying that high ambient temperature will reduce the actual power rating of the resistor by reducing permissible temperature rise. The equipment designer must realize also that the heat being produced by "hot" resistors can damage other components. Resistors usually do not fail immediately when overheated. The effect of too much heat deteriorates the component until at a later date fails. It is very easy to put a "heat bomb" in a piece of equipment that will not go off in normal production testing but will do so when the equipment gets into service and is being relied on to do its job. It is also very easy to eliminate such troubles by strict and thoughtful attention to the problem of heating.

High altitude or vacuum. With the exception of the dielectric withstanding voltage test at reduced barometric pressure, all tests in military specifications referenced herein are performed at ambient atmospheric pressure. This fact should be considered when the use of these resistors for high-altitude conditions is contemplated.

3.1 RESISTORS, GENERAL

Flammability. It should be noted that military specifications referenced herein contain no requirements concerning the flammability of the materials used in the construction of these resistors. Users should take this into consideration when a particular application involves this requirement.

Resistance tolerance versus temperature coefficient. During the selection of resistor characteristics choices must be made for resistance tolerance and temperature coefficient of resistance change. In nonwirewound film resistors the cost of obtaining low tolerance is often minimal when compared with the cost of obtaining a very low temperature coefficient. The low tolerance is obtained by careful adjustment of the width of the spiral in the resistance element; this can be done at a low cost. The temperature coefficient is controlled by processing of the film, and low temperature coefficients are expensive to obtain.

In many instances, the designer can meet circuit requirements through a trade-off of tolerance for temperature coefficient. A low tolerance selection may achieve circuit performance requirements in lieu of a low temperature coefficient with a resulting cost saving. This is particularly true when using film type resistors, MIL-R-55182.

3.1.6.8 Backlash in variable resistors. Lead screw actuated variable resistors can provide a high degree of accuracy in critical adjustments; however, the user should consider the effects of backlash in the lead screw position versus wiper position. The resistance obtained at an initial setting may change slightly under conditions of vibration and shock as the wiper settles into a new position. The magnitude of this change is allowed to be as high as 1 percent when new and can increase with age up to about 3 percent or the equivalent of one-half turn of the lead screw. In extremely critical applications, it may be desirable to decrease the resistance value of the variable resistor and add a suitable fixed resistance in series to obtain the same overall resistance, thus giving less critical adjustments but with a decrease in the adjustable range.

3.1.7 General guides and charts.

3.1.7.1 Mounting and handling. Practical guides for mounting and handling of resistors are described in the following paragraphs.

Stress mounting. Improper heat dissipation is the predominant cause of failure for any resistor type. Consequently the lowest possible resistor surface temperature should be maintained. Figure 4 illustrates the manner in which heat is dissipated from fixed resistors in free air. The intensity of radiated heat varies inversely with the square of the distance from the resistor. Maintaining maximum distance between heat-generating components serves to reduce cross-radiation heating effects and promotes better convection by increasing air flow. For optimum cooling without a heat sink, small resistors should have large diameter leads of minimum length terminating in tie points of sufficient

3.1 RESISTORS, GENERAL

mass to act as heat sinks. All resistors have a maximum surface temperature which should never be exceeded. Any temperature beyond maximum can cause the resistor to malfunction. Resistors should be mounted so that there are no abnormal hot spots on the resistor surface. When mounted, resistors should not come in contact with heat-insulating surfaces.

Resistor mounting for vibration. Resistors should be mounted so resonance does not occur within the frequency spectrum of the vibration environment to which the resistors may be subjected. Some of the most common resistor packaging methods result in large resistor noise. Resistor mounting for vibration should provide: (1) the least tension or compression between the lead and body, (2) the least excitation of the resistor with any other surface, and (3) no bending or distortion of the resistor body.

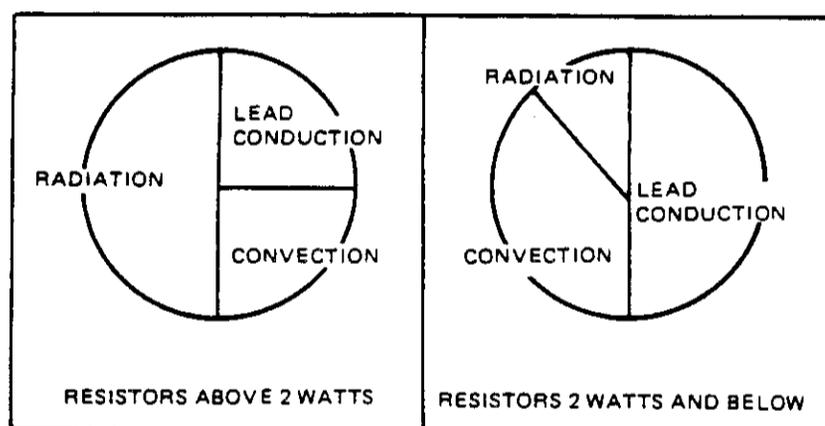


FIGURE 4. Heat dissipation of resistors under room conditions.

Circuit packaging. Even well insulated resistors that are crowded together and come into contact with each other can provide leakage paths for external current passage. This can change the resultant resistance in the circuit. Moisture traps and dirt traps are easily formed by crowding. Moisture and dirt eventually form corrosive materials which can degrade the resistors and other electronic parts. Moisture can accumulate around dirt even in an atmosphere of normal humidity. Proper space utilization of electronic parts can reduce the package size and still provide adequate spacing of parts.

Summary. The following is a guide for resistor mounting.

- a. Maintain lead length at a minimum. The mass of the tie point acts as a heat sink.

Lead should be offset (bent slightly) to allow for thermal contraction where low temperatures are present.

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3.1 RESISTORS, GENERAL

- b. Close tolerance and low-value resistors require special precautions such as short leads and good soldering techniques; the resistance of the leads and the wiring may be as much as several percent of the resistance of the resistor.
- c. Maintain maximum spacing between resistors.
- d. For resistors mounted in series, consider the heat being conducted through the leads to the next resistor.
- e. Large power units should be mounted to the chassis.
- f. Do not mount high-power units directly on terminal boards or printed circuits.
- g. To provide for the most efficient operation and even heat distribution, power resistors should be mounted in a horizontal position.
- h. Select mounting materials that will not char and can withstand strain due to expansion.
- i. Consider proximity to other heat sources as well as self-heating.
- j. Consider levels of shock and vibration to be encountered. Where large body mass is present, the body should be restrained from movement.

3.1.7.2 Effects of mechanical design and ambient conditions. Since the operation of a circuit cannot be divorced from the physical configuration it assumes when assembled, some of the points that apply herein have already been discussed. It is good, however, to check this aspect of equipment design several times, so the redundancies in the following paragraphs are deliberate for the sake of emphasis.

Mechanical design of resistors. Much trouble during the life of the equipment can be eliminated if the design engineer can be sure that the resistors specified for the circuits are soundly constructed and proper equipment assembly techniques are utilized. The resistor types listed in this handbook provide a great measure of this assurance and, in general, assure a uniform quality of workmanship.

End-caps or terminations. The connection between the resistive element and the leads attached to end-caps or terminations must be sound so that none of the stresses encountered by the device cause intermittent connections. This point is addressed in the referenced military specification. Special precautions must be taken during automatic circuit assembly to avoid damage to this connection.

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Effect of soldering. There are assembly techniques that affect resistor reliability. Resistors should never be overheated by excessive soldering-iron applications and the resistor leads should not be abraded by assembly tools. No normal soldering practice, either manual or dip soldering, should damage the resistor physically or change its resistance value appreciably.

Moisture resistance. Moisture is the greatest enemy of components and electronic equipment. Usually a resistor remains dry because of its own self-generated heat; this is true only when the equipment is turned on. If the equipment must stand for long periods under humid conditions without power applied, one should determine whether or not the circuits will operate with resistance values which have changed from the "hot" condition, and whether or not the resistance value during the warmup period will allow the equipment to work satisfactorily during this period. If it will not, he must see that a resistor adequately protected against moisture absorption is used. It is therefore up to the design engineer to analyze the need and to provide a resistor to meet these conditions. This handbook and the applicable military specifications constitute a guide as to what the various kinds of resistors will do under humid conditions.

Method of mounting. Large resistors that are not provided with some adequate means of mounting should not be considered. Under conditions of vibration or shock, lead failure can occur and the larger the mass supported by the leads the more probable a failure will be. Even when vibration or shock will not be a serious problem, ease of assembly and replaceability suggest that large components be mounted individually.

Resistor body. The body of the resistor must be sufficiently strong to withstand any handling it is likely to get. The specification should require, through workmanship and packaging requirements, that the manufacturer show that his product will not crack, chip, or break in transit, on the shelf, or in the normal assembly process.

The charts in Tables I and II will guide the selection of resistor types by comparing the order of merit for each listed characteristic.

3.1.8 Prediction model. To predict the reliability of electronic equipment, it is necessary to be able to predict the failure rate of individual parts. Failure rate factors and a procedure for calculating part failure rate are provided in MIL-HDBK-217. Different factors are required for each type of resistor. Information in this section is given to demonstrate and stress controllable factors which affect reliability. For additional discussion of this subject, see section 1.4 of this standard. To perform an actual prediction, consult the latest revision of MIL-HDBK-217.

3.1.8.1 General reliability considerations. Reliability considerations for each type of resistor are included in the detail subsection for the respective resistor type.

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3.1 RESISTORS, GENERAL

TABLE I. Selection chart for relative performance of fixed resistors

Characteristic	Order of Merit	
	Best	Poorest
Accuracy	RBR RNC RWR RER RLR RCR	
Cost/unit	RCR RLR RNC RWR RER RBR	
High-frequency performance	RNC RLR RCR RER RWR RBR	
Operating temperature range	RWR RER RNC RLR RCR RBR	
Resistance range	RCR RNC RLR RBR RWR RER	
Stability	RBR RNC RWR RER RLR RCR	
Temperature coefficient	RBR RWR RER RNC RLR RCR	
Wattage range	RER RWR RCR RLR RNC RBR	
Watts/size	RER RWR RCR RLR RNC RBR	

TABLE II. Selection chart for relative performance of variable resistors (lead screw adjustable)

Characteristic	Order of Merit	
	Better	Poorer
Accuracy	RTR	RJR
Cost/unit	RJR	RTR
High-frequency performance	RJR	RTR
Operating temperature range	RTR	RJR
Resistance range	RJR	RTR
Stability	RTR	RJR
Temperature coefficient	RTR	RJR
Wattage range	RTR	RJR
Watts/size	RTR	RJR

3.1 RESISTORS, GENERAL

Improved reliability of resistors can be obtained by the following additional efforts.

- a. Additional voltage and power derating of the devices to avoid insulation failures and accelerated deterioration due to high temperature. (Voltage and power are commonly derated to 80 and 50 percent of rated values, respectively.)
- b. Detailed visual inspection of materials and assembly operations as the resistor is manufactured, prior to the coating operation.
- c. Operating burn-in reliability screening tests on the resistor devices to screen out infant mortality failures prior to the resistors being assembled in circuits.
- d. Care in handling of resistors and inspection for damage to resistors caused during equipment manufacturing operations.

3.1.8.2 Radiation considerations. Past experimental evidence and theory have shown that resistors are less sensitive to nuclear radiation than other components such as transistors, diodes, integrated circuits, etc.

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**3.2 RESISTORS, FIXED
COMPOSITION (INSULATED)**

3.2 Fixed, composition (insulated).

3.2.1 Introduction. Resistors covered in this section are established reliability carbon composition resistors having a composition resistance element and axial leads. These resistors provide life failure rates ranging from 1.0 percent to 0.001 percent per 1,000 hours at 50 percent of full-load operation at an ambient temperature of 70°C. The failure rates are established at a 60-percent confidence level and maintained at a 10-percent producer's risk. The failure rate refers to operation at rated temperature and at rated voltage. The change in resistance between the initial measurement and any of the succeeding measurements up to and including 2000 hours shall not exceed ±15 percent. It is expected that the resistance change will not exceed ±15 percent to 10,000 hours of life testing.

3.2.1.1 Applicable military specification. MIL-R-39008, General Specification for Established Reliability, Fixed, Composition (Insulated) Resistors.

3.2.2 Usual applications. Since the fixed composition resistor is relatively inexpensive, highly reliable, and readily procurable in all standard values, it should be the first resistor considered for all applications. However, it is subject to resistance change with humidity, temperature, soldering, and shelf life and consequently, should normally be used in circuits which do not demand the stability of film or wirewound types.

The fixed composition resistor is limited in its applications because it is too noisy for many high gain circuits and its resistance falls off as frequency rises. The composition resistor has low resistance into the megahertz region.

3.2.3 Physical construction. In these resistors, the resistance element consists of a mixture of carbon, insulating material, and suitable binders either molded together or applied as a thin layer of conducting material on an insulated form. These resistors are covered by a molded jacket which is primarily intended to provide an adequate moisture barrier for the resistance element as well as mechanical protection and strength (see Figure 5). Due to the reliability requirements of MIL-R-39008, processes and controls utilized in manufacturing are stringent.

Physical dimensions. An outline drawing for RCR42 is shown in Figure 6. Refer to MIL-R-39008 for other styles.

Maximum weight. Maximum weight for the largest resistor (RCR42) is 3 grams. Refer to MIL-R-34008 for other styles.

3.2 RESISTORS, FIXED COMPOSITION (INSULATED)

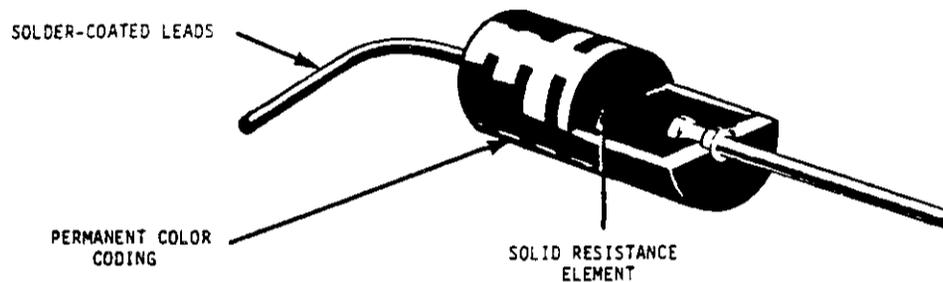


FIGURE 5. Typical construction of a carbon composition resistor.

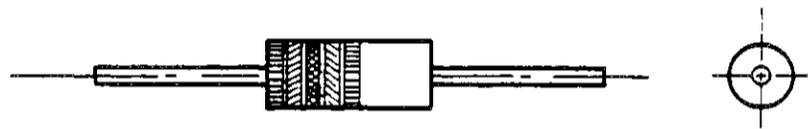
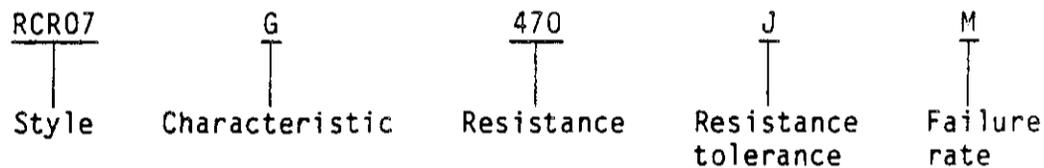


FIGURE 6. Outline drawing of a carbon composition resistor.

3.2.4 Military designation. An example of the military type designation is shown below.



3.2.5 Electrical characteristics. Electrical characteristics for established reliability, carbon composition resistors are tabulated in MIL-R-39008. A description of environmental tests with the allowable resistance changes are also tabulated in military specification MIL-R-39008.

Other electrical characteristics which must be considered in selection of the correct resistor for a particular application are as follows.

3.2.5.1 Derating. The failure rate of fixed composition resistors under operating conditions is a function of time, temperature, and applied power. Refer to MIL-STD-975 for specific derating conditions.

3.2.5.2 Peak voltages and pulsed operation. When carbon composition resistors are used under low-duty-cycle pulse conditions, the maximum permissible operating voltage is limited by breakdown rather than by heating. In such applications, the peak value of the pulse should not exceed 2 times the rated rms continuous working voltage for the type used. If the pulses are of sufficient

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**3.2 RESISTORS, FIXED
COMPOSITION (INSULATED)**

duration to raise the resistor's temperature excessively, the resistor must be derated even though the interval between pulses may be long enough to make the average heating small. In general, the above procedure must be used with caution if it permits the peak power to be more than approximately 30 to 40 times the normal power rating.

3.2.5.3 Noise. Thermal agitation or Johnson noise and resistance fluctuations or carbon noise are characteristic of carbon composition resistors. Use of these resistors in low level high-resistance (1 megohm or more) circuits should be avoided. Noise which can be expected is approximately 3 to 10 microvolts per volt. A film or wirewound resistor will usually yield more satisfactory results.

3.2.5.4 Voltage limitations. Voltage limits in the application of fixed composition resistors is often overlooked. These maximum permissible voltages, which are imposed because of insulation breakdown problems, must be taken into consideration in addition to the limitations of power dissipation. Figure 7 illustrates these boundary voltages for various size (wattage ratings) of composition resistors.

3.2.5.5 Voltage coefficient. When voltage is applied to low resistance value carbon composition resistors, resistance values may change by 2 percent, or by 0.05 percent per volt for resistors above 1,000 ohms for style RCR05, 0.035 percent per volt for resistors above 1,000 ohms for styles RCR07 and RCR20, and 0.02 percent per volt above 1,000 ohms for styles RCR32 and RCR42. The voltage coefficient for resistors below 1,000 ohms is not controlled by specifications and these resistors should not be used in circuits which are sensitive to this parameter.

3.2.5.6 Temperature-resistance. The resistance-temperature variation of carbon composition resistors cannot be defined by a temperature coefficient since the variation is nonlinear and has a different shape for different resistance values. (See Table III.)

TABLE III. Resistance-temperature characteristic

Maximum ambient operating temperature (100-percent rated wattage and 50-percent rated wattage for FR determination)	Nominal resistance	Maximum allowable change in resistance from resistance at 25°C ambient temperature	
		At -55°C (ambient)	At +105°C (ambient)
70°C	1,000Ω and under	±6.5%	±5%
	1,100Ω to 10,000 MΩ incl	±10%	±6%
	11,000Ω to 0.10 MΩ incl	±13%	±7.5%
	0.11 MΩ to 1.0 MΩ incl	±15%	±10%
	1.1 MΩ to 10 MΩ incl	±20%	±15%
	11.0 MΩ and over	±25%	±15%

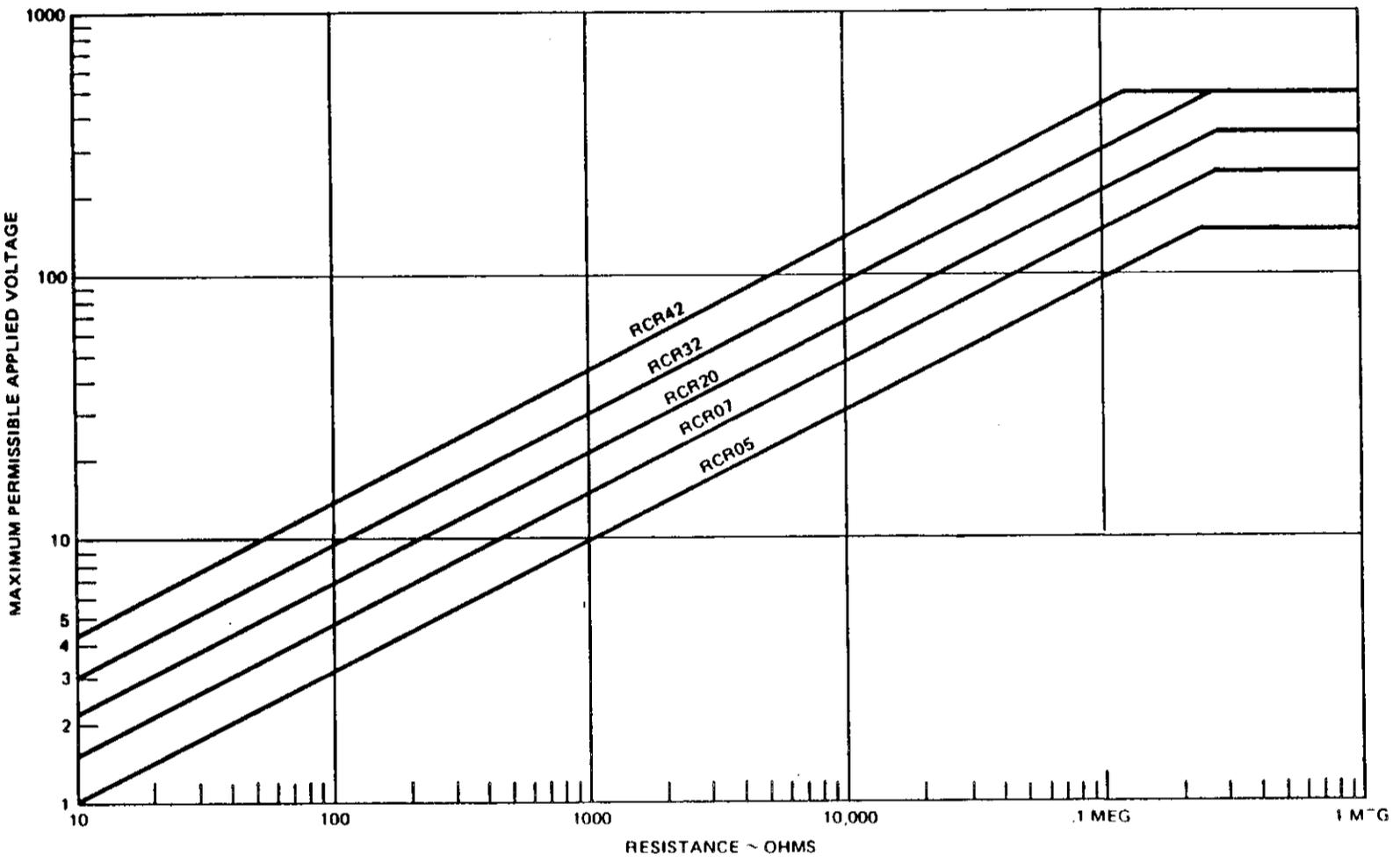


FIGURE 7. Voltage limitations by style.

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**3.2 RESISTORS, FIXED
COMPOSITION (INSULATED)**

3.2.5.7 High-frequency applications. When used in high-frequency circuits (1 megahertz and above), the effective resistance will decrease as a result of dielectric losses and shunt capacitance (both end-to-end and distributed to mounting surface). High frequency characteristics of carbon composition resistors are not controlled by specification and hence are subject to change without notice. Typical examples of changes in effective resistance are as follows.

- a. At 1 megahertz a 1/2-watt 100-kilohm resistor measures 90 percent of dc value.
- b. At 10 megahertz a 1/2-watt 100-kilohm resistor measures 55 percent of dc value.
- c. At 10 megahertz a 2-watt 1-megohm resistor measures 15 percent of dc value.

3.2.6. Environmental considerations. Established reliability, fixed, composition resistors are qualified to withstand environmental tests in accordance with Table IV of MIL-R-39008B.

Additional environmental considerations follow.

Shelf life. In general, these resistors exhibit resistance variations in shelf life as high as ± 15 percent due to moisture and temperature effects. It is recommended that after a storage period of approximately six months or more, these resistors be baked for 48 hours at 100°C (with no power applied) prior to usage. When closer life tolerance or higher accuracy is needed, resistors in accordance with MIL-R-55182 or MIL-R-39017 should be used.

Soldering. Care should be taken in soldering resistors because all properties of a composition resistor may be seriously affected when soldering irons are applied too closely to a resistor body or for too long a period. The length of lead left between the resistor body and the soldering point should not be less than 1/4 inch. Heat-dissipating clamps should be used, if necessary, when soldering resistors in close quarters. In general, if it is necessary to unsolder a resistor to make a circuit change or in maintenance, the resistor should be discarded and a new one used.

Moisture resistance. When exposed to humid atmosphere while dissipating less than 10 percent of rated power (including shelf storage, equipment nonoperating, and shipping conditions), resistance values may change up to 15 percent.

3.2.7. Reliability considerations. Fixed carbon composition resistors represent a well developed and stable technology with an inherently low failure rate. The following factors should be considered to determine the reliability of these resistors.

**3.2 RESISTORS, FIXED
COMPOSITION (INSULATED)**

3.2.7.1 Derating. Consideration must be given to the resistor's wattage rating. This is based on the materials used and is controlled by specifying a maximum hot-spot temperature. The amount of dissipation that can be developed in a resistor body at the maximum hot-spot temperature depends upon how effectively the dissipated energy is carried away and is a direct function of the ambient temperature. For operation continuously at full rating, the resistor must be connected to an adequate heat sink; this means approximately 1/2-inch leads connected to average size solder terminals with no other dissipative parts connected to the same terminals or mounted closer than one diameter. Appropriate derating must be imposed at elevated temperatures. Power dissipation capabilities of a resistor are usually lower when mounted in equipment than under test conditions. Most of the generated heat is carried away by the resistor leads. Therefore, when two resistors are connected to the same terminal, wattage ratings would be decreased approximately 25 percent. Close proximity of one resistor to another or to any other heat-generating part further reduces the wattage rating. Conformal coatings and encapsulating materials are poor heat conductors. When resistors are packaged in this manner, exercise caution in selection of the power rating. Refer to MIL-STD-975 for specific derating conditions.

Derating for optimum performance. For optimum performance, the following two "rules of thumb" have been in practice in industry for these resistors.

- a. After the anticipated maximum ambient temperature has been determined, a safety factor of 2 is applied to the wattage.
- b. Wattage is adjusted so that the hot-spot temperature does not exceed the following for the particular style.

RCR05 - 120°C
RCR07 - 120°C
RCR20 - 100°C
RCR32 - 100°C
RCR42 - 100°C

NOTE: It is recommended that either of the above techniques be considered in the application of these resistors.

3.2.7.2 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desired characteristics. Failure rate factors applicable to this specification are stated and discussed in MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures" and will differ from the failure rates established in the specification since this established failure rate is based on a "parametric failure" of ± 15 percent change in resistance from the initial measurement and any succeeding measurements taken up to and including 10,000 hours of life tests.

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**3.3 RESISTORS, FIXED, FILM
(HIGH STABILITY)**

3.3 Fixed, film (high stability).

3.3.1. Introduction. Resistors covered in this section are established reliability, fixed metal film resistors, including both hermetically and nonhermetically sealed types. These resistors possess a high degree of stability, with respect to time, under severe environmental conditions with an established reliability. These resistors provide life failure rates ranging from 1.0 percent to 0.001 percent per 1,000 hours. The failure rates are established at a 60-percent confidence level (initial qualification) and maintained at a 10-percent producer's risk. The failure rate is referred to operation at full-rated wattage and temperature with a maximum change in resistance of ± 2.0 percent from the initial measurement to any succeeding measurement up to and including 10,000 hours of life test.

3.3.1.1 Applicable military specification. MIL-R-55182, General Specification for Established Reliability, Fixed, Film, Resistors.

3.3.2. Usual applications. These resistors are designed for use in critical circuitry where high stability, long life, reliable operation, and accuracy are of prime importance. They are particularly desirable for use in circuits where high frequencies preclude the use of other types of resistors. Some of the applications for which these film-type resistors are especially suited are high-frequency, tuned circuit loaders, television side-band filters, rhombic antenna terminators, radar pulse equipment, and metering circuits such as impedance bridges and standing-wave-ratio meters.

3.3.3. Physical construction. In these resistors the resistance element consists of a metal film element on a ceramic substrate with the exception of the RNC90. The element is formed by the evaporation of a heated metal under vacuum conditions. The RNC90 consists of a metal foil on a flat substrate. Following spiraling or trimming to increase the available resistance values and the attachment of leads, the element is protected from environmental conditions by an enclosure (see Figures 8, 9, 10 and 11, and Table IV). Due to the reliability requirements of MIL-R-55182, processes and controls utilized in manufacturing are necessarily stringent.

TABLE IV. Terminal types

Characteristic	Terminal Designator (See MIL-R-55182)	Specification Indicates Weldable		Specification Indicates Solderable	
		N-Yes	R-No	N-No	R-Yes
C	N, R	N-Yes	R-No	N-No	R-Yes
H	C	Yes		Yes	
E	N, R	N-Yes	R-No	N-No	R-Yes
J	C	Yes		Yes	
K <u>1/</u>	C	Yes		Yes	
Y	C	Yes		Yes	

1/ Applicable to style RNC90 only.

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**3.3 RESISTORS, FIXED, FILM
(HIGH STABILITY)**

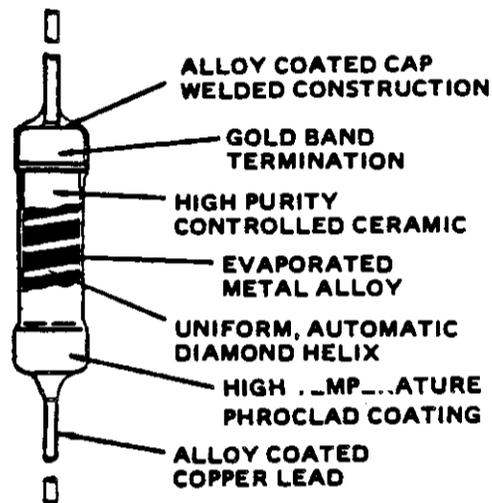


FIGURE 8. Typical construction of an axial lead type.

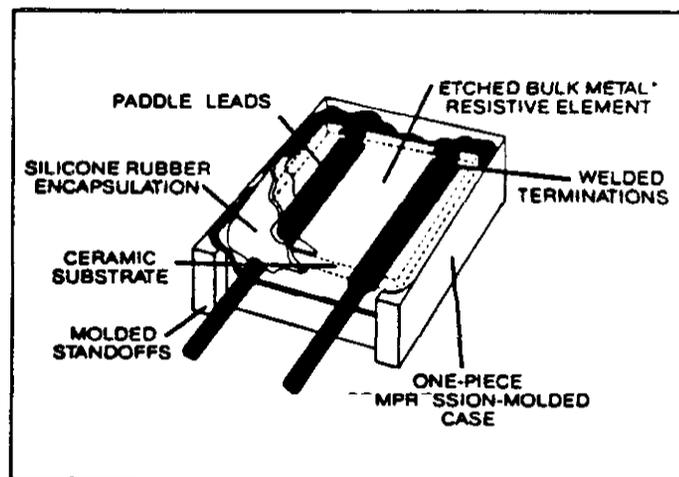


FIGURE 9. Typical construction of an RNC90 resistor.

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(HIGH STABILITY)**

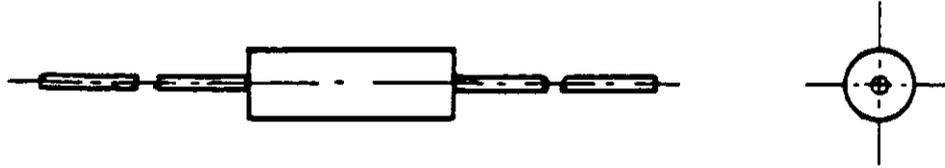


FIGURE 10. Outline drawing of styles RNR50 through RNR75 film resistors.

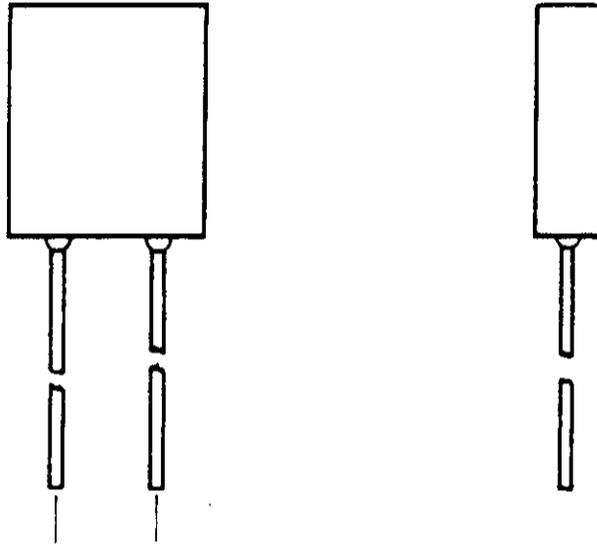


FIGURE 11. Outline drawing of a style RNC90 film resistor.

3.3.4 Military designation. Examples of the military type designation are shown below.

<u>RNR60</u>	C	<u>1003</u>	F	M
Style and terminal type	Characteristic	Resistance	Resistance tolerance	Life failure rate
 <u>RNC90</u>	 Y	 <u>162R00</u>	 B	 M
Style and terminal type	Characteristic	Resistance	Resistance tolerance	Life failure rate

3.3.5. Electrical characteristics. Electrical characteristics for established Reliability, metal film resistors are tabulated in military specification MIL-R-55182.

**3.3 RESISTORS, FIXED, FILM
(HIGH STABILITY)**

3.3.5.1 Derating. The failure rate of metal film resistors under operating conditions is a function of time, temperature, and applied power. Refer to MIL-STD-975 for specific derating conditions.

3.3.5.2 High frequency applications. When used in high frequency circuits (400 megahertz and above), the effective resistance will decrease as a result of shunt capacitance (both end-to-end and distributed capacitance to mounting surface). High frequency characteristics of metal film resistors are not controlled by specification and hence are subject to change.

3.3.5.3 Pulse applications. When metal film resistors are used in low duty cycle pulse circuits, peak voltage should not exceed 1.4 times the rated continuous working voltage (RCWV). However, if the duty cycle is high or the pulse width is appreciable, even though average power is within ratings, the instantaneous temperature rise may be excessive, requiring a resistor of higher wattage rating. Peak power dissipation should not exceed four times the maximum rating of the resistor under any conditions.

3.3.5.4 Voltage coefficient. The voltage coefficient for resistors of 1,000 ohms and above shall not exceed ± 0.005 percent per volt.

3.3.5.5 Noise. Noise output is controlled by the specification but, for metal-film resistors, noise is a negligible factor. In applications where noise is an important factor, fixed film resistors are preferable to composition types. Where noise test screening is indicated, it is recommended that the noise test procedure of MIL-STD-202, Method 308, be used for resistor screening.

3.3.6. Environmental considerations. Established reliability, metal film resistors are qualified to withstand environmental tests in accordance with Table VII of MIL-R-55182. The environmental tests are tabulated with the maximum allowable percent change in resistance value for each exposure.

Additional environmental considerations are as follows.

- a. Moisture resistance. Metal film resistors are not essentially affected by moisture except by corrosion or contamination. Coated metal film resistors may be affected if the coating is scratched or damaged and moisture is allowed to penetrate the coating.
- b. Mounting. Under conditions of severe shock or vibration (or a combination of both), resistors should be mounted in such a fashion that the body of the resistor is restrained from movement with respect to the mounting base. It should be noted that if clamps are used, certain electrical characteristics of the resistor will be altered. The heat-dissipating qualities of the resistor will be enhanced or retarded depending on whether the clamping material is a good or poor heat conductor.

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**3.3 RESISTORS, FIXED, FILM
(HIGH STABILITY)**

- c. Handling. Substrates are fragile and subject to damage during molding processes or during assembly of the resistors into circuits. Broken substrates break the film and cause open circuits.
- d. Electrostatic sensitivity. All styles, except the RNC90, are electrostatically sensitive. For tolerance B, packaging should be in accordance with MIL-R-39032.

3.3.7 Reliability considerations. Fixed metal film resistors are intended for applications requiring high stability of the resistance value during life of the circuit in which they are used. The following are additional factors in the reliability of these resistors.

3.3.7.1 Derating for optimum performance. Because all of the electrical energy dissipated by this resistor is converted into heat energy, the temperature of the surrounding air is an influencing factor when selecting a particular resistor for a specific application. The power rating of these resistors is based on operation at specific temperatures; however, in actual use, the resistor may not be operating at these temperatures. When the desired characteristic and the anticipated maximum ambient temperatures have been determined, a suitable safety factor applied to the wattage is recommended to insure the selection of a resistor having an adequate wattage-dissipation potential.

3.3.7.2 Design tolerance. Combined effects of use and environment may result in a ± 2 percent change from the "as received resistance" value. Circuits, therefore, should be designed to accept this ± 2 percent variation in resistance while continuing to operate properly.

3.3.7.3 Screening. All resistors furnished under MIL-R-55182 are subjected to conditioning through temperature cycling and overload testing.

3.3.7.4 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desired characteristics. Failure rate factors applicable to this specification are stated and discussed in MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on catastrophic failures and will differ from the failure rates established in the specification since this established failure rate is based on a parametric failure of ± 2.0 percent change in resistance from the initial measurement and any succeeding measurements up to and including 10,000 hours of life tests.

**3.4 RESISTORS, FIXED, FILM
(INSULATED)**

3.4 Fixed, film (insulated).

3.4.1 Introduction. Resistors covered in this section are established reliability, insulated film resistors having a film-type resistance element and axial leads. These resistors have resistance tolerances of ± 1.0 and ± 2.0 percent. These resistors provide life failure rates ranging from 1.0 percent to 0.001 percent per 1,000 hours. The failure rates are established at a 60-percent confidence level (initial qualification) and maintained at a 10-percent producer's risk. The failure rate is referred to operation at full rated wattage and temperature (70°C) with a maximum change in resistance of ± 4.0 percent from the initial measurement and any succeeding measurement up to and including 10,000 hours of life test.

3.4.1.1 Applicable military specification. MIL-R-39017, General Specification for Established Reliability, Fixed, Film (Insulated) Resistors.

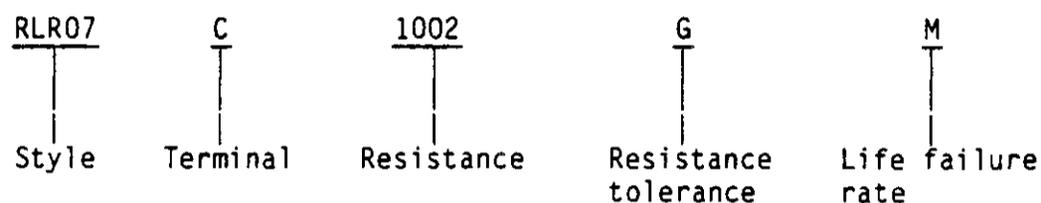
3.4.2 Usual applications. These resistor styles are used in applications requiring better stability, tolerance, and temperature coefficient requirements than carbon composition types. For applications requiring greater precision and tighter tolerances, the use of metal film or wirewound resistors is indicated.

3.4.3 Physical construction. In these resistors, the resistance element consists of a film-type resistance element (tin oxide, metal glaze, etc.). The deposition process depends on the manufacturer. The element is spiraled to achieve ranges in resistance value and, after lead attachment, the element is coated to protect it from moisture or other detrimental environmental conditions (see Figure 12). Due to the reliability requirements of MIL-R-39017, processes and controls utilized in manufacturing are necessarily stringent. Resistors furnished under MIL-R-39017 have leads conforming to type C of MIL-STD-1276. These leads are considered both solderable and weldable.

Physical dimensions. An outline drawing for RLR32 is shown in Figure 13. Refer to MIL-R-39017 for other styles.

Maximum weight. The maximum weight is 1.5 grams for the RLR32. Refer to MIL-R-39017 for other styles.

3.4.4 Military designation. An example of the military type designation is shown below.



**3.4 RESISTORS, FIXED, FILM
(INSULATED)**

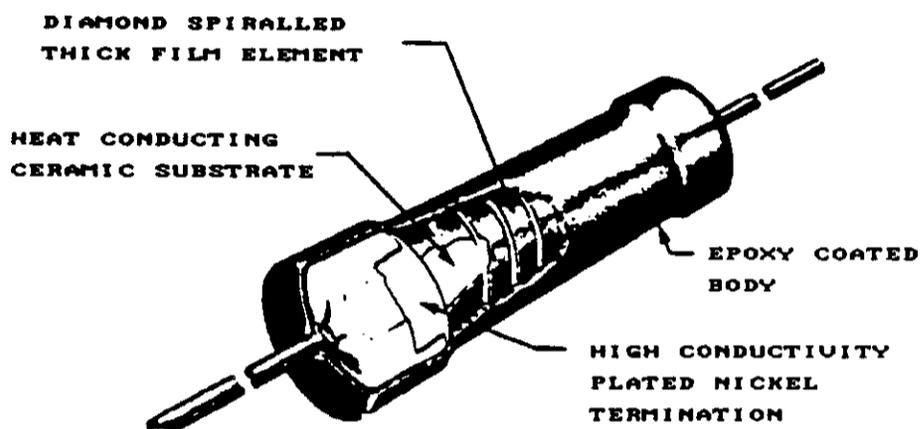


FIGURE 12. Typical construction of a metal glaze resistor.

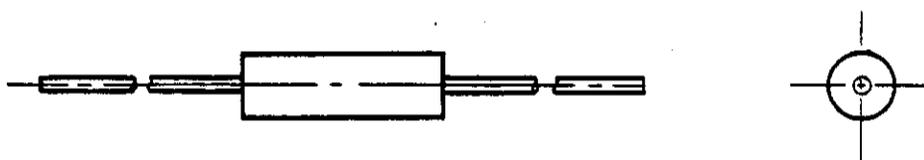


FIGURE 13. Outline drawing of a style RLR32 resistor.

3.4.5 Electrical characteristics. Electrical characteristics for established reliability, film resistors are tabulated in military specification MIL-R-39017.

Additional important electrical characteristics which must be considered in selection of the correct resistor for a particular application are as follows.

3.4.5.1 Derating. The failure rate of film resistors under operating conditions is a function of time, temperature, and applied power. Refer to MIL-STD-975 for specific derating conditions.

3.4.5.2 Maximum voltage. The maximum continuous working voltage specified for each style should in no case be exceeded regardless of the theoretically calculated rated voltage.

3.4.5.3 Noise. Noise output is uncontrolled by the specification and is considered a negligible quantity.

3.4.5.4 Frequency characteristics. These resistors are virtually noninductive. A typical response curve is illustrated in Figure 14.

3.4 RESISTORS, FIXED, FILM (INSULATED)

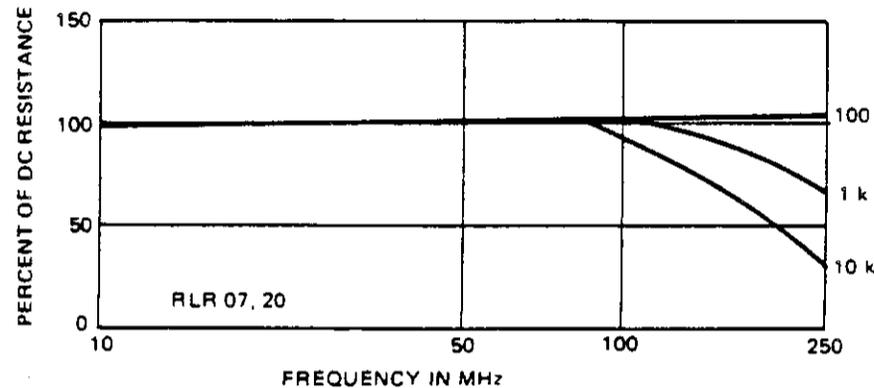


FIGURE 14. Response curve.

3.4.6 Environmental considerations. Established reliability film resistors are qualified to withstand environmental tests in accordance with Table IV of MIL-R-39017.

Additional environmental considerations are as follows.

Shelf life. MIL-R-39017 estimates a change of resistance of ± 2 percent (average) per year under normal storage conditions ($25^{\circ} \pm 10^{\circ}\text{C}$) with relative humidity not exceeding 90 percent.

3.4.7 Reliability considerations. Film resistors are designed for full power rating at ambient temperatures to $+70^{\circ}\text{C}$ and zero power rating at an ambient temperature of $+150^{\circ}\text{C}$. These resistors cost less than the fixed metal film (high stability) resistors that are in accordance with MIL-R-55182. Reliability of these resistors is excellent at ambient temperatures to $+70^{\circ}\text{C}$ when suitably derated, but they should not be considered equal to the reliability of resistors that are in accordance with MIL-R-55182 at ambient temperatures above $+70^{\circ}\text{C}$. The following are additional factors in the reliability of these resistors.

3.4.7.1 Derating for optimum performance. After the maximum ambient temperature has been determined, a suitable safety factor applied to the wattage is recommended to insure the selection of a resistor with an adequate wattage dissipation potential.

3.4.7.2 Resistance tolerance. Designers should bear in mind that operation of these resistors under the ambient conditions for which military equipment is designed may cause permanent or temporary changes in resistance sufficient to exceed their initial tolerance. In particular, operation at extreme temperatures may cause relatively large temporary changes in resistance.

3.4.7.3 Screening. All resistors furnished under MIL-R-39017 are subjected to a conditioning of $1.5 \times$ rated power for a duration of 24 hours at a test ambient temperature of 20°C to 45°C . The conditioning is followed by a total resistance check and a visual examination for evidence of arcing, burning, or charring.

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**3.4 RESISTORS, FIXED, FILM
(INSULATED)**

3.4.7.4 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desired characteristics. Failure rate factors applicable to this specification are stated in MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures" and will differ from the failure rates established in the specification since the established failure rate is based on a "parametric failure" of ± 4.0 percent change in resistance from the initial measurement to any succeeding measurement up to and including 10,000 hours of life test.

MIL-HDBK-978-B (NASA)

**3.5 RESISTORS, FIXED,
WIREWOUND (ACCURATE)**

3.5 Fixed, wirewound (accurate).

3.5.1 Introduction. Resistors covered in this section are established reliability, accurate wirewound resistors that have a maximum initial resistance tolerance of 1.0 percent and a high degree of stability with respect to time under specified environmental conditions. These resistors provide life failure rates ranging from 1.0 percent to 0.001 percent per 1,000 hours. The failure rates are established at a 60-percent confidence level (initial quantification) and maintained at a 10-percent producer's risk. The failure rate is referred to operation at full rated wattage and temperature with a maximum change in resistance of ± 0.2 percent from the initial measurement and any succeeding measurement to and including 10,000 hours of life test.

3.5.1.1 Applicable military specification. MIL-R-39005, General Specification for Established Reliability, Fixed, Wirewound (accurate) resistors.

3.5.2 Usual applications. These resistors are especially suited for use in dc amplifiers, voltmeter multipliers, electronic computers, meters, and laboratory test equipment. The resistors are not designed for high-frequency applications where ac performance is of critical performance.

3.5.3 Physical construction. In these resistors, the resistance element consists of a precisely measured (by ohmic value) length of resistance wire wound on a bobbin. The resistance wire is an alloy metal without joints, welds, or bonds (except for splicing at midpoint of a bifilar winding and at end terminals). In order to minimize inductance, resistors are wound by either reverse pi-winding or bifilar winding. The element assembly is then protected by a coating or enclosure of moisture-resistant insulating material which completely covers the exterior of the resistance element including connections and terminations (see Figure 15). Use of wire size of less than 0.001 inch in diameter is not recommended.

Physical dimensions. An outline drawing for RBR52 is shown in Figure 16A and RBR71 in Figure 16B. Refer to MIL-R-39005 for other styles.

Terminals. Weldable terminals ("U" terminals only) shall be type N-1 of MIL-STD-1276. Solderable terminals ("L" terminals only) shall meet the criteria for wire lead terminal evaluation in test method 208 of MIL-STD-202.

Maximum weight. The maximum weight for the RBR52 is 6 grams. Refer to MIL-R-39005 for other styles.

3.5.4 Military designation. An example of the military type designation is shown below.

<u>RBR52</u>	<u>L</u>	<u>12601</u>	<u>A</u>	<u>M</u>
Style	Terminal	Resistance	Resistance tolerance	Life failure rate

**3.5 RESISTORS, FIXED,
WIREWOUND (ACCURATE)**

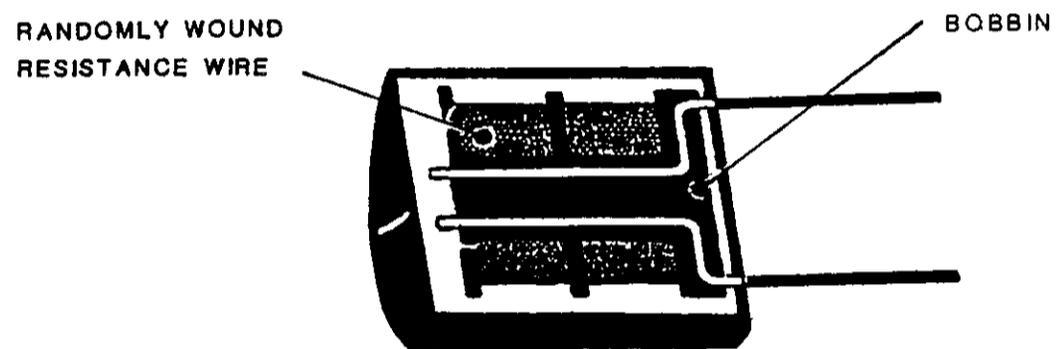
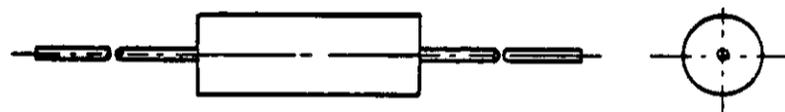
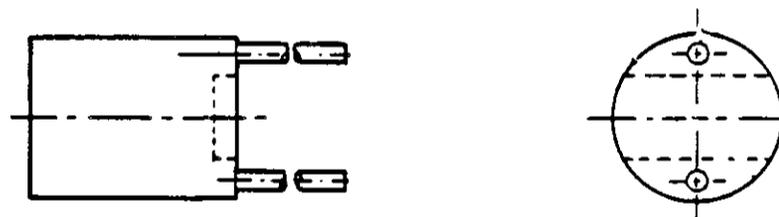


FIGURE 15. Typical construction of an accurate wirewound resistor.



A. Styles 52 through 75



B. Style 71

FIGURE 16. Outline drawing.

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**3.5 RESISTORS, FIXED,
WIREWOUND (ACCURATE)**

3.5.5 Electrical characteristics. For established reliability accurate wirewound resistors are tabulated in military specification MIL-R-39005.

Additional electrical characteristics which must be considered in selection of the correct resistor for a particular application are:

3.5.5.1 Derating. The failure rate of accurate wirewound resistors under operating conditions is a function of time, temperature, and applied power. Refer to MIL-STD-975 for specific derating conditions.

3.5.5.2 Resistance tolerance and wattage input. When using resistors with low resistance values and a tolerance of 0.1 percent or less, the design engineer must consider the fact that the resistance of the leads and other wires connected to the resistor may exceed the tolerance. Where a resistor is used in a critical application that requires the initial tolerance to be 0.1 percent or less, it is also desirable to hold resistance changes within this tolerance during operation. Since the temperature characteristic can cause the resistance to change by more than 0.1 percent, the temperature rise in the resistor must be kept to a minimum if the resistor is expected to remain within the initial tolerance during use. It is to be noted that initial nominal resistance is measured at 25°C while full-load operating temperature is 125°C. Therefore, if this close tolerance of 0.1 percent or less is to be held, the power rating of the resistors shall be reduced as indicated in Table V.

TABLE V. Resistance tolerance and wattage input

Symbol	Resistance Tolerance Percent (\pm)	Permissible Percent of Normal Wattage ^{1/}
Q	0.02	50
A	0.05	50
B	0.1	50
F	1.0	100

^{1/} These values represent the maximum wattage at which resistors should be operated at an ambient temperature up to 125°C.

3.5.6 Environmental considerations. Established reliability accurate wirewound resistors are qualified to withstand environmental tests in accordance with Table VII of MIL-R-39005C.

Additional environmental considerations are as follows.

Coating materials. These resistors are encased in nonmetallic materials. The possibility of "outgassing" at low pressures should be considered in their application.

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**3.5 RESISTORS, FIXED,
WIREWOUND (ACCURATE)**

Supplementary insulation. Where high voltages (250 volts and higher) are present between the resistor circuit and the grounded surface on which the resistor is mounted, or where resistance is so high that the insulation resistance to ground is an important factor, secondary insulation between the resistor and its mounting, or between mounting and ground, should be provided.

Soldering. Care must be exercised in soldering these resistors, particularly in the lower resistance values and tighter tolerances, since high contact resistance might cause resistance changes greater than the tolerance.

Mounting. It is suggested that resistors be mounted by restraining their bodies from movement when shock or high-frequency-vibration forces are to be encountered.

Recommended maximum ambient temperature. The maximum ambient temperature should not exceed 135°C for all styles.

3.5.7 Reliability considerations.

3.5.7.1 Derating for optimum performance. Because all of the electrical energy dissipated by a resistor is converted into heat energy, the temperature of the surrounding air becomes an influencing factor in the selection of a particular resistor for use in a specific application. After the desired resistance tolerance and the anticipated maximum ambient temperature have been determined, a suitable safety factor applied to the wattage is recommended to insure the selection of a resistor having an adequate wattage-dissipation potential, and one which will remain within specified tolerance limits.

3.5.7.2 Screening requirements. All resistors furnished under MIL-R-39005 are subjected to a 100-hour conditioning life test by cycling at rated wattage at 125°C followed by a total resistance measurement check and a visual examination for evidence of mechanical damage.

3.5.7.3 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desired characteristics. Failure rate factors applicable to this specification are stated and discussed in MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures" and will differ from the failure rates established in the specification, since the established failure rate is based on a "parametric failure" of ± 0.2 percent change in resistance from the initial measurement to any succeeding measurement to and including 10,000 hours of life test.

**3.6 RESISTORS, FIXED,
WIREFOUND (POWER TYPE)**

3.6 Fixed, wirewound (power type).

3.6.1 Introduction. Resistors covered in this section are established reliability power wirewound fixed resistors having axial leads. These resistors have a maximum initial resistance tolerance of ± 1.0 percent. These resistors provide failure rates ranging from 1.0 percent to 0.001 percent per 1,000 hours. The failure rates are established at a 60-percent confidence level (initial qualification) and maintained at a 10-percent manufacturer's risk. The failure rate is referred to operation at full rated wattage and temperature with a maximum change in resistance of ± 1.0 percent from the initial measurement to any succeeding measurement up to and including 10,000 hours of life test.

3.6.1.1 Applicable military specification. MIL-R-39007, General Specification for Established Reliability, Fixed, Wirewound (Power-Type) Resistors.

3.6.2 Usual applications. Power wirewound resistors are used where power dissipation values range from 1 to 10 watts and good permanent stability is required. They may be used in power attenuators, bridges, voltage dividers, bleeders in dc power supplies and filter networks where their poor ac characteristics will not adversely affect the circuit performance. They have the added advantage of low resistance range.

The distributed inductance and capacitance of the power wirewound resistor become increasingly important at higher frequencies. As the frequency is increased, the inductive reactance will increase and the capacitive reactance will decrease.

3.6.3 Physical construction. These resistors are constructed of a measured length of resistance wire or ribbon (of a known ohmic value) wound in a precise manner (pitch, effective wire coverage, and wire diameter are specification controlled). The continuous length of wire (wire is required to be free of joints and bonds and of uniform cross-section) is wound on a ceramic core or tube and attached to end terminations. The element is then coated or enclosed by inorganic vitreous or a silicon coating to protect it from moisture or other detrimental environmental conditions (see Figure 17). Due to the reliability requirements of MIL-R-39007, processes and controls utilized in manufacturing are necessarily stringent. Resistors may have an added requirement for non-inductive winding. Resistors which are identified by the terminal and winding designator "N" are noninductively wound using the Ayrton-Perry method.

Physical dimensions. An outline drawing for RWR78 is shown in Figure 18. Maximum weight of each style is shown in Table VI.

**3.6 RESISTORS, FIXED,
WIREWOUND (POWER TYPE)**

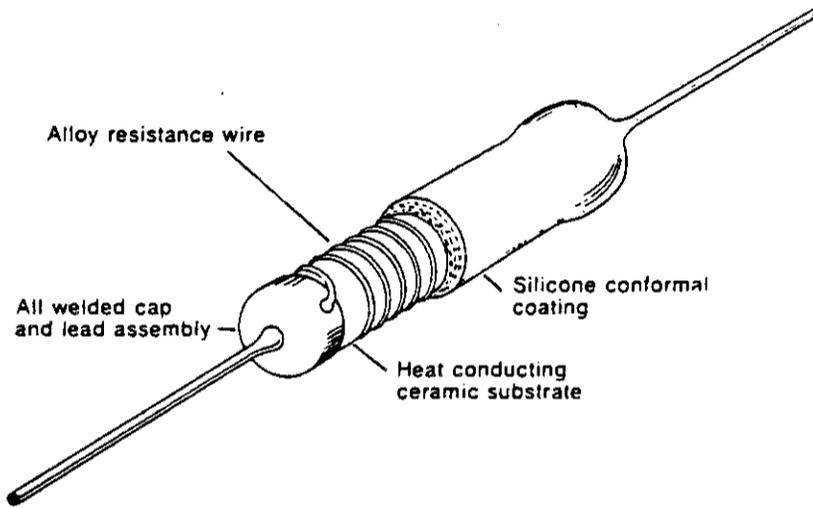


FIGURE 17. Typical construction of a power wirewound resistor.

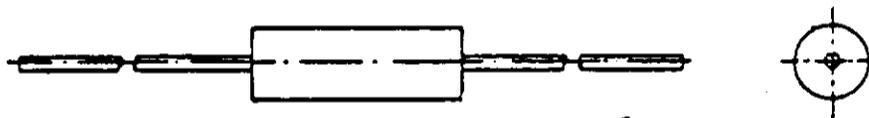


FIGURE 18. Outline drawing of a style RWR78 power wirewound resistor.

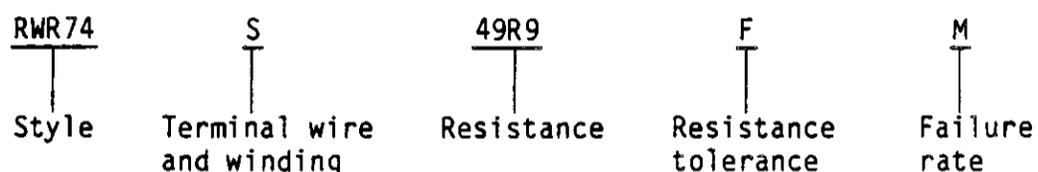
TABLE VI. Maximum weight of each style

	S and W Terminal and winding	N Terminal and winding
RWR74	5 grams	6 grams
RWR78	12 grams	13 grams
RWR80	1 gram	1 gram
RWR81	0.35 gram	0.70 gram
RWR84	5 grams	6 grams
RWR89	3 grams	4 grams

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**3.6 RESISTORS, FIXED,
WIREWOUND (POWER TYPE)**

3.6.4 Military designation. An example of the military type designation is shown below.



3.6.5 Electrical characteristics. Electrical characteristics for established reliability power wirewound resistors are tabulated in military specification MIL-R-39007.

Additional electrical characteristics which must be considered in selection of the correct resistor for a particular application are as follows.

3.6.5.1 Derating. The failure rate of power wirewound resistors under operating conditions is a function of time, temperature, and applied power. Refer to MIL-STD-975 for specific derating conditions.

3.6.6 Environmental considerations. Established reliability power wirewound resistors are qualified to withstand environmental tests in accordance with Table IX of MIL-R-39007. The environmental tests are tabulated with the maximum allowable percent change in resistance value for each exposure.

Additional environmental considerations are as follows.

Coating materials. Certain coating materials used in fabricating resistors furnished under MIL-R-39007 may be subject to "outgassing" of volatile material when operated at surface temperatures over 200°C or at low pressures. This phenomena should be taken into consideration for equipment design.

Spacing. When resistors are mounted in rows or banks, they should be spaced so that restricted ventilation and heat dissipation by nearby resistors do not cause temperatures in excess of the maximum permissible hot-spot temperature. An appropriate combination of resistor spacing and resistor power rating must be chosen if this is to be insured.

Soldering. A solder with a minimum melting temperature of 350°C should be used for soldering. Care must be exercised in soldering low value and tight tolerance resistors since high contact resistance may cause resistance changes exceeding the tolerance.

Mounting. Under conditions of severe shock or vibration, or a combination of both, resistors of all sizes described in this section should be mounted in such a fashion that the body of the resistor is restrained from movement with respect to the mounting base. It should be noted that if clamps are used,

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**3.6 RESISTORS, FIXED,
WIREWOUND (POWER TYPE)**

certain electrical characteristics of the resistor will be altered. The heat-dissipating qualities of the resistor will be enhanced or retarded depending on whether the clamping material is a good or poor heat conductor. Under less severe vibration conditions, axial lead styles may be supported by their leads only. The lead lengths should be kept as short as possible, 1/4 inch or less is preferred, but should be no longer than 5/8 inch. The longer the lead, the more likely that a mechanical failure will occur.

Secondary insulation. Where high voltages are present between resistor circuits and grounded surfaces on which resistors are mounted, secondary insulation capable of withstanding the voltage conditions should be provided between resistors and mountings or between mountings and ground.

3.6.7 Reliability considerations.

3.6.7.1 Derating. Because all of the electrical energy dissipated by a resistor is converted into heat energy, the temperature of the surrounding air becomes an influencing factor in the selection of a particular resistor for use in a specific application. The power rating for these resistors is based on operation at an ambient temperature of 25°C. However, in actual use, the resistors may not be operating at that temperature. After the desired resistance tolerance and the anticipated maximum ambient temperature have been determined a suitable safety factor applied to the wattage is recommended in order to insure the selection of a resistor. This resistor has an adequate wattage-dissipation potential and one which will remain within specified tolerance limits. Refer to MIL-STD-975 for specific derating conditions.

3.6.7.2 Screening. All resistors furnished under MIL-R-39007 are subjected to a conditioning 100-hour life test by cycling at rated continuous working voltage at 25°C and dissipating a wattage equal to the power rating (free air) of the resistor. The conditioning is followed by a total resistance measurement and a visual examination for evidence of mechanical damage.

3.6.7.3 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desired characteristics. Failure rate factors applicable to this specification are stated and discussed in MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures" and will differ from the failure rates established in the specification, since the established failure rate is based on a "parametric failure" of ± 1.0 percent change in resistance from the initial measurement to any succeeding measurement up to including 10,000 hours of life test.

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**3.7 RESISTORS, FIXED, WIREWOUND
(POWER TYPE, CHASSIS MOUNTED)**

3.7 Fixed, wirewound (power type, chassis mounted).

3.7.1. Introduction. Resistors covered in this section are established reliability chassis-mounted power wirewound resistors, having a wirewound resistance element and axial lug-type leads. These resistors utilize the principle of heat dissipation through a metal mounting surface with full rated wattage at 25°C. The initial resistance tolerance is ± 1.0 percent. These resistors provide life failure rates ranging from 1.0 percent to 0.001 percent per 1,000 hours. The failure rates are established at a 60-percent confidence level (initial qualification) and maintained at a 10-percent producer's risk. The failure rate is referred to operation at full rated wattage and temperature with a maximum change in resistance of ± 2.0 percent from the initial measurement to any succeeding measurement up to and including 10,000 hours of life test.

3.7.1.1 Applicable military specification. MIL-R-39009, General Specification for Established Reliability, Fixed, Wirewound (Power Type, Chassis Mounted) Resistors.

3.7.2. Usual applications. Chassis mounted power resistors are used in the same type of electrical applications as the axial lead, wirewound power resistors. The metal case serves as a heat sink and when suitably mounted on a metal chassis will dissipate from 5 to 30 watts for the standard sizes.

These resistors should not be used in circuits where their ac performance is of critical importance. However, provisions have been made in particular styles to minimize inductance.

3.7.3 Physical construction. These resistors are constructed of a measured length of resistance wire or ribbon (of a known ohmic value) wound in a precise manner (pitch, effective wire coverage, and wire diameter are specification controlled). Series RER45, 50, and 55 have Ayrton-Perry, or bifilar windings to reduce inductive effect. The continuous length of wire (wire required to be free of joints or bonds, and of uniform cross-section) is wound on a ceramic core or tube and attached to end terminations. The finished resistor element and termination caps are sealed by a coating material. The coated element is then inserted in a finned aluminum alloy housing which completes the sealing of the element from detrimental environments, and provides a radiator and a heat sink for heat dissipation (see Figure 19). Due to reliability requirements of MIL-R-39009, processes and controls utilized in manufacturing are stringent.

Dimensions. An outline drawing of RER75 is shown in Figure 20. Refer to MIL-R-39009 for other styles.

Maximum weight. The maximum weight for the RER75 is 32 grams. Refer to MIL-R-39009 for other styles.

**3.7 RESISTORS, FIXED, WIREWOUND
(POWER TYPE, CHASSIS MOUNTED)**

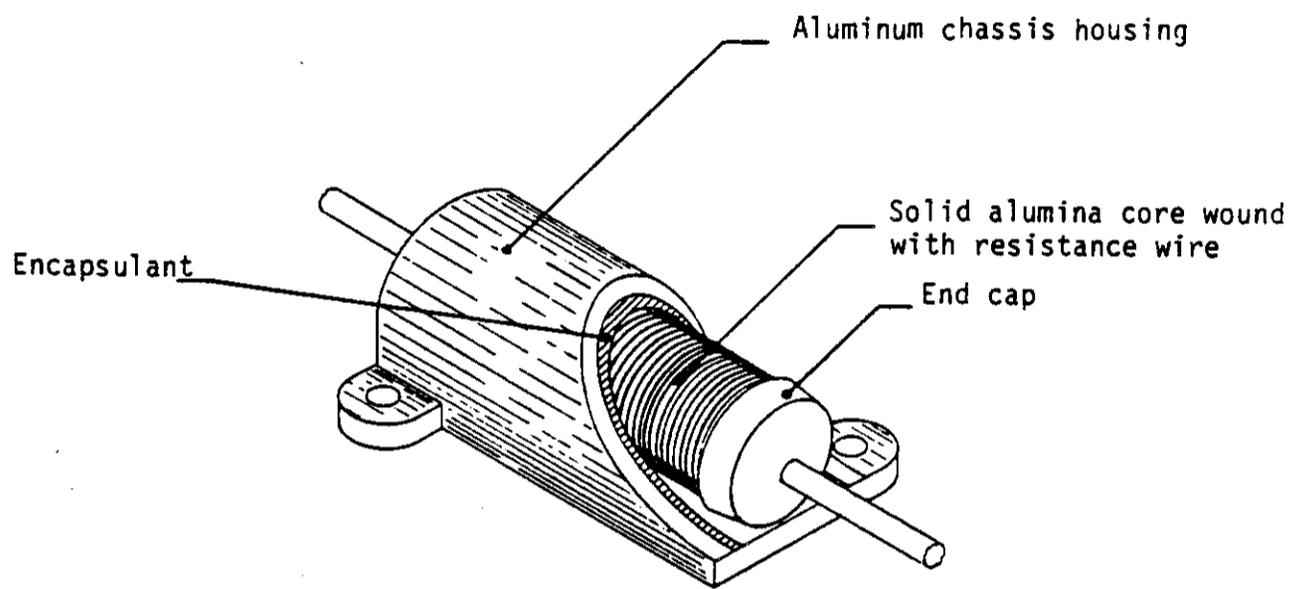


FIGURE 19. Typical construction of a wirewound (chassis mounted) resistor.

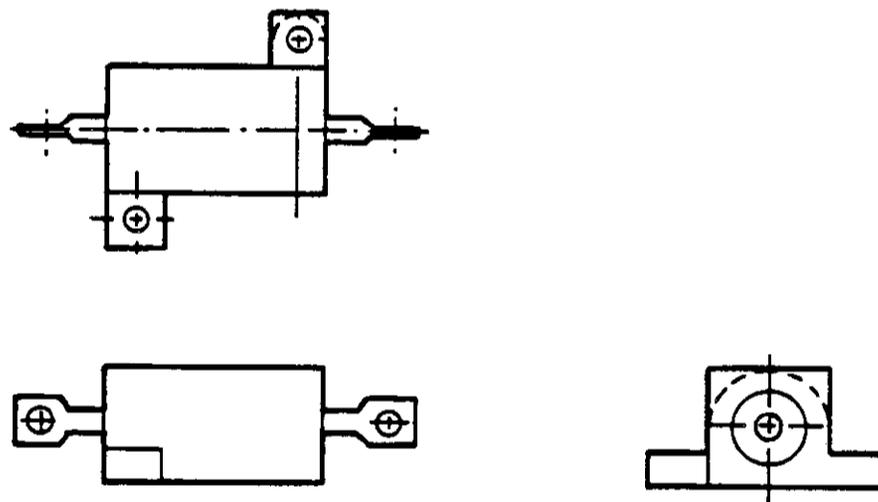
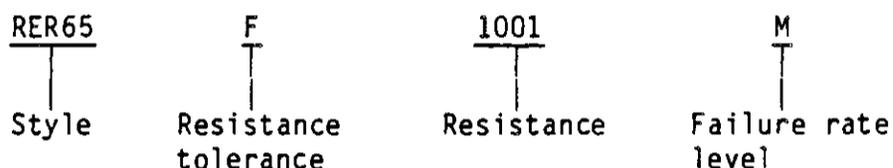


FIGURE 20. Outline drawing of a Style RER75 resistor.

3.7 RESISTORS, FIXED, WIREWOUND (POWER TYPE, CHASSIS MOUNTED)

3.7.4 Military designation. An example of the military type designation is shown below.



3.7.5 Electrical characteristics. Electrical characteristics for established reliability, fixed, wirewound power type, chassis mounted resistors are tabulated in military specification MIL-R-39009.

Additional electrical characteristics which must be considered in selection of the correct resistor for a particular application are as follows.

3.7.5.1 Derating. The failure rate of chassis mounted power wirewound resistors under operating conditions is a function of time, temperature, and applied power. Refer to MIL-STD-975 for specific derating conditions.

3.7.6 Environmental considerations. Established reliability chassis mounted power wirewound resistors are qualified to withstand environmental tests in accordance with Table IV of MIL-R-39009. The environmental tests are tabulated along with the maximum allowable percent change in resistance value for each exposure.

Spacing. When resistors are mounted in rows or banks, they should be spaced so that the restricted ventilation and heat dissipation by nearby resistors do not cause temperatures in excess of the maximum permissible continuous operating temperature. An appropriate combination of resistor spacing and resistor power rating must be chosen if this is to be assumed. In view of the chassis heat dissipation principle of these resistors, particular care must be exercised in order that the chassis temperature rise does not damage nearby components.

Soldering. A solder with a minimum melting temperature of 300°C should be used in soldering.

3.7.7 Reliability considerations.

3.7.7.1 Derating. When the chassis area and the anticipated maximum ambient temperature have been determined, a suitable factor applied to the wattage is recommended in order to insure the selection of a resistor having an adequate wattage-dissipation potential. Refer to MIL-STD-975 for specific derating conditions.

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**3.7 RESISTORS, FIXED, WIREWOUND
(POWER TYPE, CHASSIS MOUNTED)**

3.7.7.2 Screening. All resistors furnished under MIL-R-39009 are subjected to a conditioning 100-hour life test by cycling at rated continuous working voltage at 25°C and dissipating a wattage equal to the power rating (free air) of the resistor. The conditioning is followed by a total resistance measurement and a visual examination for evidence of mechanical damage.

3.7.7.3 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from initial characteristics occurring in an unpredictable manner and in too short a period of time to permit detection through normal preventive maintenance. Failure rate factors applicable to this specification are stated and discussed in MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures" and will differ from the failure rates established in the specification, since the established failure rate is based on a "parametric failure" of ± 2.0 percent change in resistance from the initial measurement to any succeeding measurement up to and including 10,000 hours of life test.

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**3.8 RESISTORS, VARIABLE, NONWIREWOUND
(ADJUSTMENT TYPE)**

3.8 Variable, nonwirewound (adjustment type).

3.8.1 Introduction. Resistors covered in this section are established reliability nonwirewound variable resistors with a contact which bears uniformly over the surface of a nonwirewound resistive element when positioned by a multi-turn lead-screw actuator. These resistors are capable of full-load operation (when maximum resistance is engaged) at a maximum ambient temperature of 85°C and are suitable for continuous operation, when properly derated, at a maximum temperature of 150°C. The resistance tolerance of these resistors is ±10 percent. These resistors possess life failure rate levels ranging from 1.0 to 0.001 percent per 1,000 hours. The failure rates are established at a 60-percent confidence level and maintained at 10-percent producer's risk on the basis of life tests. The failure rate level refers to operation at full rated voltage at 85°C, with a permissible change in resistance of ±10 percent as criteria for failure.

3.8.1.1 Applicable military specification. MIL-R-39035, General Specification for Established Reliability, Variable, Nonwirewound (Adjustment Type) Resistors.

3.8.2 Usual applications. Nonwirewound variable resistors are primarily used as trimmers for setting biases and do not dissipate much power. Because of the method by which these resistors are made, they are used in printed circuit boards and logic circuits to set voltage levels for transistors and adjust time constants of RC networks. Nonwirewound trimmers are used in applications requiring higher resistance values than available from wirewound trimmers.

3.8.3 Physical construction. These resistors have an element of continuous resistive materials (cermet, metal film, etc.) on a rectangular or arc-shaped core, depending upon the style. The sliding contact traverses the element in a circular or straight line. The element is protected from detrimental environmental conditions by a housing or enclosure. The lead screw head is insulated from the electrical portion of the resistor. Due to the reliability requirements of MIL-R-39035, processes and controls utilized in manufacturing are stringent.

Physical dimensions. Outline drawings for each style are shown in Figures 21 through 26.

Terminals. Terminal types D, P, W, X and Y are solderable not weldable. If weldable leads are required, they must be separately specified in the contract or order.

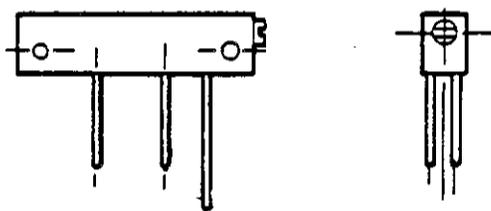
3.8.4 Military designation. An example of the military type designation is shown below.

RJR24	C	L	102	M
Style	Characteristic	Terminal	Resistance	Failure rate

**3.8 RESISTORS, VARIABLE, NONWIREWOUND
(ADJUSTMENT TYPE)**



A. Flexible lead terminal type L



B. Printed-circuit pin type Y

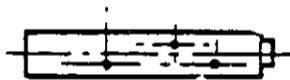
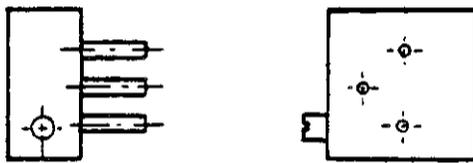


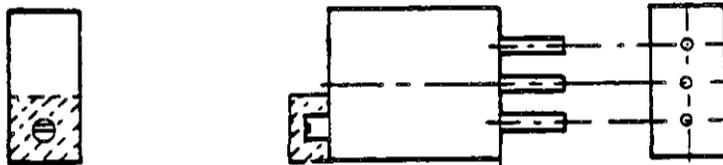
FIGURE 21. Outline drawing of a style RJR12 nonwirewound variable resistor.

MIL-HDBK-978-B (NASA)

**3.8 RESISTORS, VARIABLE, NONWIREWOUND
(ADJUSTMENT TYPE)**



A. Terminal type P



B. Terminal type W

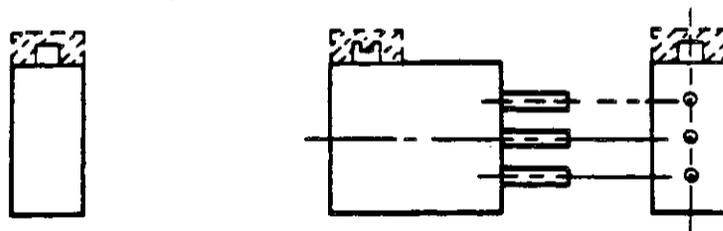
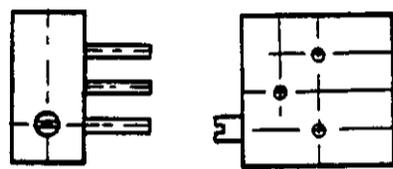
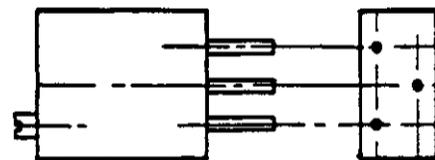
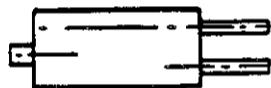


FIGURE 22. Outline drawing of a style RJR24 nonwirewound variable resistor.

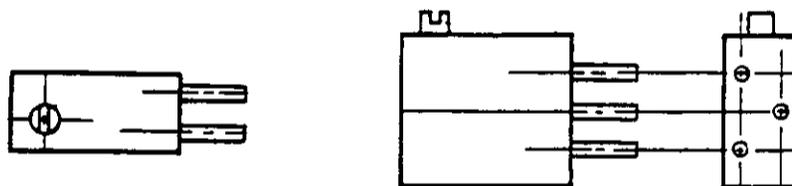
**3.8 RESISTORS, VARIABLE, NONWIREWOUND
(ADJUSTMENT TYPE)**



A. Terminal type P



B. Terminal type W



C. Terminal type X

FIGURE 23. Outline drawing of a style RJR26 nonwirewound variable resistor.

MIL-HDBK-978-B (NASA)

**3.8 RESISTORS, VARIABLE, NONWIREWOUND
(ADJUSTMENT TYPE)**

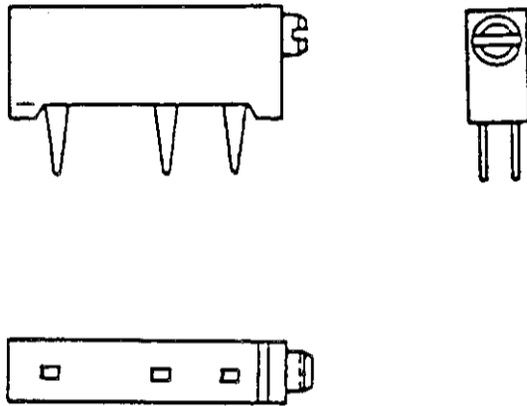


FIGURE 24. Outline drawing of a style RJR 28 nonwirewound variable resistor.

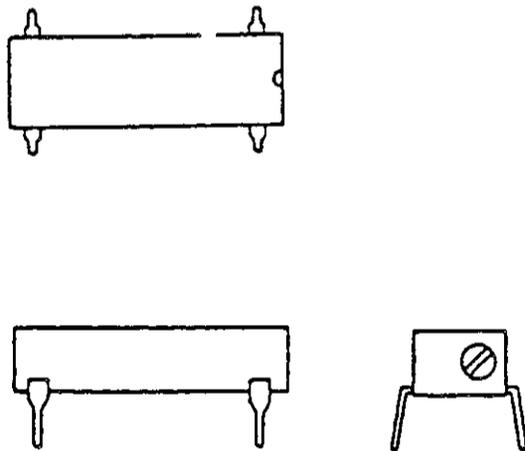


FIGURE 25. Outline drawing of a style RJR32 nonwirewound variable resistor.

**3.8 RESISTORS, VARIABLE, NONWIREWOUND
(ADJUSTMENT TYPE)**

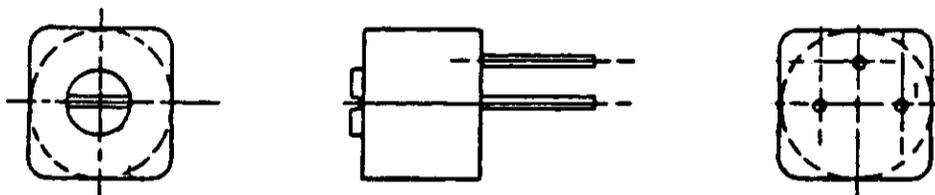


FIGURE 26. Outline drawing of a style RJR50 nonwirewound variable resistor.

3.8.5 Electrical characteristics. Electrical characteristics for established reliability, nonwirewound variable resistors are tabulated in military specification MIL-R-39035.

Other electrical characteristics which must be considered in selection of the correct resistor for a particular application are as follows.

3.8.5.1 Resistance-temperature characteristic. Consideration should be given to resistor temperature during operation to allow for the change in resistance due to the resistance-temperature characteristic. The resistor-temperature characteristic is measured between the two end terminals. When the resistance-temperature characteristic is critical, variation due to the resistance of the movable contact should be considered.

3.8.5.2 Contact-resistance variation. The contact resistance variation should not exceed ± 3 percent or 20 ohms whichever is greater for characteristic C, and ± 3 percent or 3 ohms whichever is greater for characteristics F and H.

3.8.6 Environmental consideration. Established reliability nonwirewound variable resistors are qualified to withstand environmental tests in accordance with Table IV of MIL-R-39035.

Additional environmental considerations are:

Mounting of resistors. Resistors with terminal type L should not be mounted by their flexible wire leads. Mounting hardware should be used. Printed-circuit types are frequently terminal mounted although brackets may be necessary for a high-shock and vibration environment.

Stacking of resistors. When stacking resistors, care should be taken to compensate for the rise in temperature by derating the power rating accordingly.

Selection of a safe resistor style. The wattage ratings of these resistors are based on operation at 85°C when mounted on a 1/16-inch thick, glass base, epoxy laminate. Therefore, the heat sink effect as provided by steel test plates in other specifications is not present. The wattage rating is applicable when the entire resistance element is engaged in the circuit. When only a portion is engaged, the wattage is reduced in the same proportion as the resistance.

MIL-HDBK-978-B (NASA)

**3.8 RESISTORS, VARIABLE, NONWIREWOUND
(ADJUSTMENT TYPE)**

3.8.7 Reliability considerations.

3.8.7.1 Derating. After the anticipated maximum ambient temperature has been determined, an appropriate safety factor applied to the wattage is recommended to insure the selection of a resistor style having an adequate wattage rating with optimum performance. Refer to MIL-STD-975 for specific derating conditions.

3.8.7.2 High resistances and voltages. Where voltages higher than 250 volts rms are present between the resistor circuit and grounded surface on which the resistor is mounted, or where the dc resistance is so high that the insulation resistance to ground is an important factor, secondary insulation to withstand the conditions should be provided between the resistor and mounting or between the mounting and ground.

3.8.7.3 Screening. All resistors furnished under MIL-R-39035 are subjected to a 50-hour conditioning life test by cycling at 3/4 watt at 25°C followed by contact resistance variation and total resistance measurements and a seal test for detection of leaks.

3.8.7.4 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desired characteristics. Failure rate factors applicable to this specification are stated and discussed in MIL-HDBK-217 (see MIL-R-22097 data). The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures" and will differ from the failure rates established in the specification, since the established failure rate is based on a parametric failure of ± 5 percent change in resistance encountered during the 10,000 hours life test.

**3.9 RESISTORS, VARIABLE WIREWOUND
(LEAD SCREW ACTUATED)**

3.9 Variable, wirewound (lead screw actuated).

3.9.1 Introduction. Resistors covered in this section are established reliability wirewound variable resistors having a contact which can be positioned by a multiturn lead-screw actuator over the surface of a linearly-wound resistive element. These resistors are capable of full-load operation (when maximum resistance is engaged) at a maximum ambient temperature of 85°C and are suitable for continuous operation, when properly derated, at a maximum temperature of 150°C. The resistance tolerance of these resistors is ±5.0 percent. These resistors possess life failure rate levels ranging from 1.0 to 0.001 percent per 1,000 hours. The failure rates are established at a 60-percent confidence level and maintained at 10-percent producer's risk on the basis of life tests. The failure rate level refers to operation at full rated voltage at 85°C with a permissible change in resistance of ±3.0 percent plus the specified resolution as the criteria for failure.

3.9.1.1 Applicable military specification. MIL-R-39015, General Specification for Established Reliability, Variable, Wirewound (Lead Screw Actuated) Resistors.

3.9.2 Usual applications. Wirewound variable resistors are primarily used as trimmers for setting biases and have low power dissipation characteristics. Because of the method by which these resistors are made, they are used in printed circuit boards and logic circuits to set voltage levels for transistors and adjust time constants of RC networks.

3.9.3 Physical construction. These resistors have an element of continuous-length wire, wound linearly on a rectangular or arc-shaped core, depending upon the style. The sliding contact traverses the element in a circular or straight line. The element is protected from detrimental environmental conditions by a housing or enclosure. The lead screw head is insulated from the electrical portion of the resistor. Due to the reliability requirements of MIL-R-39015, processes and controls utilized in manufacturing are necessarily stringent.

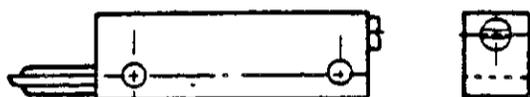
Physical Dimensions. Outline drawings for each style are shown in Figures 27 through 29.

Resistive element wire size. Use of wire sizes of less than 0.001 inch diameter is not recommended for new designs.

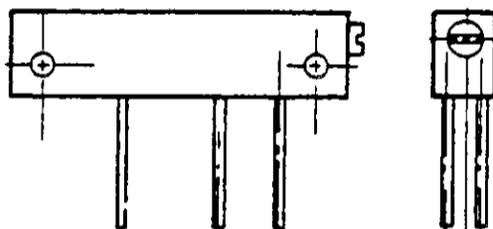
Terminals. Terminal types P, W, X, and Y are solderable not weldable. If weldable leads are required, they must be separately specified.

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**3.9 RESISTORS, VARIABLE WIREWOUND
(LEAD SCREW ACTUATED)**



A. Flexible lead terminal type L

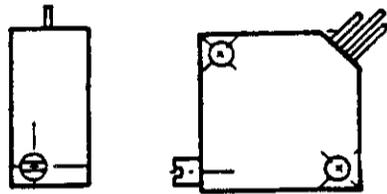


B. Printed Circuit pin type Y

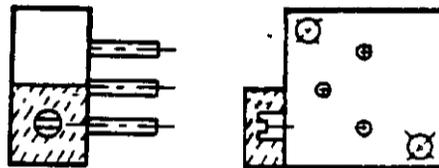


FIGURE 27. Outline drawing of a style RTR12 wirewound variable resistor.

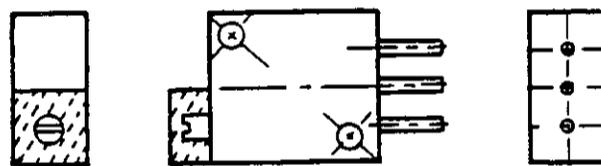
**3.9 RESISTORS, VARIABLE WIREWOUND
(LEAD SCREW ACTUATED)**



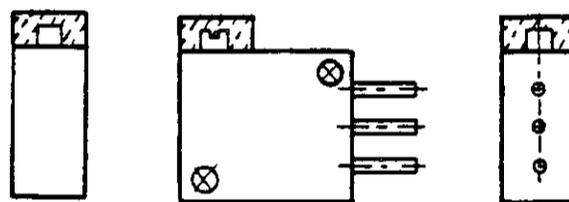
A. Flexible lead terminal type L



B. Terminal type P



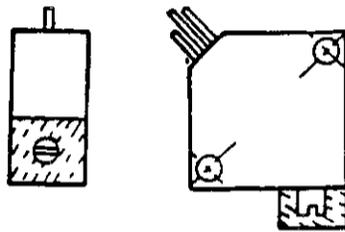
C. Terminal type W



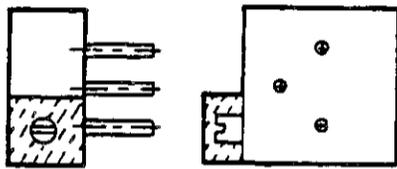
D. Terminal type X

FIGURE 28. Outline drawing of a style RTR22 wirewound variable resistor.

**3.9 RESISTORS, VARIABLE WIREWOUND
(LEAD SCREW ACTUATED)**



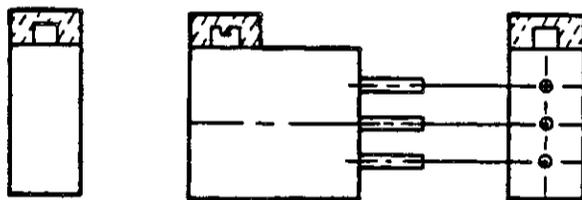
A. Terminal type L



B. Terminal type P



C. Terminal type W



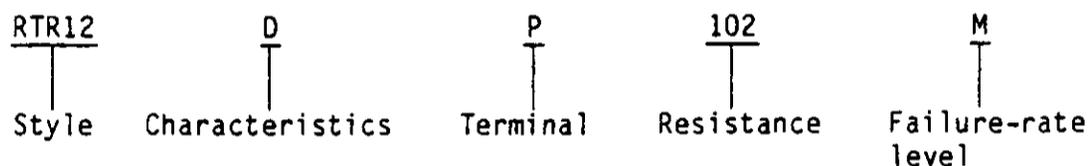
D. Terminal type X

FIGURE 29. Outline drawing of a style RTR24 wirewound variable resistor.

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**3.9 RESISTORS, VARIABLE WIREWOUND
(LEAD SCREW ACTUATED)**

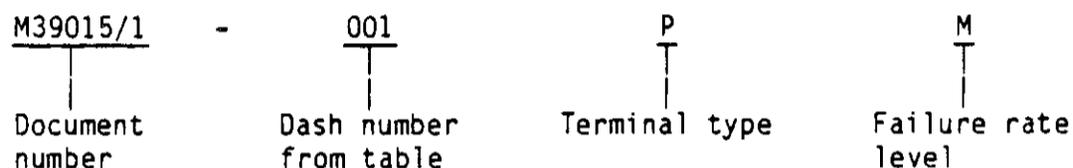
3.9.4 Military designation. An example of the military type designation is shown below.



The ordering reference for these resistors is a part number as described in MIL-R-39015. The preferred nominal total resistances values are specified in MIL-R-39015.

The part number consists of the number on this specification sheet and a dash number with letters which signify terminal and failure rate level.

Example:



3.9.5 Electrical characteristics. Electrical characteristics for established reliability wirewound variable resistors are tabulated in military specification MIL-R-39015.

Additional electrical characteristics which must be considered in selection of the correct resistor for a particular application are as follows.

3.9.5.1 Resistance-temperature characteristic. Consideration should be given to resistor temperature during operation to allow for the change in resistance due to the resistance-temperature characteristic. The resistance temperature characteristic is measured between the two end terminals. When the resistance-temperature characteristic is critical, variation due to the resistance of the movable contact should be considered.

3.9.5.2 Noise. The noise level is low compared to nonwirewound types. Peak noise is specification-controlled at an initial value of 100 ohms maximum. However, after exposure to environmental tests a degradation to 500 ohms is allowed by specification.

3.9.6 Environmental considerations. Established reliability wirewound variable resistors are qualified to withstand environmental tests in accordance with Table IV of MIL-R-39015.

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**3.9 RESISTORS, VARIABLE WIREWOUND
(LEAD SCREW ACTUATED)**

Additional environmental considerations are as follows.

Mounting of resistors. Resistors with terminal type L should not be mounted by their flexible wire leads. Mounting hardware should be used. Printed-circuit types are frequently terminal-mounted although brackets may be necessary for a high-shock and vibration environment.

Stacking of resistors. When stacking resistors, care should be taken to compensate for the added rise in temperature by derating the wattage rating accordingly.

Selection of a safe resistor style. The wattage ratings of these resistors are based on operation at 85°C when mounted on a 1/16-inch thick, glass base, epoxy laminate. Therefore the heat sink effect as provided by steel test plates in other specifications is not present. The wattage rating is applicable when the entire resistance element is engaged in the circuit. When only a portion is engaged, the wattage is reduced in the same proportion as the resistance.

High resistances and voltages. Where voltages higher than 250 volts rms are present between the resistor circuit and grounded surface on which the resistor is mounted or where the dc resistance is so high that the insulation resistance to ground is an important factor, secondary insulation to withstand the conditions should be provided between the resistor and mounting or between the mounting and ground.

3.9.7 Reliability considerations.

3.9.7.1 Derating. After the anticipated maximum ambient temperature has been determined, an appropriate safety factor applied to the wattage is recommended to insure the selection of a resistor style having an adequate wattage rating with optimum performance. Refer to MIL-STD-975 for specific derating conditions.

3.9.7.2 Screening. All resistors furnished under MIL-R-39015 are subjected to a 50-hour conditioning life test by cycling at 1 watt at 25°C, followed by peak noise and total resistance measurements, and a seal test for detection of leaks.

3.9.7.3 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desired characteristics. Failure rate factors applicable to this specification are stated and discussed in section 1.4 of MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures" and will differ from the failure rates established in the specification, since the established failure rate is based on a "parametric failure" of ± 3 percent change in resistance encountered during the 10,000 hours life test.

3.9.7.4 Failure Mechanisms. Although variable wirewound resistors are constructed in basically the same way as the fixed wirewound with respect to the resistance elements, they are not as reliable because of potential mechanical

**3.9 RESISTORS, VARIABLE WIREWOUND
(LEAD SCREW ACTUATED)**

problems associated with the wiper arm assemblies. These weaknesses fall into two categories: (1) the wirewound resistance element, and (2) the wiper assembly and enclosure.

Imperfections in the resistance wire such as reduced cross sectional area will cause hot spots to develop which may ultimately burn open. The same reduced cross section will mechanically weaken the wire which may result in a wire break. Defects also occur in the termination of the resistance element, where the resistance wires are attached to the terminal leads.

Many imperfections are inherent in a variable resistor. Again, cross sectional area is an important consideration. As the slider contact traverses the resistance element, it will cause some wear and as the cross section of the wire decreases, the resistance of the element increases. For instance, if the wiper brushes against these wires, and causes a nick of 0.00001 inch, it is more significant on a wire with a diameter of 0.00005 inch than a wire of 0.001 inch diameter. Therefore, wire diameters of less than 0.001 inch are considered reliability hazards.

There are many mechanical defects associated with the slider and screw assemblies. The defects generally result in jamming of the lead screw assembly, stripping of the threads, or improper contact of the wiper with the resistance element.

Cleanliness is important since foreign material on the resistance element may cause an open when the wiper rides over them. Faulty end stops and clutch mechanisms result in opens at the end of the travel or failure to adjust resistance value.

3.10 RESISTORS, FIXED, FILM, NETWORKS

3.10 Fixed, film, networks.

3.10.1. Introduction. This section covers fixed resistors in a resistor network configuration having a film resistance element in a dual-in-line or flat pack packages. These resistors are stable with respect to time, temperature, and humidity and are capable of full load operation at an ambient temperature of 70°C.

3.10.1.1 Applicable military specification. MIL-R-83401, General Specification for Fixed, Film, Resistor Networks.

3.10.2. Usual applications. These resistor networks are designed for use in critical circuitry where stability, long life, reliable operation, and accuracy are of prime importance. They are particularly desirable for use where miniaturization is important and where ease of assembly is desired. They are useful where a number of resistors of the same resistance values are required in the circuit.

3.10.3. Physical construction. In these resistors the resistance element consists of a film element on a ceramic substrate. The element is formed either by deposition of a vaporized metal or the printing of a metal and glass combination paste which has then been fired at a high temperature. Resistance elements are generally rectangular in shape and calibrated to the proper resistance value by trimming the element by abrasion or a laser beam. After calibration, the resistance element is protected by an enclosure or coating of insulating, moisture-resistant material (usually epoxy or a silicon compound) (see Figure 30).

Physical dimensions. Outline drawings for each style are shown in Figures 31 through 35.

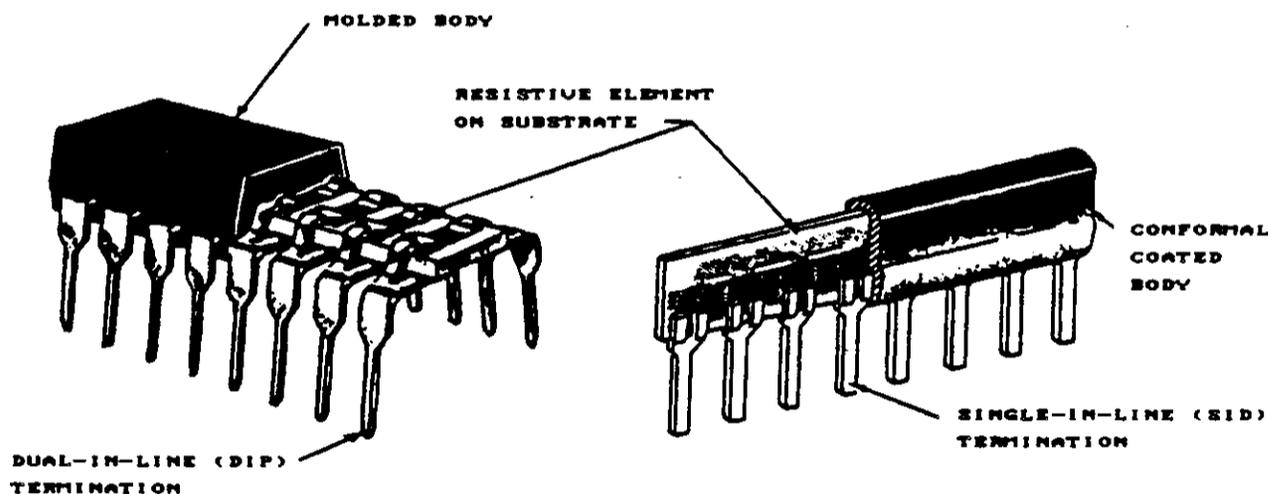


FIGURE 30. Typical construction of a typical film resistor network.

3.10 RESISTORS, FIXED, FILM, NETWORKS

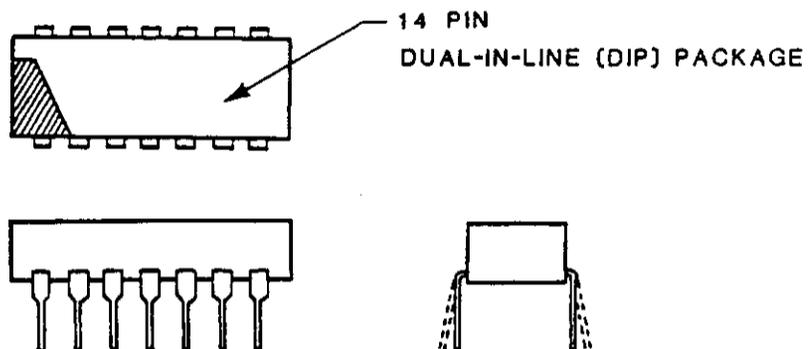


FIGURE 31. Outline drawing of a style RZ010 package configuration.

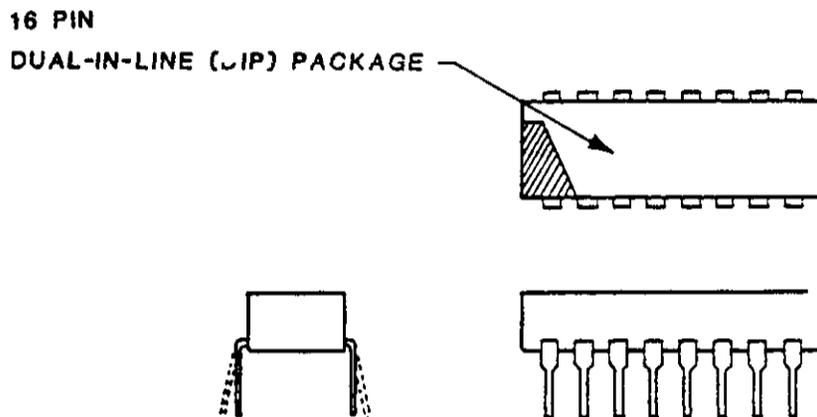


FIGURE 32. Outline drawing of a style RZ020 package configuration.

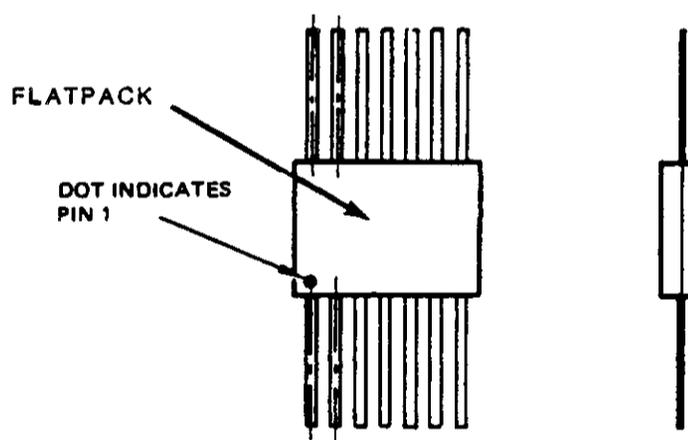


FIGURE 33. Outline drawing of a style RZ030 package configuration.

3.10 RESISTORS, FIXED, FILM, NETWORKS

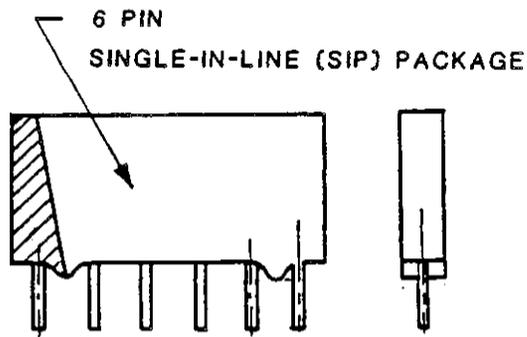


FIGURE 34. Outline drawing of a style RZ040 package configuration.

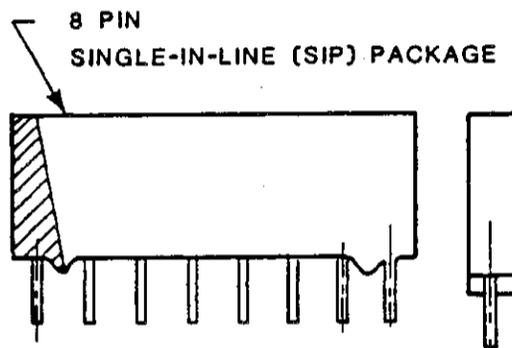
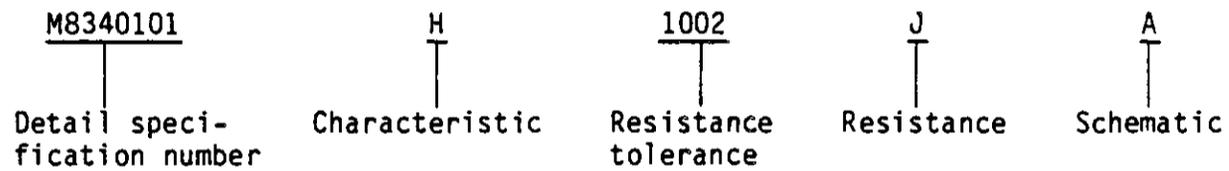


FIGURE 35. Outline drawing of a style RZ050 package configuration.

3.10.4 Military designation. An example of military part number designations is shown below.



3.10.5 Electrical characteristics. Electrical characteristics for fixed, film, resistor networks are tabulated in military specification MIL-R-83401.

Other electrical characteristics which must be considered in the application of fixed, film, resistor networks are as follows.

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3.10 RESISTORS, FIXED, FILM, NETWORKS

3.10.5.1 High frequency application. When used in high frequency circuits (200 megahertz and above), the effective resistance will be reduced as a result of shunt capacitance between resistance elements and connecting circuits. The high frequency characteristics of these networks are not controlled.

3.10.5.2 Resistance tolerance. One should bear in mind that operation of these resistor networks under the ambient conditions for which equipment is designed may cause permanent or temporary changes in resistance sufficient to exceed their initial tolerances. In particular, operation at extremely high or low ambient temperatures may cause significant temporary changes in resistance.

3.10.5.3 Voltage limitations. Because of the very small spacing between the resistance elements and the connecting circuits, maximum permissible voltages are imposed. The maximum voltage permissible for each network type is specified in MIL-R-83401.

3.10.5.4 Noise. Noise output is not controlled by specification but for these resistors noise is a negligible quantity. In an application where noise is an important factor, resistors in these networks are superior to composition types. Where noise test screening is indicated, it is recommended that MIL-STD-202, method 308 be used.

3.10.6. Environmental considerations. Film resistor networks are qualified to withstand environmental tests in accordance with Table V of MIL-R-83401.

3.10.6.1 Coating. Only hermetically sealed units (as defined in paragraph 3.10 of MIL-R-83401) should be used for space flight applications. The ceramic sandwich type construction should not be used for space flight application.

3.10.6.2 Mounting. Under severe shock or vibration conditions (or a combination of both), resistors should be mounted so that the body of the resistor network is restrained from movement with respect to the mounting base. If clamps are used, certain electrical characteristics may be altered. The heat dissipating qualities will be enhanced or retarded depending on whether the clamping material is a good or poor heat conductor.

3.10.6.3 Moisture resistance. The resistors within the networks are essentially unaffected by moisture. The specification allows only a 0.5-percent change in resistance value as a result of exposure to a standard 10-day moisture resistance test.

The environmental tests are tabulated in Table V of MIL-R-83401 along with the maximum allowable percent change in resistance value for each exposure.

3.10.7 Reliability considerations.

3.10.7.1 Derating. Because all the electrical energy dissipated by a resistor is converted into heat energy, temperature of the surrounding area is an influencing factor when selecting a particular resistor network for a specific application. The power rating of these resistor networks is based on operating at

MIL-HDBK-978-B (NASA)

3.10 RESISTORS, FIXED, FILM, NETWORKS

specific temperatures. However, a resistor network may not be operated at these temperatures. When a desired characteristic and an anticipated maximum ambient temperature have been determined, an appropriate safety factor applied to the wattage is recommended to insure the selection of a resistor network with an adequate wattage-dissipation potential. Refer to MIL-STD-975 for specific derating conditions.

3.10.7.2 Screening. All resistor networks furnished under MIL-R-83401 are subject to 100-percent screening through a 100-hour overload test plus a thermal shock test. These tests are followed by a total resistance check and a visual examination for evidence of arcing, burning, or charring.

3.10.7.3 Failure rate factors. Failures are considered to be opens, shorts, or radical departures from desirable characteristics. Failure rate factors applicable to this specification are stated and discussed in MIL-HDBK-217. The failure rate factors stated in MIL-HDBK-217 are based on "catastrophic failures." Life test failures, as defined in MIL-R-83401, are based on "parametric drift" during an operating period of 1000 hours at rated conditions.

3.11 THERMISTORS (THERMALLY SENSITIVE RESISTORS)

3.11 Thermistors (thermally sensitive resistors).

3.11.1 Introduction. The word thermistor is a contraction of thermal resistor. It is a device whose resistance varies in a significant and predictable manner with temperature. The typical thermistor is a stable, compact, and rugged two-terminal ceramic-like semiconductor manufactured by sintering mixtures of metallic oxides such as manganese, nickel, cobalt, copper, iron and uranium. In this type of nonlinear resistor, the electrical resistance varies over a wide range of temperature. In contrast with metals which have small positive temperature coefficients of resistance, thermistors are made from a class of materials known as semiconductors, most of which have relatively large negative temperature coefficients of resistance; that is, the resistance decreases markedly as the temperature increases. In positive temperature coefficient thermistors, the resistance increases with increasing temperature.

3.11.1.1 Applicable specification.

Military. MIL-T-23648, General Specifications for Thermistor (Thermally Sensitive Resistor), Insulated.

NASA. GSFC-S-311-P-18, specification for thermistor (thermally sensitive resistor), insulated, negative temperature coefficient.

3.11.2 Usual applications. Thermistors are versatile circuit elements and have many applications such as measurements and control. Some typical applications are listed in the following paragraphs.

3.11.2.1 Temperature measurements. The thermistor's large temperature coefficient of resistance is ideal for temperature measurements. In this application, the power dissipation must be so small that it does not heat the device. Precise temperature measurement is made with high-resistance thermistors in a resistance bridge. In this application sensitivity of 0.0005°C is readily attained. Lead resistance has no effect. Compensating leads and cold junctions are unnecessary. Bead thermistors are built into equipment at locations where temperature is to be measured (gear housings, bearings, cylinder heads, transformer cores) and a resistance bridge measures the temperature at a remote station.

3.11.2.2 Temperature compensation. Many electrical components have temperature coefficients which are detrimental to the temperature stability of the circuit. A properly selected thermistor in the circuit containing such a component will provide temperature compensation.

3.11.2.3 Flow-meter, vacuum gauge, and anemometer. A small voltage is applied and the current through the thermistor is measured. The amount of heat dissipated is a function of the degree of vacuum surrounding the device or the velocity of gas passing over the device. The measured current is calibrated in terms of vacuum or gas flow.

3.11 THERMISTORS (THERMALLY SENSITIVE RESISTORS)

3.11.2.4 Time delay. When a thermistor is self-heated as a result of current passing through it, its resistance varies. Due to the thermal mass, the time rate of change is fixed. The delayed buildup of circuit current can be used to introduce a fixed time delay between relay operations or to protect equipment during startup.

3.11.2.5 Power measurements, bolometer. The thermistor's resistance versus power characteristic makes it a useful power measuring device. Microwave power is measured by a bead thermistor mounted in the waveguide and biased so that bead impedance matches the cavity. When radio frequency power is applied, the bead is heated by the absorbed power. The bias current is adjusted so that the thermistor remains at the same operating temperature. The change in bias power is just equal to the radio frequency power absorbed. The thermistor also can be used to measure radiant power such as infrared or visible light.

3.11.2.6 Other applications. Thermistors are used as voltage regulators and volume limiters in communication circuits. A shunt voltage regulator is provided by shunting the circuit with a suitably chosen value of resistance in series with a thermistor. Networks of resistors and thermistors are used as compressors, expanders, and limiters in transmission circuits.

3.11.3 Physical construction. Thermistors are made by sintering mixtures of oxides of such metals as manganese, nickel, cobalt, copper, uranium, iron, zinc, titanium, and magnesium. Various mixtures of metallic oxides are formed into useful shapes. Their electrical characteristics may be controlled by varying the type of oxide used and the physical size and configuration of the thermistor. Standard forms now available are discs, beads, and probes.

Discs are made by pressing an oxide-binder mixture under several tons of pressure in a round die to produce flat pieces. These pieces are sintered, then the two flat surfaces are coated with a conducting material and leads are attached. The thin, large diameter discs have low resistance, short time constant and high power dissipation. (See Figure 36).

Beads are made by forming small ellipsoids (viscous droplets) of a metallic oxide mixture on two fine wires held tight and parallel. The material is sintered at high temperature. Upon firing, the ceramic bead cements the wires which become the leads as they become embedded tightly in the bead, making good electrical contact inside the thermistor. Bead thermistors may be coated with glass for protection or may be mounted in evacuated or gas-filled bulbs. Bead thermistors have little mass and a short time constant. (See Figures 37 and 39).

Rods are extruded through dies to make long cylindrical units of oxide-binder mixture and are then sintered. The ends are coated with conducting paste and leads are attached to the coated area. The rod type has high resistance, long time constant and moderate power dissipation. (See Figure 38).

3.11 THERMISTORS (THERMALLY SENSITIVE RESISTORS)

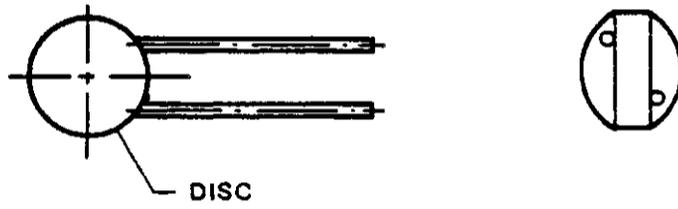


FIGURE 36. Outline drawing of a disc style thermistor.

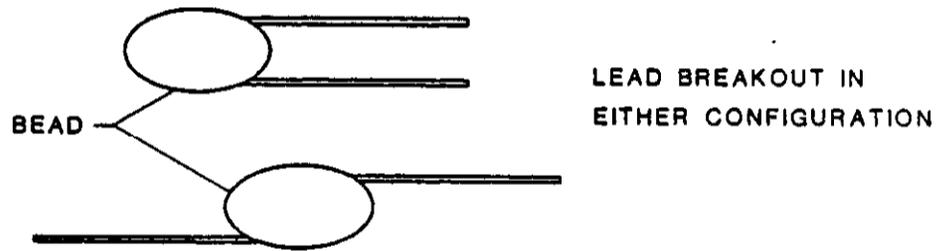


FIGURE 37. Outline drawing of a bead style thermistor.

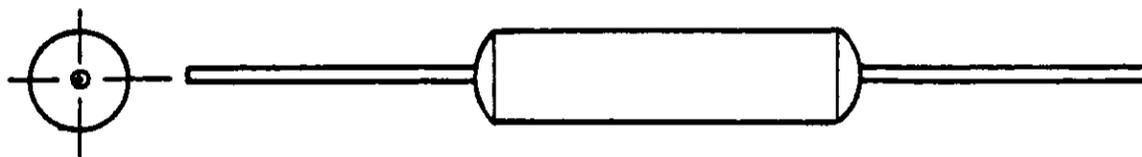


FIGURE 38. Outline drawing of a rod style thermistor.

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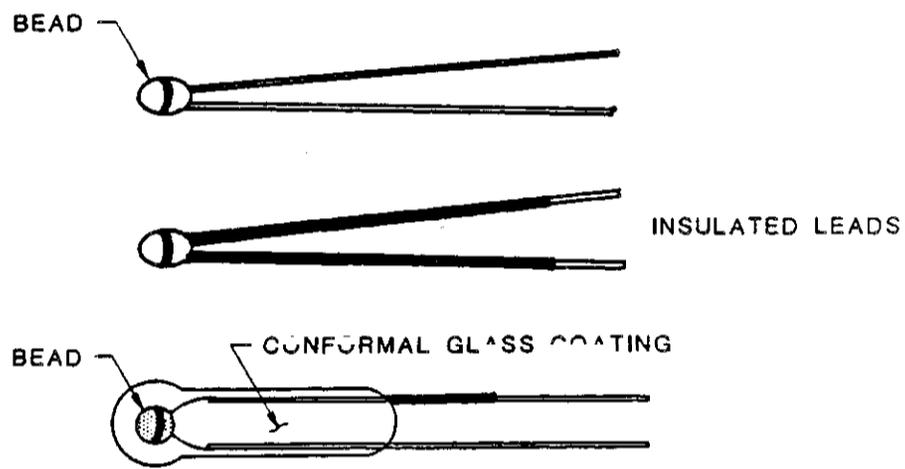
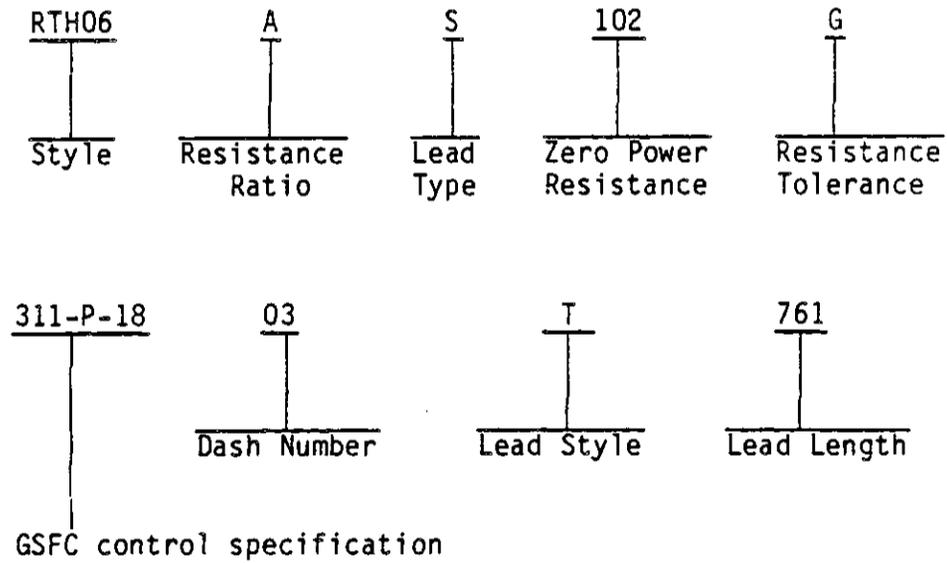


FIGURE 39. Outline drawing of a 311-P-18 style thermistor.

3.11.4 Military designation. The military type designation used for identifying and describing thermistors is shown below.



3.11 THERMISTORS (THERMALLY SENSITIVE RESISTORS)

3.11.5 Electrical characteristics. Electrical characteristics for thermistors are tabulated in military specification MIL-T-23648 and Goddard Space Flight Center Specification GSFC-S-311-P-18.

3.11.6 Environmental considerations. The thermistor has been tested for many environments and test conditions. Most testing has been done to procedures of MIL-STD-202. Currently, testing is according to procedure and format in MIL-T-23648.

3.11.7 Reliability considerations. In general, the thermistor is a highly reliable device and after installed and tested successfully, there is rarely a failure.

3.11.7.1 General reliability considerations. When using negative temperature coefficient devices, a point can be reached where the self-heating current causes a resistance drop resulting in more self-heating that increases to the destruction of the devices. This is thermal runaway. This is not a problem with positive temperature coefficient devices. Thermistor reliability is largely related to thermistor stability, which is defined as the ability of the thermistor to maintain constant resistance and resistance ratio when the thermistor is subjected to mechanical and/or thermal stress.

Another factor that may cause instability in thermistors is the change in electrical contact resistance between the thermistor material and the leads or terminal to which the thermistor is connected.

Generally, the effects of mechanical stresses such as shock, acceleration, and vibration may be alleviated by selection of the correct mounting or encapsulation for the thermistor. However, changes in contact resistance which are due to stresses that may be set up in the thermistor due to the changes in the thermistor's body temperature can affect the degree of electrical contact with a consequential shift in the thermistor's resistance characteristics. Pre-conditioning with specially developed external or internal thermal aging techniques minimizes the probability of change in contact resistance due to thermal stress.

Examples of mechanical defects which could be detected by proper visual and mechanical examination are as follows.

- a. Cracks or holes in, or chipping of, the thermistor body
- b. Wire leads broken, nicked or crushed; protective coating, if needed is missing; evidence of nonadherent areas or bare spots, chipping, flaking or peeling
- c. Terminals not suitably treated to facilitate soldering (when applicable).

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- d. Body and lead dimensions exceed the specified tolerances
- e. Identification markings missing, illegible, incomplete or incorrect
- f. Poor soldering: evidence of cold solder, joints with excessive solder, spattering, or excessive rosin residue
- g. End chipping (rod type only) exceeding 1/3 of the distance from the outside diameter to the lead.

The effect of nuclear radiation on the stability of thermistors is an important question. Tests have been made on these effects with rather encouraging results. Thermistors were exposed to gamma radiation at 1.3×10^6 roentgens/hour, fission spectrum, for more than 400 hours with no measureable damage. In another series of tests, they were exposed to neutron radiation for 2172 hours with a flux density of 4.5×10^6 N/cm²/sec for a total dose of 3.5×10^{13} N/cm² with no damage resulting. Another test provided exposure to 2 mev (million electron volts) electrons for 20 minutes in a Van de Graff generator giving a total dose of 10^9 rads and indicated no damage.

3.11.7.2 Screening. To improve thermistor reliability, it is recommended that thermistors be subjected to the following screening.

- a. Thermal shock similar to MIL-STD-202, Method 107D, Test Condition B
- b. High temperature exposure for 100 hours at 125°C per MIL-T-23648.

3.11.7.3 Failure rate factors. The predicted failure rates applicable to parts made in accordance with this specification are given and discussed in MIL-HDBK-217.

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**3.11 THERMISTORS (THERMALLY
SENSITIVE RESISTORS)**

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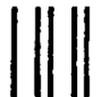
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MILITARY HANDBOOK

NASA PARTS APPLICATION HANDBOOK

(VOLUME 2 OF 5)
DIODES, TRANSISTORS, MICROWAVE DEVICES



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NASA Parts Application Handbook

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FOREWORD

This handbook provides a technological baseline for parts used throughout NASA programs. The information included will improve the utilization of the NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List (MIL-STD-975) and provide technical information to improve the selection of parts and their application, and failure analysis on all NASA projects. This handbook consists of five volumes and includes information on all parts presently included in MIL-STD-975.

This handbook (Revision B) succeeds the initial release. Revision A was not released. The content in Revision B has been extensively changed from that in the initial release.

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CONTENTS

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4.1 DIODES, GENERAL

4. DIODES

4.1 General.

4.1.1 Introduction. This section contains information on the various types of semiconductor diodes used in electronic equipment. Each type, although having essentially the same general construction, is specifically modified in die size, pellet metallization, doping, passivation, and physical configuration to accomplish a specific function. This includes small signal detection, rectification, switching, voltage reference, voltage regulation, variable capacitance, high voltage-high current rectification, transient suppression function, digital logic applications using diode arrays, and photo detection. The active portion of a diode is a semiconductor pn junction. The pn junctions are formed in various kinds of semiconductors by several techniques. The principal processes are either diffusion of a p- or an n-type impurity into a base material of the opposite type or by alloying a p-type metal into an n-type base. Another process uses a metal-semiconductor barrier type of junction (rather than the diffused type pn junction) to make a schottky diode.

Applicable military specifications.

MIL-S-19500	Semiconductor Devices, General Specification for
MIL-STD-750	Test Methods For Semiconductor Devices
MIL-STD-975 (NASA)	NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List
MIL-STD-217	Reliability Prediction of Electronic Equipment
DOD-HDBK-263	Electrostatic Discharge Control Handbook for Protection of Electrical and Electronic Parts, Assemblies and Equipment (Excluding Electrically Initiated Explosive Devices)

4.1.2 General definitions and abbreviations. This list of definitions is presented as an aid for the interpretation and understanding of the specific diode sections.

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4.1 DIODES, GENERAL

4.1.2.1 General definitions.

Anode. The electrode from which the forward current flows within the device.

Bias, forward. The bias which tends to produce current flow in the forward direction (p-type semiconductor region at a positive potential relative to the n-type region).

Bias, reverse. The bias which tends to produce current flow in the reverse direction (n-type semiconductor region at a positive potential relative to the p-type region).

Blocking. The state of a semiconductor device or junction which eventually prevents the flow of current.

Capacitance. The capacitance of a semiconductor device is the capacitance measured at designated terminals under specified conditions of bias and frequency.

Cathode. The electrode to which the forward current flows within the device.

Circuit, open. A circuit shall be considered as open circuited if halving the magnitude of the terminating impedance does not produce a change in the parameter being measured greater than the specified accuracy of the measurement.

Circuit, short. A circuit shall be considered short circuited if doubling the magnitude of the terminating impedance does not produce a change in the parameter being measured greater than the specified accuracy of the measurement.

Coefficient, temperature. The ratio of the change in a parameter to the change in temperature.

Coefficient, voltage-temperature. The change in voltage measured under specified conditions over a specified range of ambient or case temperatures. It is expressed as percent change per degree celsius.

Current, forward. The current flowing through the diode in the direction of lower resistance to the flow of steady direct current.

Current, reverse (leakage). The current flowing through the diode in the direction of higher resistance to steady direct current when a specified reverse voltage is applied.

Current, surge. The maximum current pulse which can be carried by the semiconductor diode for the length of time, repetition frequency, and waveform at the temperature specified.

4.1 DIODES, GENERAL

Diode, monolithic and multiple array. Consists of several diodes fabricated in a single monolithic chip. Monolithic arrays allow diode interconnection to form a desired circuit, whereas multiple arrays are restricted to a given circuit configuration such as a common anode circuit or a common cathode circuit.

Diode, current-regulator. A diode which limits current to an essentially constant value over a specified voltage range.

Diode, photo. A diode that is responsive to radiant energy.

Diode, semiconductor. A semiconductor device having two terminals and exhibiting a nonlinear voltage-current characteristic.

Diode, transient voltage suppressor. Transient voltage suppressors are characterized by two zener diodes oriented back to back and are capable of high voltage transient suppression.

Diode, tuning. A varactor diode used for rf tuning including functions such as automatic frequency control (afc) and automatic fine tuning (aft).

Diode, varactor. A two terminal semiconductor device in which use is made of the property that its capacitance varies with the applied voltage.

Diode, voltage-reference. A diode which is normally biased to operate in the breakdown region of its voltage-current characteristic, and which develops across its terminals a reference voltage of specified accuracy when biased to operate throughout a specified current and temperature range.

Diode, voltage-regulator. A diode which is normally biased to operate in the breakdown region of its voltage-current characteristic and which develops across its terminals an essentially constant voltage throughout a specified current range.

Efficiency, conversion. The ratio of the product of the average values of the direct voltage and direct current output to the total alternating current power input.

Efficiency, rectification. The ratio, expressed as a percentage, of the dc load voltage to the peak ac input voltage in a half-wave rectifier circuit with a resistive load.

Equilibrium, thermal. Thermal equilibrium is considered to have been reached when doubling the test time interval does not produce a change, due to thermal effects, in the parameter being measured that is greater than the specified accuracy of the measurement.

Impedance, dynamic. The ratio of the change in voltage to the corresponding change in current under the specified test conditions.

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Impedance, small-signal breakdown. The ratio of the small signal ac voltage to the ac current in the breakdown voltage region of the V-I characteristic under the specified test conditions.

Impedance, small-signal forward. The ratio of the small-signal ac voltage to the ac current in the forward region of the V-I characteristic under the specified conditions.

Junction, semiconductor. A region of transition between semiconductor regions of different electrical properties (e.g., p⁺-n, n⁺-p semiconductors) or between a metal and a semiconductor.

Matched pair. A pair of diodes identical in outline dimensions and with matched electrical characteristics. The two diodes may both be forward polarity, one forward and one reverse polarity, or both reverse polarity.

Metallurgical bond. A metallurgical bond occurs when two or more materials (metal or semiconductor) are brought into contact under temperature and pressure to form a eutectic melt, melt solution, or solid diffusion of the materials. The bond solidifies to form a regrowth or recrystallization region which contains material from both segments of the bond. A good bond is mechanically cohesive and able to withstand a predefined level of tensile stress.

Noise figure. At a selected input frequency, the noise figure is the ratio of the total noise power per unit bandwidth (at a corresponding output frequency) delivered to the output termination to the portion thereof contributed at the input frequency by the input termination, whose noise temperature is standard (293 ± 5 K) at all frequencies.

Package type. A category in which all packages have the same case outline, configuration, materials (including bonding, wire, or ribbon and die attach) piece parts (excluding preforms which differ only in size), and assembly processes.

Rating. The nominal value of any electrical, thermal, mechanical, or environmental quantity assigned to define the operating conditions under which a component, machine, apparatus, electronic device, etc. is expected to give satisfactory service.

Rating, absolute maximum. The values specified on data or specification sheets for "maximum ratings," or "absolute maximum ratings" are based on the "absolute system" and unless otherwise required for a specific test method are not to be exceeded under any application or test conditions. These ratings are limiting values beyond which the serviceability of any individual semiconductor device may be impaired. Unless otherwise specified, the voltage, current, and power ratings are based on continuous dc power conditions at free air ambient temperature of 25 °C. For pulsed or other conditions or operation of a similar nature, the current, voltage, and power dissipation ratings are a function of time and duty cycle.

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4.1 DIODES, GENERAL

In order not to exceed absolute ratings, the equipment designer has the responsibility of determining an average design value for each rating below the absolute value of that rating by a safety factor, so that the absolute values will never be exceeded under any usual conditions of supply-voltage variation, load variation, or manufacturing variation in the equipment itself.

Regulation, breakdown-voltage. The change in breakdown voltage between two specified values of reverse current.

Rectifiers, silicon controlled. A bistable semiconductor device that comprises three or more junctions and can be switched between conducting and nonconducting states.

Resistance, thermal. Thermal resistance is the temperature rise, per unit power dissipation, of a junction above the temperature of a stated external reference point under conditions of thermal equilibrium.

Small signal. A signal shall be considered small if doubling its magnitude does not produce a change in the measured parameter greater than the specified accuracy of the measurement.

Source, constant-current. A current source is considered constant if halving the generator impedance does not produce a change in the measured parameter greater than the required precision of the measurement.

Source, constant-voltage. A voltage source is considered constant if doubling the generator impedance does not produce a change in the measured parameter greater than the required precision of the measurement.

Surge Current. See Current, Surge.

Temperature, Ambient. The air temperature measured below a semiconductor device, in an environment of substantially uniform temperature, cooled only by natural air convection and not materially affected by reflective and radiant surfaces.

Time, delay. The delay time of a pulse is the time interval from a point at which the leading edge of the input pulse has risen to 10 percent of its maximum amplitude to a point at which the leading edge of the output pulse has risen to 10 percent of its maximum amplitude.

Time, recovery. See "Time, recovery, reverse" and/or "Time, recovery, forward."

Time, recovery, forward. The time required for the current or voltage to reach a specified condition after instantaneous switching from zero or a specified reverse voltage to a specified forward biased condition.

Time, recovery, reverse. The time required for the current or voltage to reach a specified condition after instantaneous switching from a specified forward current condition to a specified reversed bias condition.

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Time, rise. The rise time of a pulse is the time duration in which the amplitude of the leading edge of the pulse is increasing from 10 to 90 percent of the maximum amplitude.

Time, turn-on. Turn-on time is equal to delay time plus the rise time of the output pulse.

Voltage, breakdown. The breakdown voltage is the maximum instantaneous voltage that can be applied across a junction in the reverse direction without an external means of limiting the current. It is also the instantaneous value of reverse voltage at which a transition commences from a region of high small-signal impedance to a region of substantially lower small-signal impedance.

Voltage, forward. The voltage drop resulting from the flow of forward current through the semiconductor diode.

Voltage, reverse. The voltage drop resulting from the flow of reverse current through the semiconductor diode.

4.1.2.2 Abbreviations. Refer to MIL-STD-19500, Appendix B for abbreviations and symbols generally accepted by the electronic industry.

4.1.3 NASA standard parts. See subsection 1.1 Introduction for a complete description of the NASA Standard Parts Program. In addition to this handbook, the principal elements of this program include MIL-STD-975.

MIL-STD-975 is a standard parts list for NASA equipment; Section 4 contains a summary of standard diodes.

4.1.4 General device characteristics. The general device characteristics of switching, rectifier and power, voltage regulator, voltage reference, current regulator, voltage variable capacitor diodes, transient voltage suppressors, multiple and monolithic diode arrays, silicon controlled rectifiers, and photo diodes are briefly discussed in this section.

4.1.4.1 Switching diodes. Switching diodes are normally made of silicon or germanium. The silicon switching diode has a higher forward voltage drop than the germanium diode, but the silicon device has lower leakage currents and higher maximum junction temperatures. Silicon devices exhibit normal rectifying characteristics and have very fast reverse recovery times.

The fast reverse recovery time is accomplished either through the use of gold or platinum doping or by high energy (13 MeV) electron irradiation of the lightly doped region of the device, which greatly reduces the minority carrier lifetime in both that region and at the junction.

4.1.4.2 Rectifier and power diodes. This section includes three general types of rectifiers: power diodes, fast switching power rectifiers, and power Schottky barrier rectifiers.

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4.1 DIODES, GENERAL

- a. Power diodes. These devices are general purpose low voltage rectifiers available in both axial leaded and stud mounted packages. The dc output reverse working voltages vary from 50 to 1000 V, with the dc output currents varying from 1 to 125 A (depending upon the type of package used). These devices are used extensively in ac power supplies for rectification such as half-wave and full-wave rectifiers, half-wave voltage doublers and full-wave voltage doublers.
- b. Fast switching power rectifiers. These devices have reverse working voltages varying from 50 to 1000 V and dc output currents of 1 to 3000 A. They have fast recovery times, typically about 150 ns. This type of rectifier is suitable for use in such circuit applications as inverters, choppers, low rf interference, free-wheeling rectifiers, and dc power supplies.
- c. Schottky barrier rectifiers. These devices make use of the rectification effect of a metal to a silicon semiconductor barrier. These devices have better forward conductivity characteristics than conventional pn-junction rectifiers (because the forward voltage is about 0.25 V) but higher reverse leakage currents. However, the lower forward voltage results in higher rectification efficiency, which is somewhat offset by a greater reverse power dissipation. Power Schottky diodes are used primarily in low voltage applications where pn-junction rectifiers would result in excessive power dissipation due to forward losses and/or commutation losses. Because the forward conduction is mainly by majority carriers, the usual recovery characteristics caused by minority carrier storage in pn junctions is completely absent. Schottky diodes can, therefore, switch very rapidly. For this reason they are ideally suited to rectification of high-frequency ac power for switching regulators and converters.

In designing with Schottky power diodes, the reverse power dissipation is a significant factor (unlike pn-junction rectifiers). Therefore, the thermal design of the equipment must be sufficiently conservative to prevent thermal runaway and destruction of Schottky rectifiers.

4.1.4.3 Voltage regulators. The voltage regulator is a silicon junction diode possessing a very high back resistance up to its critical reverse breakdown, or zener, voltage. At this point, the back resistance drops to a very small value. In this region, the current increases very rapidly, whereas the voltage drop across the diode remains almost constant. Voltage regulators (commonly referred to as zener diodes), when biased in the reverse direction, can be used as voltage regulators, or reference elements because the reverse voltage at V_z will remain essentially constant for a wide range of current. The power dissipation capabilities are usually in the range of 0.4 to 20 W. The temperature coefficient ranges from typically $-2\text{mV}/^\circ\text{C}$ for a $V_z = 3.0\text{ V}$ to $+175\text{ mc}/^\circ\text{C}$ for a $V_z = 200\text{ V}$.

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4.1.4.4 Voltage reference diodes. The silicon voltage reference diode consists of two or more silicon junctions encapsulated into a single package but designed to exhibit a relatively fixed voltage and a fixed current with a very low temperature coefficient (such as 0.0005%/°C for a $V_Z = 6.35$ V). The effective voltage time stability initial-to-peak varies from 10 to 50 ppm/1000 hours for these devices.

The voltage reference diode can be used in any circuit that requires a stable reference that is insensitive to shock vibration or position. Their inherent stability allows them to be used in circuits requiring an extremely high degree of voltage time stability.

4.1.4.5 Current regulators. The current regulator diode is basically a field effect transistor that has its gate and source connected. This particular device is useful in such applications as over-current protection, transistor biasing, linear ramp or stairstep generators, differential amplifiers, and precision reference voltage sources. The 1N5285 through 1N5314 family series covers a current range from 200 μ A to 4.7 mA in 32 different current steps.

4.1.4.6 Voltage variable capacitance diodes. The voltage variable capacitance diode is a silicon pn-junction diode designed for use as a voltage variable capacitor. The capacitance varies essentially as $1/\sqrt{V}$ as the voltage across its terminals is varied. This type of device maintains constant characteristics over a wide temperature range but is less temperature sensitive when used at higher operating voltages. Therefore, if capacitance variations must be minimized, it is advisable to operate the device at higher bias voltages.

4.1.4.7 Transient voltage suppressors. Silicon transient voltage suppressors are used in applications where large voltage transients can permanently damage voltage-sensitive components.

Transient voltage suppressor (TVS) devices have a high current surge capability, extremely fast response time and a low impedance. Because of the unpredictable nature of transients, impedance is not specified as a parametric value. However, a minimum voltage at low current conditions (BV) and a maximum clamping voltage (V_C) at a maximum peak pulse current are specified in the manufacturer's performance specification.

TVS devices are designed to absorb a peak pulse power of 500, 1500 or 15,000 W dissipation for 1 ms. The response time of the clamping action of the TVS device is theoretically instantaneous (1×10^{-12} s): Therefore, they can protect integrated circuits, MOS and MSI integrated circuits, hybrids, and other voltage-sensitive semiconductors and components.

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TVS devices can be used in series or parallel to increase the peak power rating. TVS devices have proven to be effective in airborne avionics and controls, mobile communication equipment, computer power supplies and in many other applications where inductors and switching transients are present.

4.1.4.8 Multiple arrays and monolithic arrays. Diode arrays consist of several diodes which have been fabricated in an integrated circuit form. The diode array is not a unique device with characteristics that differ from discrete diodes; it is simply a packaging technique that permits a significant reduction in the size of electrical systems.

Monolithic arrays allow interconnection of diodes to form a desired circuit, whereas multiple arrays are restricted to a given circuit configuration, such as a common anode or common cathode circuit.

4.1.4.9 Silicon-controlled rectifiers (thyristors). The silicon-controlled rectifier (SCR) is a semiconductor device that has characteristics similar to those of a thyratron gas tube. The device can be switched between states by a current or polarized voltage pulse. Unlike conventional transistors, the device lends itself to use as a high current, high voltage rectifier or a static latch (limited to microseconds), or a sensitive high gain amplifier control.

The silicon-controlled rectifier is well adapted for use as a latching switch or high-power gain amplifier. The SCR can be turned on by a momentary application of control current applied to the control gate, whereas tubes or transistors require a continual signal. The turn-on time is about 1 μ s, and turn-off time is about 10 to 20 μ s. This latching action can be controlled by signals of only a few microwatts but can switch up to 200 V. With associated transistor-triggering circuits, unusual current gain on the order of 10^9 can be obtained. Because of this advantage there are many applications to which the SCR can be adapted.

4.1.4.10 Photodiodes. Photodiodes are used as light detectors in a variety of applications, such as card and tape readers, pattern and character recognition, shaft encoders, position sensors, and counters.

Photodetectors may be subdivided into four categories: conventional pn junction photodiodes, PIN photodiodes, avalanche photodiodes, and Schottky barrier photodiodes.

Conventional pn junction photodiodes are used in the applications outlined above in the photoconductive mode. The PIN photodiode and the avalanche photodiode are used as visible or infrared detectors in fiber optic system applications usually with a suitable optical source such as a laser diode for long distances and a LED device for shorter distances (1 km).

The Schottky photodiodes are recommended when a high blue response (less than 500 nm) or larger areas (greater than 1 cm²) are required in the system application.

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The main parameters of the photodiode are response time, responsivity, spectral response, noise, and dark current. The peak wavelength of the spectral response curve (R_A vs λ) at a given temperature will largely dictate the circuit application requirements. The peak wavelength is a strong function of the semiconductor material band gap and the doping used to make the junction.

4.1.5 General parameter information. The basic function of the semiconductor diode and general parameter information is presented in the following discussion. For parameter information on transient voltage suppressors (see subsection 4.8), diodes arrays (4.9), silicon-controlled rectifiers (4.10), and photodetectors (4.11), refer to the appropriate individual paragraphs.

4.1.5.1 The pn junction. A pn junction can be formed only by a chemical process within a single crystal. If separate p- and n-type crystals were joined mechanically, a polycrystalline semiconductor would be the subsequent result. This type device will not furnish rectification.

Accompanying the formation of the pn junction is a region known as the depletion zone. This zone is so called because within it there is an absence of holes and excess electrons. This depletion zone is also referred to as the space charge zone because the acceptor and donor ions are fixed and charged electrically. This space charge forms a barrier to current flow.

When an external battery is applied with the positive output attached to the p-layer, the junction is in the forward-biased state. The external voltage source causes the holes in the p-region to move from the positive source potential to the negative potential. An opposite action occurs in the n-material. These two actions cause the depletion zone to shrink, thereby causing less resistance to majority carrier current flow.

When the applied battery or source potential is placed with the negative potential attached to the p-layer, the junction is in the reverse-biased state. The holes in the p-region are drawn toward the negative terminal, while the electrons in the n-region are drawn to the positive terminal. This resultant action causes the depletion zone to increase in thickness, subsequently creating a large resistance for majority carrier flow.

The reverse current tends to maintain a relatively constant value at all voltages up to a voltage called junction breakdown voltage. In this voltage region, current conduction across the junction interface increases rapidly and the diode is often destroyed by heating. There are two causes of voltage breakdown in semiconductor diodes: avalanche breakdown and zener breakdown.

Avalanche voltage breakdown can be thought of as an electrical multiplication process.

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In the avalanche process, a free electron acquires enough energy from the applied voltage (which exists across the very narrow junction interface) to accelerate it sufficiently such that when it collides with a fixed electron, it knocks it free. These electrons are again accelerated until each undergoes a second collision, resulting in further electron multiplication. The higher the applied voltage, the more rapid the electron multiplication. The voltage across the junction does not increase substantially, because the energy of the avalanche electron is limited by the critical impact velocity.

Under zener breakdown, the actual breakdown of the semiconductor is initiated through the direct rupture of the covalent bonds due to an exceptionally strong electric field which is developed across the junction. Zener breakdown is more prevalent in a very narrow junction where the field intensity becomes very high. In wider junctions, avalanche breakdown is more prevalent, because the impressed voltage is not confined to such a narrow region. In diodes, the junction will recover when the magnitude of the reverse voltage is reduced below the breakdown voltage provided the diode junction has not been damaged by excessive temperatures while operating in the breakdown region.

4.1.5.2 Diode voltage-current relationship. The ideal pn junction follows the voltage-current relationship as predicted by the simple first-order theory as developed by Schottky. This relationship is expressed by the following equations:

$$I_F = I_{Sat} \left(e^{qV/nkT} - 1 \right)$$

or

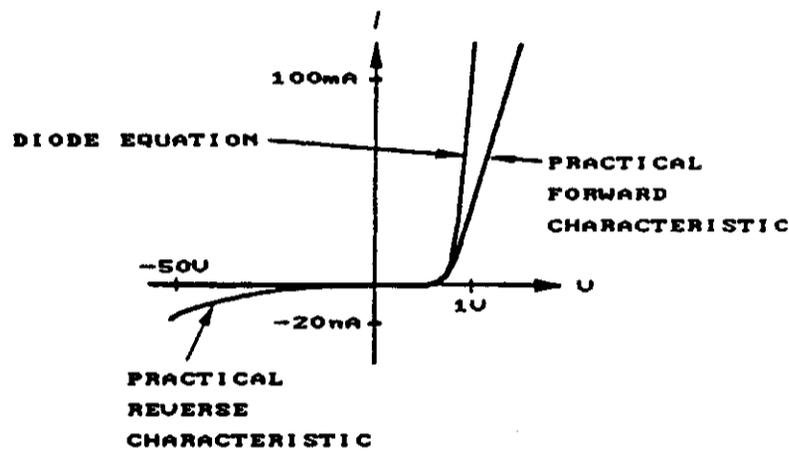
$$V = \frac{nkT}{q} \ln \left(1 + \frac{I_F}{I_{Sat}} \right)$$

where

- I_F = Forward junction current
- I_{Sat} = Reverse saturation junction current (due to bulk silicon only)
- k = Boltzman's constant
- T = Absolute temperature (degrees Kelvin)
- q = Electronic charge
- V = Voltage across the junction
- n = Ideality factor, varies from 1.0 to 2.0, depending upon device design features.

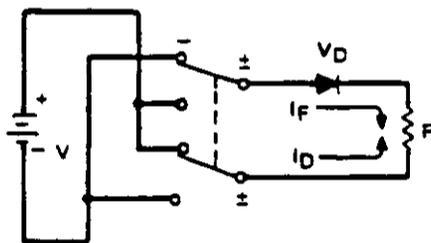
A plot of the typical silicon diode characteristic is shown in Figure 1. This shows that for any reverse voltage, in excess of a few tenths of a volt, a small reverse current is produced which remains essentially constant. When a forward voltage is applied, the forward current increases exponentially.

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FIGURE 1. Typical silicon diode characteristics.

An ideal rectifier diode has a very low forward resistance and a very high reverse resistance. To create a diode with a low forward drop and a high reverse voltage capability, it is necessary for the semiconductor layer on one side of the junction to be highly doped with impurities (low resistivity) and the opposite layer to have a low doping level (high resistivity). If the p-region is more heavily doped with impurities than the n-region, it then has a greater number of current carriers and becomes the anode.

4.1.5.3 The pn junction turn-off theory. The characteristics of a diode under turn-off conditions may be explained by using a turn-off test circuit such as the one shown in Figure 2, which is a diode in series with a resistor placed across some fixed power supply. The polarity of the power supply may be reversed with the aid of a double-pole double-throw switch. If, initially, the switch is connected so that the diode is biased in the forward direction, the induced voltage causes a steady-state current flow I_F in the diode and load as shown in Figure 2. Consequently, $I_F = (V - V_D) / R_L$.

FIGURE 2. Diode turn-off time test circuit.

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4.1.5.4 Reverse Transient. The reverse transient occurs when a diode is switched from a forward-conducting state to a reverse-biased condition. The reverse impedance of a diode is very high. However, this high impedance condition does not appear instantaneously when the diode is reverse-biased. A typical reverse recovery waveform is given in Figure 3.

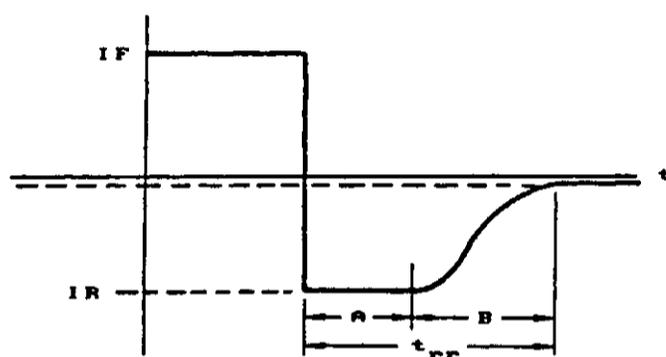


FIGURE 3. Typical reverse recovery waveform.

The delay in the appearance of the high reverse impedance is the result of two factors: minority carrier storage and junction capacitance. As a first approximation, region A can be attributed to minority carrier storage and region B to the junction capacitance.

4.1.5.5 Minority carrier storage. When the junction is forward biased, an excess of minority carriers builds up on either side of the junction region. From pn junction theory, the density of holes in the n region in the vicinity of the junction is given by:

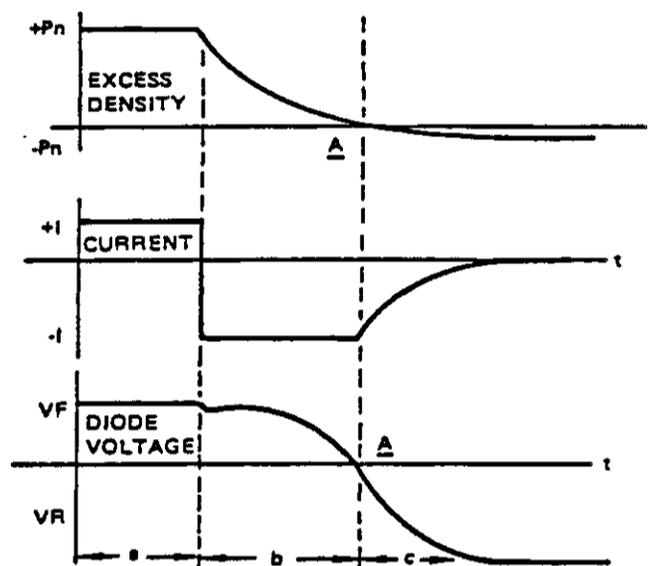
$$P(x=0) = P_n e^{qV/KT}$$

From the definition of excess density,

$$P(\text{excess}) = P(x=0) - P_n(e^{qV/KT})$$

Therefore, the excess density is a function of the voltage across the junction of the device. When the diode is forward-biased, the term, $e^{qV/KT}$, is much greater than 1 because q/KT at room temperature is approximately 40. When the voltage across the device drops to zero, all excess carriers will have been swept from the junction region as shown in Figure 4 (point A). However, the current at point "A" does not instantaneously drop to zero because of the junction capacitance of the device.

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- a = Forward bias
 b = Minority carrier storage
 c = Charging of the Junction Capacitance

FIGURE 4. Switching characteristics.

4.1.5.6 Storage time. When the pn junction is biased in the reverse direction for the majority carriers, it is biased in the forward direction for the minority carriers. The application of forward bias will, due to carrier action, increase the minority carrier density in both the anode and cathode sides of the junction. The minority carrier density in the highly doped p layer will be greater than in the n layer. In order to stop the I_F by reversing the external source voltage, it is necessary for the minority carriers to return to the junction. This travel of minority carriers takes a finite time. The diode will remain forward-biased until the minority carrier density at the junction becomes less than the equilibrium value. The finite time that it takes to return to equilibrium is known as storage time (t_s).

4.1.5.7 Junction capacitance. The voltage across the junction of the device creates a space charge on either side of the junction. This space charge is equivalent to a capacitor with a cross-sectional area and plate separation equal to that of the junction, and having a permittivity equal to that of the semiconductor material. The junction capacitance is a nonlinear function of the junction voltage, increasing with forward bias and decreasing as reverse bias increases. This capacitance is given by:

$$C = \frac{K}{(V + \phi)^n}$$

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where: K = semiconductor doping constant
 V = applied voltage
 ϕ = built-in junction potential
 n = diode constant ($n = 1/3$ for graded junction)
 (diffused) and $n = 1/2$ for abrupt junction (step).

A typical curve of junction capacitance versus voltage is shown in Figure 5.

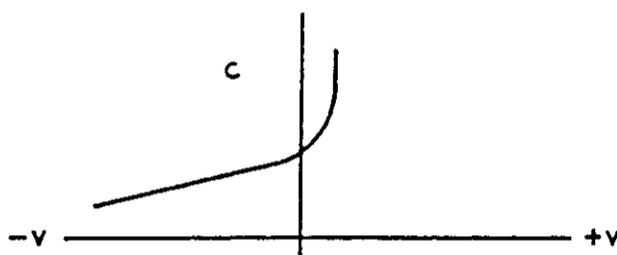


FIGURE 5. Typical junction capacitance vs voltage.

As stated previously, the recovery region C shown in Figure 4 is attributed to the junction capacitance. This capacitance in series with circuit resistance determines the effective time constant of the reverse voltage across the device.

Several factors are responsible for the magnitude and duration of the reverse recovery transient. The magnitude of the transient is primarily controlled by the external circuit in which the diode is used. Typically, this magnitude is given by V_R/R_L (V_R = reverse voltage, R_L - limiting resistance). The duration of the transient is dependent upon the forward current (magnitude and duration), reverse current, junction capacitance, series resistance, and temperature.

4.1.5.8 Metal barrier Schottky junction. A Schottky barrier diode consists of a metal semiconductor junction formed between aluminum, gold, silver or platinum metallization, and lightly doped n-type silicon or gallium arsenide with an N/N⁺ epitaxial structure. Refer to Figure 6A for a typical structure.

In both materials, the electron is the majority carrier. In the metal, the level of minority carriers (holes) is insignificant. When the materials are joined the electrons in the n-type silicon semiconductor material immediately flow into the adjoining metal, establishing a heavy flow of majority carriers. Because the injected carriers have a very high kinetic energy level compared with the electrons of the metal, they are commonly called "hot carriers." In the conventional pn junction minority carriers are injected into the adjoining region. Here, the electrons are injected into a region of the same electron plurality. Schottky diodes are unique, in that conduction is entirely by majority carriers. The heavy flow of electrons into the metal creates a region near the junction surface depleted of carriers in the silicon material--much

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like the depletion region in the pn junction diode. The additional carriers in the metal establish a negative wall in the metal at the boundary between the two materials. The net result is a "surface barrier" between the two materials, preventing further current flow. That is, any electrons (negatively charged) in silicon material face a carrier-free region and a negative wall at the surface of the metal.

The application of a forward bias as shown in Figure 6 will reduce the strength of the negative barrier through the attraction of the applied positive potential for electrons from this region. The result is a return to the heavy flow of electrons across the boundary, the magnitude of which is controlled by the level of the applied bias potential. The barrier at the junction for a Schottky diode is less than that of the pn junction device in both the forward- and reverse-bias regions. The result is a higher current at the same applied bias in the forward- and reverse-bias regions. This is a desirable effect in the forward-bias region but highly undesirable in the reverse-bias region.

As the forward current flow is by the injection of electrons into the metal which already contains a high density of free electrons, the storage effect which limits the switching speed of pn junction diodes, does not occur in Schottky diodes. In addition, by making the junction area (surface metallization) small, the depletion capacitance due to space charge can be kept small. The overall small capacitance provides an extremely fast switching speed of typically few picoseconds.

Due to the high electron concentration in the metal, the drift of the electrons across the junction into the semiconductor is considerably higher than in the pn junction case, resulting in reverse leakage current at least an order of magnitude greater than that for a corresponding silicon pn junction (Figure 6).

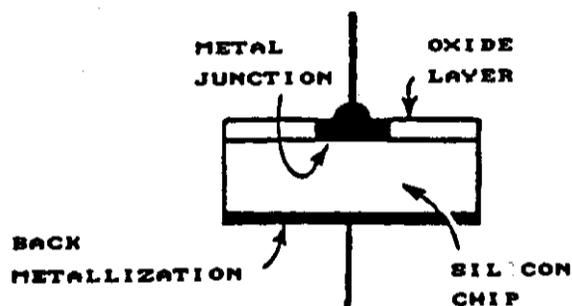
4.1.6 General reliability considerations. The ultimate goal of a circuit designer is to produce circuits, which when assembled into a system, will enable the equipment to perform its intended function with less than a specified percentage of down time due to equipment malfunction. To do this, the designer must have a knowledge of all facets of reliability that contribute to system reliability, some of which follow:

- a. The relationship of component reliability to system reliability
- b. The causes of component failure
- c. How reliability is measured
- d. The various methods of specifying reliability assurance
- e. The factors involved in selecting components
- f. The effect of circuit design upon overall system reliability.

4.1 DIODES, GENERAL

Reliability is usually expressed as failure rate in percent per thousand hours. Because failure rate is the reciprocal of MTBF, it is also possible to express diode failure rates in terms of MTBF, but such an expression can be extremely misleading. For example, failure rates on the order of 1.0 to 0.1% per thousand hours of life testing at maximum rated conditions are available from the semiconductor industry today. A failure rate of 0.1%/1000 hours is equivalent to an MTBF of 1,000,000 hours (over 114 years).

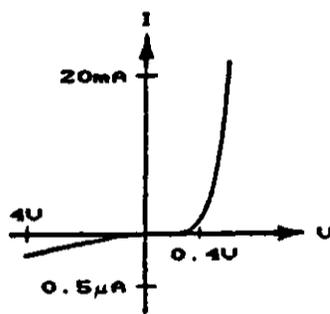
This is a rather meaningless figure for a number of reasons, some of which will be discussed in the following paragraphs.



A. Typical structure



B. Circuit symbol type



C. I-V characteristics type

FIGURE 6. Schottky-barrier diode.

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4.1 DIODES, GENERAL

4.1.6.1 Achieving diode reliability. Three major factors contribute to diode reliability:

- a. Good basic device design and material design as it relates to packaging
- b. Good manufacturing processes
- c. Quality and reliability control.

Only when all three factors are optimized will diode reliability be at a maximum.

4.1.6.2 Causes of failure. A general knowledge of the causes of semiconductor device failure is essential to an understanding of diode reliability. Diode failure mechanisms can be broadly classified as follows:

- a. Surface defects
- b. Mechanical defects
- c. Bulk defects
- d. Wire/bond defects
- e. Contamination defects.

4.1.6.2.1 Surface defects. Most diode failures are related to the condition of the semiconductor surface. A poor surface condition may be caused either by imperfections within the encapsulated diode, or by failure of the package (which causes the semiconductor surface to be subject to the external environment), or a combination of the two. During diode fabrication, every precaution is taken to assure stability of semiconductor surfaces. This is particularly true for the fabrication steps just prior to encapsulation.

Such techniques as (1) the encapsulation of the diodes in an inert atmosphere (such as nitrogen) to reduce the possibility of chemical reaction with the semiconductor surface, (2) the use of getters which absorb moisture to maintain low partial vapor pressure within the package, and (3) the use of surface passivation for silicon devices to form a chemically bonded film for surface deactivation are all designed to stabilize or isolate the semiconductor surface from the surrounding environment.

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4.1 DIODES, GENERAL

Stresses, which cause a change in the state of the semiconductor surface during diode life, are another potential source of diode failure. Among the factors which can introduce mechanisms to change the state of the diode surface are the following:

- a. Entrapment of moisture or other contaminants within the diode during encapsulation
- b. Loss of the hermetic seal due to improper encapsulation; i.e., leaks present at the time the diode is manufactured or occurring during subsequent diode life.

One of the mechanisms of failure is the creation of conductive shunt paths, which can be above or below the oxide surface of passivated silicon diodes.

The surface passivation of silicon diodes by the growth of silicon dioxide or nitride, which are chemically grown on the surface, affords a greater degree of surface protection than has previously been available. However, unless properly designed and manufactured, even this class of diodes may have surface instability problems. Among the causes of these problems are contaminants sealed beneath the passivated surface, pin holes in the passivated film, and ionized conductive paths on the surface of the passivated film. As a result, hermetic encapsulation is still desirable, even for passivated diodes when maximum reliability is required.

Surface defects are most often detected by reverse current (I_R) instability over periods of life stressing. Fabrication techniques are not identical for all device types, and these differences can create different levels of I_R between the device types. The magnitude of I_R , therefore, becomes significant only when compared with the mean I_R for that device. (See Table I).

4.1.6.3 Mechanical defects. The mechanical defects which can occur in diffused diode include:

- a. Poor bonding of die to header
- b. Poor lead-to-die contact
- c. Lack of hermetic seal.

Poor contact of the die to the header may increase the thermal resistance of the diode, resulting in high junction temperatures during high-power operation. Poor lead-to-die contacts may cause hot spots, but this is of secondary importance for relatively low level applications (See Table II). The effects of poor package sealing on surface stability have been reviewed in previous paragraphs.

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TABLE I. Failure mechanism analysis surface defects

Failure Mechanism	Description/Cause	Detection Method	Method to Minimize or Eliminate Cause
Ion migration through and across the oxide	Inversion of the n region near the anode contact, surface contamination and lack of an oxide sealant	High temperature, reverse bias, reverse current deltas	Improved control of surface cleanliness, application of phosphosilicate glass over the thermal oxide
Hole trapping in the oxide	Donor states in the oxide become positively charged resulting in an accumulation layer and premature breakdown	High temperature, forward bias, reverse current, deltas	Improved control of surface cleanliness, application of phosphosilicate glass over the oxide
Holes in oxide	Pinholes in the oxide result in shorting paths from anode to cathode	Electrical test	Improved process control

4.1.6.4 Bulk defects. Bulk defects in diodes are generally a less frequent cause of poor reliability than surface or mechanical defects. Bulk defects are often difficult to detect by in-process controls during the diode fabrication process, although they are usually detected at the final electrical test.

Included in this classification of defects are crystal imperfections and undesired impurities. Crystal defects can cause nonuniform diffusion, resulting in high current concentrations and hot spots. Undesired impurities can result in uneven voltage gradients. These uneven voltage gradients can cause, in a worst case, failure due to punch through. A second class of bulk defects results from diffusion of impurities and metal contacts into the bulk material at normal operating temperatures. This problem is generally minimized in a well designed and fabricated diode. (See Table III.)

4.1.6.5 Wirebond/interconnect defects. Diodes come in a variety of packages, some of which are similar to a transistor TO-5 or a microcircuit dual-in-line package. For these types, the wirebond failures will be similar to the transistor wirebond failures.

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Gold-aluminum systems exhibit failures caused by excessive intermetallic formations or voids under the bond. Aluminum-aluminum systems exhibit failures in the wire adjacent to the bond area due to excessive pressure during bonding which causes the wire to crack.

Established, verified, and controlled bond schedules are a requirement for reliable bonds. The part manufacturer should have in-line control to assure that bond quality is being maintained.

TABLE II. Failure mechanism analysis mechanical defects

Failure Mechanism	Description/Cause	Detection Method	Method to Minimize or Eliminate Cause
<u>A. Double slug construction</u>			
Loss of hermetic seal	Fracture of glass case	Visual, dye penetrant	Handling techniques, mounting techniques, 100% visual inspection
	Seal fracture due to insufficient sealing area, low heat, short time, contamination	Visual	100% visual, better controls on preseal and seal operations
Poor solderability	Contamination of plating or base metal during plating	Hot oil dip of solder coated leads on sample basis, visual	Control of cleaning and plating operation, 100% visual
Intermittent	Variation of conductivity during temperature cycling, lack of control of assembly process	Monitor V_f during temperature cycling	Process Control, use monolithic double slug construction

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4.1 DIODES, GENERAL

TABLE II. Failure mechanism analysis mechanical defects (Continued)

Failure Mechanism	Description/Cause	Detection Method	Method to Minimize or Eliminate Cause
B. <u>Whisker (DO-7 type) construction</u>			
Loss of hermetic seal	Fracture of the glass to seal	Hermetic seal test	Handling techniques, mounting techniques
Die lifted	Excessive voids in the eutectic bond	X-ray, thermal impedance test	Vacuum during eutectic bonding
Poor solderability	Contamination of plating or base material prior to or during plating	Hot oil dip of solder coated leads on a sample basis, visual inspection	Control of cleaning and plating operations
	Oxidation or contamination of the plated metal due to handling	Visual	Application of a protective coating that is compatible with normal fluxing and soldering operations
C. <u>Whisker construction</u>			
Loss of hermetic seal	Fracture of glass case	Visual, seal test	Handling techniques, mounting techniques
	Seal fracture due to insufficient sealing area, low heat, short time, contamination	Visual, seal test	100% visual, better controls on pre-seal and seal operations
Poor solderability	Contamination of plating or base metal during plating	Hot oil dip of solder coated leads on a sample basis	Control of cleaning and plating operation, 100% visual

4.1 DIODES, GENERAL

TABLE II. Failure mechanism analysis mechanical defects (Continued)

Failure Mechanism	Description/Cause	Detection Method	Method to Minimize or Eliminate Cause
D. <u>Voidless mono-lithic construction</u>			
Loss of hermetic seal	Fracture of glass case	Visual, dye penetrant, electrical	Handling techniques, mounting techniques
External lead separation	Poor braze	Visual, pull test	Single element metallurgical bond, better control of brazing operation, vacuum brazing
Poor solderability	Contamination of plating or base metal prior to or during plating	Hot oil dip of solder coated leads on sample basis, visual inspection	Control of cleaning and plating operation, use of single element leads, 100% visual

TABLE III. Failure mechanism analysis bulk defects

Failure Mechanism	Description/Cause	Detection Method	Method to Minimize or Eliminate Cause
Metal precipitation	Gold precipitates during manufacturing operations such as glass sealing, resulting in soft breakdown characteristics	Power cycling with electrical test (breakdown voltage)	Not known, possibly related to dislocations in the silicon
Dislocations	Imperfections in the silicon that provide nucleation centers for metal precipitates that result in soft breakdown characteristics	Power cycling with an electrical test (breakdown voltage)	Monitor wafer processing using X-ray topography technique
Cracked die	Scribe crack propagates into the junction	Internal visual Electrical test	Ultrasonic or laser scribing

4.1 DIODES, GENERAL

Most of the S or C bend type of interconnect defects are easily detected by a visual inspection of the completed package prior to painting. These are the misplaced, excessive bend, or touching die type of defect. A more difficult type of defect to detect is insufficient pressure against the die contact or mechanical contact only.

Power stud diodes use a variety of anode/cathode interconnect techniques to effect low anode resistance ohmic contact. Some use an ultrasonic bond (aluminum-to-aluminum), some use an alloy type, and others use solder. The solder type of interconnect is desirable because of a thermal fatigue problem. The alloy or eutectic type has proven to be an acceptable method because it has been shown to be free from metal thermal fatigue problems. The ultrasonic bond is a reliable technique if properly established and controlled. Some parts may either have an alloy and an ultrasonic bond or ultrasonic bond and electrical weld.

The prime concerns should be that the process has been established, verified, is controlled, and uses materials that are metallurgically compatible as manufactured.

Some double-slug diodes depend on the glass case to provide the compression to maintain electrical contact between the die and the leads. This can result in parameter variation during temperature cycling. After a stabilization period, the parameters are within specification limits. The change occurs during the temperature change because of a difference in thermal coefficient of expansion between the glass and the heat sink materials. Also, a part can operate in an intermittent fashion, operating properly only when axial compression is applied. In these cases the glass case is not providing enough compression to hold the leads in contact with the die.

Some double-slug diodes use a construction technique that forms a monolithic structure of the heat sinks and the die. This is achieved by plating the die and the heat sinks with compatible materials and elevating them to a high temperature (900 °C) to form a bond. The bond interface is across the total die surface.

Methods of external lead attachment to the diode heat sinks vary. One method is to form the lead attachment directly to the heat sink using only the heat sink plating and the lead material to form a eutectic bond. This results in a bond that is as strong as the tensile strength of the wire. The bond is formed at about 700 °C and is relatively void free. Another method is to use a third metal as the eutectic. This method has a tendency to cause voids similar to those encountered in eutectic die mounting. A great many voids could result in a weak lead connection that could fail in shock or vibration during service.

4.1.6.6 Contamination defects. Contamination discussed in this paragraph refers to a material in the package, loose or attached, that is not part of the design. Examples are weld splatter, silicon chips, solder balls, and pieces of internal lead wire.

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4.1 DIODES, GENERAL

Sealing of metal packages by welding causes weld splatter. These particles of weld splatter may or may not be located so as to be detected during electrical testing. Those that are not may relocate during shock or vibration in usage and cause shorts.

Silicon chips can result from diamond scribing. Scribing may cause cracks in the silicon that will cause a chip of silicon to separate from the die. The propagation rate is dependent on time, temperature, and environment exposure. The chip could cause an electrical short or mechanical damage during shock or vibration in usage.

Some diodes are assembled using solder to attach the die to the header. During the soldering process, solder balls may be formed. This introduces a potential failure mechanism of the same type as weld splatter. The balls may break loose during environmental stress and cause shorts.

Diodes assembled with internal lead wires are exposed to potential failure mechanisms similar to transistors and integrated circuits. Wire particles and pieces may remain in the package to cause shorts or mechanical damage. These kinds of diodes lend themselves to internal visual inspections prior to capping, to screen out many of the defects.

4.1.6.7 Failure analysis. Although complete diode failure analysis is quite complex, preliminary analysis can prove very helpful in improving reliability.

If electrical testing shows the diode to be inoperative, the failure may be mechanical in nature, and the device should be X-rayed in an attempt to see the cause of failure before the diode is cut open. Opening a diode case should be the last operation in failure analysis because additional damage may be done which may mask the original cause of failure. After the diode is opened, the cause of mechanical failure will usually become apparent under microscopic examination.

If, when tested, the diode shows little or no deviation from the specification, it is well to observe its characteristics on a curve tracer where any irregularity in characteristic curves will be apparent. The diode should be tapped, while its characteristics are being observed, to detect any intermittent condition.

If the diode shows excessive leakage, the case should be thoroughly washed to remove any conductive paths that have formed externally. If the diode is from the low leakage series, it is necessary to assure that the body paint is still intact.

The investigation may be carried further by increasing and decreasing the diode temperature to the limits of the diode rating while observing the device characteristics on a curve tracer for irregularities.

4.1 DIODES, GENERAL

With the possible addition of a leak detection test, this is probably as far as failure analysis can be practically conducted outside of a semiconductor laboratory. Even for this preliminary analysis, thoroughly trained personnel and complete facilities are necessary.

4.1.6.8 Failure rate as a function of time. An idealized curve of component failure rate versus time is shown in Figure 7. Several features of this familiar "bathtub" curve are important in any consideration of diode reliability. The first portion of this curve indicates a high failure rate and then a steadily decreasing failure rate during the screening portion of diode life. This region is often referred to as "the infant mortality region." The portion of this curve which shows a decreasing failure rate for diodes has been demonstrated. These early life failures are generally classified as a result of poor workmanship.

The failure rate during the very early life of a diode depends upon a number of factors. Among these are the actual zero time in the life of a diode, the definition of failure, and the inherent reliability of the diode. Actually, the life of a diode begins when the encapsulating process is completed. On high-reliability diode lines a period of stressing at elevated temperature is often standard operating procedure to stabilize the diode's characteristics. The time and the stress applied during this stabilization process will affect the early life failure rate, and thus will significantly affect the shape of the curve.

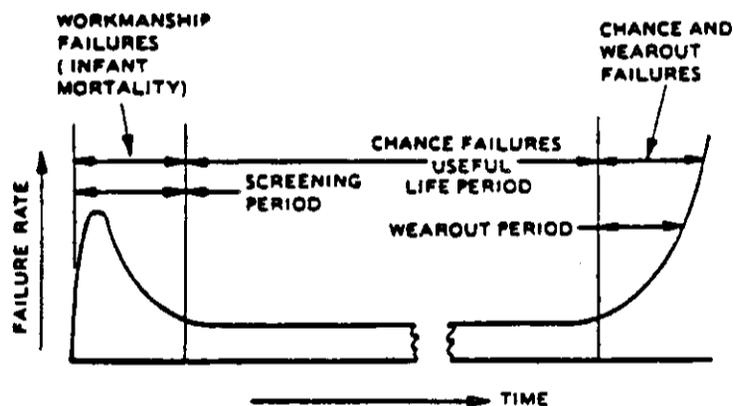


FIGURE 7. Failure rate as a function of time.

In any discussion of failure rate, the criteria used to define a failure will affect the failure rate for any given period of time. For example, a diode which has a certain amount of instability of characteristics early in life can exhibit different failure rates depending upon the relationship of initial limits and limits after a specified period of time. When tested to a life test specification, which defines a failure as a device exceeding the initial electrical parameter limits, these diodes will have a higher early life failure rate than they would have had if tested to a specification with life test limits relaxed from initial limits. If the diode parameters continue to drift

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with time, even the relaxed life test limits would be exceeded and the total number of failures would be the same, regardless of the specified limits. However, if the diodes should stabilize after a short period of time, as is often the case, then the failure rate would be less to the relaxed life test points than to the tighter limits.

Figure 7 shows that after the initial high and decreasing failure rate period, which can be attributed to workmanship faults not detected during the manufacturing process, a period of relatively constant failure rate at a low level commences. This is the period of random failures.

The final portion of Figure 7 shows an increasing failure rate indicated as "wearout." This portion of the failure rate versus time curve is extremely difficult to define and will vary depending on diode method of fabrication and stress applied. This increasing failure rate can be introduced by such mechanisms as thermal fatigue of the solders between the silicon die and the mount due to repeated cycling of junction temperature while the case is at more or less a fixed temperature, by glass hermetic seal failure due to environmental cycling, by fatigue of internal construction due to mechanical stress, or by bulk defects. Little data is available from either diode life tests or system field tests to permit an accurate picture of this portion of the failure rate versus time curve. Contrary to the early life failure which may be characterized as workmanship faults, the failures which occur in the wearout period are believed to be a result of the basic design limitations of a diode.

The fact that failure rate is not constant with time throughout diode life dictates that any statement of failure rate must refer to the time period considered. Normally, failure rates are based upon the first 1,000 hours of diode life tests unless otherwise stated. This changing failure rate during diode life is a reason for not using MTBF as a measure of reliability on an individual diode basis.

4.1.6.9 Environmental considerations. All MIL-STD-975 diodes require vigorous screening and conformance testing in accordance with MIL-S-19500 and its applicable slash sheet. These tests are done to assure that the devices can withstand certain levels of environmental conditions such as pressure, vibration, moisture, temperature cycling, mechanical stress, and, if required, radiation. The use of these devices should take precautions so that these levels are not exceeded in the design of a system.

4.1.6.9.1 Electrostatic discharge. Because the damage caused by electrostatic discharges (ESD) is not always detectable, the user must take precaution when the user must take precaution when handling ESD sensitive devices. Refer to the Electrostatic Discharge Control Handbook, MIL-HDBK-263, for guidance.

4.1.6.10 Screening procedures. Because many early-life diode failures are the result of manufacturing flaws, it is possible to develop screening procedures to improve diode reliability. Actually, most diode manufacturers use screening procedures as a regular part of the diode fabrication process. The effectiveness of any screen procedure must be carefully verified for the particular type of diode under consideration.

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Screening tests are applied to uncover manufacturing defects and potential early-life failures. Depending on the manufacturing process used, screening tests such as power-applied burn-in, temperature cycling, and high temperature storage may be used.

All diodes are measured for significant electrical characteristics to detect devices with abnormalities, which may cause poor reliability. Most bulk and surface defects are detected at electrical characteristics screening.

Additional screening processes may be used to improve reliability, but unless properly selected, they may have the opposite effect of actually reducing diode life. For example, extreme mechanical stresses may not only destroy weak units, but may weaken good units.

Two conclusions, basic to diode reliability, may be drawn from the curve of failure rate versus time. These are as follows:

- a. Relatively short term life tests (e.g., 1,000 hours) are sufficient to assure diode reliability for long time use.
- b. Diode quality can be enhanced by the use of screening procedures to eliminate workmanship failures.

Item b above is best exemplified by the TXV process indicated in Figure 8 and the JANS process as found in Figure 9 in accordance with MIL-S-19500, General Specification for Semiconductor Devices.

The highest level of military product assurance is JANS (grade 1). The screens and process controls are of the latest technology. The product assurance level of JANS has been established for devices that are intended for space applications.

4.1.6.11 Power derating. The objective of power derating is to hold the worst-case junction temperature to a value lower than the normal permissible rating. The typical diode specification for thermal derating expresses the change in junction temperature with power for the worst case of the devices. The actual temperature rise per unit of power will be considerably less, but is not a value which can be readily determined for each unit. The user should refer to MIL-STD-975 for derating factor guidelines.

4.1.6.11.1 Junction temperature derating. Junction temperature derating requires that the ambient or case temperature for the part not be exceeded in the application.

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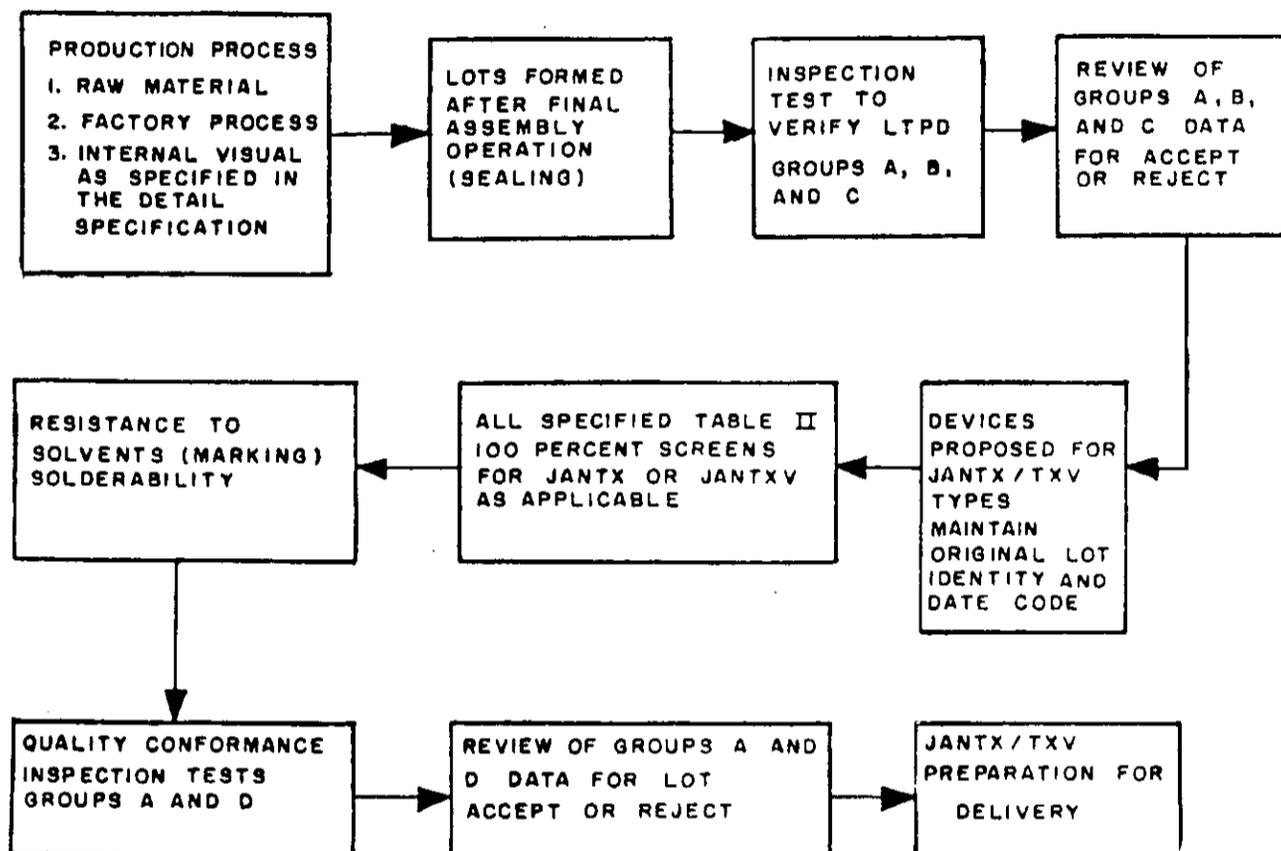


FIGURE 8. Order of procedure diagram for JANTXV devices.

The ambient temperature for a device that does not have some means of thermal connection to a mounting surface, includes the temperature rise due to the device, adjacent devices, and any heating effect which can be encountered in service. For space applications, the junction temperature should be limited to +125 °C maximum.

4.1.6.11.2 Thermal resistivity to air. The thermal resistivity to air is expressed in °C/W (or mW) or its reciprocal derating factor, which is usually expressed in mW/°C. These terms are based on the following assumptions:

- a. The device is mounted in air at normal atmospheric pressures
- b. Air is free to circulate by normal convection
- c. The device is attached by its leads to an infinite heat sink at a controlled distance from the device

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- d. The ability of the device to dissipate heat will be decreased by:
 - 1. Longer leads
 - 2. Restricted air circulation
 - 3. Radiant heat reflection from adjacent areas.
- e. The ability of the device to dissipate heat will be increased by:
 - 1. Increase in air flow
 - 2. Increase in air pressure
 - 3. Decrease in the lead length
 - 4. Surrounding the device with potting compounds having a thermal conductivity greater than air.

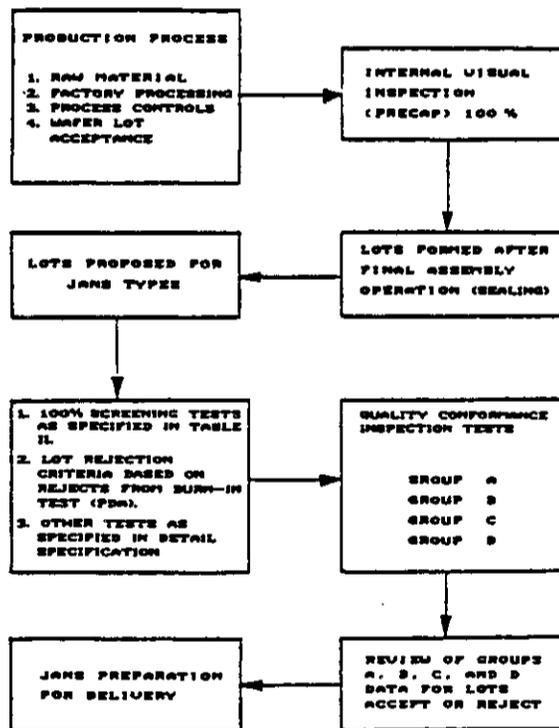


FIGURE 9. Order of procedure diagram for JANS devices.

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4.1.6.11.3 Maximum allowable power for a device rated to air. This is established by controlling the thermal conductance of the medium surrounding the device under operational conditions so as to equal or exceed the normal conductance of air at one foot per second velocity.

The power may be determined from the following equations:

$$P_w(\text{max}) = \frac{T_{jo} - T_a}{\theta_{JA}}$$

where

$P_w(\text{max})$ = Maximum allowable derated power
 T_{jo} = Derated junction temperature from derating table
 T_a = The temperature immediately surrounding the semiconductor in the most severe operating condition with all adjacent heat producing devices in operation
 θ_{JA} = Manufacturer's thermal resistivity rating.

Where thermal conductance is expressed as a derating factor or 1/thermal resistivity, the above equation becomes:

$$P_w = (T_{jo} - T_a) (\text{Derating factor in air})$$

4.1.6.11.4 Maximum power for device rated to case temperature. This is established as follows:

- a. Devices which have an established means for making thermal connection to a heat sink are rated for a case temperature instead of air temperature. The case temperature can be controlled by the derated power. The thermal resistivity, junction to case (θ_{JC}), or the thermal derating factor, is used to determine acceptable power for a maximum junction temperature when the case temperature is controlled.
- b. Maximum power is determined from the case temperature of the device measured under the most severe operating conditions. The equation to be used is:

$$P_w(\text{max}) = \frac{T_{jo} - T_c}{\theta_{JC}}$$

where

θ_{JC} = Thermal resistance of the device rated to case
 T_c = Measured case temperature
 T_{jo} = Derated junction temperature.

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4.1.6.12 Voltage derating. The voltage rating of a device can vary with temperature, frequency, or bias condition. The rated voltage is the voltage compensated for all factors determined from the manufacturer's data sheet. The reliability derating consists of the application of a percentage figure to the voltage determined from all factors of the rating. For space applications voltages should be derated to 50 percent of the limits specified on the device data sheets. The user should refer to MIL-STD-975 for derating factor guidelines.

Instantaneous peak voltage derating. This derating is the most important and least understood derating and is required to protect the device against the high voltage transient spike of voltages which can occur as a result of magnetic energy stored in inductors, transformers, or relay coils. Transient spikes also can result from momentary unstable conditions which cause high amplitude oscillation during switching turn on or turn off.

Transient spike or oscillating conditions due to the discharge of leakage or static electricity will cause minute breakdown of surfaces within the body of the semiconductor. The minute breakdown may not cause failure but can cause a substantial increase in the probability of failure during service.

Circuit clamping by zener diodes, transient suppressors, or resistive voltage dividers, and current limiting by means of limiting resistors are required to maintain instantaneous transient voltages within the allowable limits.

Continuous peak voltage is the voltage at the peak of any signal or continuous condition which is a normal part of the design conditions. A continuous peak voltage is the highest voltage which can be observed on an oscilloscope under any normal operating condition.

Design maximum voltage is the highest average voltage. This is essentially the dc voltage as read by a dc meter. The ac signals can be superimposed on the dc voltage allowing a higher peak voltage providing the continuous peak voltage is not exceeded.

4.1.6.13 Selecting diodes. A number of factors must be considered in the choice of a diode type when reliability is of prime importance. These are the following:

- a. The reliability of the device under consideration should be proven. New diode types with better electrical characteristics are constantly being announced. There is too often a tendency on the part of circuit designers to select these devices because of their high performance capabilities. It must be remembered that it takes time to adequately prove the reliability of a diode, and that generally the reliability of newer types of diodes has not been verified to the extent of older types.

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- b. The diode should have been in production long enough for any problems that may adversely affect reliability to have been eliminated. Early in the production phase of a diode type, major emphasis is often given to process improvement to optimize electrical characteristics. As the production process and yields improve, reliability will generally also improve.
- c. The diode type under consideration should be a major portion of the manufacturer's yield. A characteristic of the semiconductor industry has been that a number of diode types of varying electrical characteristics are simultaneously produced on the same line. As manufacturing experience is gained, the process can be adjusted to optimize production of the most desired types. However, it is often true that a diode type which represents a small percentage of the yield of a production line may have some abnormality which will make its reliability different from the majority of the line output.

4.1.6.14 Circuit design considerations. The selection of the most reliable diode to perform the required function is basic, but is only the first step in assuring reliable circuit operation. Several circuit design considerations to assure reliable diode performance follow:

- a. When possible, circuit performance should be based upon the most stable diode parameters.
- b. Realistic limits for component variations due to tolerance, temperature, and time should be used. Wider limits must be allowed for characteristics which are less stable than those which show good stability with life.
- c. Circuit design dependent upon diode characteristics that are uncontrolled can lead to poor reliability and should be avoided. If circuit performance is dependent upon unspecified diode characteristics and thus not controlled, there is no assurance that subsequent production will have consistent characteristics.
- d. The use of derated operating conditions can be a factor to secure reliable circuit performance. The conditions to be derated and the amount of derating must be carefully determined to insure reliable circuit operation and still maintain required performance. Circuit performance and reliability, in this sense, compromise each other.
- e. The environment which the diode, circuit, and system encounters during assembly, testing, and use must be controlled to assure maximum reliability.

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4.1 DIODES, GENERAL

4.1.6.15 Radiation considerations. To ensure that a circuit functions properly in space applications, the design engineer must consider the effects of radiation exposure. Survivability requirements for systems that will encounter a radiation environment add to the challenge. The design engineer must know how radiation affects his circuit and components. When semiconductor devices are exposed to radiation environments, changes occur in their rated electrical parameters. The magnitude of the changes is a function of such things as the type of radiation, neutron or gamma rays, and time or duration. Generally, permanent damage is associated with displacement effects resulting from neutron radiation, and transient effects are the result of ionization from gamma ray radiation. The user should refer to MIL-STD-19500 for the four levels (M, D, R, and H) of radiation hardness assurance (RHA) and requirements. The letters M, D, R, and H following the S or TXV portion of the JAN prefix indicate the level of RHA. Parts without an RHA prefix either have not been tested for RHA or have not passed the testing.

Diodes exposed to radiation will display changes in the following electrical characteristics:

- a. Breakdown voltage increases
- b. Leakage current increases
- c. Forward voltage increases
- d. Reverse recovery time.

Each of these changes in the electrical characteristics of a diode is a result of radiation-induced changes in the semiconductor material itself. The increase in breakdown voltage can be an improvement, except for zeners, of the electrical characteristics over the value prior to radiation. This change is the result of an increase in the bulk resistivity of the base region of the diode. These increases in leakage current and forward voltage are undesirable electrical characteristic changes. When designing for the use of zener diodes, the design engineer must take into consideration the fact that the zener voltage will change. Zener diodes with a $V_z > 10$ V will have an increase in breakdown voltage as a function of neutron fluences, whereas those with a $V_z < 10$ V will have a decrease in breakdown voltage. These increases or decreases are caused by the two mechanisms involved in the breakdown. The low reverse breakdown voltage devices exhibit true zener internal field emission breakdown whereas the higher voltage devices break down as a result of the avalanche process. The amount of change or effect on a diode device is also dependent on the manufacturer's processes and the technology used. This varies from vendor to vendor and device to device. It is recommended that, for critical applications, the manufacturer be contacted and the specifics be discussed or resolved. Consideration should also be given to the type and amount of material shielding required for a given application.

4.2 DIODES, SWITCHING

4.2 Switching

4.2.1 Introduction. Switching diodes are normally made of silicon or germanium. The silicon diode has a higher forward voltage drop than the germanium diode; however, the silicon device has lower leakage currents and higher maximum operating junction temperatures. Silicon devices exhibit normal rectifying characteristics and perform the rectifying function very rapidly. They are classified according to the general groups below:

- a. Ultrahigh speed switching types such as the 1N4148-1 with speeds of 1.0 to 50 ns
- b. Medium speed types in the range of 50 to 150 ns (1N5417)
- c. Standard switching types of 150 ns and up (1N645-1).

4.2.2 Usual applications. Switching diodes are useful in many types of computer circuits. They can perform a variety of functions in a particular circuit; specifically, clamping, saturation prevention, logic, and blocking. Care should be taken to make the forward current pulse as short in duration as possible. The magnitude of the forward current should be as small as the circuit application will allow. Some consideration also should be given to providing the maximum attainable reverse current pulse while holding the impedance of the external circuit to a minimum. Finally, the temperature at which the device operates should be as low as possible. Examples of some switching diode applications follow.

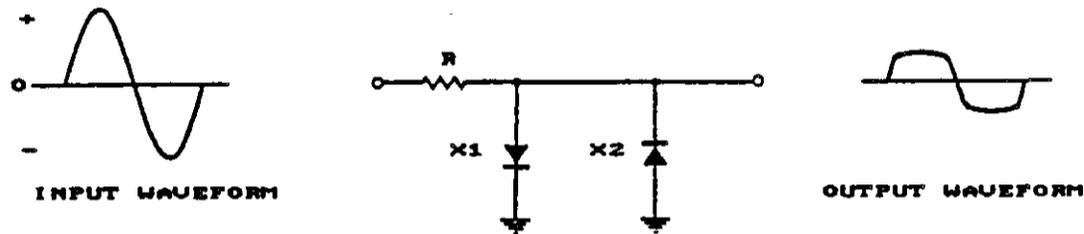
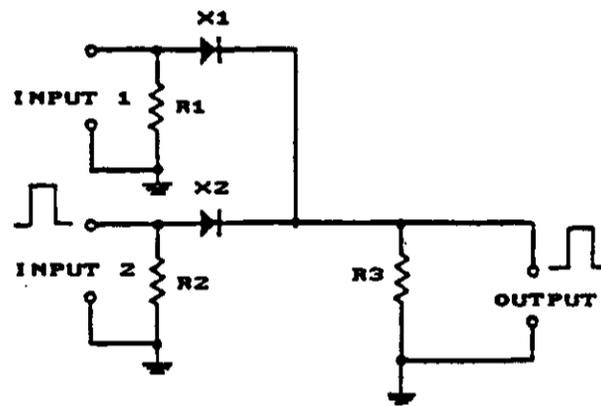
4.2.2.1 Clamping circuit. A clamping circuit is used to limit voltage pulses passing through the circuit. A simple clamping circuit is shown in Figure 10. When the magnitude of the input waveform is less than the threshold voltage, the diodes do not conduct. As the input signal becomes larger than the threshold voltage, the diodes begin to conduct. Diode X1 clamps the voltage during the positive cycle of the input, and diode X2 clamps during the negative cycle.

The resistance of the diodes and the series resistor form a voltage divider circuit which attenuates the signal. As the input signal and diode current increases, the diode resistance decreases, varying the voltage divider ratio to keep the output voltage near the threshold voltage. This variation in output voltage is called "soft limiting."

4.2.2.2 Logic circuit. Diodes may be used to implement Boolean logic functions. Of these, the OR (+) function is one of the simplest logic functions. In Figure 11 the function $Y = A + B$ is implemented. The circuit will produce a positive output pulse across common-load resistor R3 if there is a positive pulse at any one of the inputs; that is if A or B is energized. The reason is that the input pulse forward biases the diode, and the resulting pulse current develops a pulse voltage drop across R3.

The resistance of input resistors R1 and R2 and output resistor R3 will depend on the diode characteristics and the characteristics of the driver and load circuits.

4.2 DIODES, SWITCHING

FIGURE 10. Clamping circuit.FIGURE 11. Logic circuit.

4.2.2.3 Saturation prevention circuit. When a transistor is used in a switching application, the transistor is driven from cutoff to saturation, the inherent storage time of the transistor limits the time required for it to turn off. This problem can be minimized by using a diode to keep the transistor in the operating region, thus preventing it from going into saturation. In Figure 12, the diode is used to reduce the turnoff time of the transistor. The transistor base voltage, 0.6 V for silicon transistors, is the anode voltage for the diode. If the diode is silicon, it will start to conduct when the voltage drop reaches about 0.35 V. Any additional reduction in collector voltage will turn the diode on harder and prevent the transistor from going into hard saturation.

4.2 DIODES, SWITCHING

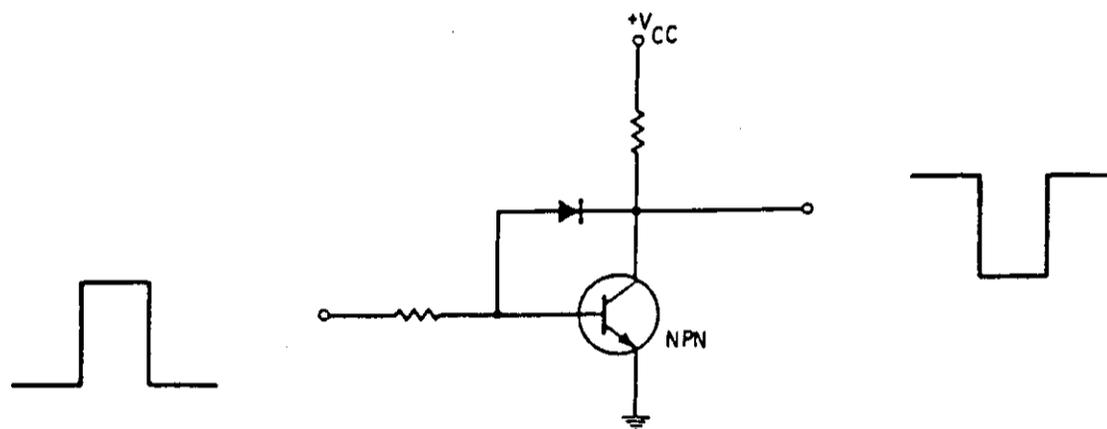
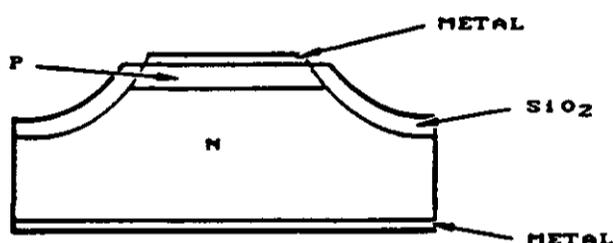


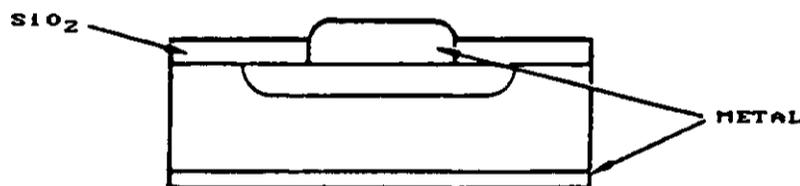
FIGURE 12. Saturation prevention circuit.

4.2.3 Physical construction. The axial lead switching diode generally utilizes a passivated diffused or planar process die in a glass-to-metal or double-slug metallurgical bond construction. The two basic die processes are shown in Figure 13 and 14.



METALLIZE MESA & BACK TO FORM CONTACTS

FIGURE 13. Passivated diffused process.



METALLIZATION IN HOLD AND ON BACK TO FORM CONTACT

FIGURE 14. Planar process.

4.2.3.1 Back contact. Illustrated in Figure 15 below are three back contact techniques used in constructing axial lead switching diodes. Although the alloyed back contact is the strongest, the high-temperature alloyed back (associated with the double-slug construction) is most commonly used.

4.2 DIODES, SWITCHING

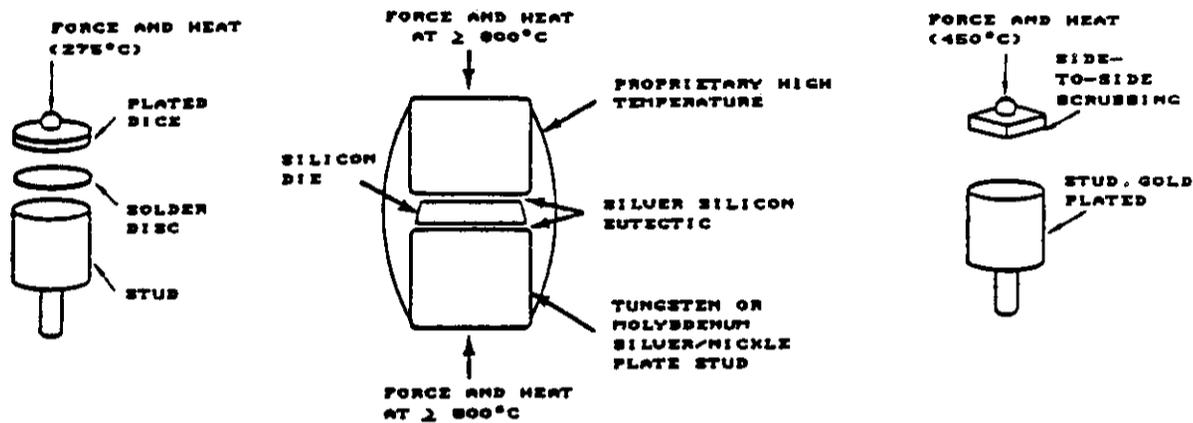


FIGURE 15. High-temperature alloyed back contact.

The solder contact has good mechanical strength and thermal conductivity. The silicon-silver alloy contact has good mechanical strength and thermal conductivity. The silicon chip is metallurgically brazed to plated molybdenum or tungsten slugs at temperatures greater than 700 °C. The silicon-gold alloy contact is the strongest and most thermally conductive of the above techniques. Because of the high melting temperature of the silicon-gold eutectic alloy (377 °C), the contact will withstand high temperatures during overload.

Molybdenum, tungsten, and dumet are usually used for studs in glass packages because their thermal expansion coefficients match those of certain glasses, allowing sealing of these glasses to them. Copper and copper alloys are usually used in devices which handle appreciable power because they have much better thermal conductivity than molybdenum and dumet.

4.2.3.2 Front contact. Electrical contact is made to the die in three ways as shown in Figure 16. Although three contact techniques are mentioned, the stud or lead contact associated with the double-slug construction is the most common and reliable.

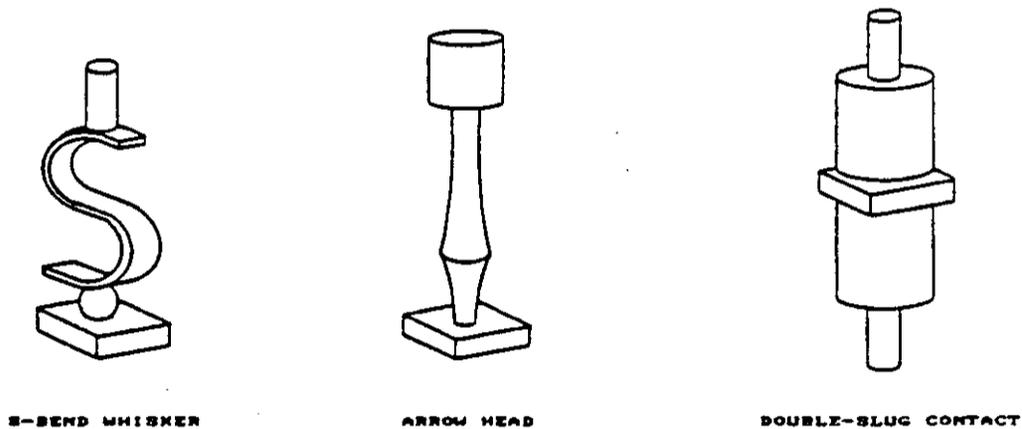


FIGURE 16. Various styles of electrical contact.

4.2 DIODES, SWITCHING

The whisker is held under compression in contact with the top of the die. In this way, electrical contact is made to the die while the force on it is limited by the spring constant of the whisker. The arrowhead contact is made using solder between the wire and the top of the die. The stud contact is usually soldered to the top of a diffused die, making good electrical and mechanical contact but not limiting the stress on the die.

The whisker contact has good shock and vibration tolerance. The arrowhead contact is somewhat superior due to its lower mass due to imbedding of the end of the contact in the metal on top of the die. The whisker structures provide good mechanical isolation of the die from mechanical strain induced by lead flexure and thermal expansion and contraction of the package.

The double-slug contact transmits more stress to the die than the whisker structure, but is used in spite of this because it provides good heat transfer, thus raising the power rating of the package.

4.2.3.3 Seals. Generally, two sealing techniques are used: metal-to-glass or double-slug construction. The package in Figure 17 uses glass-to-metal seals. These glass-to-metal seals are formed prior to final sealing and the final seal occurs by fusing glass-to-glass at the point noted in the figure.

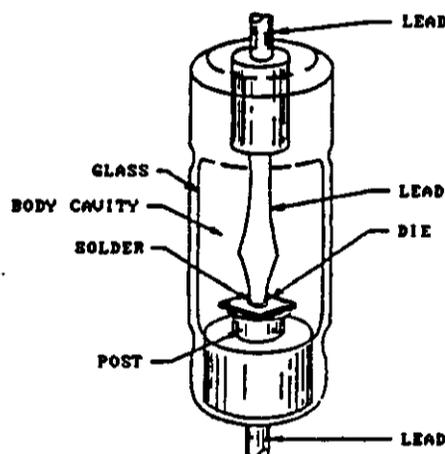


FIGURE 17. Glass-to-metal construction.

The double-slug package in Figure 18 is made by metallurgically bonding both sides of the die to either molybdenum or tungsten slugs to which a glass sleeve or glass slurry is fused for final sealing. The final step is a lead-brazing operation. This double-slug construction is most commonly used for producing axial leaded switching diodes for military applications.

4.2 DIODES, SWITCHING

Figure 19 shows a typical switching diode package.

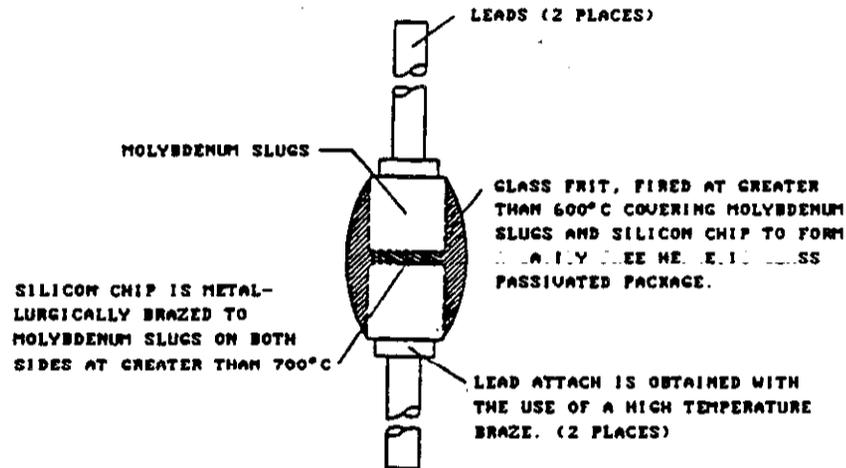


FIGURE 18. Typical double-slug construction.

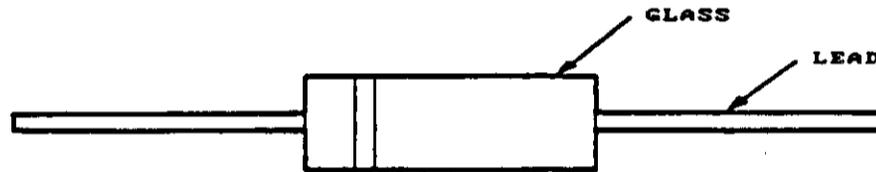
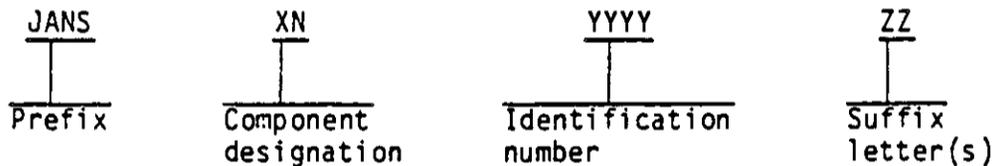


FIGURE 19. Outline drawing of a typical switching diode package.

4.2.4 Military designation. The military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

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4.2 DIODES, SWITCHING

4.2.5 Electrical characteristics. Table IV shows electrical ratings for various switching diodes. The values are for general reference only. Specific values should be found in the applicable MIL-S-19500 reference sheets.

TABLE IV. Typical electrical ratings for switching diodes

Device Type	V _R (V)	V _F (V)	I _F (mA)	Time t _{rr} maximum (ns)	19500 Slash Sheet
1N645-1	225	1.0	400	-	240
1N647-1	400	1.0	400	-	240
1N649-1	600	1.0	400	-	240
1N4148-1	75	1.0	10	5	116
1N5712	16	1.0	35	-	445

NOTE: This table is not to be used for part selection. Use MIL-STD-975 for that purpose.

4.2.5.1 Junction capacitance. In the p-n semiconductor switching diode, there are two capacitive effects to be considered. Both types of capacitance are present in the forward- and reverse-bias regions, but one so outweighs the other in each region that only that one is considered in each region. In the reverse-bias region there is the transition- or depletion-region capacitance (C_T), while in the forward-bias region there is the diffusion (C_D) or storage capacitance.

The basic equation for the capacitance of a parallel plate capacitor is defined by $C = \epsilon A/d$, where ϵ is the permittivity of the dielectric (insulator) between the plates of area A separated by a distance d. In the reverse-bias region there is a depletion region (free of carriers) that behaves essentially like an insulator between the layers of opposite charge. Because the depletion region will increase with increased reverse-bias potential, the resulting transition capacitance will decrease, as shown in Figure 20. The fact that the capacitance is dependent on the applied reverse-bias potential has application in a number of electronic systems.

4.2 DIODES, SWITCHING

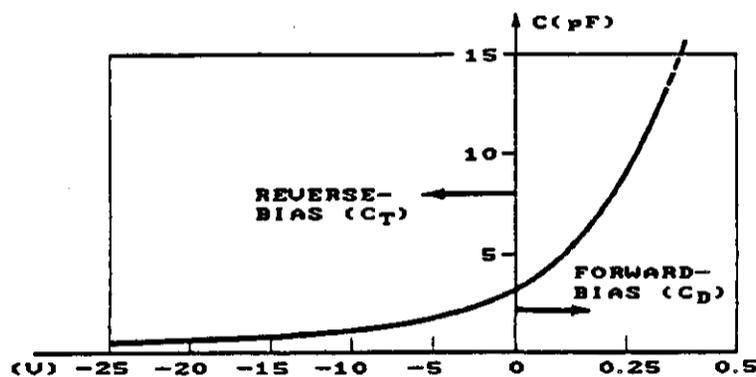


FIGURE 20. Transition and diffusion capacitance vs applied bias for a silicon diode.

Although the effect described above will also be present in the forward-bias region, it is overshadowed by a capacitance effect directly dependent on the rate at which charge is injected into the regions just outside the depletion region.

In other words, the capacitance effect is directly dependent on the resulting current of the diode. Increased levels of current will result in increased levels of diffusion capacitance. However, increased levels of current result in reduced levels of associated resistance, and the resulting time constant ($\tau = RC$), which is very important in high-speed applications, does not become excessive.

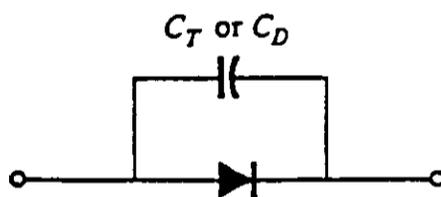


FIGURE 21. Including the effect of the transition or diffusion capacitance on the semiconductor diode.

The capacitive effects described above are represented by a capacitor in parallel with the ideal diode, as shown in Figure 21. For low- or midfrequency applications (except in the power area), however, the capacitor is normally not included in the diode symbol.

4.2 DIODES, SWITCHING

4.2.5.2 Reverse recovery time. In the forward-bias state, it has been shown in an earlier section that there are a large number of electrons from the n-type material progressing through the p-type material and a large number of holes in the n-type--a requirement for conduction. The electrons in the p-type and of holes progressing through the n-type material establish a large number of minority carriers in each material. If the applied voltage should be reversed to establish a reverse-bias situation, ideally the diode will change instantaneously from the conduction state to the nonconduction state. However, because of the large number of minority carriers in each material, the diode will simply reverse as shown in Figure 22.

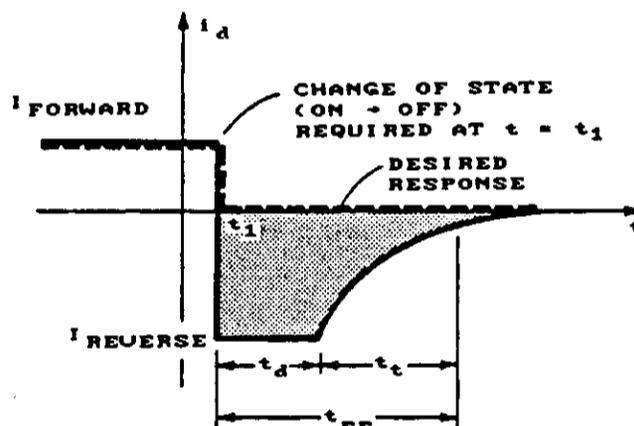


FIGURE 22. Defining the reverse recovery time.

Thus, the diode will stay at this measurable level for the period of time t_d (storage time) required for the minority carriers to return to their majority-carrier state in the opposite material. Eventually, when this storage phase has passed, the current will reduce in level to that associated with the non-conduction state. This second period of time is denoted by t_t (transition interval). The reverse recovery time is the sum of these two intervals: $t_{rr} = t_d + t_t$. Naturally, it is an important consideration in high-speed switching applications. Most industrially available switching diodes have a t_{rr} in the range of a few nanoseconds to 1 μ s.

4.2.6 Environmental considerations. Typical environmental conditions and screening tests that switching diodes are capable of withstanding are not substantially different from those given in paragraphs 4.1.6.9 Environmental considerations, and 4.1.6.10 Screening procedures in subsection 4.1 Diodes, General. For the specific device selected consult the applicable MIL-S-19500 reference slash sheets.

4.2 DIODES, SWITCHING

4.2.7 Reliability considerations.

4.2.7.1 Failure modes. Predominant failure modes of the switching diode are found in the reverse-current and forward-voltage performance parameters. The failure mechanisms such as (1) mechanical stress, (2) thermal stress, and (3) operational life tests, account for the more significant incidences of failure, particularly the variable frequency exposure and longterm life tests. As noted previously, the TXV process with attendant V_f and I_R parameters should indicate any device that is beginning to show gradual deterioration. Other reliability considerations in device design are presented in the recommended approach to general diode construction, given in paragraph 4.1.6 Diodes, General.

4.2.7.2 Derating. History has shown that the largest single cause of diode failure is operation above allowable levels of electrical and thermal stress. Accordingly, it is imperative that derating of parts be performed to increase the reliability of electronic systems. Users should refer to MIL-STD-975 for derating factor guidelines. The design and use of a part should always be within the thermal and electrical stresses defined. High temperature operation is the most destructive stress for a semiconductor. It will result in electrical parameter drift, and a general degradation of the electrical and mechanical characteristics of the device.

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**4.3 DIODES, RECTIFIERS AND
POWER DIODES**

4.3 Rectifiers and power diodes.

4.3.1 Introduction. This section includes three general types of rectifiers: power diodes, fast switching power rectifiers, and power Schottky rectifiers. This general group does not include all available rectifiers but includes the most commonly used types.

4.3.1.1 Power diodes. The devices in this series are general purpose low-voltage rectifiers available in both axial-leaded and stud-mounted packages. Individual dc output voltages range from 50 to 1000 V. Maximum dc output currents are commonly supplied in values ranging from 1 to 6 amperes for axial-leaded packages and as high as 125 A for stud-mounted packages.

4.3.1.2 Fast switching power rectifiers. These rectifiers, with reverse voltages ranging from 50 to 1000 V and dc output currents of 1 to 300 A, have fast recovery times that are typically around 150 ns. Maximum dc output current ratings range from 1 to 6 amperes for axial-leaded packages and up to 300 A for studmounted packages and other high current packages.

4.3.1.3 Power Schottky rectifiers. These rectifiers are lower voltage devices with reverse voltages ranging from 15 V to 100 V and output currents up to 200 A. Schottky rectifiers come in axial-leaded, dual TO-3, and stud-mounted packages. Because they are majority carrier devices, their reverse recovery time is insignificant, which makes them excellent devices for high frequency applications.

4.3.2 Usual applications.

4.3.2.1 Power diodes. These devices are ideally suited for applications requiring rectification in the 1 to 125 amperes region, where low power loss and minimum physical size are important. In addition to breakdown voltage, and forward voltage, forward-surge current, and peak-reverse power, are of importance when selecting a power diode. A high-junction temperature and extreme low forward voltage drop and thermal impedance permit high current operation with minimum space requirements. Some types are available with negative polarity for use in bridge circuits or circuits that need a negative heat sink in half-wave and center-tap applications. Consideration should be given to absolute maximum ratings and derating characteristics.

4.3.2.2 Fast switching power rectifiers. This type of rectifier exhibits fast recovery time, typically below 150 ns, and is suitable for use in such circuit applications as inverters, choppers, low rf interference, free-wheeling rectifiers, and dc power supplies. In most of these applications, dissipation of heat is important, and therefore circuit design should take into consideration the specific device derating characteristics. See paragraph 4.3.7.

Consideration should also be given to the performance of the reverse recovery time, t_{rr} , at elevated temperatures because most of the existing MIL-S-19500 rectifier specifications specify t_{rr} at ambient temperature only.

4.3 DIODES, RECTIFIERS AND POWER DIODES

Evaluation of fast-switching power rectifiers on t_{rr} at high temperatures has shown a large variance in the percent change of t_{rr} from ambient to elevated temperatures among manufacturers. Because higher t_{rr} at elevated temperatures increases power dissipation losses and can possibly cause a timing problem in certain applications, the designer should evaluate the performance of t_{rr} in the circuit at the operating temperature of the application.

A typical reverse recovery time characteristic of a rectifier is shown in Figure 23.

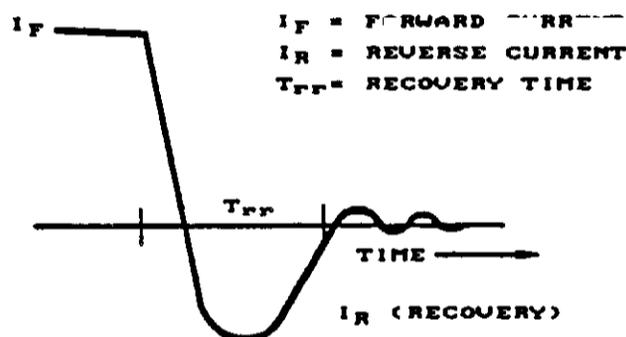


FIGURE 23. Reverse recovery time of a typical fast switching power rectifier.

4.3.2.3 Power Schottky rectifiers. The Schottky rectifier has a metal barrier type junction rather than the diffused pn-junction of standard power diodes. The most commonly used barrier metals or alloys are chromium, platinum, nickel platinum, molybdenum, and tungsten. The chromium barrier provides low-forward voltage with a very high-leakage current. However, the tungsten barrier provides low-leakage current with high-forward voltage. Because efficiency is a major consideration, the nickel platinum barrier provides the best choice due to its low-forward drop with a minimum of leakage current. The Schottky rectifier has two characteristics that make it useful as a rectifier in switching power supply circuits: (1) the low-forward voltage drop, 0.3 V, results in considerably less heat dissipation, compared with pn-junction rectifiers; therefore, it has a higher operating efficiency in the circuit (see Figure 24), and (2) the fast recovery time is also a significant advantage. The reverse recovery time of the power Schottky is generally less than 10 ns, compared with 150 ns for a standard fast recovery pn-junction diode.

The Schottky rectifier has some disadvantages which should also be considered. The most significant disadvantage is the reverse bias electrical characteristic. The peak reverse voltage is lower than that of the pn-junction diode, which limits its use to low voltage applications, and the reverse leakage current is higher than the pn types. The second disadvantage is that Schottky rectifiers have a lower operating temperature.

4.3 DIODES, RECTIFIERS AND POWER DIODES

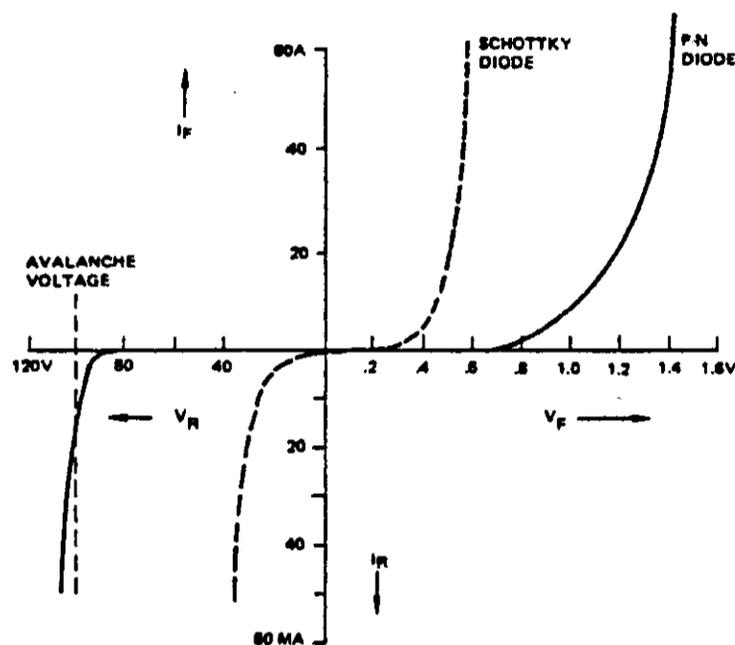


FIGURE 24. PN junction and Schottky rectifier I/V characteristics.

4.3.3 Physical construction. The rectifier and power rectifier are generally made using a passivated diffused or planar process whereas the power Schottky rectifier is made with a metal barrier Schottky die. Rectifiers commonly are of glass-to-metal or double-slug construction whereas high-power rectifiers and high-power Schottkies use a stud-mounted package. The three basic die processes are shown in Figures 25, 26, and 27.

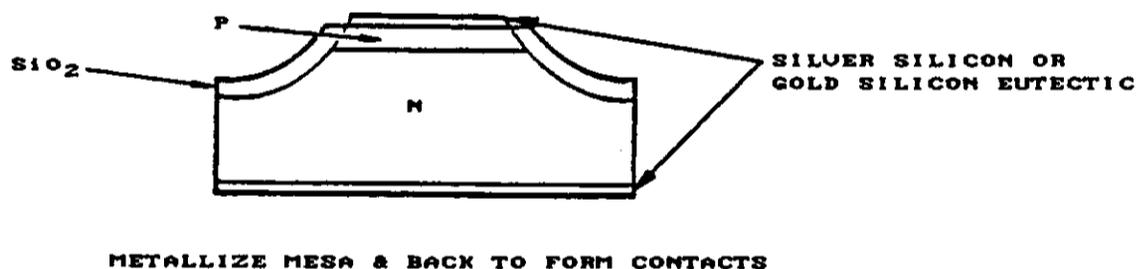


FIGURE 25. Passivated diffused process.

**4.3 DIODES, RECTIFIERS AND
POWER DIODES**

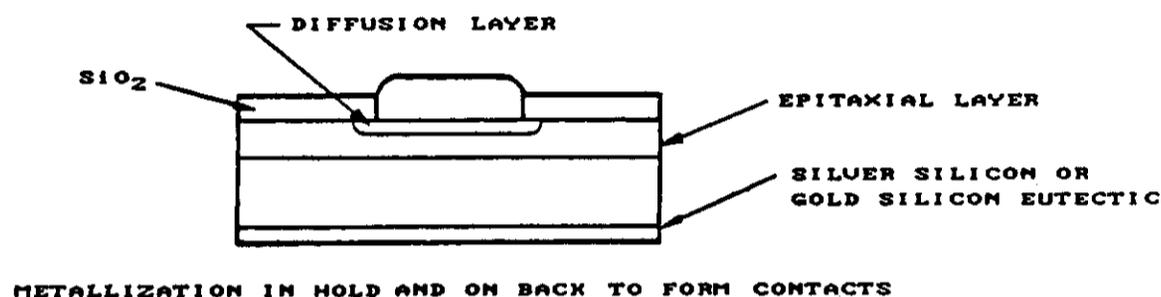


FIGURE 26. Planar process.

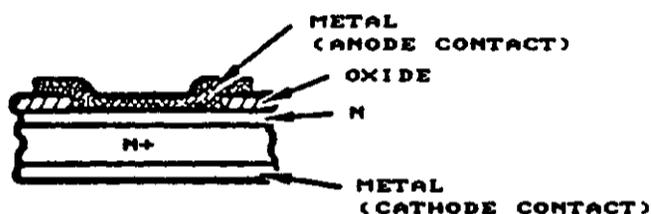


FIGURE 27. Metal barrier Schottky junction.

4.3.3.1 Back contacts. Illustrated in Figure 28 below are three back contact techniques used in constructing rectifier, power, and Schottky diodes. Although the alloyed back contact is the strongest contact, rectifier and power diodes generally made with high temperature alloy and solder back contacts, respectively, for high reliability applications.

The solder contact has good mechanical strength and thermal conductivity. The silicon-silver alloy contact, which is generally associated with a high temperature alloy, has good mechanical strength and thermal conductivity. The silicon chip is metallurgically brazed to plated molybdenum or tungsten slugs at temperatures exceeding 700 °C. The silicon-gold alloy contact is the strongest and most thermally conductive of the above techniques. Because of the high melting temperature of the silicon-gold eutectic alloy (377 °C), the contact will withstand high temperatures during overload.

Molybdenum, tungsten, and dumet are usually used for studs in glass packages because their thermal expansion coefficients match those of certain glasses for sealing. Copper and copper alloys are usually used in devices that handle appreciable power because they have much better thermal conductivity than molybdenum, tungsten, or dumet.

4.3 DIODES, RECTIFIERS AND POWER DIODES

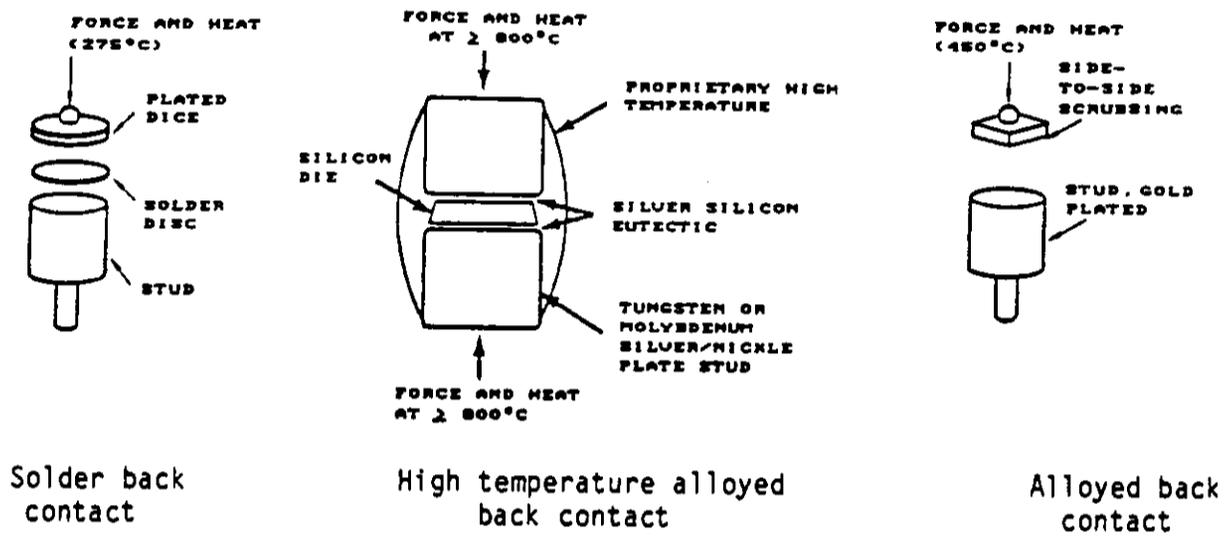


FIGURE 28. Back contact formation techniques.

4.3.3.2 Front contacts. Electrical contact is made to the die in three ways as shown in Figure 29. Although three contact techniques are mentioned, the stud construction is the most common and reliable.

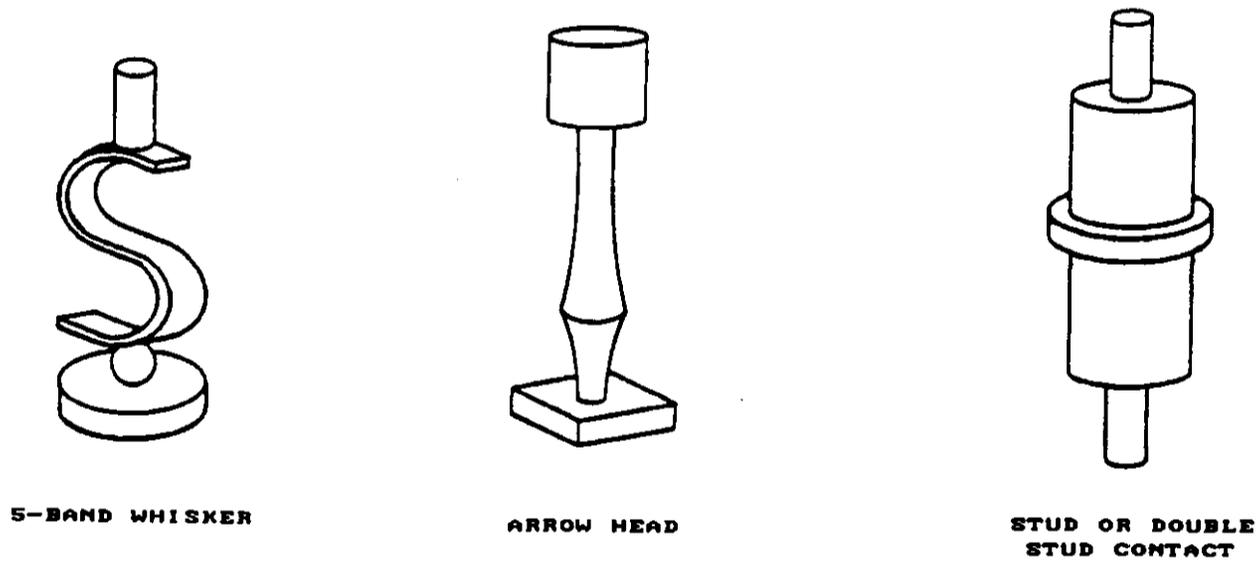


FIGURE 29. Three various styles of electrical contact.

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4.3 DIODES, RECTIFIERS AND POWER DIODES

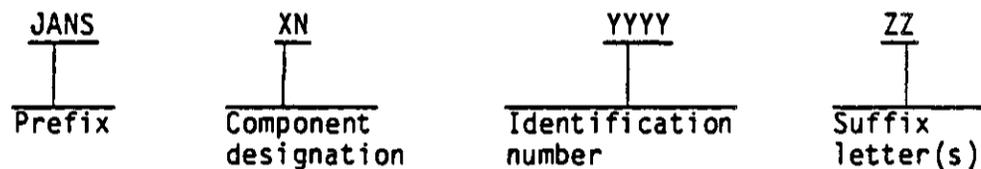
The whisker is held under compression in contact with the top of the die. In this way, electrical contact is made to the die while the force on it is limited by the spring constant of the whisker. The arrowhead contact is made using solder between the end of the wire and the top of the die. The stud contact is usually soldered to the top of a diffused die which makes good electrical and mechanical contact but does not limit the stress on the die.

The whisker contact has good shock and vibration tolerance. The arrowhead contact is somewhat superior due to the imbedding of the end of the contact in the metal on top of the die which prevents sliding of the contact. Both whisker structures provide good mechanical isolation of the die from mechanical strain induced by lead flexure and thermal expansion and contraction of the package.

The stud and double-slug contact transmits more stress to the die than the whisker structure but is used in spite of this because it provides good heat transfer, thus raising the power rating of the package.

4.3.3.3 Seals. Generally, three sealing techniques are used: metal-to-glass, double-slug, and stud mounted. The package (A) in Figure 30 uses glass-to-metal seals. These glass-to-metal seals are formed prior to final sealing. The final seal occurs by fusing glass-to-glass at the point noted in the figure. The double-slug package (B) in Figure 30 is made by metallurgically bonding both sides of the die to either molybdenum or tungsten slugs to which a glass sleeve or glass slurry is fused for final sealing. The final step is a lead-brazing operation. This type of construction is most often associated with rectifier diodes. The stud-mounted package (C) assembly sequence illustrated in Figure 30 has solder-back and band-front contacts. This construction is most often associated with power rectifiers and power Schottky rectifiers. Figures 31 32 show a typical double-slug rectifier and power diode package, respectively.

4.3.4 Military designation. Military designation for diodes is formulated as follows:



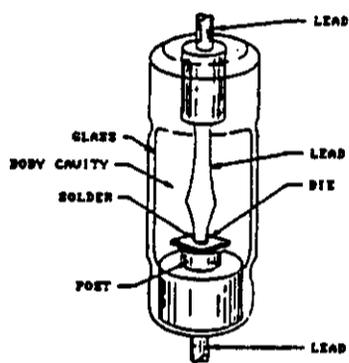
The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

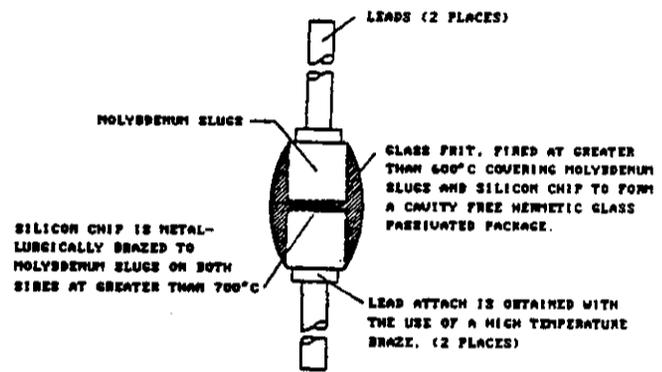
The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

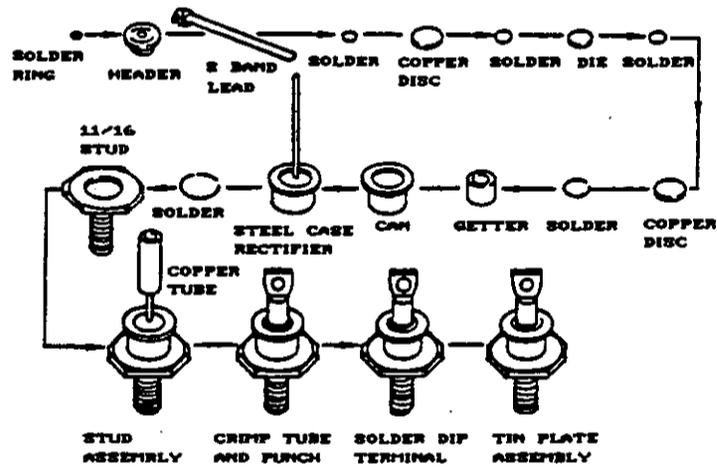
4.3 DIODES, RECTIFIERS AND POWER DIODES



A. Glass-to-metal



B. Double-slug



C. Stud

FIGURE 30. Package sealing techniques.

**4.3 DIODES, RECTIFIERS AND
POWER DIODES**

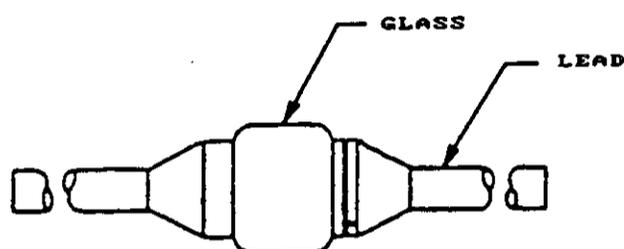


FIGURE 31. Outer drawing of a typical double-slug rectifier.

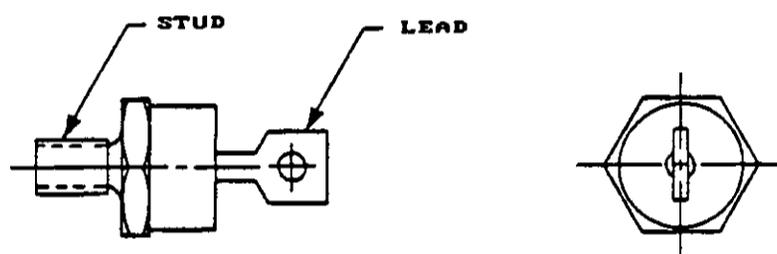


FIGURE 32. Outer drawing of a typical power diode.

4.3.5 Electrical characteristics.

4.3.5.1 Power diodes and switching rectifiers. Table V illustrates the maximum ratings of typical power diodes and switching rectifiers. Average output (rectified) current is referred to as I_o and surge current refers to the maximum current pulse that can be carried by the device for the length of time, repetition frequency, and waveform specified. Basic characteristics and parameter information on diodes and rectifiers are discussed in the "Diodes, General" section.

TABLE V. Power diode and switching rectifier maximum ratings

Device Type	V_R	I_o $T_A = 55^\circ C$	I_f (Surge) 1/120 Sec.	MIL-S-19500 Slash Sheet
1N5415	50V	3A	80A	411
1N5416	100V	3A	80A	411
1N5417	200V	3A	80A	411
1N5418	400V	3A	80A	411
1N5419	500V	3A	80A	411
1N5420	600V	3A	80A	411
1N5615	200V	1A	25A	429
1N5617	400V	1A	25A	429
1N5619	600V	1A	25A	429
1N5621	800V	1A	25A	429
1N5623	1000V	1A	25A	429

NOTE: This table is not to be used for part selection, use MIL-STD-975.

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4.3 DIODES, RECTIFIERS AND POWER DIODES

The operating characteristics shown above are for general reference only. For any given type of device, refer to applicable MIL-S-19500 slash sheet.

4.3.5.2 Power Schottky rectifiers. For electrical characteristics on Schottky rectifiers refer to applicable MIL-S-19500 slash sheet.

4.3.6 Environmental considerations. Typical environmental conditions and screens which rectifier diodes are capable of withstanding are not substantially different from those given in paragraphs 4.1.6.9 Environmental considerations and 4.1.6.10 Screening procedures in subsection 4.1, Diodes, general. For the specific device selected, consult the applicable MIL-S-19500 reference slash sheet.

4.3.7 Reliability considerations.

4.3.7.1 Power diodes and fast switching rectifiers. Fast switching power rectifiers, as with other silicon diodes of this type, may exhibit localized mechanical weakness because of a large junction area, due to either device die and material design including ohmic contact metallurgy and package design, or defects and flaws that result from the fabrication process itself.

The former type of weakness would include the use of lead-rich solder for either or both anode and cathode solder ohmic contact regions. The resulting recrystallization (plastic deformation) of these regions results in progressively higher $R_{\theta JC}$ (thermal impedance) values, leading to higher junction temperatures that will permanently degrade and eventually destroy the device. The use of the standard gold-silicon eutectic alloy or the gold-germanium eutectic alloy will eliminate this failure phenomenon. If a lead-rich solder is used for a device contact, a power cycling test under applicable system conditions should be used to determine the thermal cycle capability of the device.

The purpose of high temperature reverse bias and power burn-in tests, as specified in the JANTEXV or JANS specifications, is to eliminate defective devices caused by surface ionic contamination of the primary passivation, large shifts in the reverse leakage current, microcracks extending into the junction area, and poor metallurgical bonding in the die attach area.

4.3.7.2 Power Schottky rectifier. Although power Schottky diodes and rectifiers are extremely reliable when used properly in the application, catastrophic failures can occur if incorrect test methods are used.

A mistake commonly made is to measure reverse leakage current by applying reverse voltage until a specified current is reached. This can result in inadvertently applying excess reverse voltage to the rectifier which causes dielectric failure of the barrier.

The test must be made by applying the specified reverse voltage and measuring the resulting reverse current. The rectifier must not be stressed beyond the peak reverse voltage specified on the applicable electrical specification.

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4.3 DIODES, RECTIFIERS AND POWER DIODES

Another mistake which causes an increase in high-leakage currents is to measure the leakage at an elevated ambient temperature by means of a dc reverse bias.

Due to its high-leakage current, the diode is actually heated during the test which causes the leakage current to increase still further. This can result in a thermal runaway condition and destruction of the diode. The solution is to use a reverse voltage pulsed at a low-duty cycle, about 1 percent, to minimize the power dissipation.

Schottky barrier diodes and rectifiers also have a dv/dt failure mode. If the test fixture is allowed to deliver the specified reverse voltage at a low-source impedance, some diodes may fail when inserted one after the other into the test socket. These failures occur because the reverse voltage application rate exceeds the limit of 700 V/ μ sec. However, by using a 1000-ohm series resistor, the junction capacitance of the diode under test is sufficient to limit the rate of applied voltage to about 20 V/ μ sec.

The question of whether to use insulating hardware or to mount the Schottky devices on an electrically isolated heat sink frequently arises. In terms of overall efficiency, the isolated rectifier with a grounded heat sink is better. However, if excessive junction temperature is a problem, due to high-ambient conditions or inadequate heat sinking, the isolated heat sink is the best approach.

4.3.7.3 Derating. History has shown that the largest single cause of diode failure is operation above allowable levels of thermal and electrical stress. Accordingly, it is imperative that derating of parts be performed to increase the reliability of systems. Users should refer to MIL-STD-975 for derating factor guidelines. In general, derating for voltage, current, and surge current, should be 50 percent of the value published on the vendor data sheet. The junction temperature should be limited to +125 °C maximum. The design and use of a part should always be within the thermal and electrical stresses defined. High temperature operation is the most destructive stress a semiconductor device could be subjected to. It will result in electrical parameter drift, and a general degradation of the electrical and mechanical characteristics of the device.

4.4 DIODES, VOLTAGE REGULATOR

4.4 Voltage regulator.

4.4.1 Introduction. Technological advances in the manufacture of silicon junction diodes have made them extremely valuable components for use in accurate voltage regulator reference designs. The silicon junction diode is a semiconductor device possessing a very high reverse bias resistance up to its critical reverse breakdown, or zener, voltage. At this point, the back resistance drops to a very small value. In this region, the current will increase very rapidly, while the voltage drop across the diode remains almost constant. Figure 33 illustrates that over a wide range of current, an essentially constant voltage will be maintained. Zener diodes, therefore, when biased in the reverse direction, can be used as a voltage regulator, or reference element.

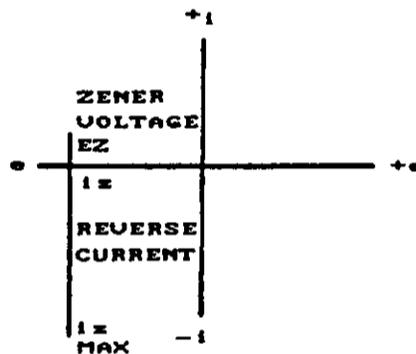


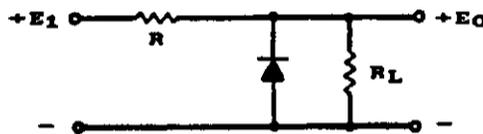
FIGURE 33. Idealized reverse breakdown characteristic.

The silicon diode regulator possesses definite advantages over other types of reference elements. It has a longer life expectancy because of mechanical ruggedness and does not suffer from deterioration under storage. There is essentially no aging during its operating life, unlike other regulating devices. Small size and light weight make its use especially useful in airborne or portable equipment. Moreover, the silicon regulator or combination thereof, can be supplied in values at any desired voltage and can operate over a wide range of current whereas other regulators are restricted to specific voltages and very limited current ranges.

4.4.2 Usual applications. Beyond breakdown, the characteristics of the silicon zener diode are almost identical to the gas voltage-regulator tube and may be considered to be the equivalent to the semiconductor. The zener diode can be used in exactly the same manner to provide a constant voltage output.

The simple shunt regulator, Figure 34, is a circuit in which a shunt element draws variable current through a resistor that is also in series with the load.

4.4 DIODES, VOLTAGE REGULATOR

FIGURE 34. Shunt voltage regulator.

The current through resistor R is dependent upon the load requirements. As the load increases or decreases, the zener shunt element will draw more or less current. The net result is a constant output voltage across R_L .

4.4.2.1 Thermally induced resistance. Resistor R should be selected so that the current in the diode does not exceed $I_Z \text{ max}$, or exceed the maximum power dissipation rating if the load R_L were removed. In practice, I_Z is chosen as approximately 20 percent of $I_Z \text{ max}$, and as a shunt regulator, it will absorb current variations between the I_Z and $I_Z \text{ max}$. limits. Referring to Figure 34, it can be seen that as I_Z increases, dynamic resistance decreases.

The greater the permissible current through the shunt diode, the larger the ripple reduction will be and the better the voltage regulation. However, as the diode current is increased, the junction temperature will rise to the point where the dynamic resistance will increase due to a thermally induced resistance in the element.

This thermally induced resistance will tend to limit regulation at high currents, whereas the dynamic resistance will dominate at lower currents. Therefore, when using a diode, a compromise operating point must be selected. The alternative would be to use a regulator of higher power dissipation capabilities possessing lower thermal resistance.

The maximum permissible diode current is limited by the temperature rise of the junction and by the heat sink provided to dissipate the heat. Thus, the use of a zener diode as a voltage regulator element is limited only by its rated current handling capabilities. Zener regulators are available over a wide range of power dissipations.

4.4.2.2 Multiple junctions. In many instances, the circuit design engineer requires regulation at higher voltages. Three alternatives are available. A single junction unit rated for the desired voltage can be used but it suffers from a high positive temperature coefficient and high dynamic resistance.

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4.4.2.2.1 Series arrangements. A second alternative to achieve optimum performance is to use a number of lower voltage units in series. The resultant temperature coefficient, dynamic resistance, and thermal resistance will be much reduced for the series combination.

An example of how a high voltage multiple junction assembly compares with a single high voltage unit can be shown when six 5-V zeners are connected in series. A single 30-V zener diode would possess a temperature coefficient of close to $+0.1\%/^{\circ}\text{C}$ and a dynamic resistance of $60\ \Omega$, and might handle only 30 mA for a particular style diode. The series combination, on the other hand, would have essentially zero coefficient, resistance of only $6\ \Omega$, and the ability to handle over 200 mA of reverse current.

Regulation of a high voltage multiple junction assembly will be superior to a single high voltage area in all respects at any given current. It is also possible to obtain a much closer tolerance by such a series combination. If similar units are used, the series regulator assembly will also have higher current ratings as a result of its increased heat dissipation capability. The higher wattage rating is directly proportional to the number of similar series units used.

4.4.2.2.2 Packaged series assemblies. The third alternative is the use of packaged series assemblies. Here, six selected zener diodes are assembled in series to provide voltage regulators from 24 through 160 V. The total dissipation of the package is 5 W without heat sink and it allows the designer to place the assembly in a convenient location on the chassis without having to provide for mounting of individual diodes.

The power dissipation per diode in such a series combination will be inversely proportional to the number of units used; in this case, six. For a given current, there will be a significant lowering in the operating junction temperatures.

As the junction temperature is lowered, life expectancy of each diode will increase.

It is possible, therefore, to achieve an increase in overall reliability despite the fact that more individual components are involved. In general, multiple junction operation is preferred in all ways.

4.4.2.3 Double anode zener diodes. In the 6- to 8-V region, it is possible, to compensate for the reverse positive temperature coefficient by taking advantage of the negative coefficient of another diode operating in the forward direction. Two such diodes, when connected in series with reversed polarities make possible a stability of ± 1 percent or better over the temperature range of -55 to $+100\ ^{\circ}\text{C}$. Optimum compensation and stability will occur at currents of approximately 10 mA.

4.4 DIODES, VOLTAGE REGULATOR

These double anode zener diodes can also be used to provide a means of simple ac regulation because the diodes are constructed in a symmetrical manner; the reverse voltages of both diodes are matched to within ± 5 percent.

Typical applications for such devices include: calibration source for oscilloscopes, ac limiting in servo control systems, and spike clipping. The basic ac regulator circuit using double anode zener diodes is shown in Figure 35. The idealized characteristics of a double anode-type zener diode is shown in Figure 36.

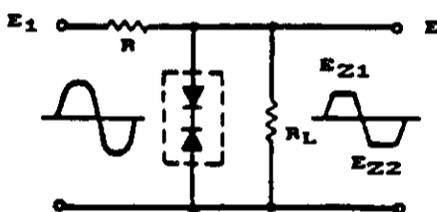


FIGURE 35. Basic ac regulator.

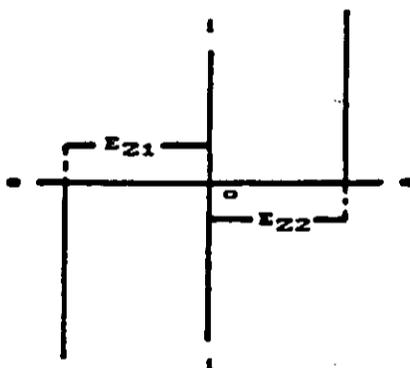


FIGURE 36. Idealized characteristics of double anode type diode.

4.4.2.4 Zener diode clamp for inductive spikes in MOSFETS. For inductive loads, as is the case in switching regulators, overvoltage transients can be produced when the device is switched off, even though the dc supply voltage for the drain circuit is well below the V_{DS} rating of the MOSFET. The faster the MOSFET is switched, the higher the overvoltage will be. If the device cannot absorb this extra inductive energy produced by the overvoltage spikes ($1/2 LI^2$ or $VI t$) a catastrophic failure (drain-source short) may occur.

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Inductance is always present to some extent in a practical circuit, and therefore, there is always danger of inducing overvoltage transients when switching off. Usually the main inductive component of the load will be clamped. Stray circuit inductance still exists, however, and overvoltage transients will still be produced as a result.

The first approach to this problem is to minimize stray circuit inductance, by means of careful attention to circuit layout, to the point that whatever residual inductance is left in the circuit can be tolerated. However, if this cannot be entirely accomplished, and for added safety against spikes by the remaining stray circuit inductance, a zener diode should be connected physically as close as possible to the drain and source terminals, as shown in Figure 37.

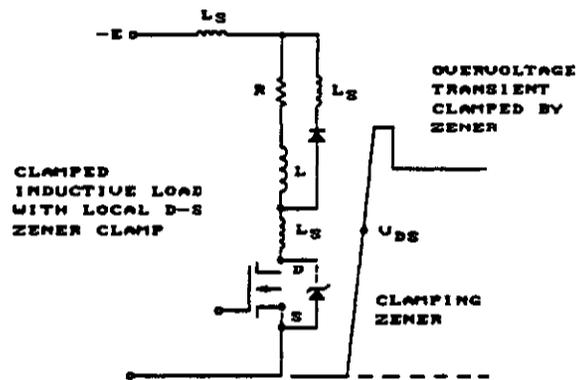


FIGURE 37. Zener voltage regulator.

4.4.2.5 Bias method for a relay amplifier. Figure 38 illustrates a method of bias for a relay amplifier that can be held in the off condition until a pre-determined input level has been reached. The zener diode provides a bias for the amplifier that is close to cut-off, and current through the JFET will be insufficient to energize the relay. When the gate is made more positive, it can be seen that the bias will remain constant even though the drain current increases.

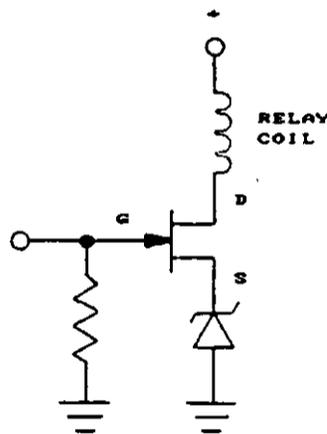


FIGURE 38. Relay amplifier bias supply.

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4.4.2.6 Zener diodes for fixed biasing purposes. The zener diode may be considered as an equivalent constant voltage source when used for biasing purposes. Figure 39 illustrates how a form of fixed bias for two JFETs in a single-ended push-pull servo amplifier may be provided. The result is a simple, highly efficient circuit with a minimum of components.

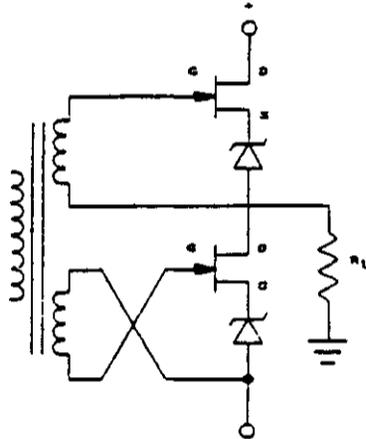


FIGURE 39. Zener diode as fixed bias.

4.4.2.7 Zener diodes as coupling devices. A silicon zener diode may be used as a coupling device between two amplifier stages in much the same manner as a capacitor; it is illustrated in Figure 40.

The dc level will be reduced only by an amount equal to the zener voltage drop. The zener element permits frequency response down to dc due to its extremely low resistance. Although the diode may be replaced by a resistor to achieve the required level change, there could be a loss in signal due to the voltage-divider action of the resistors.

4.4.2.8 Reference element application. Fortunately, the zener diode is basically a low voltage device. This fact makes it particularly attractive in the design of transistorized equipment. Of particular interest is its application as the reference element in a series-regulated power supply as shown in Figure 41.

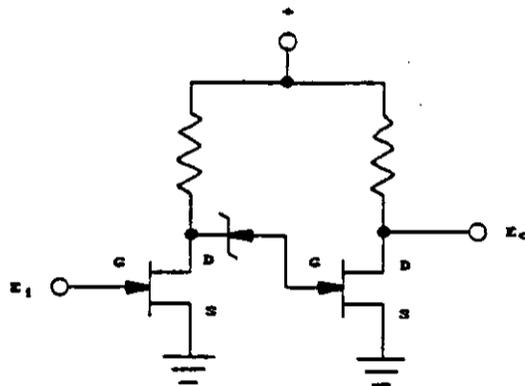
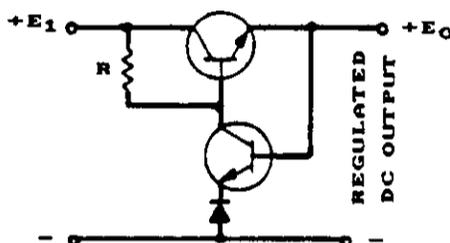
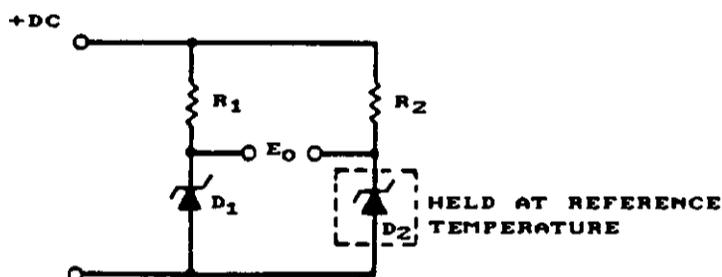


FIGURE 40. Zener diodes as dc coupling devices.

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FIGURE 41. Zener reference element in transistor power supply.

4.4.2.9 Temperature sensing devices. The apparent disadvantage of a zener diode's temperature coefficient may be put to a useful purpose in the form of a temperature-sensing device. A bridge composed of two resistors and two similar diodes, Figure 42, can be constructed so as to indicate a temperature level when one of the diodes is held at a reference temperature and the other is subjected to the varying environment. A typical 10-V zener diode has a temperature coefficient of $+0.07\%/^{\circ}\text{C}$ which corresponds to $7\text{ mV}/^{\circ}\text{C}$ change. The sensing element will, therefore, indicate an imbalance of 0.7 V when undergoing a $100\text{ }^{\circ}\text{C}$ temperature change. The output can be read directly or fed to a recording galvanometer for permanent records.

FIGURE 42. Temperature-sensitive bridge.

4.4.2.10 Selective signaling circuit. A precision selective signaling circuit can be made to operate a series of relays in a sequential manner corresponding to the value of applied voltage. In Figure 43, as the input voltage increases, the zener diode with the lower voltage will turn on first and will remain on, thereby energizing the appropriate relay coil. Subsequently, as the input voltage reaches the voltage level of each individual zener diode they too will energize the appropriate relay.

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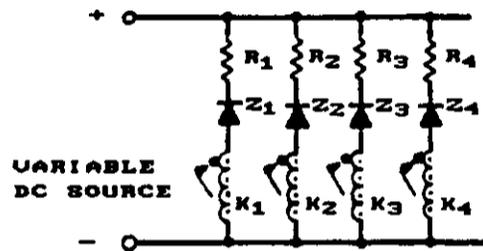


FIGURE 43. Selective relay signaling circuit.

The principal advantage over other means is that the pull-in of each relay is virtually independent of coil characteristics and is dependent only on the sharp reverse characteristic of the zener diode. The exact voltage value for relay operation can be set by preselection of the proper zener diode, rather than the often impossible task of choosing relays of dissimilar characteristics.

4.4.3 Physical construction. The voltage regulator diode generally uses a passivated diffused, planar or diffused planar junction or epitaxial substrate process die in a glass-to-metal or double-slug package. The three basic die processes are shown in Figures 44, 45, and 46.

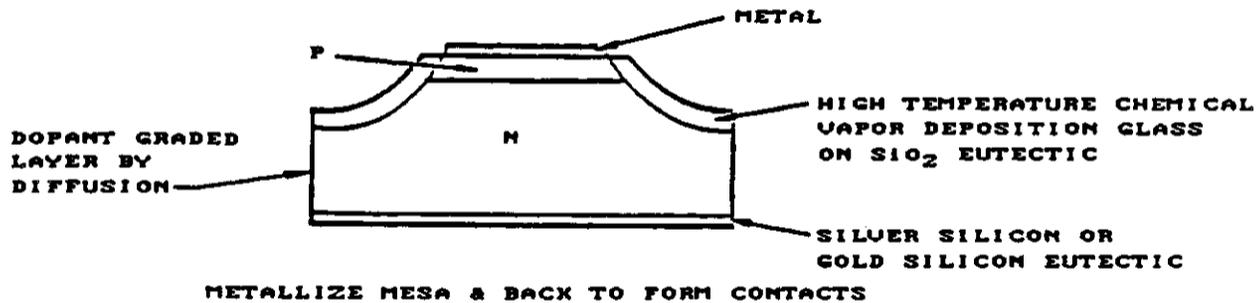


FIGURE 44. Passivated diffused process.

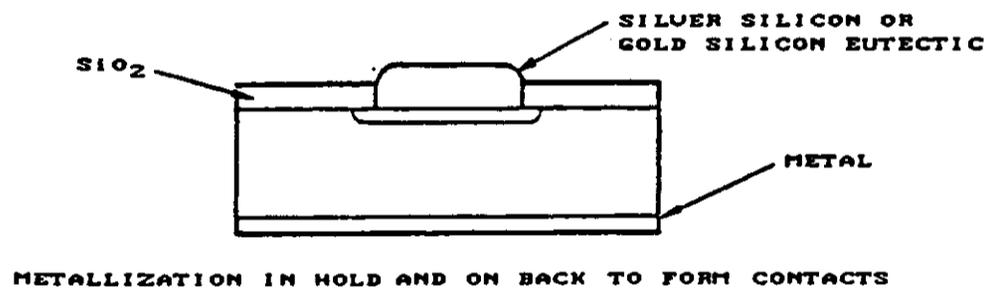


FIGURE 45. Planar process.

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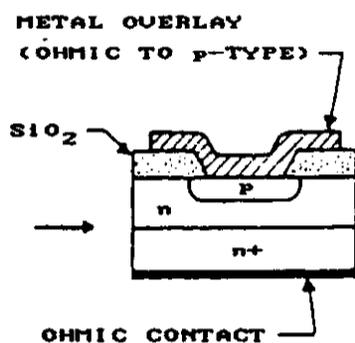


FIGURE 46. Diffused planar junction on epitaxial substrate.

4.4.3.1 Back contacts. Illustrated in Figure 47 are three back contact techniques used to construct voltage regulator diodes. Although the alloyed back contact is the strongest contact, voltage regulator diodes generally use high temperature alloy back contacts, associated with double-slug construction, for high reliability applications.

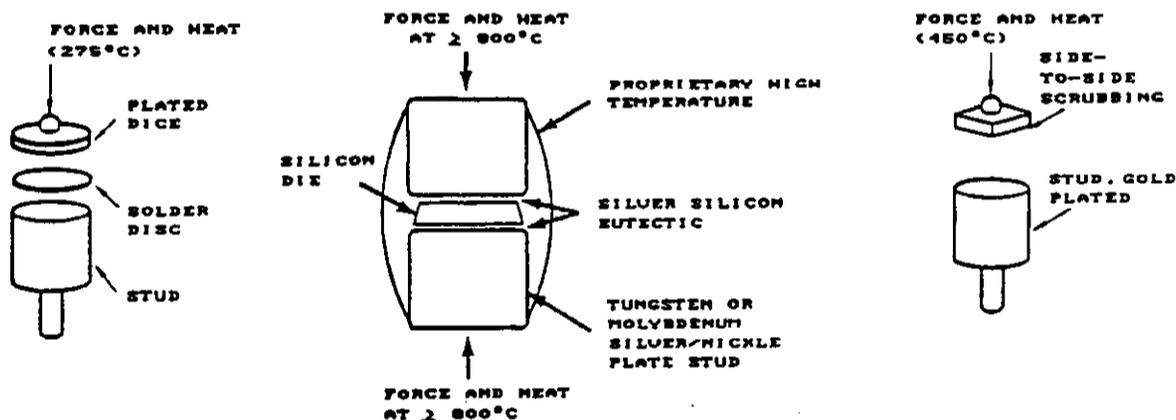


FIGURE 47. Back contact formation techniques.

The solder contact has good mechanical strength and thermal conductivity. The silicon-silver alloy contact, which is generally associated with the high temperature alloy, has good mechanical strength and thermal conductivity. The silicon chip is metallurgically brazed to plated molybdenum or tungsten slugs at temperatures exceeding 700 °C.

4.4 DIODES, VOLTAGE REGULATOR

The silicon-gold alloy contact is the strongest and most thermally conductive of the above techniques. Because of the high melting temperature of the silicon-gold eutectic alloy (377 °C), the contact will withstand high temperatures during overload.

Molybdenum, tungsten, and dumet are usually used for studs in glass packages because their thermal expansion coefficients match those of certain glasses, allowing sealing of these glasses to them. Copper and copper alloys are usually used in devices which handle appreciable power because they have much better thermal conductivity than molybdenum, tungsten, and dumet.

4.4.3.2 Front contact. Electrical contact is made to the die in three ways as shown in Figure 48.

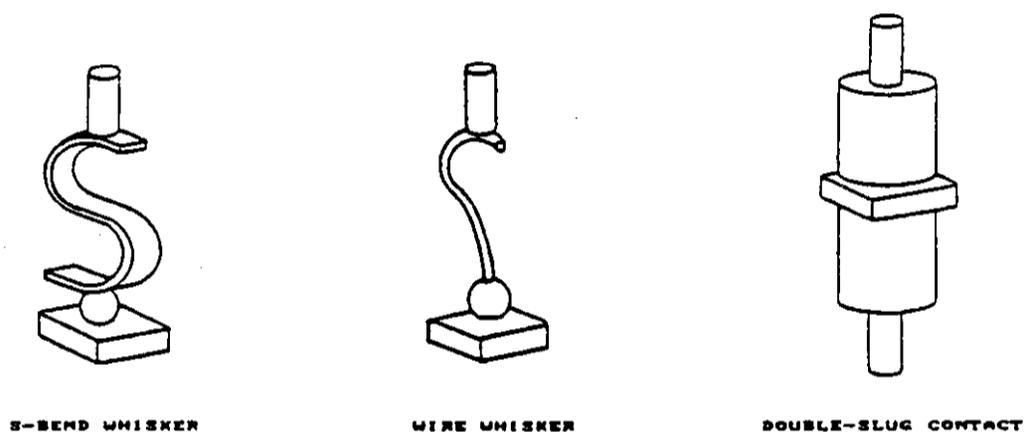


FIGURE 48. Various styles of electrical contact.

The S-bend and wire whisker are held under compression in contact with the top of the die. In this way, electrical contact is made to the die while the force on it is limited by the spring constant of the whisker. The stud contact is usually soldered to the top of a diffused die which provides good electrical and mechanical contact, but does not limit the stress on the die.

The whisker contacts have good shock and vibration tolerance. The wire whisker, however, is somewhat superior in this respect due to its lower mass and the imbedding of the whisker in the metal on top of the die which prevents sliding of the contact. Both whisker structures provide good mechanical isolation of the die from mechanical strain induced by lead flexure and thermal expansion and contraction of the package.

The double-slug stud contact transmits more stress to the die than the whisker structure, but is used despite this because it provides good heat transfer, thus raising the power rating of the package.

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4.4.3.3 Seals. Generally, two sealing techniques are used: glass-to-metal and double-slug construction. The glass-to-metal package, in Figure 49A, uses only glass-to-metal seals. These glass-to-metal seals are formed prior to final sealing and the final seal occurs by fusing glass-to-glass at the point noted in the figure. The double-slug package in Figure 49B is made by metal-lurgically bonding (>700 °C) both sides of the die to either molybdenum or tungsten slugs. To this a glass sleeve or glass slurry is fused for final sealing. The final step is a lead brazing operation. This type of construction is most often associated with voltage regulator diodes.

Figure 50 shows a typical voltage regulator diode package.

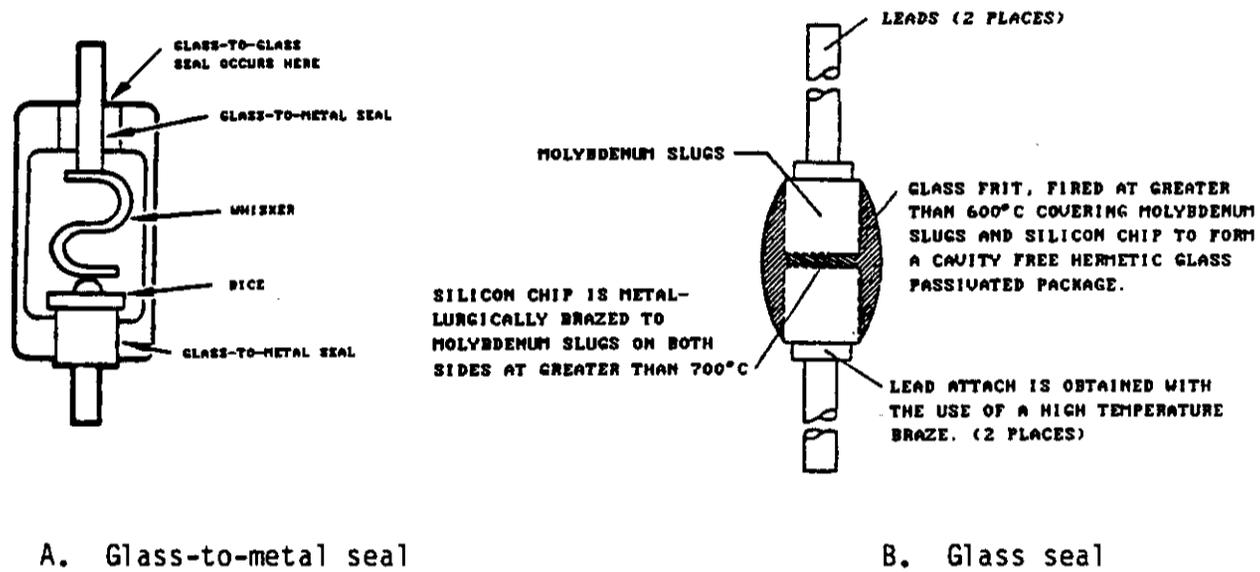


FIGURE 49. Typical construction of sealing techniques.

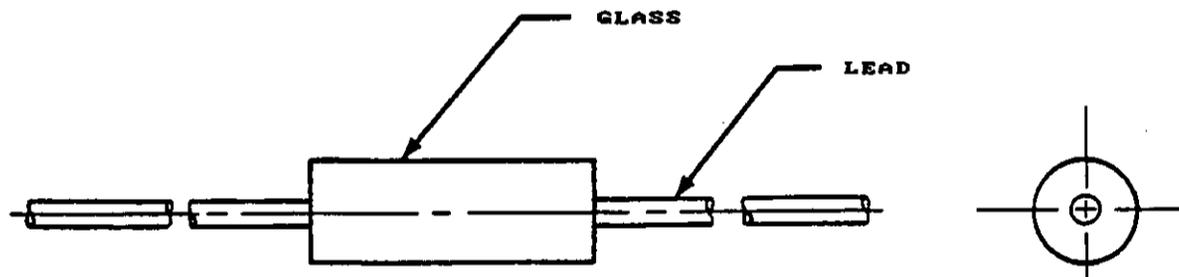
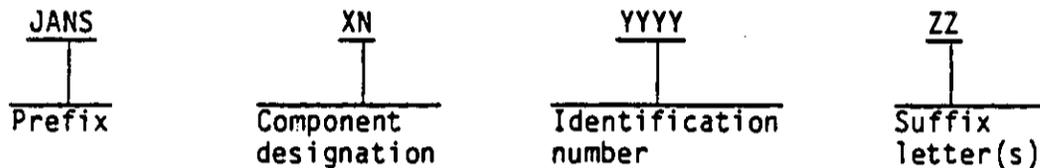


FIGURE 50. Outline drawing of voltage regulator diode.

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4.4.4 Military designation. The military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code; if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

4.4.5 Electrical characteristics.

4.4.5.1 Voltage-ampere characteristics. The volt-ampere characteristics for a typical zener diode, given in Figure 51, show that it can conduct current in both directions. The forward current, I_F , is a function of the forward voltage, V_F , and is often useful for biasing applications.

The normal operating mode of the regulating diode is a function of its reverse characteristics. In this region, V_Z is the nominal zener voltage which, in various types, may range from 2.4 to 200 V in selected 5, 10 or 20 percent groupings. The zener knee is represented by a sharp break from virtual non-conductance to conductance at the nominal zener voltage. A small zener current I_{ZK} , is the minimum current required to establish the characteristics over the knee and essentially on the flat zener plateau. This minimum regulator current varies among the various types of diodes and is given in the specification sheets. The maximum zener current I_{ZM} , is limited by the power dissipation of the diode and is a function of the junction temperature. Values of I_{ZM} are also given in the specifications.

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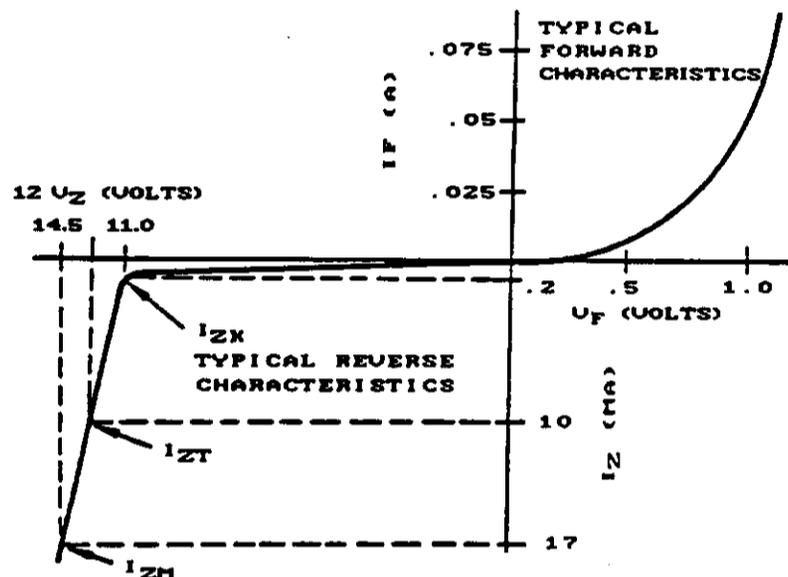


FIGURE 51. Typical forward and reverse characteristics for a 12 volt device.

V_Z is essentially constant between the limits of I_{ZK} and I_{ZM} . The plateau has, however, a small positive slope in which the precise value of V_Z will change as a function of I_Z . This change arises from the volt-ampere characteristics zener impedance, A , for fast changes I_Z , and the thermal effect due to junction temperature changes with d-c levels of I_Z . The latter will be discussed under temperature effects. The maximum zener impedance is usually specified at two selected test points for each type device. The first point is near the knee of the zener plateau and the second is near the midrange of the usable zener current excursion.

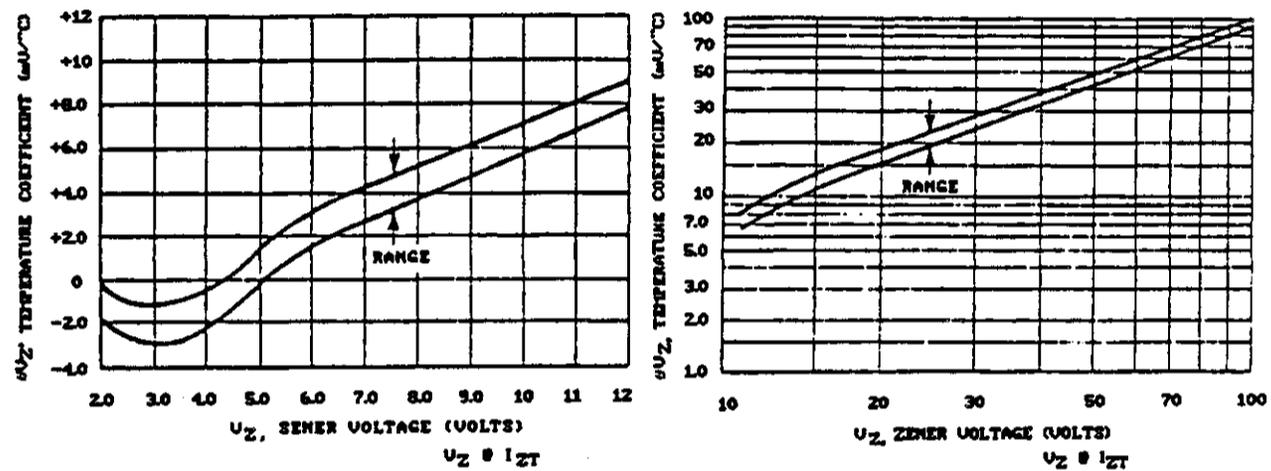
The temperature coefficient of the zener regulator diode for $V_Z \leq 5$ V via zener breakdown is negative and is somewhat dependent on the bias current as shown in graph C of Figure 52. The temperature coefficient of $V_Z = 5.0$ V may be somewhat positive or somewhat negative depending upon the bias amount as shown in graph C of Figure 52. The temperature coefficient for $V_Z > 6.5$ V is positive and is relatively independent of bias current levels for devices with V_Z up to 100 V and beyond. The variation in temperature coefficient (TC) is shown in graph A and graph B of Figure 52.

Temperature coefficient, often abbreviated as TC or represented by the symbol K_T , may be defined in either of the two following ways:

$$K_T = \frac{\Delta V_B (\text{mV})}{\Delta T (^\circ\text{C})} \text{ mV}/^\circ\text{C},$$

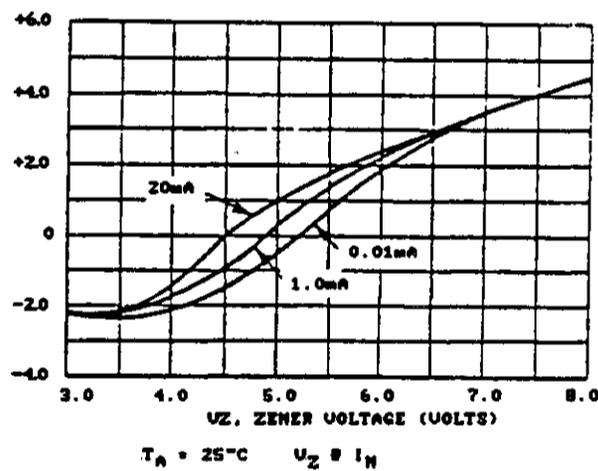
$$K_T^* = \frac{(\Delta V_B)/V_B}{100(\Delta T)} \text{ \%}/^\circ\text{C}.$$

4.4 DIODES, VOLTAGE REGULATOR



A. Range for units to 12 V.

B. Range for units 12 to 100 V.



Below 3 and above 8 V, changes in zener current do not affect temperature coefficients

C. Effect of zener current

FIGURE 52. Zener diode temperature coefficient graphs.

Some designers prefer working with the %/°C figure especially when working with power supplies in which the output voltage is usually a constant multiplied by the V_b of the voltage regulator (VR) diode used as a reference. The TC of the output voltage due to the VR diode is equal to the value K_T^* of the voltage regulator diode if expressed in %/°C.

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Other designers are more concerned about the actual voltage change and prefer to express the value of K_T in $\text{mV}/^\circ\text{C}$. To compute the total expected change in V_B caused by a temperature change, the value of V_T expressed in $\text{mV}/^\circ\text{C}$ is multiplied by the temperature difference. The result is in millivolts.

There may be a need to occasionally change back and forth between K_T as expressed in $\text{mV}/^\circ\text{C}$ and K_T^* expressed in $\%/^\circ\text{C}$. The following equations make this a simple task:

$$K_T^* = \frac{K_T}{10 V_B}$$

$$K_T = 10 V_B K_T^*$$

VR diodes which are in the mid-voltage range and exhibit effects of both zener and avalanche breakdown may have either a positive or negative temperature coefficient, depending on which effect is predominant. At low current levels, the zener effect is strongest and the TC is negative as mentioned previously.

At rather high current levels the avalanche effect becomes more evident and the VR diode takes on a positive TC. Because both zener and avalanche breakdown modes are actually at work simultaneously, with the mixture controlled by the relative current level, there is a variation in the value of TC as the current is changed.

At some specific current, the negative TC of the zener effect is exactly equal to the positive TC of the avalanche effect and the net TC is zero. In Figure 53 this condition of zero TC occurs at the point in which all three of the characteristic curves intersect.

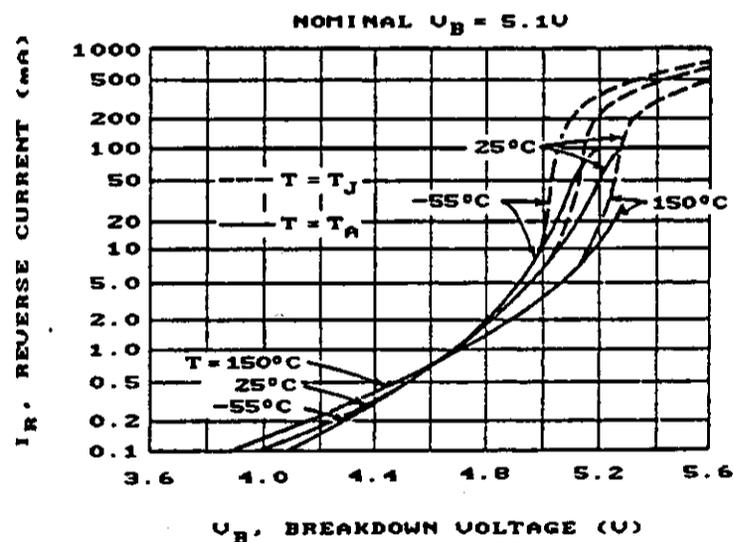


FIGURE 53. Reverse breakdown characteristics for a 1N5231 voltage regulator both zener and avalanche breakdown effects are exhibited.

4.4 DIODES, VOLTAGE REGULATOR

4.4.5.2 Zero temperature coefficient. The zero temperature coefficient characteristic is not limited to diodes of exactly a 5-V breakdown only, but can be found at various operating current points in regulators in the voltage range from 4.5 to 6.5 V. Below 4.5 V, the high current value necessary to achieve the condition of zero coefficient is prohibitive and approaches the maximum I_z of the device. Above 6 V, the intersection of zero coefficient has reached the zero reverse current line.

In certain applications, it may be necessary to place the diode in a temperature-stable environment. A small oven system can control the temperature within about a degree. Due to the increased cost, larger size, and additional circuit complexity, this must be considered an inefficient method of achieving stability.

On the other hand, there are applications which demand a reference at a discrete voltage, usually high, where the use of an oven may be the only solution. Where operation at an arbitrarily lower voltage value is possible, either of the following alternatives should be considered:

- a. A series of temperature-compensated regulator diodes is available. For example, the 1N827 with a $V_z = 5.9$ to 6.5 V, has a temperature coefficient of only $\pm 0.001\%/^{\circ}\text{C}$ or 0.062 mV/ $^{\circ}\text{C}$ at $I_z = 7.5$ mA, over a temperature range of -55 to $+100$ $^{\circ}\text{C}$.
- b. For optimum performance, a number of lower voltage units can be connected in series. The resultant temperature coefficient, dynamic resistance, and thermal resistance will be considerably reduced by using a series configuration; see paragraph 4.4.2.2.1 for a discussion of this approach. A single junction unit rated for the desired voltage exhibits a high positive temperature coefficient and a high dynamic resistance.

4.4.5.3 Maximum current limits. Any silicon zener regulator has a maximum limit of current range over which it can operate. This upper limit is established by the heat dissipation capability of the regulator. The maximum current which can flow through the diode is limited in practice by the junction temperature, which must not rise above some critical value. Heat generated internally at the crystal junction, together with the ambient temperature, contributes to the determination of the junction temperature. Refer to paragraph 4.4.7.3 for recommended reliability derating criteria.

Various power classes of regulators ranging from several hundred milliwatts to 100 W. are in production.

It should be noted that although I_z is the recommended operating point, usually 20 percent of I_z maximum, any current beyond the zener-breakdown curvature may be arbitrarily selected. If operation near the knee of the curve is desired, diodes exhibiting a V_z of above approximately 7 V should be chosen.

4.4 DIODES, VOLTAGE REGULATOR

4.4.5.4 Zener capacitance. Zener diodes are basically pn junctions operated in the reverse direction; therefore, they display a capacitance that decreases with increasing reverse voltage because the effective width of the pn junction is increased by the removal of charges as reverse voltage is increased. This decrease in capacitance continues until the zener breakdown region is entered; very little further capacitance change takes place, owing to the now fixed voltage across the junction. The value of this capacitance is a function of the material resistivity (ρ) (amount of doping--which determines V_Z nominal), the diameter (D) of junction or die size (determines amount of power dissipation), the voltage across the junction, V_R , and some constant, K. This relationship can be expressed as:

$$C_d = \frac{KA}{(\rho V_R)^n}$$

A = Effective Junction Area
 C_d = Junction Capacitance
 n = -0.5 for Alloy Junction
 -0.33 for Graded Junction

After the junction enters the zener region, capacitance remains relatively fixed and the ac resistance then decreases with increasing zener current. Typically this value is between 10 and 10000 picofarad.

4.4.5.5 High frequency and switching considerations. At frequencies above 100 kHz, and switching speeds less than 10 μ s, shunt capacitance of zener diodes begins to seriously affect their usefulness. In any application where the signal is recurrent, the shunt capacitance limitations can be overcome by operating a fast diode in series with the zener. Upon application of a signal, the fast diode conducts in the forward direction, charging the shunt zener capacitance to the level where the zener conducts and limits the peak. When the signal swings to the opposite direction, the fast diode becomes backbiased and prevents fast discharge of shunt capacitance. The fast diode remains backbiased when the signal rises and exceeds the charge level of the capacitor. When the signal exceeds this level, the fast diode conducts as does the zener.

Thus, between successive cycles or pulses, the charge in the shunt capacitor holds off the fast diode, preventing capacitive loading of the signal until zener breakdown is reached.

4.4.5.6 Noise density considerations. The breakdown voltage of the junction in a voltage regulator diode does not occur simultaneously across the entire area but it will take step-like functions. This contributes to a ragged characteristic in the knee region which is shown in an exaggerated form in Figure 54. When the current reaches a certain value, the entire junction takes part in the breakdown process and no further steps in breakdown current occur.

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FIGURE 54. Exaggerated breakdown characteristics in the knee region.

If the voltage regulator diode is biased in the region of the knee, even very slight variation in current will cause jumps in the breakdown voltage which will appear as a noise voltage.

A small part of this noise is due to the internal resistance associated with the device. A larger part of zener noise is a result of the zener breakdown phenomenon and is called microplasma noise. This microplasma noise is generally considered white noise with equal amplitude for all frequencies from about zero cycles to approximately 200,000 cycles. To eliminate the higher frequency component of noise a small shunting capacitor can be used. The lower frequency noise generally must be tolerated, because a capacitor required to eliminate the lower frequencies would degrade the regulation properties of the zener device in many applications.

Most manufacturers rate the low power zener diode series, such as the 1N5221 and 1N4115 families, with a maximum noise density at 250 μA . The rating of microvolts rms, root-mean-square, per square root cycle enables the calculation of the maximum noise density, volts per square root bandwidth.

Noise density decreases as zener current increases. This can be seen by the graph in Figure 55, where a typical noise density is plotted as a function of zener current for some typical low noise zener diodes, 1N4115 and 1N4124.

4.4 DIODES, VOLTAGE REGULATOR

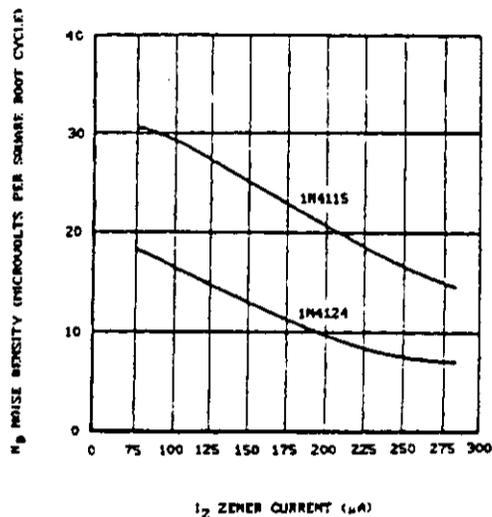


FIGURE 55. Typical noise density vs zener current.

The junction temperature will also change the zener noise levels. Thus the noise density rating must indicate bandwidth, current level, and temperature.

The circuit given in Figure 56 is used to measure noise density. The input voltage and load resistance is high so that the zener diode is driven from a constant current source. The amplifier noise must be negligible compared with the zener device under test. The filter bandpass is known so that the noise density in volts rms per square root cycle can be calculated.

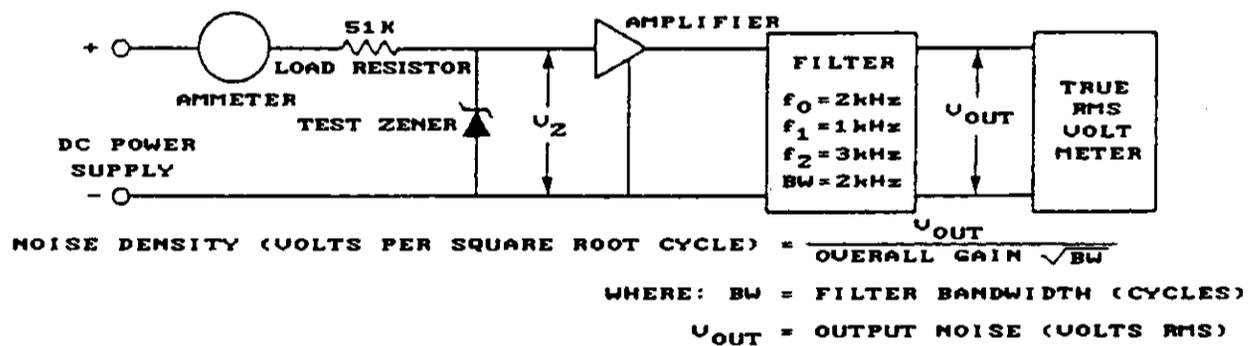


FIGURE 56. Noise density measurement method.

4.4 DIODES, VOLTAGE REGULATOR

The very low-voltage zener regulator diodes operate primarily on field emission or zener breakdown have very little noise. When avalanche breakdown becomes more dominant ($V_Z > 6$ volts), the noise density likewise increases as shown in Figure 57. In addition, noise levels in zener regulator diodes are greatly process-dependent and rather wide variations are probable among different manufacturers of the same type.

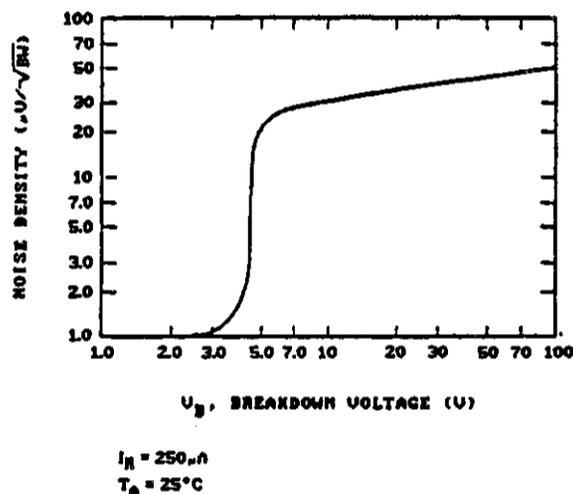


FIGURE 57. Typical noise density vs breakdown voltage.

4.4.6 Environmental considerations. Typical environmental conditions and screens which voltage regulator diodes are capable of withstanding are not substantially different from those given in paragraphs 4.1.6.9 Environmental considerations and 4.1.6.10 Screening procedures in subsection 4.1 Diodes-general. For the specific device selected consult the MIL-S-19500 reference slash sheet.

4.4.7 Reliability considerations.

4.4.7.1 Failure mechanisms and data. In extensive tests of zener voltage regulator diodes, the two most critical parameters were zener impedance and reverse current. Zener impedance is a measure of the slope of the diode's voltage-current characteristics in the breakdown or voltage-regulation region, and reverse current (leakage) is a measure of the diode's characteristics in the blocking or nonconducting direction at stated conditions of voltage and temperature. Long term life tests have yielded a significant number of failures in the reverse current and zener impedance parameters. Failure mechanisms of zener diodes are similar to those of other diode types discussed in paragraph 4.1.6 General reliability considerations, subsection 4.1 Diodes-general.

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4.4 DIODES, VOLTAGE REGULATOR

4.4.7.2 Screening. As with previous recommendations for other types of diodes, the voltage regulator procured as a JANS device offers the best protection against the above mentioned problems. It should be noted that the two characteristics which are usually measured to verify zener diode quality are the zener diode breakdown voltage (V_Z) and reverse current (I_R). Measurement of these characteristics during burn-in screening verifies the reliability of the diode. The magnitude of reverse current is largely dependent upon surface conditions, die size and thickness, and junction area of the semiconductor. Therefore, delta measurements during burn-in for I_R are a useful criterion of diode reliability.

JANTXV Zener diodes are not subjected to high temperature reverse bias (HTRB). This could allow some zener diodes to escape with surface contamination. JANS zener diodes are subjected to HTRB at 80 percent of nominal V_Z for parts with V_Z greater than 10 V. For parts with V_Z less than 10 V the test is not necessary.

4.4.7.3 Derating. History has shown that the largest single cause of diode failure is operating above allowable levels of thermal and electrical stress. Accordingly, it is imperative that derating of parts be performed to enhance the reliability of systems. Users should refer to MIL-STD-975 for derating factor guidelines. The design and use of a part should always be within the thermal and electrical stresses defined. High temperature operation is the most destructive stress a semiconductor could be subjected to. It will result in electrical parameter drift and general degradation of the electrical and mechanical characteristics of the device.

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4.5 DIODES, VOLTAGE REFERENCE

4.5 Voltage reference.

4.5.1 Introduction. A silicon voltage-reference diode is an assembly of two or more silicon junctions encapsulated in a single package. This device exhibits an extremely good voltage regulating ability over wide variations in ambient temperature for long periods of time.

A term often confused with the voltage reference diode is the zener diode. A zener diode is designed to operate under reverse bias for purposes of voltage regulation.

4.5.2 Usual applications. Application of the reference diode should take into consideration the following points: (1) current through the device should not vary widely because the diode then loses much of its value as a precise voltage reference; (2) anticipated operating temperature ranges should govern choice of current selected for device application. For control of this all important constant current over widely varying temperatures, the circuits of Figure 58 are suggested.

Circuit A of Figure 58 is usable only if a source of well-regulated high voltage is available. It must have sufficient reserve current available to supply the additional 10 mA required by the reference element. This power supply should be almost impervious to temperature effects. Variations in the value of R will, of course, result in undesirable current shifts.

In cases where no such separate regulated source is available, or where isolation is required, an independent supply may be designed using two series-connected reference diodes as preregulators, as shown in circuit B of Figure 58. Because these are also highly stable devices themselves, the end reference element is assured of an extremely constant current over a wide range of temperatures. The chief disadvantage is the cost of the additional reference devices.

Less costly zener diodes may be used as the preregulator instead of the reference diode as diagrammed in circuit C of Figure 58. The three 5.6-V diodes operating at their zero temperature coefficient current point provide excellent stability.

A fourth method involves only the use of a single zener diode in the range of 12 to 20 V as shown in circuit D of Figure 58. Such a diode, however, will suffer from an inherently positive temperature coefficient; i.e., its voltage will rise with temperature, which increases the current through the reference element. This disadvantage can be overcome by compensating with resistor R, which also has a positive coefficient. As the voltage across the zener regulator increases with temperature, the resistor will tend to increase in resistance so as to maintain a relatively constant current through the reference element.

4.5 DIODES, VOLTAGE REFERENCE

In all the circuits shown in Figure 58, resistor R is selected to provide the 10 mA reference current, or less as discussed earlier.

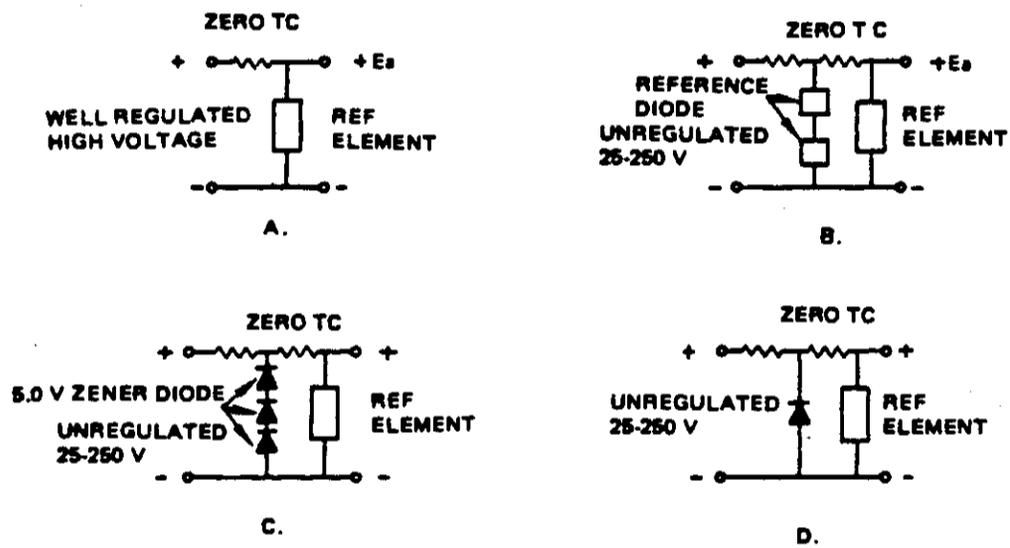


Figure 58. Typical Bias Current Systems.

4.5.3 Physical construction. The voltage reference diode is generally made with passivated diffused, planar or diffused planar junction on an epitaxial substrate process die in a single glass-to-metal or double-slug package. The multiple junctions are assembled back-to-back to achieve the desired temperature coefficient (TC). One junction operates in a reverse-bias condition and exhibits a positive TC; other junction(s) operate in the forward-biased condition and exhibit a negative TC. The net result is a near-perfect cancellation of voltage drift versus temperature change, ΔV . The three basic die processes are shown in Figures 59, 60 and 61 respectively.

4.5 DIODES, VOLTAGE REFERENCE

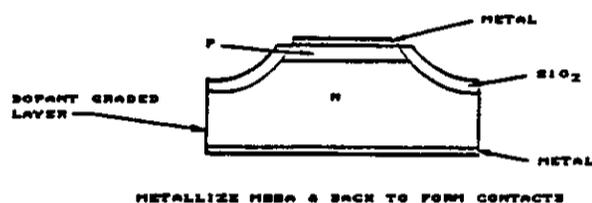


FIGURE 59. Passivated diffused process.

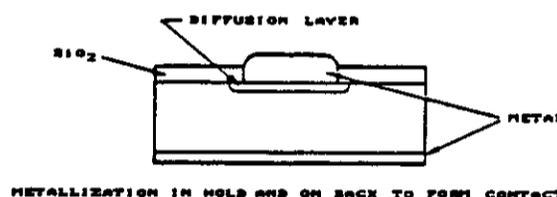


FIGURE 60. Planar process.

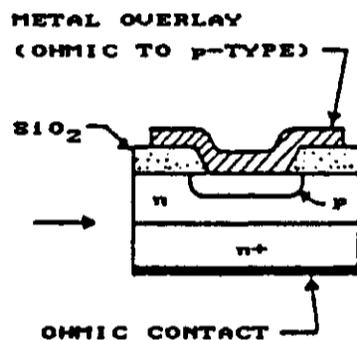


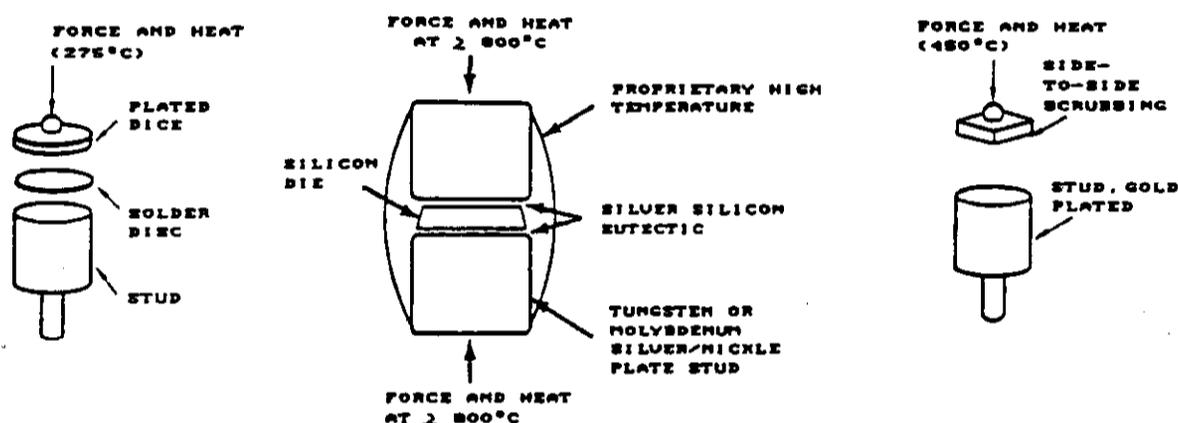
FIGURE 61. Diffused planar junction on an epitaxial substrate.

4.5.3.1 Back contacts. Illustrated in Figure 62 below are three back contact techniques used in constructing voltage reference diodes. Although the alloyed back contact is the strongest contact, voltage reference diodes generally use high temperature alloy back contacts, associated with double-slug construction, for high reliability applications.

The solder contact has good mechanical strength and thermal conductivity. The silicon-silver alloy contact which is generally associated with the high temperature alloy has good mechanical strength and thermal conductivity. The silicon chip is metallurgically brazed to plated molybdenum or tungsten slugs at temperatures exceeding 700 °C.

The silicon-gold alloy contact is the strongest and most thermally conductive of the above techniques. Because of the high eutectic temperature of silicon-gold eutectic alloy, 377 °C, the contact will withstand high temperatures during overload.

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FIGURE 62. Back contact formation techniques.

Molybdenum, tungsten, and dumet are usually used for studs in glass packages because their thermal expansion coefficients match those of certain glasses, allowing sealing of these glasses to them. Copper and copper alloys are usually used in devices which handle appreciable power because they have much better thermal conductivity than molybdenum, tungsten and dumet.

4.5.3.2 Front contacts. Electrical contact is made to the die in three ways as shown in Figure 63.

The C bend and S bend whiskers are held under compression in contact with the top of the die. In this way, electrical contact is made to the die while the force on it is limited by the spring constant of the whisker. The stud contact is usually soldered to the top of a diffused die which makes good electrical and mechanical contact but does not limit the stress on the die.

The whisker contacts have good shock and vibration tolerance. Both whisker structures provide good mechanical isolation of the die from mechanical strain induced by lead flexure and thermal expansion and contraction of the package.

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The double-slug contact transmits more stress to the die than the whisker structure, but is used in spite of this because it provides good heat transfer, thus raising the power rating of the package.

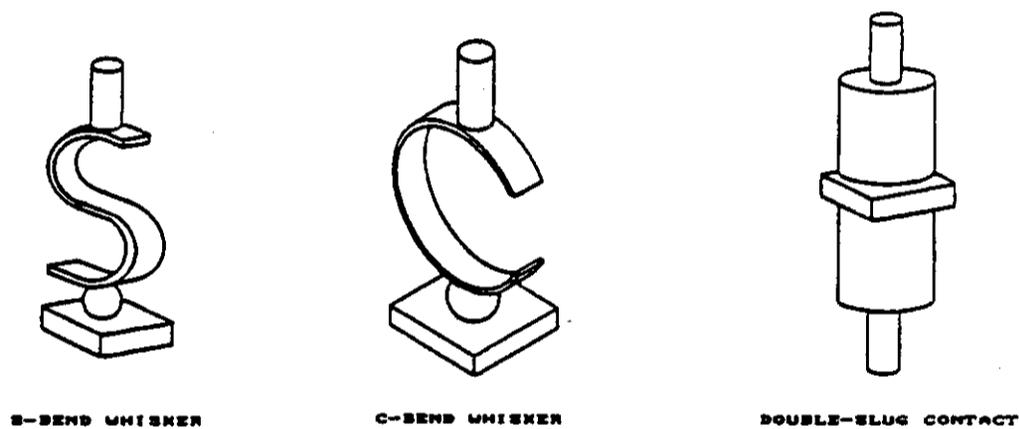


FIGURE 63. Various styles of electrical contact.

4.5.3.3 Seals. Generally, two sealing techniques are used: metal-to-glass and double-slug construction. The glass-to-metal package in Figure 64A uses only glass-to-metal seals. These glass-to-metal seals are formed prior to final sealing; the final seal occurs by fusing glass-to-glass at the point noted in the figure.

The double-slug package of Figure 64B is made by metallurgically bonding, > 700 °C, both sides of the die to either molybdenum or tungsten slugs to which a glass sleeve or slurry is fused for final sealing. The final step is a lead brazing operation. This type of construction is most often associated with voltage reference diodes.

Figure 65 shows a typical voltage reference diode package.

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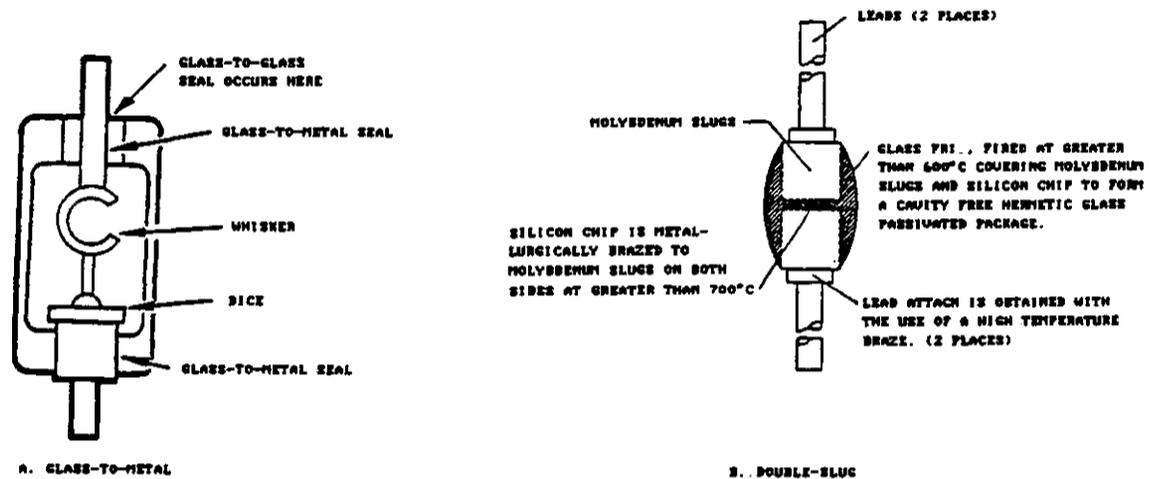


FIGURE 64. Various sealing techniques.

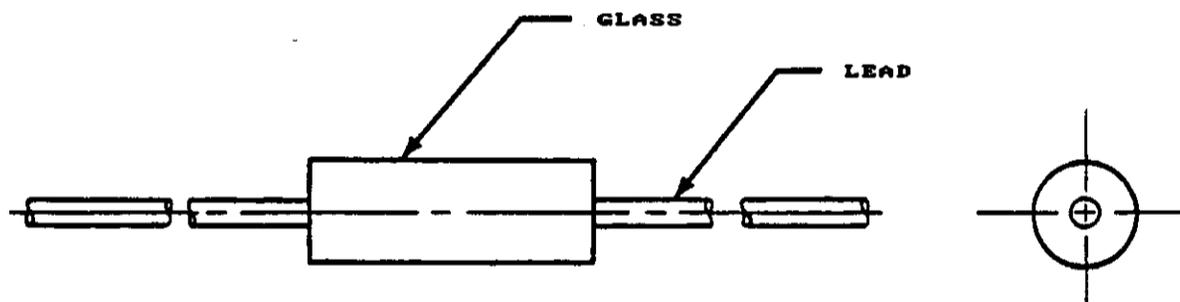
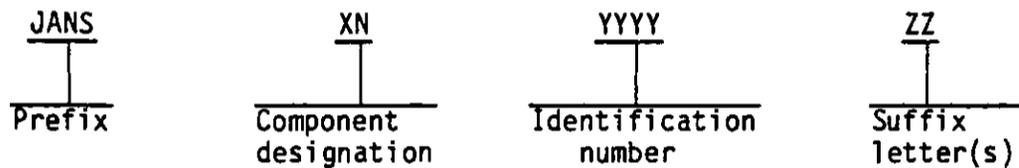


FIGURE 65. Outline drawing reference diode.

4.5.4 Military designation. The military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable is placed after the prefix. See MIL-M-19500 for details.

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The component designation is 1N for diodes.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

4.5.5 Electrical characteristics.

4.5.5.1 Stability. In the reference diode, the stability, often called absolute or time stability, is defined as the change in zener voltage over a period of time (operating or shelf life). This change is due to some characteristic of the device itself, rather than external conditions such as changes in operating temperature or current. This time stability is often confused with temperature stability.

Figure 66 is a stability curve for an 8.4-V reference diode conducting 10 mA current with $TC = \pm 0.001\%/^{\circ}C$. It can be seen that the reverse current can vary ± 1 mA without exceeding the TC requirement. However, if the reverse current varies, the resultant variation in voltage drop, as shown in Figure 67, must be considered. If the element has a drop of 8.4 V at 10.0 mA, the drop at 9 and 11 mA will be 8.389 V and 8.411 V, respectively. This is a net change of ± 11 mV and represents a change of about 1 mV per 100 mA deviation from the design center of 10 mA. Referring again to Figure 66, it can be seen that if, in addition to the normal allowable drift with temperature, a simultaneous change in reverse current from 10 mA to 9 mA occurs, an additional voltage variation corresponding to -11 mV will result. At 100 $^{\circ}C$, the voltage across the reference element could possibly have varied from its original value at +25 $^{\circ}C$ by as much as -17.3 mV, 11 plus 6.3 mV. This represents a stability factor of only about $\pm 0.003\%/^{\circ}C$; three times more than the specification allows. Therefore, if the current through the element is allowed to vary during operation, this otherwise stable device loses much of its value as a precise voltage reference.

To investigate what happens to the stability of this reference element at reverse currents other than those specified, measurements were taken at a number of discrete current values from 8 mA through 12 mA. From Figure 68, it can be seen that current values above and below the 9 to 11 mA range result in the element exhibiting thermal stability characteristics that no longer conform to the $\pm 0.001\%/^{\circ}C$ requirements. Note that at 12 mA the stability is poor at both high and low temperatures. Operation at 8 mA shows excellent low temperature characteristics, but at high elevated temperatures the drift is excessively negative. However, if a particular application calls for good voltage stability only over a limited temperature range (i.e., -25 $^{\circ}C$ through

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+75 °C) it would seem advantageous to limit the current through the element to 9 mA, 10 percent lower than its specified rating. For laboratory use, and in industrial equipment, where the temperature seldom exceeds +55 °C, a stability of at least $\pm 0.0005\%/^{\circ}\text{C}$ could be expected. If operation at only very low temperatures is contemplated, a current of even less than 8 mA would prove optimum. The choice of current should be set by the expected operating environmental conditions.

Figure 69 illustrates thermal voltage stability versus reverse operating currents, and shows how anticipated operating temperature ranges should govern the choice of current.

4.5.5.2 Forward Characteristics. Excellent stabilization of transistor collector current for variations in both supply voltage and temperature can be obtained by the use of a compensating diode operating in the forward direction in the bias network of amplifier or oscillator circuits. Figures 70 and 71 show the forward characteristics of a transistor and a compensating diode, respectively. The operating point is represented on the diode characteristics by the dashed horizontal line. The diode current at this point determines a bias voltage which establishes the transistor idling current. This bias voltage shifts with varying temperature in the same direction and magnitude as the transistor characteristic, and thus provides an idling current that is essentially independent of temperature.

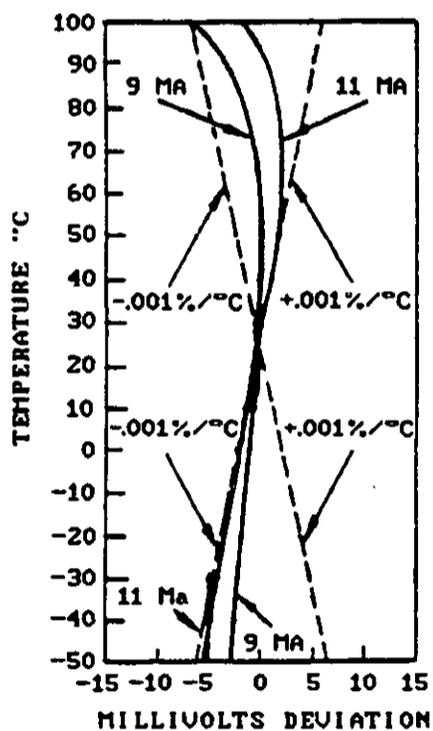


FIGURE 66. Typical stability factor at 9 and 11 mA reverse current.

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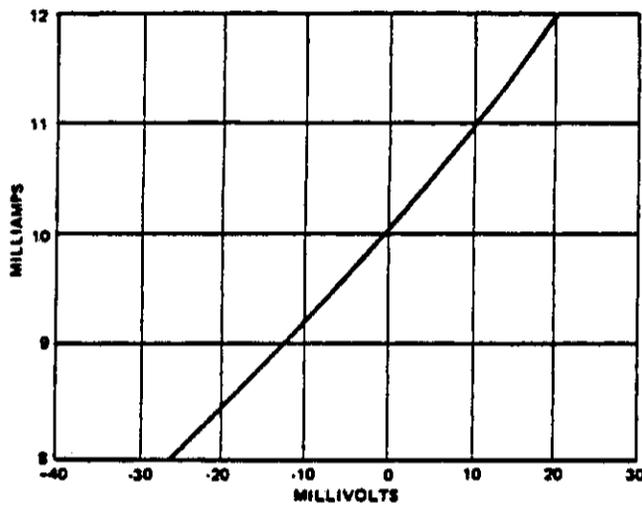


FIGURE 67. Typical voltage drop variation from a 10 mA reference point.

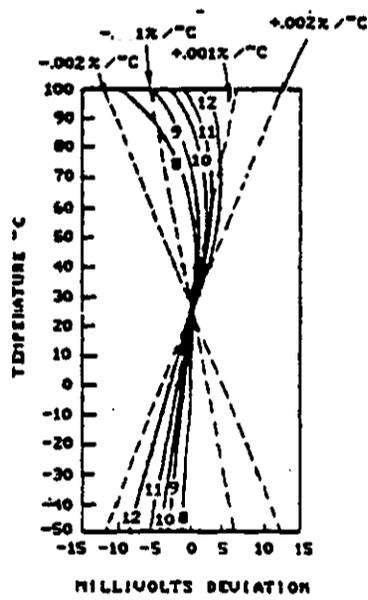


FIGURE 68. Typical stability at 8 through 12 mA.

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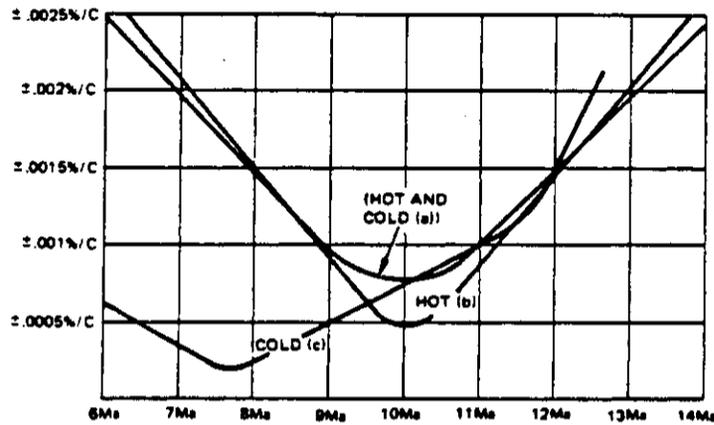


FIGURE 69. Typical thermal voltage stability vs reverse current.

The use of a compensating diode also reduces the variation in transistor idling current as a result of supply-voltage variations. Because the diode current changes in proportion with the supply voltage, the bias voltage to the transistor changes in the same proportion and idling-current changes are minimized.

4.5.5.3 Temperature effects. Figure 72 shows the effect of temperature on voltage reference diode characteristics. The forward characteristic does not vary significantly with reverse breakdown (zener voltage) rating. A change in ambient temperature from 25 to 100 °C produces a shift in the forward curve in the direction of lower voltage.

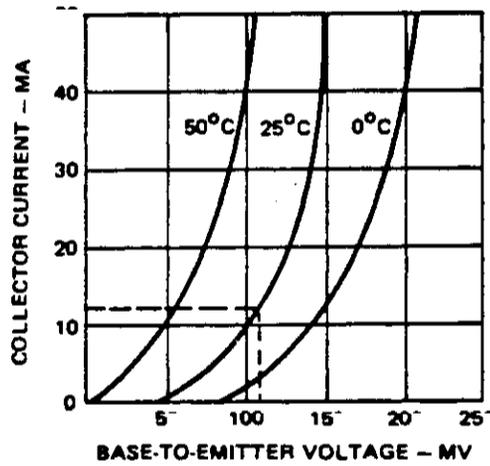


FIGURE 70. Transfer characteristics of transistor.

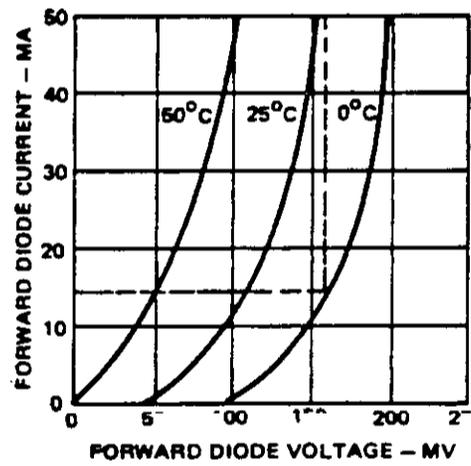


FIGURE 71. Forward characteristics of compensating diode.

4.5 DIODES, VOLTAGE REFERENCE

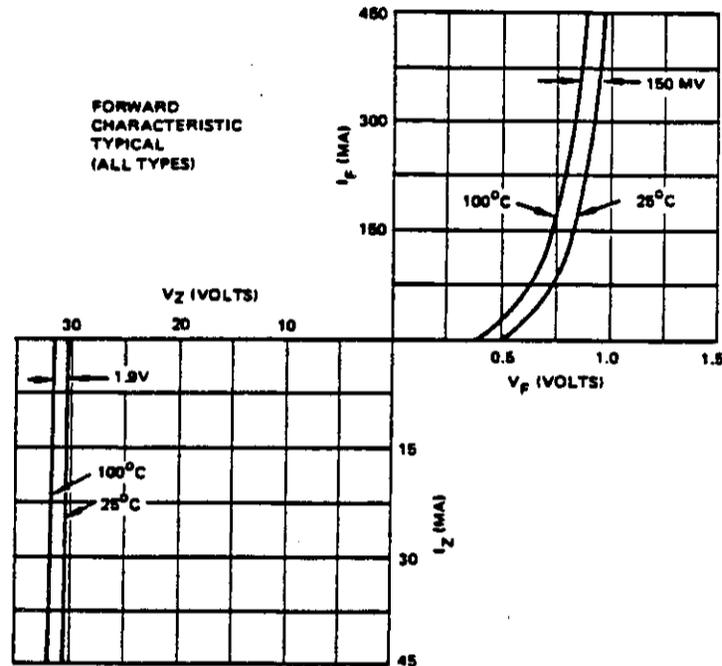


FIGURE 72. Effects of temperature on voltage reference diode characteristics.

4.5.5.4 Temperature compensation. Various combinations of forward biased junctions and reverse biased junctions may be arranged to achieve compensation. From Figure 73, it can be seen that if the absolute value of voltage change (ΔV) is the same for both the forward biased diode and the zener diode where temperature has gone from 25 °C to 100 °C, the total voltage across the combination will be the same at both temperatures since one ΔV is negative and the other positive. Furthermore, if the rate of increase or decrease is the same throughout the temperature change, the voltage will remain constant. The nonlinearity associated with the voltage temperature characteristics is a result of this rate of change not being a perfect match, i.e.,

$$V_{REF} = V_Z + \Delta V_Z + V_D - \Delta V_D$$

Figure 74 shows the voltage temperature characteristics of a typical voltage reference diode. It can be seen that the voltage drops slightly with increasing temperature.

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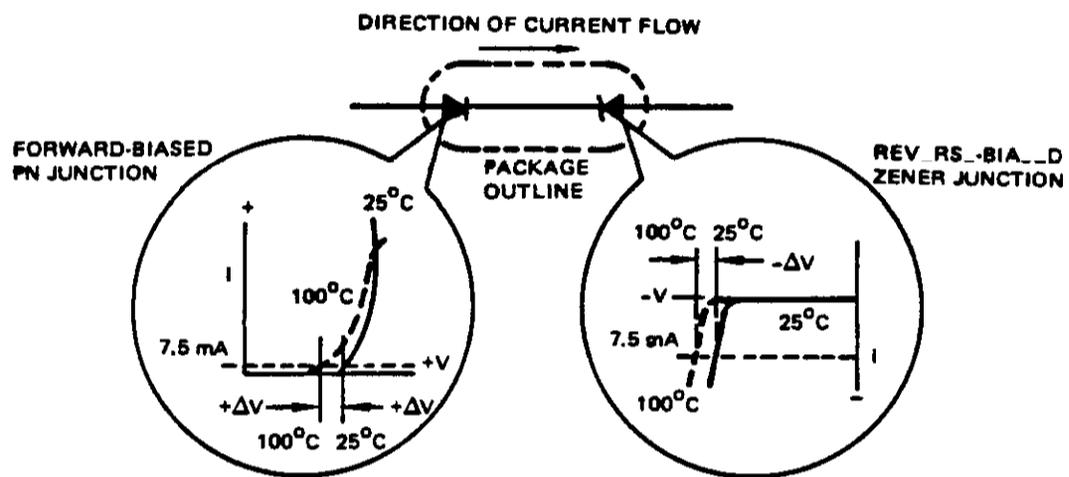


FIGURE 73. Principle of temperature compensation.

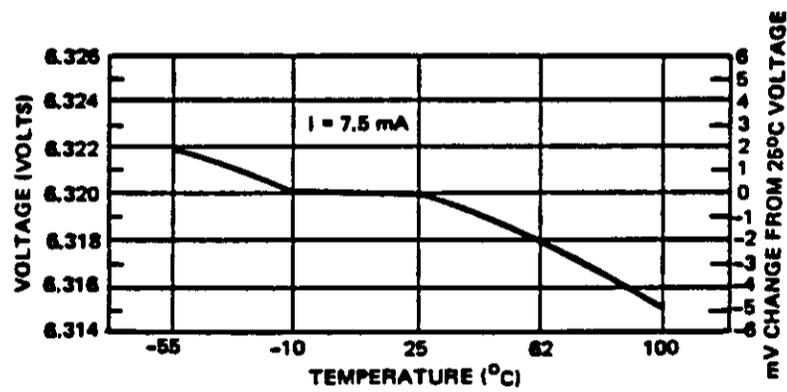


FIGURE 74. Voltage vs temperature for a typical voltage reference diode (1N827).

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Figure 75 illustrates that there is a significant change in the temperature coefficient of a unit depending on how much above or below the test current the device is operated. If the three curves intersect at "A," operation at I_A results in the least amount of voltage deviation due to temperature from the +25 °C voltage. At I_B and I_C there are greater excursions, $\Delta V_B + \Delta V_C$, from the 25 °C voltage as temperature increases or decreases.

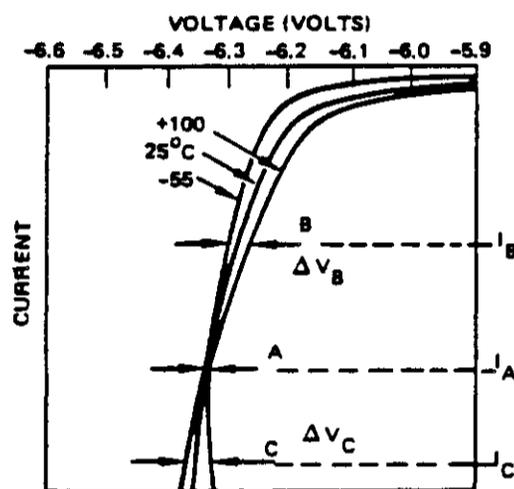


FIGURE 75. Voltage-ampere curves showing crossover at A.

To summarize, the following are significant considerations for reference diode applications:

- a. Current regulation is probably the most critical consideration when using temperature compensated units.
- b. Zener impedance is defined as the slope of the V-I curve at the test point corresponding to the test current. Impedance changes with temperature, but the variation is usually small and it can be assumed that the amount of current regulation needed at +25 °C will be the same for other temperatures.
- c. Standard voltage reference diodes typically exhibit a time stability of 200-500 ppm per 1000 hours.

4.5.5.5 Electrical ratings. Table VI gives the electrical ratings for several typical voltage regulator diodes.

The information is for general reference only. For the specific device selected, consult the applicable MIL-S-19500 specification.

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4.5 DIODES, VOLTAGE REFERENCE

TABLE VI. Electrical ratings

Device Type No.***	Reference Voltage Range			Dynamic Impedance		Voltage Temperature Stability BV <u>1/</u>
	BV (Min. Volts)	BV (Max. Volts)	@ I _z (mA)	Z Ohms	@ I _z (mA)	
1N827-1*	5.90	6.50	7.5	15	7.5	0.009
1N829-1*	5.90	6.50	7.5	15	7.5	0.005
1N937B-1**	8.55	9.45	7.5	20	7.5	0.037
1N938B-1**	8.55	9.45	7.5	20	7.5	0.018
1N939B-1**	8.55	9.45	7.5	20	7.5	0.009
1N940B-1**	8.55	9.45	7.5	20	7.5	0.0037

* From MIL-S-19500/159, -1 means metallurgically bonded, ±5 percent V_z tolerance.

** From MIL-S-19500/156, -1 means metallurgically bonded ±5 percent V_z tolerance.

*** All devices tabulated have military outline D07.

1/ BV is measured in volts per span of temperature. The particular span of temperature varies for different devices and may be found in the applicable MIL-S-19500 slash sheet.

NOTE: This table is not to be used for part selection; use MIL-STD-975 for that purpose.

4.5.6 Environmental considerations. Typical environmental conditions and screens which voltage reference diodes are capable of withstanding are not substantially different from that given in paragraphs 4.1.6.9 Environmental considerations, and 4.1.6.10 Screening procedures, in subsection 4.1 Diodes-general. For the specific device selected, consult the applicable MIL-S-19500 reference slash sheet.

4.5.7 Reliability considerations. There are many failure modes associated with voltage reference diodes. The majority of failures are due to zener voltage and short and open circuits. The mechanical and thermal stress tests identify most of the conditions of these failures. As with other diode families, it is suggested that the JANTXV and JANS process would eliminate all or most of these failure types.

4.5.7.1 Failure mechanisms. Because the failure mechanisms for this type of diode are similar to those of other diode types, paragraph 4.1.6 Reliability considerations should be consulted for a discussion of diode failure mechanisms.

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4.5.7.2 Derating. History has shown that the largest single cause of diode failure is operation above allowable thermal and electrical stress levels. Accordingly, it is imperative that derating of parts be performed to increase the reliability of military systems. Users should refer to MIL-STD-975 for derating factor guidelines. In general, derating for zener current should be limited to no more than $I_z = I_z \text{ nom} + .5 (I_z \text{ max.} - I_z \text{ nom})$ of the values published on the vendor data sheet. The junction temperature should be limited to +125 °C maximum. The design and use of a part should always be within the thermal and electrical stresses defined. High temperature operation is the most destructive stress a semiconductor device could be subjected to. It will result in early end of life, electrical parameter drift, and a general degradation of the device's electrical and mechanical characteristics.

4.6 DIODES, CURRENT REGULATOR

4.6 Current regulator.

4.6.1 Introduction. The current-regulator diode is basically a field-effect transistor that has its gate and source terminals connected together. It is called either a constant-current diode or field-effect current regulator diode. It presents a constant current independent of the terminal voltage over a wide operating range and exhibits a very high circuit impedance.

4.6.2 Usual applications. The current-regulator diode is useful in such applications as over-current protection, transistor biasing, linear ramp or stairstep generators, differential amplifiers, and precision reference voltage sources. It is always used in the forward bias direction between the pinch-off voltage and the breakdown voltage. The symbol and polarity of a current regulator diode are shown in Figure 76.



FIGURE 76. Symbol and polarity of current regulator diode.

Devices are available that cover the current range from 220 μA to 4.7 mA in 32 different current steps. The family series of devices that cover this range are the 1N5285 through 1N5314 diodes. The current-regulator diodes in this series have a tolerance of plus or minus 10 percent and are available as JANTXV versions in a DO-7 package.

To use the current-regulator diode as a constant-current bias source in a differential amplifier circuit, the device would be connected as shown in Figure 77.

The current-regulator diode replaces the transistor, zener diode, and resistors that would normally be used in a conventional circuit. To use the current-regulator diode as a low-voltage reference source, it would be connected in the circuit as shown in Figure 78.

The constant-current regulator diode drives a known resistor value that produces an output reference voltage whose value is determined by Ohm's law. This results in a precision millivolt reference source when it is desirable to generate a stable and accurate voltage reference at voltages lower than those offered by zener diodes.

4.6 DIODES, CURRENT REGULATOR

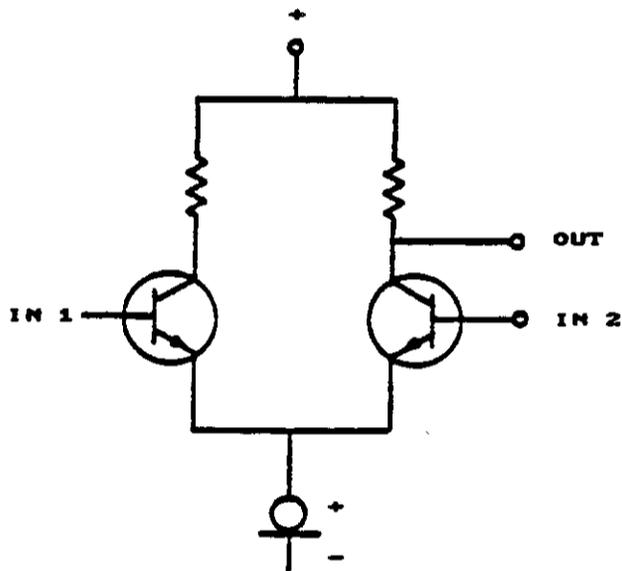


FIGURE 77. Constant current bias source in a differential amplifier circuit.

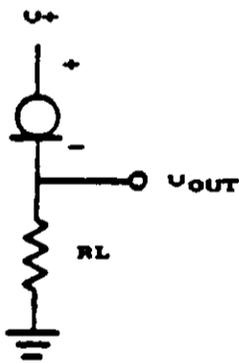


FIGURE 78. Circuit showing current regulator diode as a low-voltage reference source.

The current-regulator diode can also be combined with a zener diode to obtain better voltage and temperature performance than a zener diode could provide by itself. See Figure 79 for the circuit connection. This combination may be used when an accurate temperature-stable reference voltage is required.

4.6 DIODES, CURRENT REGULATOR

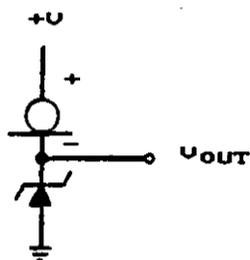


FIGURE 79. Current regulator diode combined with a zener diode.

4.6.3 Physical construction. The current-regulator diode generally has a planar process die in a glass-to-metal package. The basic die process is shown in Figure 80.

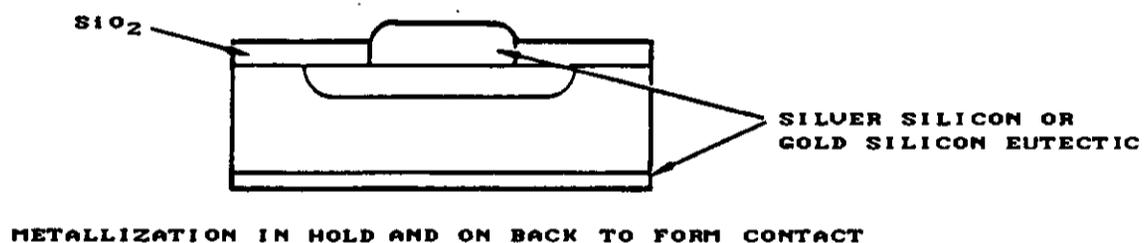


FIGURE 80. Planar process.

4.6.3.1 Back contacts. Illustrated in Figure 81 is the solder back contact used in constructing current regulator diodes.

The solder contact has good mechanical strength and thermal conductivity. The stud material is generally molybdenum, tungsten, or dumet because their coefficients of expansion closely match those of glass.

4.6 DIODES, CURRENT REGULATOR

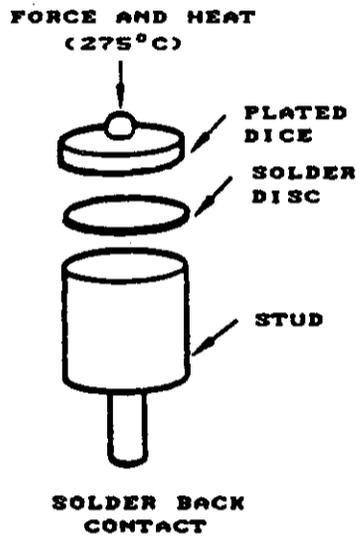


FIGURE 81. Solder back contact.

4.6.3.2 Front contacts. Electrical contact is made to the die in two ways as shown in Figure 82.

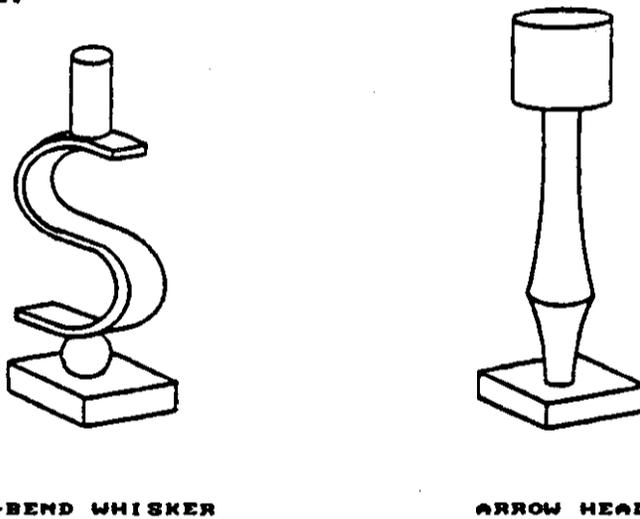


FIGURE 82. Two styles of electrical contact.

The whisker is held under compression in contact with the top of the die. In this way, electrical contact is made to the die while the force on it is limited by the spring constant of the whisker. The arrowhead contact is made using solder between the wire and the top of the die.

4.6 DIODES, CURRENT REGULATOR

The whisker contact has good shock and vibration tolerance. The arrowhead contact is somewhat superior due to imbedding of the end of the contact in the metal on top of the die. The whisker structure provides good mechanical isolation of the die from mechanical strain induced by lead flexure and thermal expansion as well as contraction of the package.

4.6.3.3 Seals. Figure 83 shows the sealing technique for glass-to-metal seals.

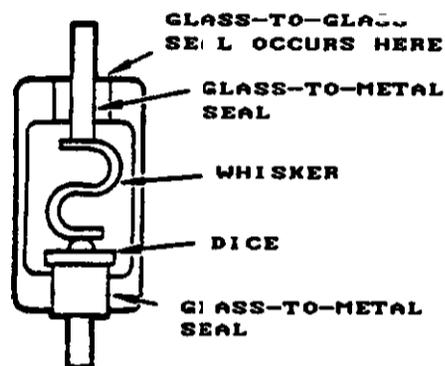


FIGURE 83. Typical construction of glass-to-metal seal technique.

The sealing technique generally used for a current-regulator diode is a glass-to-metal seal. The glass-to-metal seals in the above package are formed prior to final sealing, which is a glass-to-glass seal.

Figure 84 shows a typical current regulator diode package.

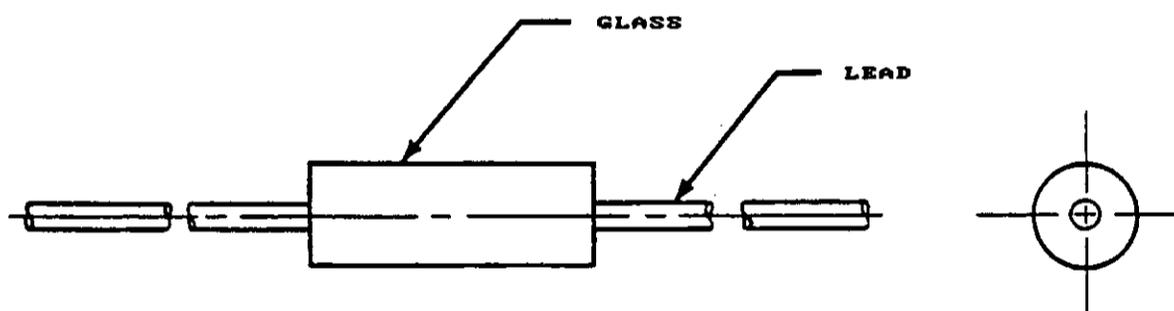
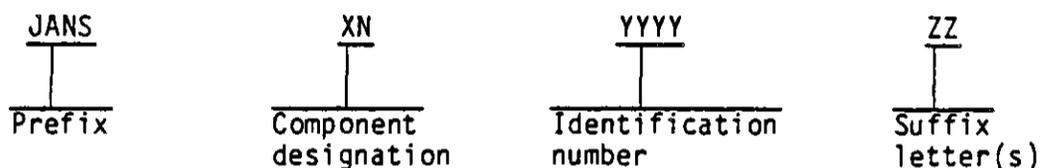


FIGURE 84. Outline drawing of a typical current-regulator diode (DO-7 package).

4.6 DIODES, CURRENT REGULATOR

4.6.4 Military designation. The military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

The identification number is assigned in order by registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

4.6.5 Electrical characteristics. The current-regulator diode is basically an FET with its gate and source terminals connected. An FET connected in this way acts as a constant-current device over a wide voltage range. This range is from the pinch-off voltage to the forward breakdown voltage point. In this region, the power supply or circuit voltage can vary widely with very little change, if any, in the current the device allows to pass. Therefore, the device is biased in the forward direction between the pinch-off voltage and the forward breakdown voltage. Any circuit voltage within this range will result in a constant drain current. The ideal constant-current diode will have a very low pinch-off voltage, a high breakdown voltage, and as high a dynamic impedance between these limits as possible. The current-regulator diode should not be used in the reverse bias direction.

4.6.6 Environmental considerations. Typical environmental conditions and screening tests which current regulator diodes are capable of withstanding are not substantially different from those given in paragraphs 4.1.6.9 Environmental considerations and 4.1.6.10 Screening procedures in subsection 4.1 Diodes, general. For the specific device selected, consult the applicable MIL-S-19500 reference slash sheet.

4.6.7 Reliability considerations.

4.6.7.1 Failure modes. The major electrical failure modes for current-regulator diodes are a shift in the regulator current (I_p) or a failure of the device to

4.6 DIODES, CURRENT REGULATOR

regulate. Break-down voltage (BV) is not a major electrical failure mode with current-regulator diodes because of the high voltage. Because of the above catastrophic failure modes, the devices are more likely to have shorts than opens.

Because the failure mechanisms for this type of diode are similar to those of other diode types, paragraph 4.1.6 Reliability considerations should be consulted for problems related to diode failure mechanisms.

4.6.7.2 Derating. History has shown that the largest single cause of failure for diodes is operation above allowable levels of thermal and electrical stress. Accordingly, it is imperative that derating of parts be performed done to increase the reliability of systems. Users should refer to MIL-STD-975 for derating factor guidelines.

The design and use of a part should always be within the thermal and electrical stresses defined. High temperature operation is the most destructive stress a semiconductor could be subjected to. It will result in electrical parameter drift, and a general degradation of the electrical and mechanical characteristics of the device.

4.7 DIODES, VOLTAGE VARIABLE CAPACITOR

4.7. Voltage-variable capacitor.

4.7.1 Introduction. Voltage-variable capacitance diodes are not included in MIL-STD-975, however, they are included in this handbook to provide a technical understanding of this part.

The voltage-variable capacitance diode is a silicon pn-junction diode designed for use as a voltage-variable capacitor. This device provides a means of changing circuit capacitance through bias voltage control. The capacitance varies essentially as $1/\sqrt{V}$ as the voltage across its terminals is varied. This device maintains constant characteristics over a wide temperature range but is less temperature sensitive when used at higher operating voltages. Therefore, if capacitance variation must be minimized, it is advisable to operate at higher bias voltages.

With voltage-variable capacitance diodes, it is possible to construct very low noise amplifiers. In principle, the noise figure of a parametric amplifier is equal to 0 dB if the amplifier is considered to be ideal; i.e., one that generates no internal noise. This is not possible in a conventional amplifier because of thermal and shot noise inherent in semiconductor devices. Parametric amplifiers more closely approach the ideal and are thus useful in the front end of receivers because their low noise increases the detection range by several factors.

4.7.2 Usual applications. The voltage-variable capacitance diode is useful from low audio to ultra high frequencies and from $-65\text{ }^{\circ}\text{C}$ to $+150\text{ }^{\circ}\text{C}$. It is useful in automatic frequency control, voltage tuning, frequency modulation, and harmonic multiplication. An example of a high-power harmonic multiplier application is provided in Figure 85.

In the circuit, an input of 50 W at 50 MHz is converted to an output of 32.5 W at 150 MHz. This is accomplished by the varactor diode which builds up a waveform with flattened positive peaks because its capacitance is high at positive and low-negative bias; also, the negative peaks are exaggerated because the capacitance is low at high reverse bias. This distorted wave contains strong harmonics which can be picked up by the output tuned circuit, which in this case is tuned to the third harmonic.

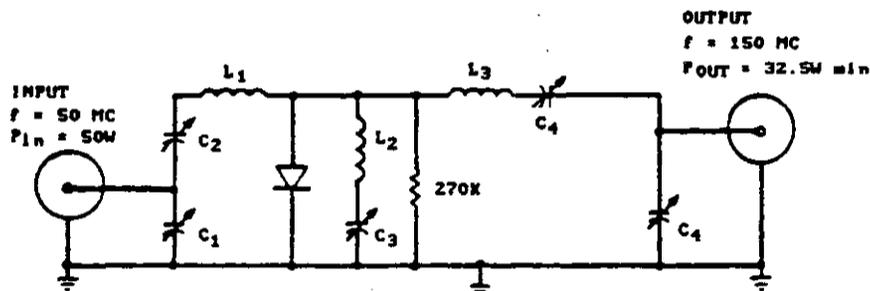


FIGURE 85. 50 to 150 MHz tripler test circuit.

4.7 DIODES, VOLTAGE VARIABLE CAPACITOR

In the design of harmonic multipliers, lumped circuit techniques are useful up to 450 MHz with little performance degradation provided coil and capacitor "Q" values of 200 to 300 are maintained.

Above 450 MHz, coaxial, stripline, or helical coil resonators are recommended. Component values are not particularly critical; however, excessive inductance or insufficient coupling can cause low efficiency, and insufficient inductance or excessive coupling can cause poor filtering. Simple experimentation with well constructed and shielded breadboards is generally sufficient for circuit optimization. Note that an adequate tuning range must be provided to insure input match over normal voltage variable capacitance diode variations, and that spurious signals between stages should be kept below 30 dB by suitable filter circuits.

In typical applications, doubling efficiency is 5 percent greater than that for tripling, and, quadrupling efficiency 5 percent less than that for tripling.

4.7.3 Physical construction. The voltage-variable capacitance diode is generally made with a passivated diffused process die in a glass-to-metal package. The basic die process is shown in Figure 86.

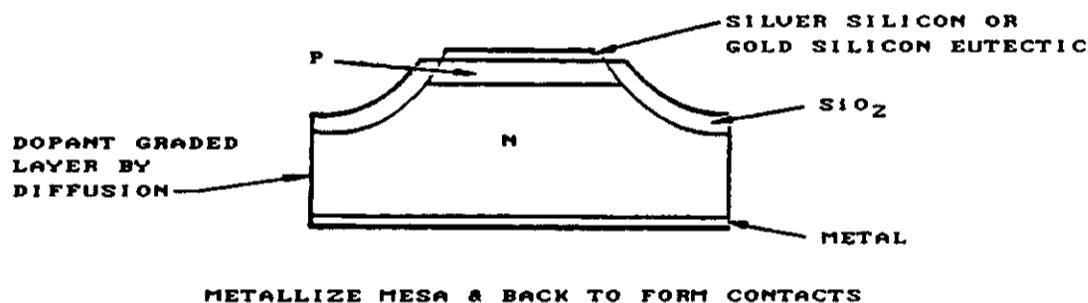


FIGURE 86. Passivated diffused process.

The following general description of construction characteristics is typical of most voltage-variable capacitance diodes. The high power varicap incorporates the high Q varicap crystals in a high power package. The alloyed abrupt junction gives the user a device which closely approaches the theoretical formula $C = 1/\sqrt{V}$ for the capacitance versus voltage curve. The capacitance nonlinearity of the high power varicap can be defined by the following formula:

$$C_V = (C - C_{pkg}) \cdot \left(\frac{4 + V_0}{V + V_0} \right)^n + C_{pkg}$$

where

voltage stock sensitivity (n) = 0.48; package capacitance = 1.5 pF, C = capacitance at -4 reverse voltage at -4 V barrier potential (V_0) = 0.7 @ 25 °C (TC of $V_0 \cong 2$ mV/°C).

4.7 DIODES, VOLTAGE VARIABLE CAPACITOR

The high-power varicap has broad area contacts which are metallurgically bonded at 320 °C. These positive contacts and surface passivation on both sides of the crystal give the high-power varicap its increased power handling capabilities and high reliability characteristics.

As previously noted, the above typical physical characteristics are for reference only. For details of the specific device selected, consult the applicable MIL-S-19500 specification sheet.

4.7.3.1 Back contact. Illustrated in Figure 87 is the solder back contact technique used in the construction of voltage-variable capacitor diodes.

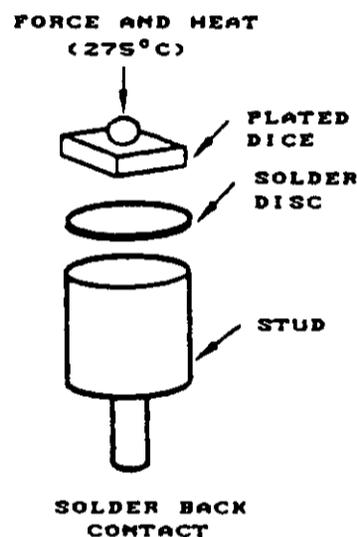


FIGURE 87. Back contact formation techniques.

The solder contact has good mechanical strength and thermal conductivity. The stud material is generally molybdenum, tungsten, or dumet because their coefficients of expansion closely match those of glass.

4.7.3.2 Front contact. Electrical contact is made to the die in two ways as shown in Figure 88.

The S bend and wire whiskers are held under compression in contact with the top of the die. In this way, electrical contact is made to the die while the force on it is limited by the spring constant of the whisker.

The whisker contacts have good shock and vibration tolerance. The wire whisker is somewhat superior in this respect due to its lower mass and the imbedding of the whisker in the metal on top of the die, which prevents sliding of the

4.7 DIODES, VOLTAGE VARIABLE CAPACITOR

contact. Both whisker structures provide good mechanical isolation of the die from mechanical strain induced by lead flexure and thermal expansion and contraction of the package.

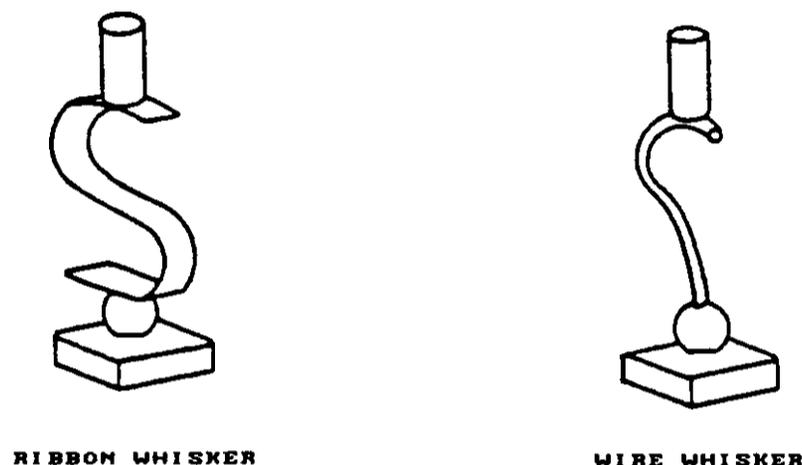


FIGURE 88. Two styles of electrical contact.

4.7.3.3 Seals. Figure 89 shows the sealing technique for glass-to-metal seals.

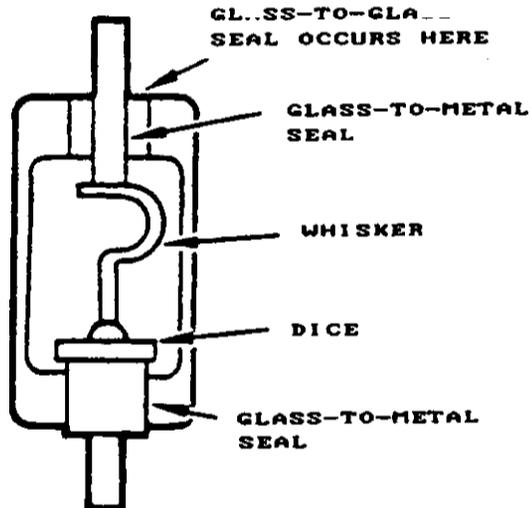


FIGURE 89. Glass-to-metal seal technique.

The sealing technique generally used for a voltage variable capacitor diode is a glass-to-metal seal. The glass-to-metal seals in the above package are formed prior to final sealing, which is a glass-to-glass seal.

4.7 DIODES, VOLTAGE VARIABLE CAPACITOR

Figure 90 below shows a typical voltage-variable capacitor diode package.

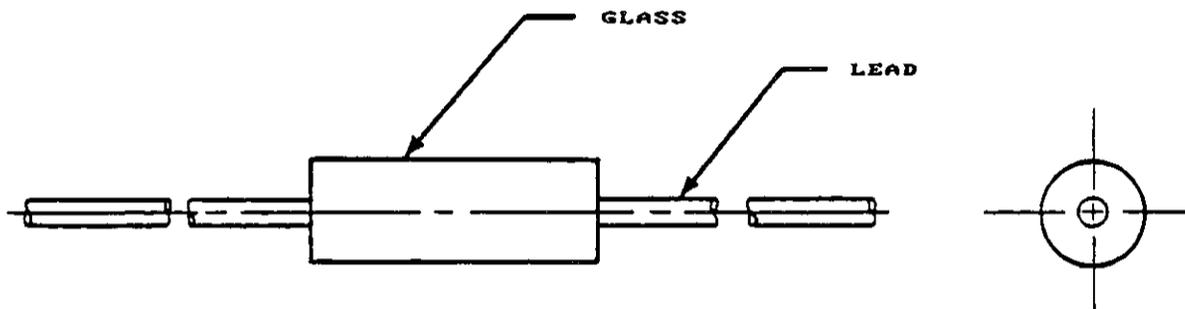


FIGURE 90. A typical voltage-variable capacitor diode package.

4.7.4 Military designation. The military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

4.7.5 Electrical characteristics. Table VII illustrates typical electrical characteristics of voltage-variable capacitors (60-V series).

The values listed in Table VII are typical and for general reference only. For the specific device selected, consult the applicable MIL-S-19500 specification sheet. Diodes are listed by 1N numbers. Refer to subsection 4.1 Diodes, general for definition of electrical terms.

4.7 DIODES, VOLTAGE VARIABLE CAPACITOR

In considering the values of Table VII, reverse current, reverse voltage, capacitance and "Q" (quality factor) are significant parameters in determining performance of the voltage-variable capacitance diode. Of these parameters, the quality factor or Q is perhaps the most important, hence the following discussion is presented to further describe its significance.

Table VII. Voltage-variable capacitance diodes

Device Type	Cap. at V_R pF	V_{dc}	Cap. Ratio from Min.	I_R (μA)	P_T (max)	Q (Min)
1N5139A	6.8	4	2.7	10	400	350
1N5140A	10.0	4	2.8	10	400	300
1N5141A	12.0	4	2.8	10	400	300
1N5142A	15.0	4	2.8	10	400	300
1N5143A	18.0	4	2.8	10	400	250
1N5144A	22.0	4	3.2	10	400	200
1N5145A	27.0	4	3.2	10	400	200
1N5146A	33.0	4	3.2	10	400	200
1N5147A	39.0	4	3.2	10	400	200
1N5148A	47.0	4	3.2	10	400	200

NOTE: These parts are not included in MIL-STD-975. This table is not intended to be a part selection list.

The equivalent circuit of a reverse biased diode in the microwave bands can be represented as shown in Figure 91.

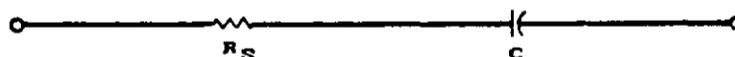


Figure 91. Equivalent circuit.

Where R_S is the series resistance or the spreading resistance of the diode due to the bulk resistivity of the semiconductor material; C is the transition capacitance of the reverse biased junction and is a function of the bias voltage.

Because a small dissipated power in the diode compared with the stored energy is desired, the usual concept of the figure of merit, Q , can be used to characterize the quality of the voltage-variable capacitance diode. By definition:

$$Q = 2\pi \frac{(\text{Average energy stored per cycle})}{(\text{Average energy lost per cycle})} \cong \frac{\text{Reactive power}}{\text{Active power}}$$

4.7 DIODES, VOLTAGE VARIABLE CAPACITOR

For a series RC circuit this becomes:

$$Q = \frac{1}{2\pi f RC}$$

where f is the frequency at which the Q is measured.

A few notes pertinent to this formula when it is applied to characterize voltage-variable capacitance diodes are as follows:

- a. It is a standard practice to measure microwave capacitance diodes at 1 MHz either at the voltage breakdown or at zero bias.
- b. The $Q \cong 1/f$ simple relationship is not followed entirely by these diodes. The deviation increases as the frequency decreases. This fact makes it difficult to extrapolate the Q to other frequencies.
- c. It must be made clear whether or not the influence of cartridge capacitance is included in the quoted Q of a diode. The inclusion of the cartridge capacitance increases the Q , thereby giving false information about the quality of the junction capacitance.

The Q of the best quality voltage-variable capacitance diodes when measured under standard conditions (i.e., 1 MHz, at V_{BR} , junction capacitance only) are between 10 and 16.

C_{min} means the minimum junction capacitance; i.e., the capacitance measured at the breakdown voltage.

4.7.6 Environmental considerations. Typical environmental conditions and screening tests which voltage variable capacitance diodes can withstand are not substantially different from those given in paragraphs 4.1.6.9 Environmental conditions and 4.1.6.10 Screening procedures in subsection 4.1 Diodes, general. For specific devices selected consult the applicable MIL-S-19500 reference slash sheets.

4.7.7 Reliability considerations.

4.7.7.1 Failure modes. Recent development of the JANTXV varicap specification for the IN5139A through 5148A has done much to reduce the incidence of failures. Particular attention should be given to the merits of a reverse bias bake at elevated temperatures. This test screens out diodes with leakage paths on junction surface areas which contribute greatly to many operational life failures.

4.7.7.2 Derating. History has shown that the largest single cause of failure of diodes is operation above allowable levels of thermal and electrical stress. Accordingly, it is imperative that derating of parts be performed to increase the reliability of the system. Users should refer to MIL-STD-975 for specific derating factor guidelines on diodes.

**4.8 DIODES, BIPOLAR TRANSIENT
VOLTAGE SUPPRESSORS (TVSs)**

4.8 Bipolar transient voltage suppressors (TVSs).

4.8.1 Introduction. Transient voltage suppressors (TVSs) consist of two zener diodes oriented back-to-back and sealed in a package that provides adequate thermal dissipation. They exhibit extremely fast response times, low series resistance (R_{ON}), and very high surge voltage handling capabilities. Unlike a zener diode, whose function is voltage regulation, a transient voltage suppressor is designed to protect high voltage sensitive circuits by suppressing transient voltages.

4.8.2 Usual applications. Transients are a major concern of designers. They result from a variety of reasons. The most common are normal switching operations, power supply switching, and circuit disturbances caused by load switching, magnetic coupling and voltage spikes. Voltage transients are a major cause of component failures. Random voltage transient spikes can damage voltage sensitive devices such as ICs, hybrids and MOS devices. Because TVSs exhibit a fast response time and low clamping factor, they can protect these voltage sensitive devices.

TVSs have been used effectively in mobile communication equipment, computer power supplies, airborne avionics, and other applications where inductive and switching transients are present.

When choosing a TVS for a particular application, the following important factors should be considered:

- a. The maximum clamping voltage (V_C) should be determined in order to provide adequate protection for a circuit or component. Once V_C has been determined, it will be used to calculate the power for worst case designs for a given application.
- b. The TVS selected should exhibit a reverse stand-off voltage (V_R) equal to or greater than the circuit operating voltage (maximum ac or dc peak voltage with tolerances).
- c. To select the appropriate TVS one must also determine the transient pulse power (P_p). This can be accomplished by using the simple definition; transient pulse power (P_p) equals the peak pulse current (I_{pp}) multiplied by the clamping voltage (V_C).

$$P_p = V_C \times I_{pp}$$

4.8 DIODES, BIPOLAR TRANSIENT VOLTAGE SUPPRESSORS (TVSs)

4.8.2.1 Microprocessor system TVS applications. The TVS is placed on the signal and input power lines to prevent system failures caused by the effect of switching power supplies, ac power surges and transients, such as electrostatic discharges as illustrated in Figure 92. A TVS across the signal line to ground will prevent transients from entering the data and control buses. TVSs shunted across the power lines ensure a transient-free operating voltage.

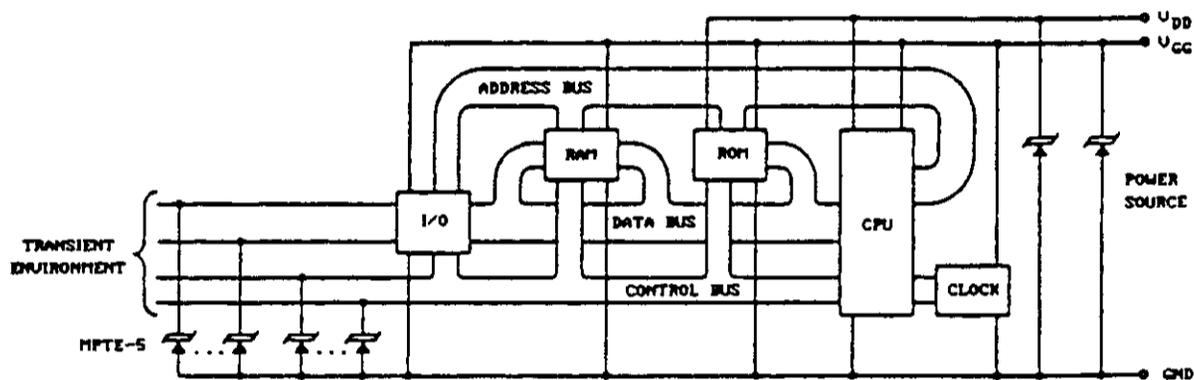


FIGURE 92. Microprocessor system TVS application.

4.8.2.2 DC line TVS applications. A TVS in the output of a voltage regulator can replace many components used as protection circuits such as the crowbar circuit illustrated in Figure 93. It may also be used to protect the bypass transistor from voltage spikes across the collector to emitter terminals.

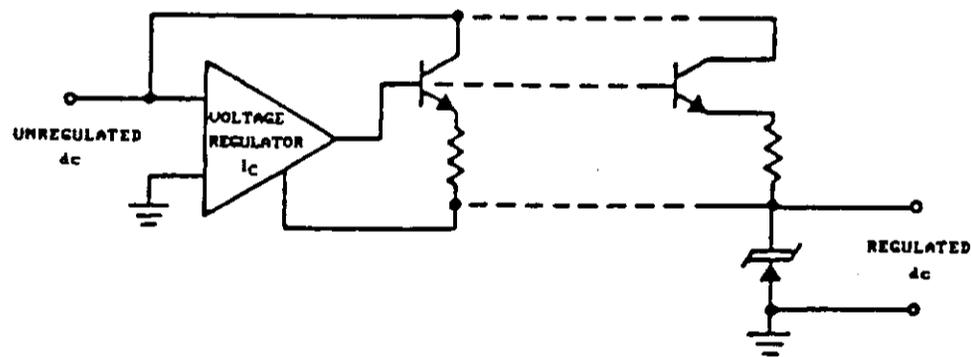


FIGURE 93. DC line TVS application.

4.8 DIODES, BIPOLAR TRANSIENT VOLTAGE SUPPRESSORS (TVSs)

4.8.2.3 Relay and solenoid TVS applications. The coil inductance of a solenoid or relay can release energy that can damage contacts or drive transistors. A TVS used as shown in Figure 94 would provide adequate protection.

The proper TVS can be selected by determining peak pulse power (P_p) and pulse time (t_p). Knowing the values of V_{CC} , L , and R_L the following equations can be used to determine P_p and t_p .

$$I_0 = \frac{V_{CC}}{R_L}$$

$$P_p = I_p \times V_C$$

$$t_1 = \frac{V_{CC}/R_L}{V_C/L}$$

$$t_p = \frac{t_1}{2}$$

Figure 95 shows peak pulse power, P_p versus pulse duration, t_p for a given TVS device.

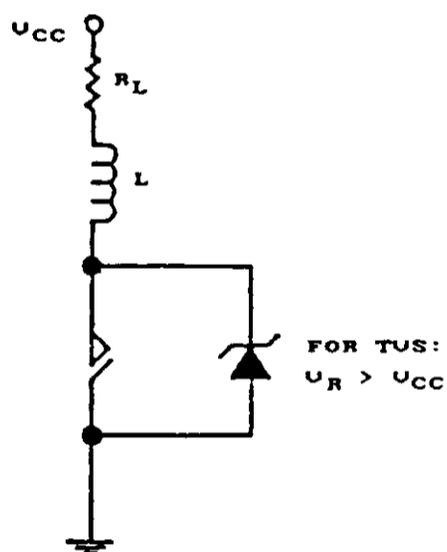
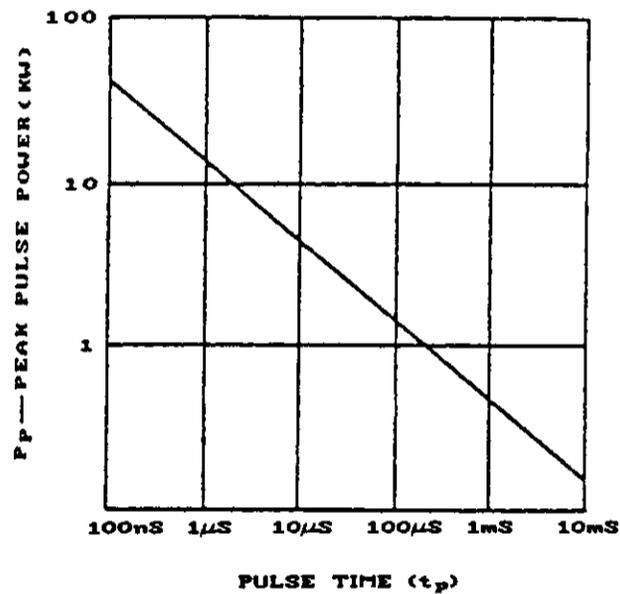


FIGURE 94. Coil and contacts, dc.

**4.8 DIODES, BIPOLAR TRANSIENT
VOLTAGE SUPPRESSORS (TVSs)**



Max. duty cycle = 0.1%.

FIGURE 95. Peak pulse power vs pulse duration.

The applications described herein are but a few of the many possible uses of the transient voltage suppressor.

4.8.2.4 Protecting switching power supplies. A designer needs to protect against three types of transients: load transients, line transients, and internally generated transients. Because transients have high energy levels they cause improper operation and component failure. Protecting a switching power supply is an example of utilizing TVS devices. The following are typical components that need to be protected in a switching power supply:

- a. High voltage switching transistors
- b. Rectifiers
- c. Output rectifiers
- d. Control circuitry.

Figure 96 shows a typical switching power supply with TVS devices used for protection in voltage sensitive areas.

4.8 DIODES, BIPOLAR TRANSIENT VOLTAGE SUPPRESSORS (TVSs)

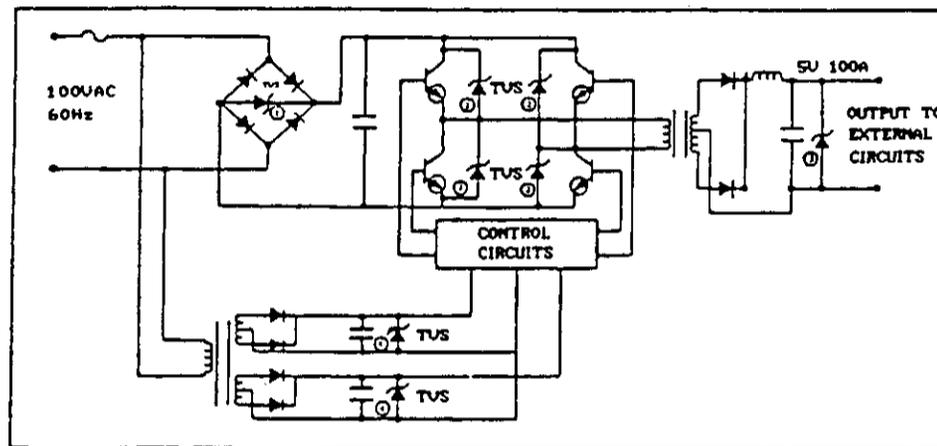


FIGURE 96. Typical switching power supply.

4.8.3 Physical construction. The transient voltage suppressor diode is generally made using two passivated diffused, planar or diffused planar junction on an epitaxial substrate process die placed back-to-back in a single glass-to-metal or double-slug package. The three basic die processes are passivated diffused, planar, and diffused planar junction on epitaxial substrate, as shown in Figures 97, 98, and 99, respectively.

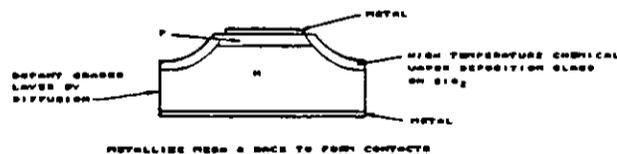


FIGURE 97. Passivated diffused process.

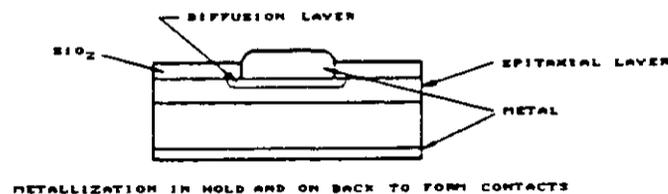


FIGURE 98. Planar process.

**4.8 DIODES, BIPOLAR TRANSIENT
VOLTAGE SUPPRESSORS (TVSs)**

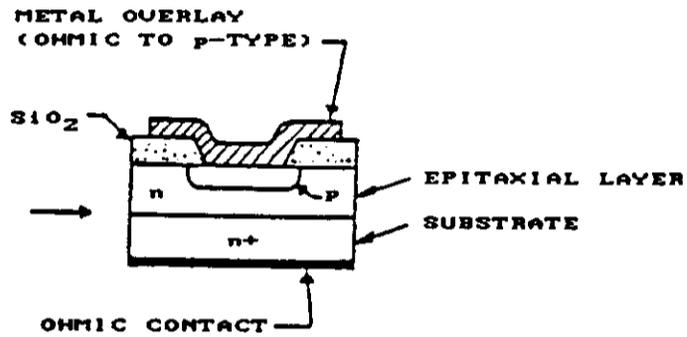


FIGURE 99. Diffused planar junction on epitaxial substrate.

4.8.3.1 Back contact. Illustrated in Figure 100 below is the back contact technique used in constructing transient voltage suppressor (TVS) diodes. TVS diodes generally use high temperature alloy back contacts, associated with double-slug construction.

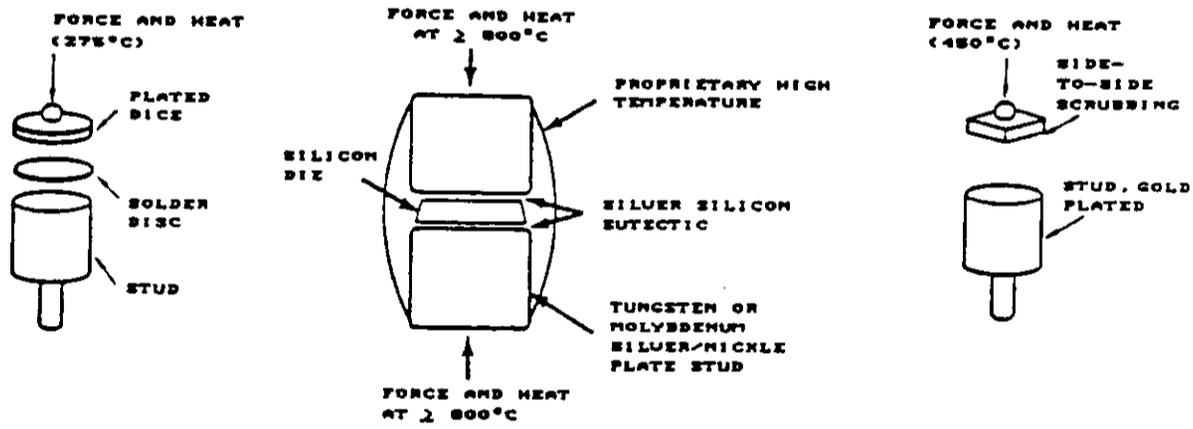


FIGURE 100. Back contact formation techniques.

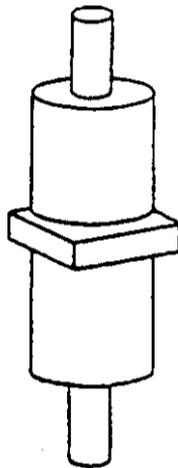
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4.8 DIODES, BIPOLAR TRANSIENT VOLTAGE SUPPRESSORS (TVSs)

The silicon-silver alloy contact which is generally associated with the high temperature alloy has good mechanical strength and thermal conductivity. The silicon chip is metallurgically brazed to plated molybdenum or tungsten slugs at temperatures greater than 700 °C.

Molybdenum, tungsten, and dumet are used for studs in glass packages because their thermal expansion coefficients match those of certain glasses, allowing sealing of these glasses to them.

4.8.3.2 Front contacts. Electrical contact is made to the die as shown in Figure 101.



DOUBLE-SLUG CONTACT

FIGURE 101. Various styles of electrical contact.

The double-slug contact provides good heat transfer, thus raising the power rating of the package.

4.8.3.3 Seals. The double-slug package in Figure 102 is made by metallurgical bonding (700 °C) both sides of the die to either molybdenum or tungsten slugs to which a glass sleeve or glass slurry is fused for final sealing. The final step is a lead brazing operation. This type of construction is most often associated with transient voltage suppressor diodes.

4.8 DIODES, BIPOLAR TRANSIENT VOLTAGE SUPPRESSORS (TVSs)

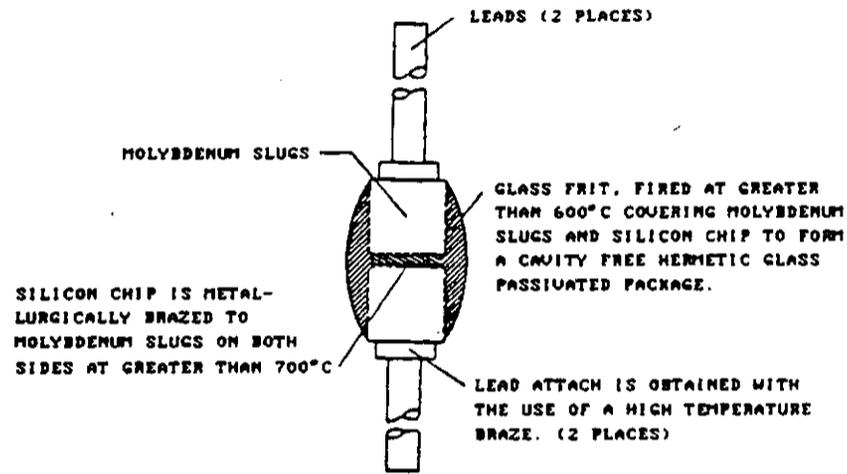


FIGURE 102. Sealing techniques.

Figure 103 shows a typical transient voltage suppressor package.

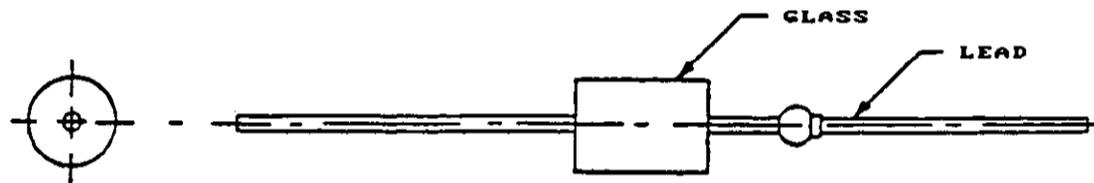
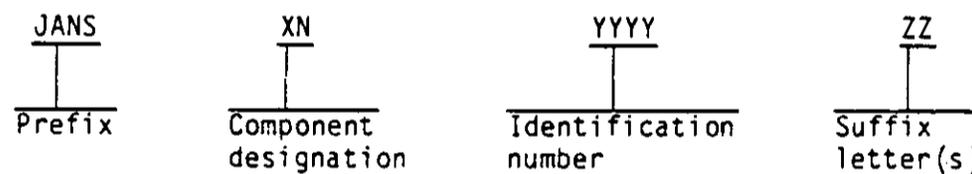


FIGURE 103. Typical transient voltage suppressor (D0-13 package).

4.8.4 Military designation. Military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

The identification number is assigned in order of registration and has no other significance.

4.8 DIODES, BIPOLAR TRANSIENT VOLTAGE SUPPRESSORS (TVSs)

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R) or any other letter to indicate a modified version.

4.8.5 Electrical characteristics.

- a. Voltage-ampere characteristics. The volt-ampere characteristic for a typical transient voltage suppressor, TVS, given in Figure 104 shows that the device can conduct current in both directions. Characterized by its two zener die placed back to back, the TVS is useful for bi-directional transient suppression.

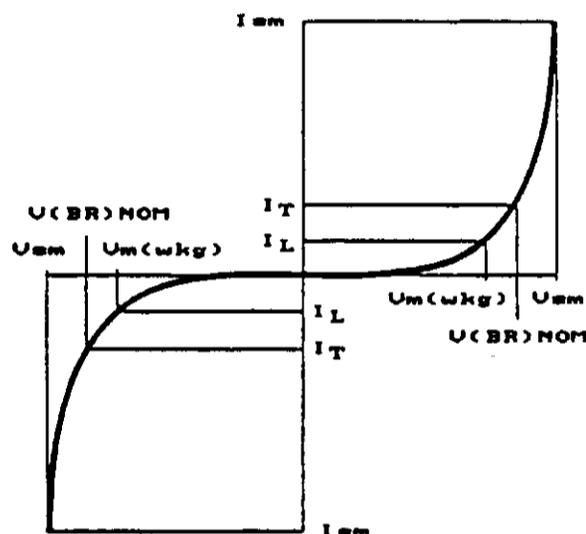


FIGURE 104. Typical characteristic curve for bi-directional transient suppressor.

- b. Working peak voltage (V_M , WKG) or stand-off voltage (V_R). This parameter is the maximum permissible dc working voltage. It is the highest reverse voltage at which the TVS will be nonconducting.
- c. Maximum peak surge voltage (V_{SM}) or maximum clamping voltage ($V_{C,max}$). This is the maximum voltage drop across the TVS while it is subjected to the peak pulse current, usually for 1 ms.
- d. Minimum breakdown voltage (BV_{min}). The breakdown voltage is the reverse voltage at which the TVS conducts. This is the point where the TVS becomes a low impedance path for the transient.

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**4.8 DIODES, BIPOLAR TRANSIENT
VOLTAGE SUPPRESSORS (TVSs)**

- e. Test current (I_T). This current is the zener current at which the nominal breakdown voltage is measured.
- f. Maximum leakage current (I_L). This current is the current leakage measured at the maximum dc working voltage (V_M or V_R).
- g. Maximum peak surge current (I_{SM}). This current is the maximum permissible surge current.

4.8.6 Environmental considerations. Typical environmental conditions and screening tests which transient voltage suppressors are capable of withstanding are not substantially different from those given in paragraphs 4.1.6.9 Environmental considerations and 4.1.6.10 Screening procedures in subsection 4.1 Diodes, general. For the specific device selected, consult the MIL-S-19500 reference slash sheet.

4.8.7 Reliability considerations.

4.8.7.1 Failure mechanisms and data. In extensive tests of transient voltage suppressors, the three most critical parameters were peak pulse power, peak surge voltage and reverse leakage current. Because of its inherent use as a transient suppressor, the TVS must sustain high voltage pulse power. Opens or shorts caused by cracking of the die are main-failure modes if peak reverse power rating is exceeded. Long term life tests have yielded a significant number of failures in these areas. Failure mechanisms of TVSs are similar to those of other diode types discussed in paragraph 4.1.6 General reliability considerations in subsection 4.1 Diodes, general.

4.8.7.2 Screening. As with previous recommendations for other diode types, the transient voltage suppressor procured as a JANTXV or JANS device offers the best protection against the above problems. The two characteristics which are usually measured to verify TVS quality are the breakdown voltage (BV) and reverse leakage current (I_R). Measurement of these characteristics during burn-in screening will verify the reliability of the diode. The magnitude of reverse leakage current is largely dependent upon surface conditions, die size, thickness, and junction area of the semiconductor. Delta measurements of reverse leakage during burn-in is a useful criterion in determining the long term reliability of the device.

4.8.7.3 Derating. History has shown that the largest single cause of failure of TVSs is operation above allowable levels of thermal and electrical stress. Accordingly, it is imperative that derating of parts be performed to enhance reliability. Users should refer to MIL-STD-975 for specific derating factor guidelines. The design and use of a part should always be within the thermal and electrical stresses defined. High temperature operation is the most destructive stress a semiconductor device could be subjected to. It will result in electrical parameter drift and general degradation of the electrical and mechanical characteristics of the device.

4.9 DIODES, MONOLITHIC AND MULTIPLE ARRAYS

4.9 Monolithic and multiple arrays.

4.9.1 Introduction. Diode arrays consist of several diode circuits fabricated in an integrated circuit form. The diode array is not a unique device with characteristics that differ from discrete diodes. It is simply a packaging technique that permits a significant reduction in the size of electrical systems. The arrays are available in either 10- or 14-ceramic flat packs or ceramic 14- and 16-lead dual-in-line packages, for high density packaging and reliability.

Although not complex in operation, these arrays are widely used in applications requiring moderate current, fast switching, or digital logic operations.

The military designation, electrical characteristics, environmental conditioning and reliability of diode arrays are similar to those of their discrete diode counterpart.

4.9.2 Usual applications. Digital logic circuits commonly use diode arrays that can be used to construct logic gates. Figure 105 is an example of such a circuit; it is referred to as an AND gate.

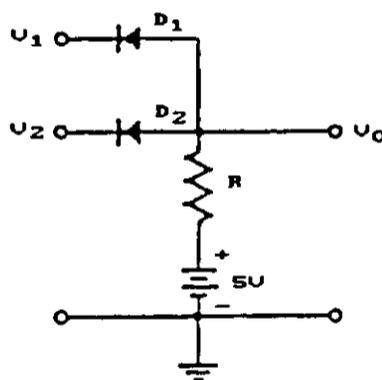


FIGURE 105. Diode logic gate, AND gate.

In the operation of an AND gate, all the signals are binary digital signals and have only one of two possible states (low and high). In Figure 105, the output voltage (V_0) becomes high only if V_1 and V_2 are high. However, if either of these voltages are zero, the corresponding diode is forward-biased, and V_0 becomes short-circuited by the diode and becomes zero (low state).

4.9 DIODES, MONOLITHIC AND MULTIPLE ARRAYS

Diode arrays are also used to construct read only memories (ROMs) where the data to be stored is in the form of a table. These are called "look-up tables." See Table VIII, which shows an e^{-x} data table using 4-bit words.

The e^{-x} look-up table can be implemented by using a 4- to 16-line demultiplexer such as a 74154 silicon microcircuit. This device responds to a 4-bit binary input by making its appropriate output low. To convert the 74154 into a ROM, diode arrays are connected as shown in Figure 106. When a given output line goes low and a diode is connected from V_{CC} to the output line of the 74154, the output bit goes low because the diode conducts. On the other hand, if a diode is not connected from V_{CC} to the 74154 output line, that output bit remains high. The diode array forms the ROM portion of the circuit.

TABLE VIII. e^{-x} data table

x	Binary	e^{-x}	Binary	Decimal Equivalent
0.0	0000	1.00	1111	15
0.1	0001	0.90	1110	14
0.2	0010	0.82	1100	12
0.3	0011	0.74	1011	11
0.4	0100	0.67	1010	10
0.5	0101	0.61	1001	9
0.6	0110	0.55	1000	8
0.7	0111	0.50	1000	8
0.8	1000	0.45	0111	7
0.9	1001	0.41	0110	6
1.0	1010	0.37	0110	6
1.1	1011	0.33	0101	5
1.2	1100	0.30	0101	5
1.3	1101	0.27	0100	4
1.4	1110	0.25	0100	4
1.5	1111	0.22	0011	3

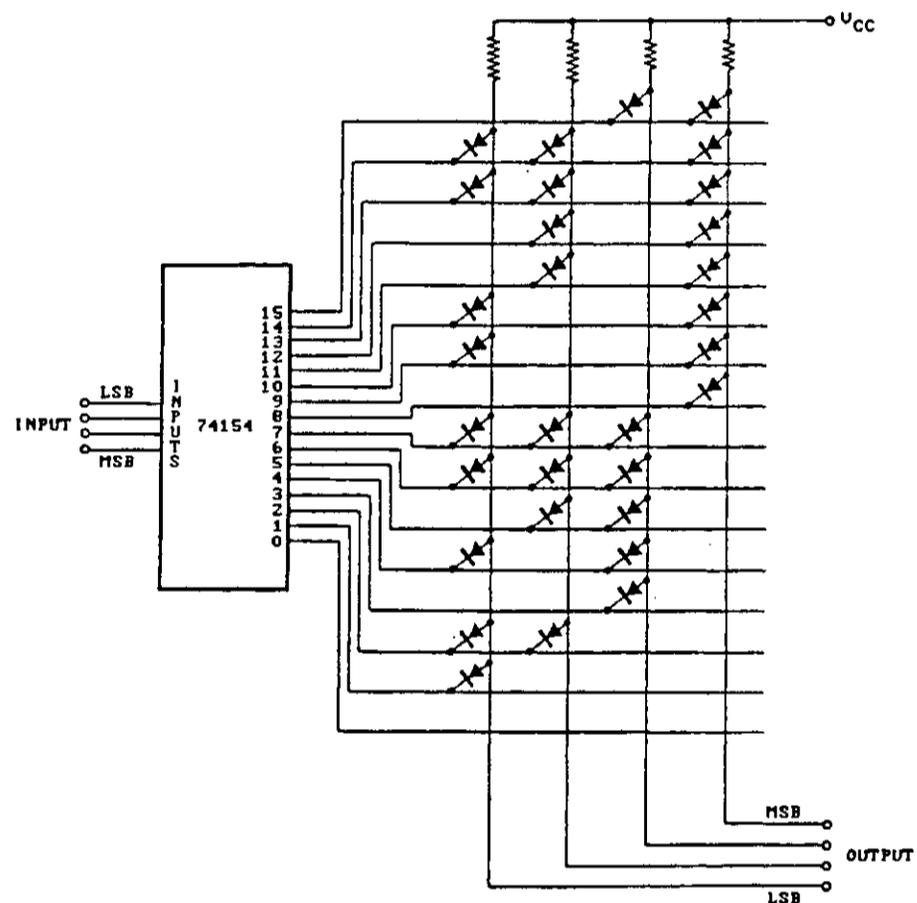
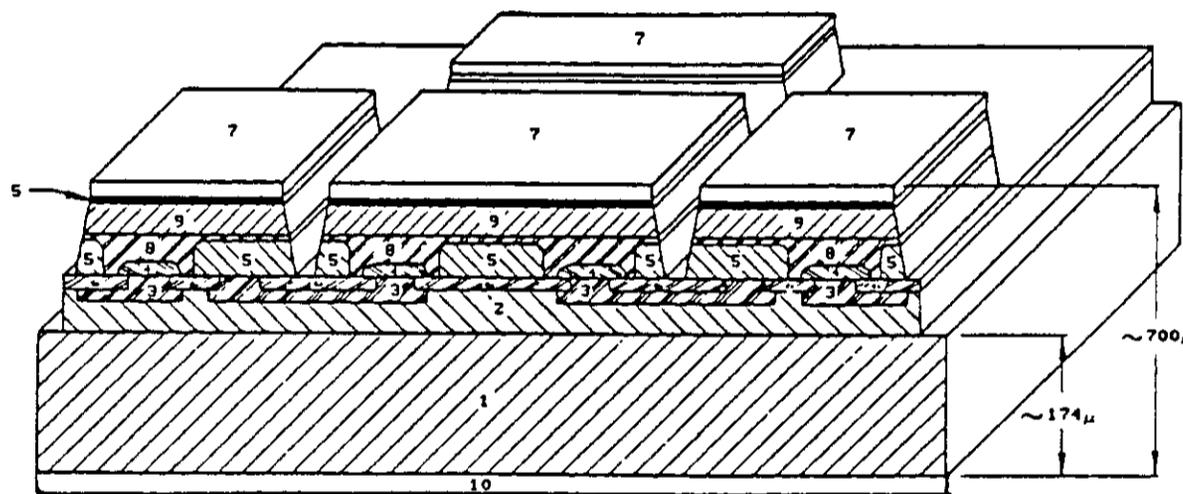
4.9 DIODES, MONOLITHIC AND
MULTIPLE ARRAYS

FIGURE 106. A ROM synthesized from a 74154 demultiplexer and the appropriate diode connections as a look-up for exponentiation (see Table XI).

4.9.3 Physical construction. All air-isolated monolithic planar or multiple diode arrays use the same fabrication process to facilitate the manufacture of diode arrays in integrated circuit form. The elimination of parasitic capacitance (which is a very important consideration in overall switching speed) is accomplished through air isolation of the diode circuits. Air isolation is obtained by selective chemical etching of the whole device wafer from the backside of the silicon wafer as shown in Figure 107.

4.9 DIODES, MONOLITHIC AND MULTIPLE ARRAYS

The process used by the manufacturer is a very general one, capable of fabricating a wide range of diode arrays. The process supplies a hermetically sealed structure at the chip level, but of integrated circuit size as compared with discrete diodes with similar electrical characteristics. The excellent hermeticity is the result of using silicon nitride over the device structure in conjunction with a unique fused glass sandwich construction. A physical diagram representing a typical diode array is shown in Figure 107, with Arabic numerals identifying the materials used in its construction.



- | | |
|--|---|
| 1. Support wafer (silicon backing wafer) | 6. Passivation |
| 2. Glass dielectric layer | 7. Aluminum bonding pads (6- μ thick) |
| 3. Aluminum Interconnects (3 μ thick) | 8. N type epitaxial layer |
| 4. P ⁺ Junction | 9. N ⁺ substrate |
| 5. N ⁺⁺ Cathode contact (channel stops) | 10. Gold film backing for mounting |

FIGURE 107. Typical construction of an air isolated monolithic or multiple diode array.

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4.9 DIODES, MONOLITHIC AND MULTIPLE ARRAYS

In the process two silicon wafers are used, one containing the diffused junction configuration desired for the device, and the other the mechanical support during hot processing. The passive wafer contains no device structures, nor is it called upon to provide any electrical contact function. All contacts to the devices are provided from the "top" side of the finished chip, which is the "bottom" side of the active device wafer. The two wafers are sandwiched and sealed together using a fused glass frit between the wafers. Prior to sandwich sealing, the diffused side of the device wafer is first metallized with aluminum to provide the necessary interconnects between junctions, and it is subsequently coated with a glass powder slurry. The second mechanical support wafer is then applied to the device wafer and the two are sealed together under moderate compressive force by heating both wafers to an elevated temperature at which the glass frit fuses to form a glass seal between the two wafers. The assembled "sandwich" is then etched using an aluminum metal masking pattern to produce localized mesas. The mesas are inverted relative to the original diffusion orientation. These metallized mesa tops provide bonding pads for termination of the chip. A nitric acid rich, oxidizing, silicon-etch solution containing hydrogen fluoride (HF) is used for this purpose. The oxidizing nitric acid-rich etchant insures no attack on the aluminum back contact mask pattern. The mesa etching is then stopped when it "bottoms out" at the silicon nitride layer.

In this way, the manufacturer is able to create a wide range of diode device arrays using this process without resorting to material other than an N/N⁺ epitaxial device wafer. Therefore, a great variety of diode array structures may be produced by judicious selection of mesa etch masking and back contact configurations. The flexibility of the process in producing common anode and common cathode diode configurations is demonstrated in the generalized schematic device cross section shown in Figure 107.

For cathode contacts, an ultrasonic bond is made using a 1.5 mil aluminum wire connecting the chip bond pad areas (item 7 in Figure 107) to the lead posts inside the package. For anode contacts, a similar aluminum-to-aluminum ultrasonic bond is made to the aluminum bonding pads on the glass dielectric (item 2, Figure 107) which are exposed after etching. Pad configuration is not shown.

The diode arrays come in various hermetically sealed ceramic flat packages as illustrated in Figures 108 and 109.

4.9 DIODES, MONOLITHIC AND
MULTIPLE ARRAYS

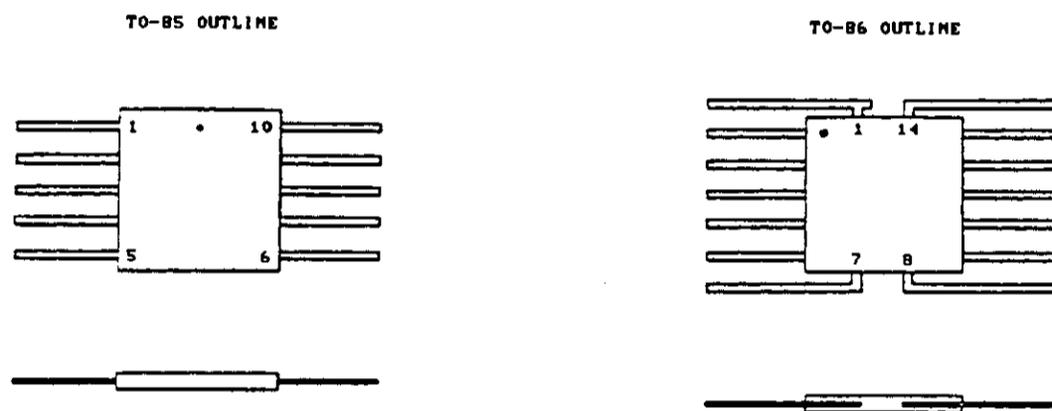


FIGURE 108. Package configurations for multiple arrays.

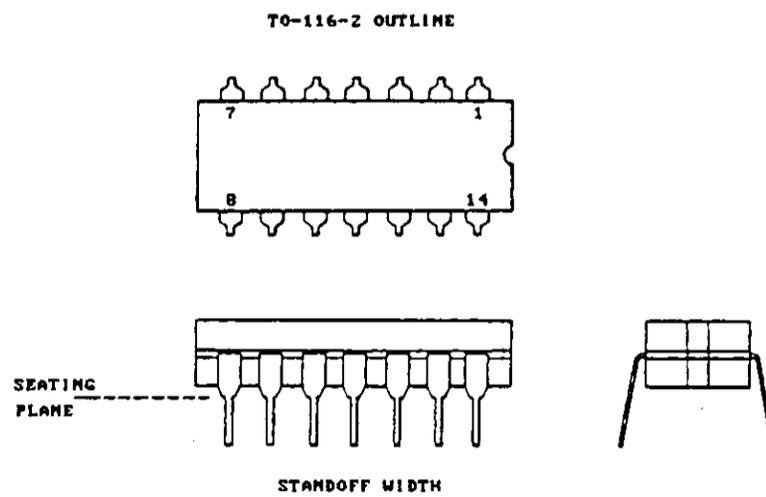
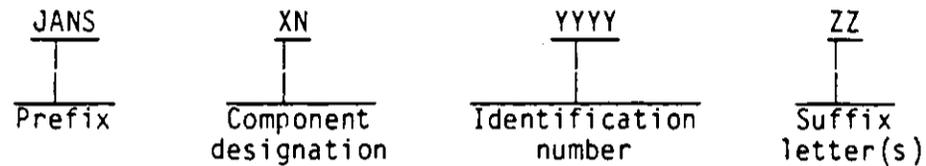


FIGURE 109. Package configuration for monolithic array.

**4.9 DIODES, MONOLITHIC AND
MULTIPLE ARRAYS**

4.9.4 Military designation. Military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

4.9.5 Electrical characteristics.

4.9.5.1 Multiple arrays. The basic diode arrays are the 1N5768, 1N5770, 1N5772 and 1N5774. These arrays consist of four different diode circuit configurations

which come in either a T0-85 or T0-86 hermetically sealed ceramic flat package, as outlined in Figure 110.

Basic device characteristics and parameter information on diode arrays are similar to those discussed in 4.1 section, paragraph 4.1.5 General parameter information in subsection 4.1 Diodes, general.

4.9 DIODES, MONOLITHIC AND MULTIPLE ARRAYS

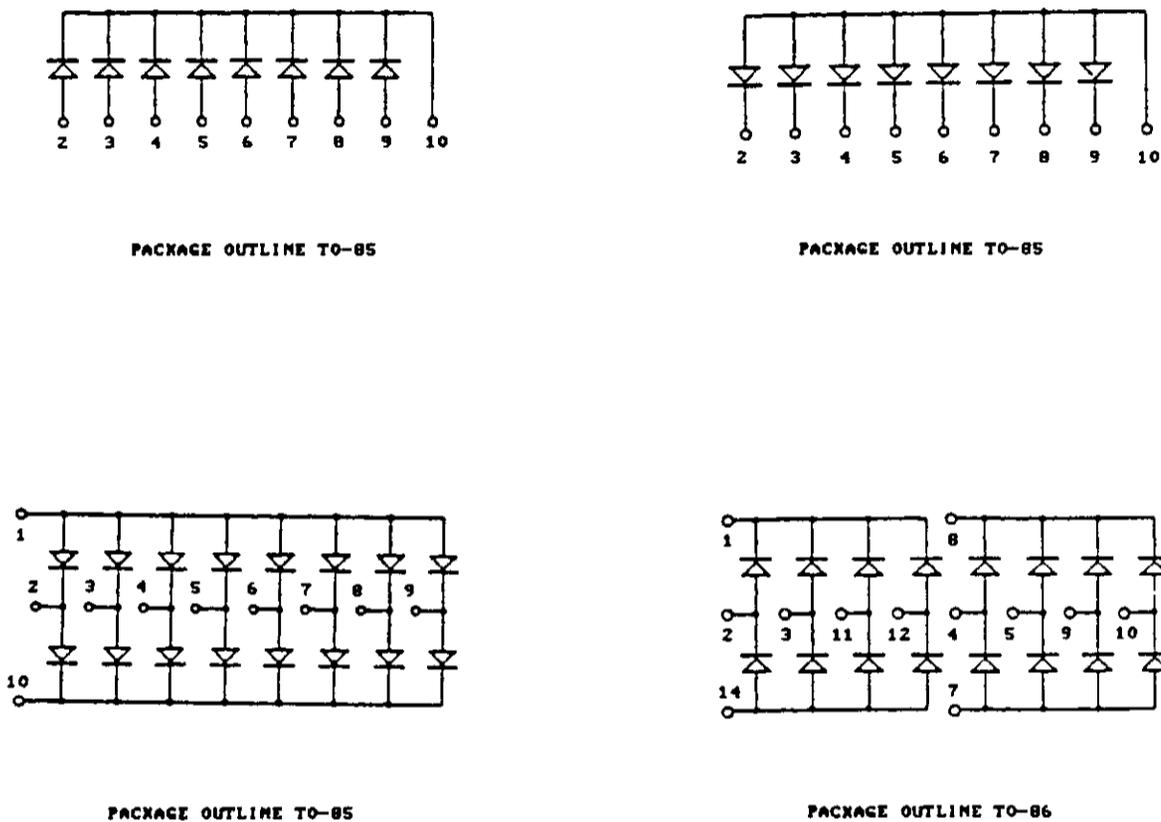


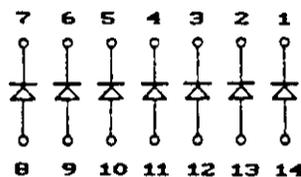
FIGURE 110. Multiple diode array circuit configurations.

4.9.5.2 Monolithic arrays. The 1N6100 and 1N6101 devices are functionally identical but come in different package configurations as shown in Figure 111. Using the same manufacturing process as outlined under paragraph 4.9.3 Physical construction, the diode array can be wired for the application at hand while conserving valuable space on the printed circuit board. Also, the reverse recovery time (max) is 5 ns and not 20 ns as for the 1N5768, 1N5770, 1N5772 and 1N5774 series.

Basic device characteristics and parameter information on diode arrays are similar to those discussed in subsection 4.1, Diode-general and paragraph 4.1.5.1.

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**4.9 DIODES, MONOLITHIC AND
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PACKAGE OUTLINES

TO-86	1N6100
TO-116-2	1N6101

FIGURE 111. Monolithic diode array.

4.9.6 Environmental considerations. Typical environmental test conditions and screens which switching diode arrays are capable of withstanding are not substantially different from those given in paragraphs 4.1.6.9 Environmental consideration and 4.1.6.10 Screening procedure, is in subsection 4.1, Diodes-general.

4.9.7 Reliability considerations.

4.9.7.1 Failure modes. The statements of paragraph 4.2.7.1 Reliability considerations for switching diodes are applicable to switching diode arrays.

Considering the material design and construction of the diode array (see paragraph 4.9.3 Physical construction), the entire active portion of the device is buried in the glass dielectric layer and is inaccessible by any contaminant once the device is properly wire bonded and packaged.

4.9.7.2 Derating. The derating for voltage, current and forward surge current should be 50 percent of the value published in the vendor data sheet.

Users should refer to MIL-STD-975 for derating factor guidelines.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

4.10 Silicon controlled rectifiers (SCR)

4.10.1 Introduction. The silicon-controlled rectifier (SCR) belongs to the thyristor family. It is a semiconductor device that can be switched between two states, from off to on, or from on to off, by a current or polarized voltage pulse. Unlike conventional transistors, the device lends itself to use as a high-current high-voltage rectifier, a static latch limited to microseconds, and a sensitive high-gain amplifier control. As shown in Quadrant I of Figure 112, under forward-bias conditions (anode positive with respect to the cathode) the SCR has two states. At low values of forward-bias, the SCR exhibits a very high impedance; in the forward blocking or off state, a small forward current, called the forward off-state current, flows through the device. As the forward-bias is increased, however, a voltage point is reached at which the forward current increases rapidly, and the SCR switches to the on state. This value of voltage is called the breakover voltage. When the SCR is in the on state, the forward current is limited primarily by the impedance of the external circuit.

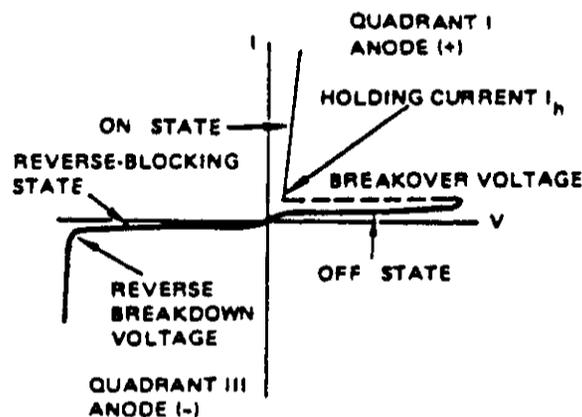


FIGURE 112. Characteristics of a silicon controlled rectifier.

Under reverse bias (anode negative with respect to cathode) as shown in quadrant III of Figure 112, the SCR exhibits a very high internal impedance, and only a small amount of current, called the reverse blocking current, flows through the device. The current remains very small and the device remains "off" unless the reverse voltage exceeds the reverse breakdown voltage limitation. At this point, the reverse current increases rapidly and the SCR undergoes thermal runaway, a condition that normally causes irreversible damage. The value of the reverse breakdown voltage generally is greater than the forward breakover voltage for most types of SCRs.

The breakover voltage of the SCR can be controlled or varied by application of a pulse at the gate electrode. Figure 113 demonstrates that as the amplitude of the pulse is increased, the breakover voltage decreases until the curve

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

closely resembles that of a rectifier. In normal operation, the SCR is operated with critical values well below the breakover voltage and is made to switch on by gate signals of sufficient magnitude to assure that the device is switched to the "on" state at the desired instant.

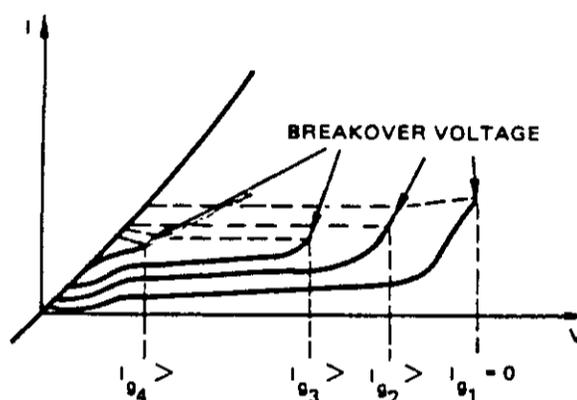


FIGURE 113. Curves showing the forward-voltage characteristics of an SCR for different values of gate current.

After the device is switched on (triggered) by the gate signal, the current through the device is independent of the gate voltage or current. The SCR remains in the on state until the principal current is reduced to a level below that required to sustain conduction.

Basic SCR types are offered with current ratings extending from 0.5 to 2000 A rms with voltage ratings spanning the range of 25 to 3000 V peak. Many specialized SCR types also are available, including high speed inverter SCRs, SCRs for use over very wide temperature ranges, light-activated SCRs, very high voltage SCRs, and SCRs tested to very rigid quality levels for high reliability applications. All graphs and charts are for typical SCR devices and not for a specific device. Reference should be made to the component-detailed specification for particular characteristic parameters.

4.10.2 Usual applications. The silicon-controlled rectifier is well adapted for use as a latching switch or high power gain amplifier. The SCR can be turned on by a momentary application of control current applied to the control gate, while tubes or transistors require a continual on signal. The turn-on time is about 1 μ s, and turn-off time about 10 to 20 μ s. This latching action can be controlled by signals of only a few microwatts and can switch up to 200 V. With associated transistor triggering circuits, unusual current gain of many billions can be obtained. Because of this advantage there are many applications to which the SCR can be adapted. A few of these applications follow.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

4.10.2.1 The dc static switch. Figure 114 illustrates a static SCR switch for use in a dc circuit. When a low power signal is applied to the gate of SCR₁, this SCR is triggered and voltage is applied to the load. The right hand plate of C charges positively with respect to the left hand plate through R₁.

When SCR₂ is triggered, capacitor C is connected across SCR₁, so that this SCR is momentarily reverse biased between anode and cathode. This reverse voltage turns SCR₁ off and interrupts the load current, provided that the gate signal is not applied simultaneously to both gates.

SCR₁ should be selected so that the maximum load current is within its rating. SCR₂ need conduct only momentarily during the turn-off action; it can be smaller in rating than SCR₁. The minimum value of commutating capacitance C can be determined by the following equations:

$$\text{for resistive load: } C = 1.5t_{\text{off}}I/E \text{ } \mu\text{F}$$

$$\text{for inductive load: } C = t_{\text{off}}I/E \text{ } \mu\text{F}$$

where

t_{off} = turn-off time of SCR in microseconds

I = maximum load current, including possible overloads, in amperes at time of commutation

E = minimum dc supply voltage

The resistance of R₁ should be ten to one hundred times less than the minimum effective value of the forward blocking resistance of SCR₂. This latter value can be derived from the published leakage current curves for the SCR under consideration.

In some cases, a mechanical switch may be substituted for SCR₂ to turn off SCR₁ when it (the switch) is momentarily closed. Many other useful variations of this basic dc static switch can be devised.

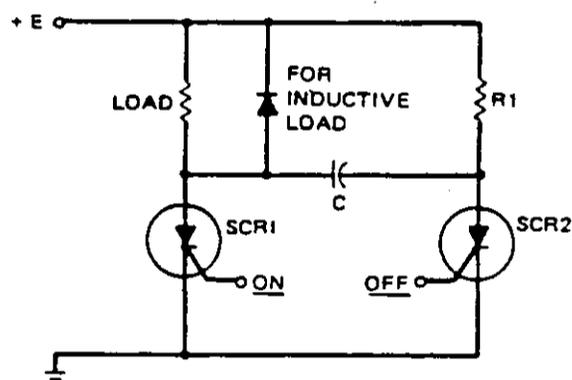
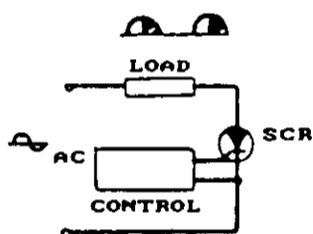


FIGURE 114. The dc static switch.

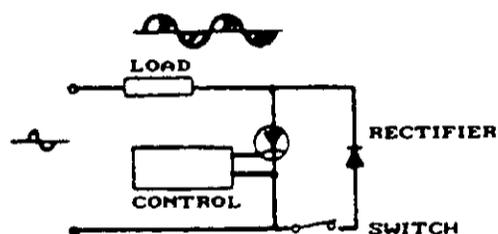
4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

4.10.2.2 Principle of phase control. Phase Control is the process of rapid on-off switching which connects an ac supply to a load for a controlled fraction of each cycle. This is a highly efficient means of controlling the average power to loads such as lamps, heaters, motors, and dc supplies. Control is accomplished by governing the phase angle of the ac wave at which the SCR is triggered. The SCR will then conduct for the remainder of that half-cycle.

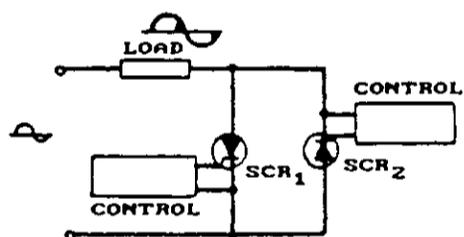
There are many forms of phase control possible with the SCR as shown in Figure 115. The simplest form is the half-wave control of Figure 115A, which uses one SCR for control of current flow in one direction only.



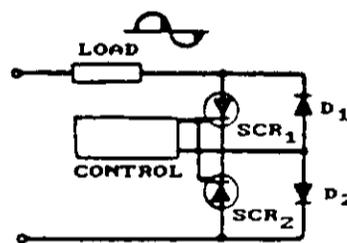
A. Controlled half-wave



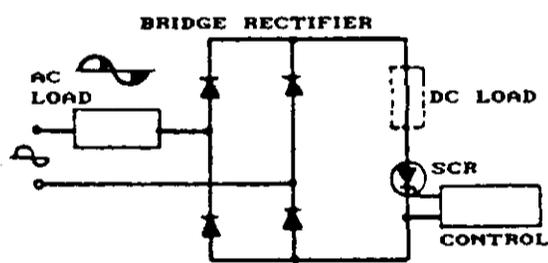
B. Controlled half plus fixed half-wave



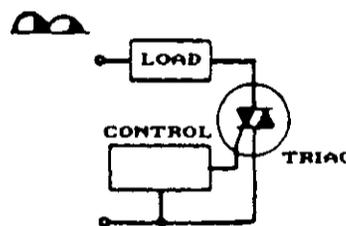
C. Controlled full wave



D. Controlled full wave



E. Controlled full wave for ac or dc



F. Full wave control with triac

FIGURE 115. Basic forms of ac phase control.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

This circuit is used for loads which require power control from zero to one-half of full-wave maximum and permit or require dc. The addition of one rectifier, as shown in Figure 115B, provides a fixed half-cycle of power which shifts the power control range to half-power minimum and full-power maximum but with a strong dc component. The use of two SCRs, shown Figure 115C, provides control from zero to full-power and requires isolated gate signals, either as two control circuits or pulse-transformer coupling from a single control. Equal triggering angles of the two SCRs produce a symmetrical output wave with no dc component.

Reversible half-wave dc output is obtained by controlling the symmetry of the triggering angle.

An alternate form of full-wave control is shown in Figure 115D. This circuit has the advantage of a common cathode and gate connection for the two SCRs. Although the two rectifiers prevent reverse voltage from appearing across the SCRs, they reduce circuit efficiency by their added power loss during conduction.

The most flexible circuit, shown Figure 115E, uses one SCR inside a bridge rectifier and may be used for control of either ac or full-wave rectified dc. Losses in the rectifiers, however, make this the least efficient circuit form and commutation is sometimes a problem. On the other hand, using one SCR on both halves of the ac wave is a more efficient utilization of SCR capacity, hence the choice of circuit form is based on economic factors as well as performance requirements.

4.10.2.3 Inverter configurations. Rectifier circuits occur in several configurations such as half-wave, full-wave, and bridge. Inverter circuits may be grouped in an analogous manner. Figure 116 shows the different types of configurations. Methods of triggering and commutation have been left out for clarity.

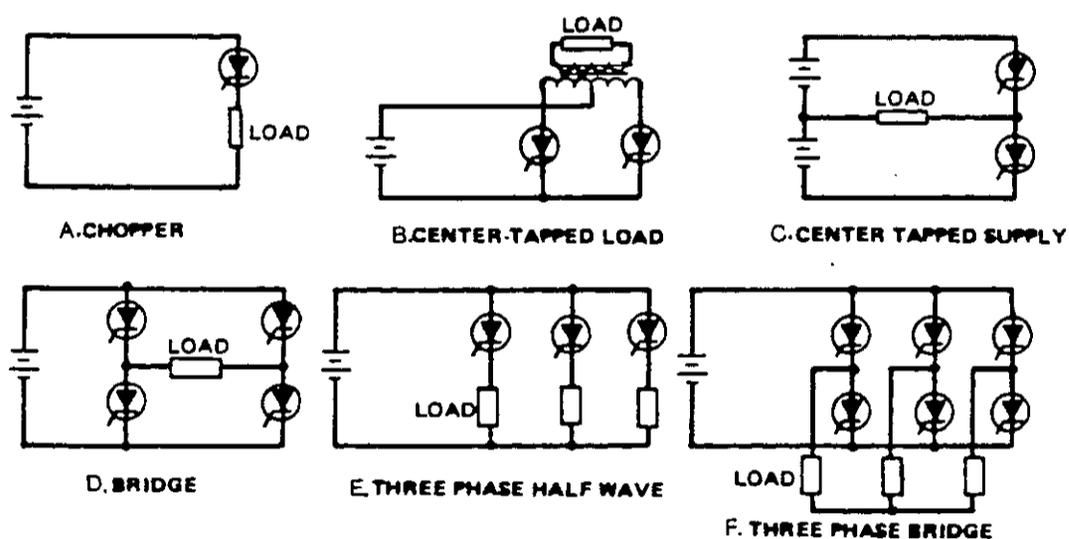


FIGURE 116. Inverter configurations.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

4.10.2.4 Pulse modulator switches. The conventional pulse modulator circuit using SCRs as switches operates as a Class A inverter, as shown in Figure 117.

The current pulses through the SCR are narrow, ranging typically from 0.1 to 10 μ s base width. The repetition rate is from several hundred to several thousand cycles per second.

Turn-off time and dv/dt are usually not critical characteristics of SCRs for this application. The most critical static characteristic is blocking voltage and the most critical dynamic characteristic is a high value of di/dt . The stress of high values of di/dt may be alleviated by using a saturating reactor in series with the SCR.

To minimize jitter and the delay and rise time, the gate of the SCR should be driven as hard as the ratings permit. The rise time of the gate pulse should be much less than 1 μ s.

High frequency SCRs make ideal pulse modulator switches.

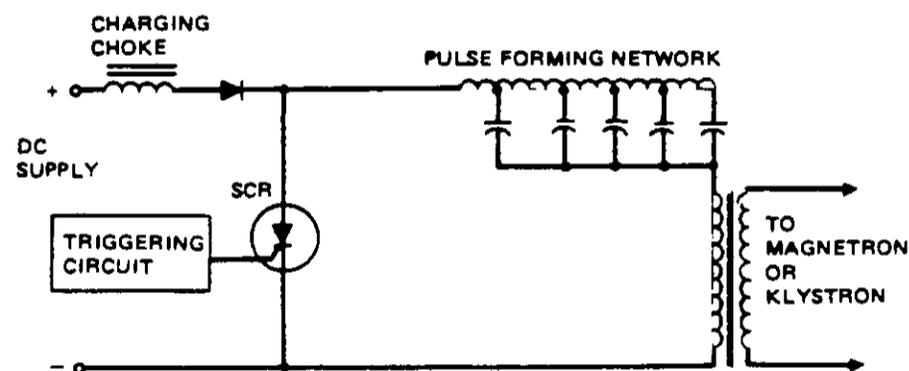


FIGURE 117. Basic pulse modulator circuit.

4.10.3 Physical construction. An SCR is a four-layer p-n-p-n device that has three electrodes, a cathode, anode, and control gate. See Figures 118, 119 and 120 for junction diagrams, typical cross-section views, and typical DO-5 package outline drawing, respectively.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

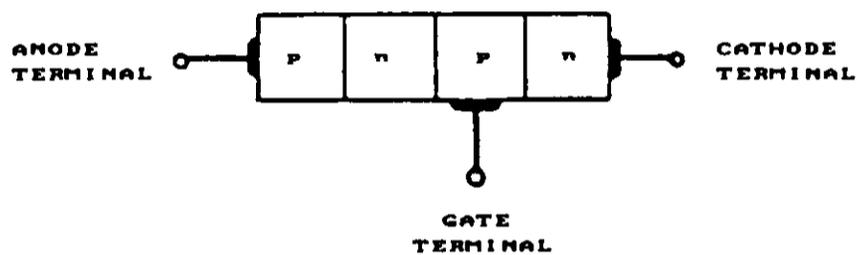


FIGURE 118. Typical p-n-p-n junction diagram.

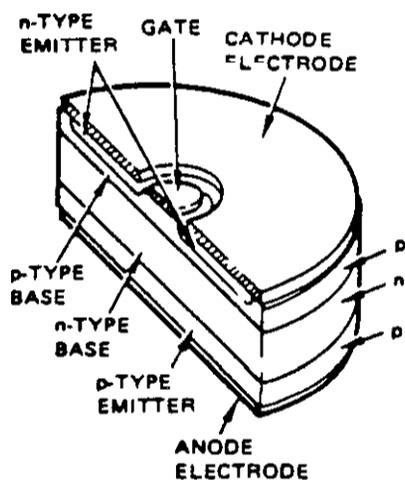


FIGURE 119. Cross-section of a typical SCR pellet.

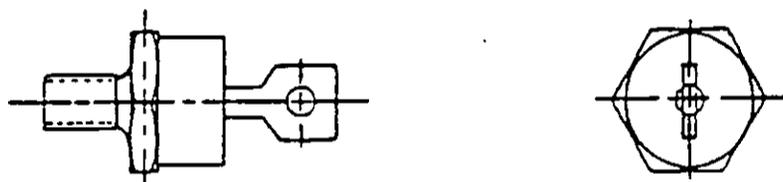


FIGURE 120. Dimensions of a typical SCR (D0-5 package).

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

4.10.4 Military designation. The military designation for diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 1N for diodes.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

4.10.5 Electrical characteristics

4.10.5.1 Surge and I^2t ratings, nonrecurrent. In the event that a type of overload or short circuit can be classified as nonrecurrent, the rated junction temperature can be exceeded for a brief instant, thereby allowing additional overcurrent rating. Ratings for this type of nonrecurrent duty are given by the surge current curve and by the I^2t rating.

Figure 121 shows the maximum allowable nonrecurrent surge current at rated load conditions for a high current SCR. The junction temperature is assumed to be at its maximum rated value immediately prior to surge; it is therefore apparent that the junction temperature will exceed its rated value for a short time.

The data shown in this curve are values of peak rectified sinusoidal waveforms (60 Hz) in a half-wave circuit. The one-cycle point, therefore, gives an allowable nonrecurrent half sine wave of 0.00834-second duration (half-period of 60 Hz frequency) of a peak amplitude of 150 A. The 20-cycle point shows that 20 rectified half-sine waves are permissible, separated by equal off times and each of an equal amplitude of 80 A.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

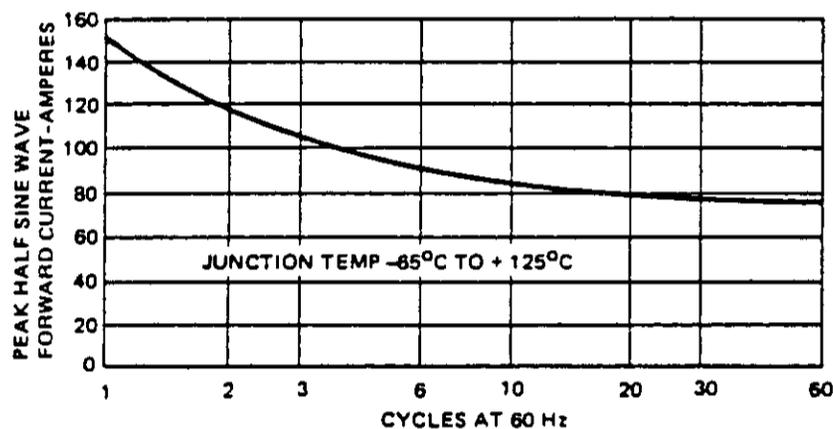


FIGURE 121. Maximum allowable nonrecurrent peak surge forward current at rated load conditions.

I^2t ratings apply for nonrecurrent overloads shorter than one half cycle. For such times, the SCR behaves essentially like a resistance with a fixed thermal capacity and negligible power dissipating means, and it displays a current capability which can be expressed as a constant, I^2t , where I is the rms value of current over an interval, t . The I^2t rating of the SCR is given in the specifications. This rating assumes that the SCR is already in the conducting state. If the SCR is turned on into a fault, the current-time relationships (di/dt) during the turn-on interval must be within the device's switching capabilities.

4.10.5.2 Holding and latching current. As with the solenoid of an electro-mechanical relay, an SCR requires a certain minimum anode current to maintain it in the closed or conducting state. If the anode current drops below this minimum level, designated as the holding current, the SCR reverts to the forward blocking or open state. The holding current for a typical SCR has a negative temperature coefficient; that is, as its junction temperature rises, its holding current requirement decreases.

A somewhat higher value of anode current than the holding current is required for the SCR to initially "pickup." If this higher value of anode latching current is not reached, the SCR will revert to the blocking state as soon as the gate signal is removed. After this initial pickup action, however, the anode current may be reduced to the holding current level. Where circuit inductance limits the rate of rise of anode current and thereby prevents the SCR from switching solidly into the conducting state, it may be necessary to make alterations in the circuit.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

A meaningful test for the combined effects of holding and latching current is shown in Figure 122. The SCR under test is triggered by a specified gate signal. The test circuit allows the SCR to latch into conduction at a current level I_{F1} . The test circuit then reduces the current to a continuously variable level I_{F2} . The current I_{F2} at which the SCR reverts to the off state is the desired value of holding current.

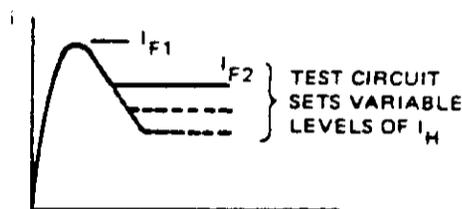


FIGURE 122. Holding current test waveform.

4.10.5.3 Rate of rise of forward voltage (dv/dt). A high rate of rise of forward (anode-to-cathode) voltage may cause an SCR to switch into the on, or low impedance, forward conducting state. In the interest of circuit reliability, it is important to characterize the device with respect to its dv/dt withstand capability.

SCRs are characterized with respect to dv/dt withstand capability in two ways:

- a. The so-called static dv/dt withstand capability. This specification covers the case of initially energizing the circuit or operating the device from an anode voltage source that has superimposed fast rise-time transients. Such transients may arise from the operation of circuit switching devices or result from other SCRs' operation of adjacent circuits.
- b. The maximum allowable rate of reapplication of forward blocking voltage while the SCR is regaining its rated forward blocking voltage (V_{FXM}) following the device turn-off time (t_{off}) under stated circuit and temperature conditions. In this context dv/dt is an important part of the overall SCR turn-off characterization.

Figure 123 shows the typical static dv/dt as a function of temperature for a medium current SCR with its gate open. The rate of rise of anode voltage shown on the ordinate is the slope of a straight line starting at zero anode voltage and extending through the one-time constant (τ) point on an exponentially rising voltage. The upper right hand portion of the figure illustrates this definition.

The definition shown in Figure 123 has become accepted as a standard by the industry. Some specification sheets give the time constant under specified conditions rather than by a curve as in Figure 123. It should be noted that: $\tau = 0.632 \times \text{rated SCR voltage } (V_{FXM}) / dv/dt$.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

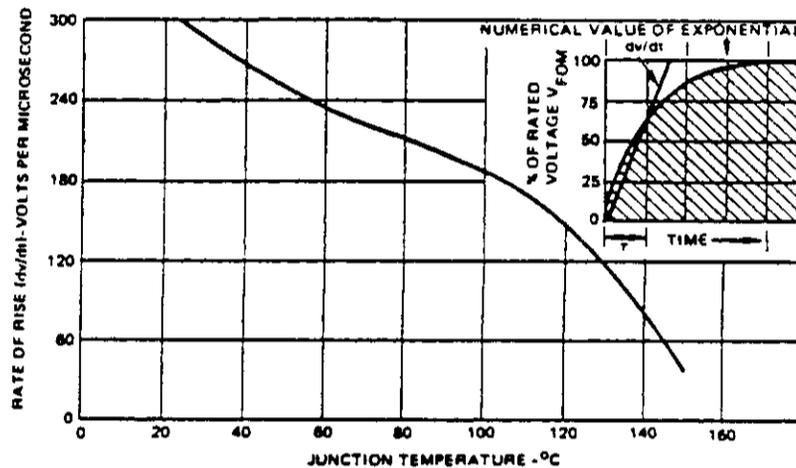


FIGURE 123. Rate of rise (dv/dt) of forward voltage that will not turn on SCR.

The initial dv/dt withstand capability will be recognized as being greater than the value defined. In terms of specified minimum time constant it is

$$dv/dt \Big|_{t=0+} = \frac{\text{rated SCR voltage } (V_{FXM})}{\tau}$$

In terms of specified maximum dv/dt capability, the allowable initial dv/dt withstand capability is

$$dv/dt \Big|_{t=0+} = (1/0.632)dv/dt = 1.58 dv/dt$$

The shaded area shown in the insert of Figure 123 represents the area of dv/dt values that will not trigger the SCR. These data enable the circuit designer to tailor his circuitry in such a manner that reliable circuit operation is assured. Because a high circuit-imposed dv/dt effectively reduces $V_{(BR)FX}$ (the actual anode voltage at which the particular device being observed switches into the on state) under given temperature conditions, a higher voltage classification unit will allow a higher rate of rise of forward voltage for a given peak circuit voltage.

As an illustration of this, consider a SCR with $V_{FOM} = 200$ volts at $T_j = 130$ °C operating at a peak circuit voltage of $E = 150$ volts, or $(150/200) \times 100 = 75$ percent of rated voltage. Figure 123 shows that a $dv/dt = 120$ V/μs can be applied to the one time constant (τ) point of an exponentially rising voltage. In this example, $\tau_1 = (0.632 \times 200)/120 = 1.06$ μs. Accordingly, from Figure 123, the time in which the anode of the SCR may reach 150 V, or 75 percent of rated voltage is $t_1 = 1.5 \times 1.06 \approx 1.6$ μs.

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If a C38D type is selected with $V_{FOM} = 400$ V, under the same conditions it may be found by using the ratio of rated voltages that $t_2 = \tau_1 400/200 = 2 \tau_1 = 2 \times 1.06 = 2.12 \mu\text{s}$ to reach 63.2 percent of 400 or 253 V. Because in this case a maximum circuit voltage of 150 V or $150/400 = 37.5$ percent of the rated 400 V V_{FOM} is the design requirement, it is seen from the insert of Figure 123 that $t_2 = 0.5 \tau_2 = 0.5 (2.12) = 1.06 \mu\text{s}$.

By selecting a 400-volt device the time to reach the operating circuit voltage in this example can be reduced to $t_2/t_1 = 1.06/1.6 (100) = 67$ percent.

Reverse biasing of the gate with respect to the cathode may increase dv/dt beyond that shown in Figure 123.

The circuit shown in Figure 124 can be used to suppress excessive rate of rise of anode voltage. The time constant of the load resistance, R_L , in ohms and capacitor, C , in microfarads should be selected so that

$$\tau < R_L C \mu\text{s}$$

where

τ = minimum time constant of exponential forward voltage rise specified for SCR

Referring to Figure 124, SCR discharges capacitor, C , via resistor, R_d . R_d should be selected on the basis of limiting the peak capacitor discharge current E/R_d during the SCR turn-on interval to a value within the device capability. For best results, the circuit of Figure 124 should be wired and placed in close proximity to the SCR in order to minimize inductive effects. Also, the capacitor should have good high-frequency characteristics.

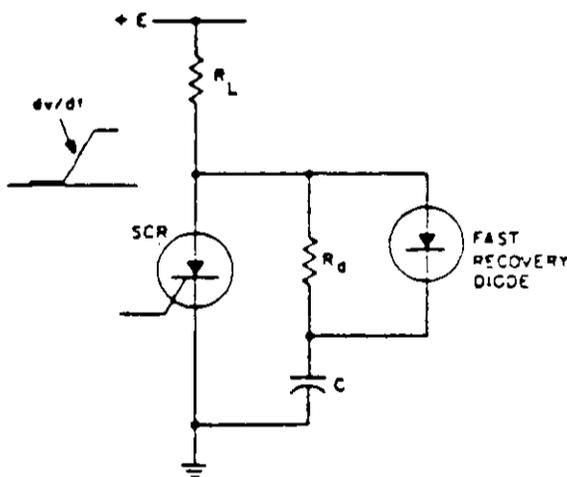


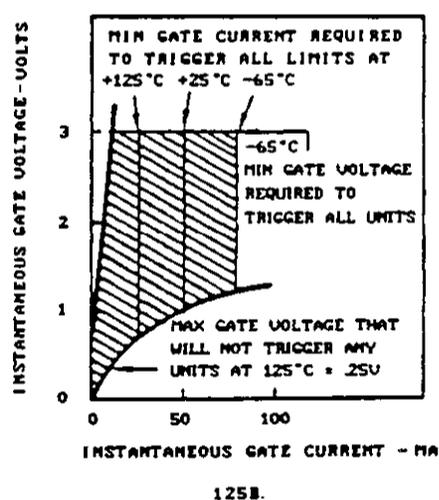
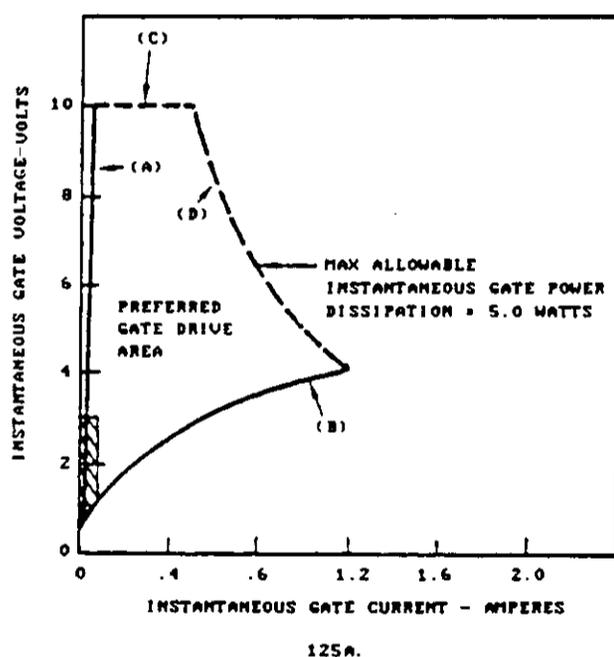
FIGURE 124. Rate of rise of anode voltage (dv/dt) suppression circuit.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

4.10.5.4 Triggering specifications for the dc gate. The typical dc gate triggering characteristics of an SCR are presented in the form of a graph illustrated in Figure 125A. The graph shows gate-to-cathode voltage as a function of positive gate current (flow from gate to cathode) between limit lines (A) and (B) for all SCRs of the type indicated. These data apply to a zero-anode-current condition (anode open).

The basic function of the triggering circuit is to simultaneously supply the gate current to trigger I_{GT} and its associated gate voltage to trigger V_{GT} . The shaded area shown in Figure 125B contains all the possible trigger points (I_{GT} , V_{GT}) of typical SCRs. The trigger circuit must therefore provide a signal (I_{GT} , V_{GT}) outside the shaded area to reliably trigger these SCRs.

The area of trigger circuit-SCR gate operation is indicated as the "preferred gate drive area." It is bounded by the shaded area in Figure 125B which represents the locus of all specified triggering points (I_{GT} , V_{GT}), the limit lines A and B, line C representing rated peak allowable forward gate voltage V_{GF} , and line D representing rated peak power dissipation P_{GM} , Figure 125A. Some SCRs may also have a rated peak gate current, I_{GFM} , which would appear as a vertical line joining curves B and D, Figure 125A.



Note: Junction temperature -65 to +125 °C

FIGURE 125. Characteristics of dc gate triggering.

4.10 DIODES, SILICON CONTROLLED RECTIFIERS (SCR)

Figure 125B shows the detail of the locus of all specified trigger points and the temperature dependence of the minimum gate current to trigger I_{GTmin} . The lower the junction temperature, the more gate drive is required for triggering. Some specifications may also show the effect of forward anode voltage on trigger sensitivity. Increased anode voltage, particularly with small SCRs, tends to reduce the gate drive requirement. Also shown is the small positive value of gate voltage below which no SCR of the particular type will trigger. The reverse quadrant of the gate characteristic is usually specified in terms of maximum voltage and power ratings.

4.10.5.5 Load lines. The trigger circuit load line is the diagonal line connecting the properly derated values of current and voltage in Figure 125B and must intersect the individual SCR gate characteristic in the region indicated as preferred gate drive area. The intersection, or maximum operating point, should also be located as close to the maximum applicable (peak, average, etc.) gate power dissipation curve as possible. Gate current rise times should be in the order of several amperes per microsecond to minimize anode turn-on time, particularly when switching into high currents. This results in minimum turn-on anode switching dissipation and minimum jitter.

Construction of a load line is a convenient means of placing the maximum operating point of the trigger circuit-SCR gate combination into the preferred triggering area. Figure 126A illustrates a basic trigger circuit of source voltage (e_s) and internal resistance (R_G) driving an SCR gate. Figure 126B shows the placement of the maximum operating point well into the preferred trigger area close to the rated dissipation curve. The load line is constructed by connecting a straight line between the trigger circuit open circuit voltage (E_{OC}) entered on the ordinate and the trigger circuit short circuit current, $I_{SC} = E_{OC}/R_G$, entered on the abscissa.

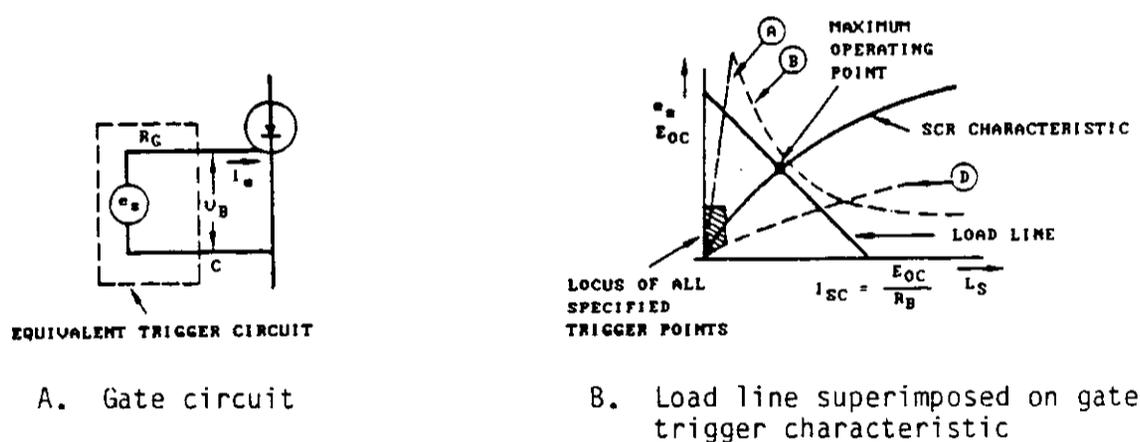


FIGURE 126. Gate circuit and construction of load line.

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**4.10 DIODES, SILICON CONTROLLED
RECTIFIERS (SCR)**

If the trigger circuit source voltage, $e_s(t)$, is a function of time, the load line sweeps across the graph, starting as a point at the origin and reaching its maximum position at the peak trigger circuit output voltage.

The applicable gate power curve is selected on the basis of whether the average or peak allowable gate power dissipation is the limiting factor. For example, if a dc trigger is used, the average maximum allowable gate dissipation (0.5 W) must not be exceeded. If a trigger pulse is used, the peak gate power curve is applicable (the 5-Watt peak power curve labeled D in Figure 125). For intermediate gate trigger waveforms, the limiting allowable gate power dissipation curve is determined by the duty cycle of the trigger signal according to the equation:

average allowable gate power = peak gate drive pwv x pulse width x pulse repetition rate.

4.10.6 Environmental considerations. Typical environmental conditions and screens which silicon control rectifiers are capable of withstanding are not substantially different from those given in paragraphs 4.1.6.9 Environmental considerations and 4.1.6.10 Screening procedures in subsection 4.1, Diodes, general. For the specific device selected, consult the applicable MIL-S-19500 reference sheet.

4.10.7 Reliability considerations. Refer to the subsection 4.11 Diode, general, paragraph 4.1.6 Reliability section for a complete list of failure mechanisms. Some of the more pertinent failure mechanisms and how they relate to SCRs are explained below.

4.10.7.1 Structural flaws. Structural flaws are generally considered to be the result of weak parts, discrepancies in fabrication or inadequate mechanical design. Various in-process tests performed on the device, such as forward voltage drop at high current density levels and thermal resistance measurement, provide effective means for the monitoring of controls against such flaws.

The modes of failure generally associated with mechanical flaws for an SCR are excessive on-voltage drop, failure to turn on when properly triggered, and open circuit between the anode and cathode terminals. These types of failure mechanisms are relatively rare.

4.10.7.2 Encapsulation flaws. Encapsulation flaws are deficiencies in the hermetic seal that will allow undesirable atmospheric impurities to reach the semiconductor element. Foreign atmospheres, such as oxygen and moisture, can react in such a way as to permanently alter the surface characteristics of the silicon.

A change in surface conductivity is evidenced by a gradual increase of the forward and reverse blocking current characteristics. Because the SCR is a current-actuated device, it will lose its capacity to block rated voltage if blocking current degrades beyond some critical point. This type of mechanism

**4.10 DIODES, SILICON CONTROLLED
RECTIFIERS (SCR)**

may eventually result in catastrophic failure. The rate of degradation is dependent mostly on the size of the leak and the level of stress, particularly thermal stress, that is applied.

4.10.7.3 Internal contaminants. The inclusion of a source of ionizable material inside the sealed package can result in semiconductor surface inversion, layer formation. This leads to permanent electrical instability in the device.

The occurrence of such defects can cause discrepancies in the turn-on and turn-off stages in an SCR. For example, a variation can occur in the voltage at which the SCR is turned on. On the other hand, a defect such as the one mentioned above, could cause a device not to be turned off completely.

In certain cases, it is possible, using external circuitry (negative gate or resistor biasing), to suppress the effects of such contaminants to the point of rendering the device still usable in the circuit.

4.10.7.4 Material electrical flaws. In certain uses, device failure can result from defective junction formation in the semiconductor. Crystal dislocations at a semiconductor p-n junction can result in the formation of diffusion pipes or diffusion spikes which could cause emitter-collector shorts in the given device. These pipes or spikes become very important as the depletion width of the lightly doped region decreases. The probability of occurrence of oxidation-induced stacking faults and other defects can be reduced by careful processing of the semiconductor device.

4.10.7.5 Metal diffusion. Impure metal atoms lodged in a semiconductor device can, under certain circumstances, diffuse into the p-n junction, thus causing device degradation.

The probability and extent of device degradation and their effects on device lifetime are all functions of the temperature and the diffusing metal atom species.

Lithium, copper, silver, gold, zinc, nickel, and iron are examples of metals that diffuse rapidly in silicon.

4.10.7.6 Derating. History has shown that the largest single cause of SCR failure is operating above allowable levels of electrical and thermal stress. Accordingly, it is imperative that derating of parts be performed to enhance the reliability of electronic systems. Users should refer to MIL-STD-975 for derating factor guidelines. The design and use of a part should always be within the thermal and electrical stresses defined. High temperature operation is the most destructive stress for a semiconductor. It will result in early electrical parameter drift, and a general degradation of electrical and mechanical characteristics of the device.

4.11 DIODES, PHOTODIODES

4.11 Photodiodes.

4.11.1 Introduction. One of the most basic of all the optoelectronic solid-state devices is the photodiode. The photodiode may be used in the photoconductive mode or the photovoltaic mode. In the photovoltaic mode, the device is used to convert radiant energy to a generated voltage under zero bias conditions. In this mode, the output voltage generated across its electrodes will vary with the intensity of the light striking the thin top P⁺ or N⁺ layer. In the photoconductive mode, the device is reverse biased and the photocurrent generated (I_L) will vary linearly with the intensity of the light striking the top surface of the pn junction.

Figure 127 illustrates the equivalent circuit which is applicable for both modes of operation and defines the various terms used.

Figure 128 illustrates with separate circuit diagrams the photoconductive and photovoltaic modes of operation for photodiodes in general, whether the basic semiconductor material used is silicon, germanium, or some III-V semiconductor alloys.

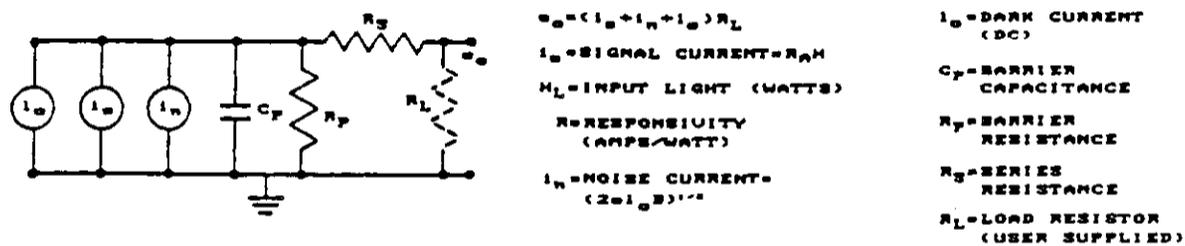


FIGURE 127. Equivalent circuit for photodiodes (all models).

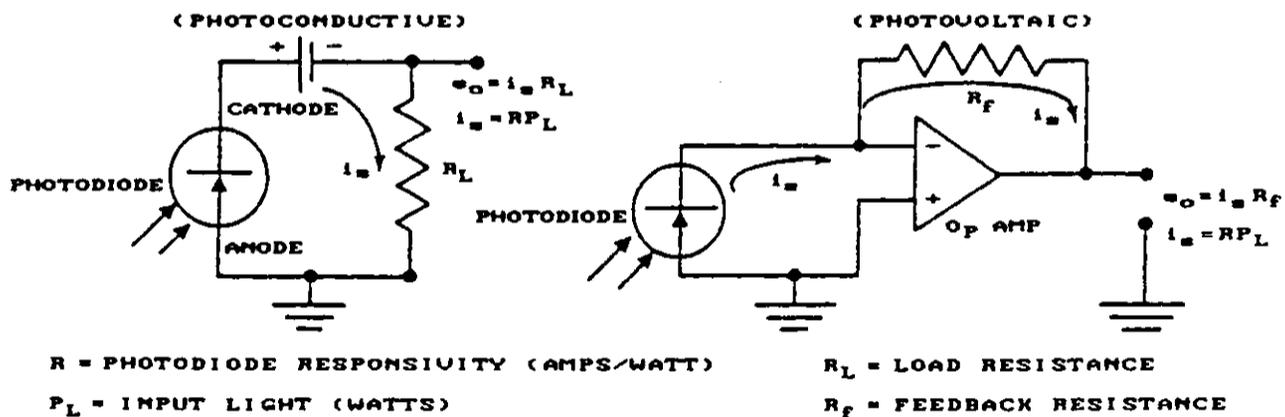


FIGURE 128. Typical hookup circuits for various photodiode modes.

4.11 DIODES, PHOTODIODES

The pn junction photodiode has a slow response time, being as slow as 30 μ s. Responsivity can be good, however, depending somewhat on the construction of the device. Responsivity of the pn junction photodiode can be in the neighborhood of 0.3 to 0.7 A/W (amperes/watt) at the wavelength of peak sensitivity. The spectral response of most silicon pn junction photodiodes is from about 500 to 1000 nm. Again, this can be adjusted by varying the type of semiconductor material used. The pn junction photodiodes are also characterized by a high thermal noise level and high dark current, which make them useful only for short distance communication. In practice, most conventional pn junction photodiodes are used more for industrial sensing than for communication purposes.

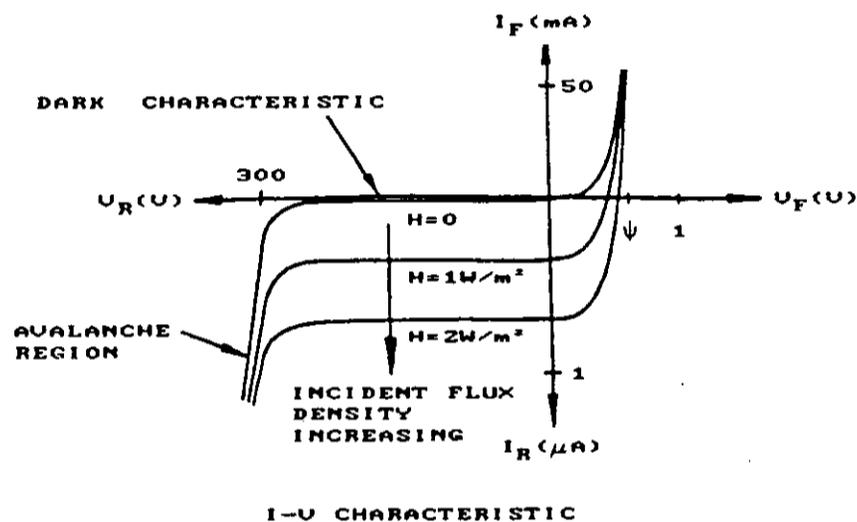


FIGURE 129. Typical photodiode I-V diagram.

Another way of describing a photodiode is with the familiar current-voltage, or I-V, diagram. Figure 129 is the I-V diagram for a typical photodiode. The arrow points to the voltage at which the photodiode is reverse biased. When the photodiode is illuminated, the I-V curve drops down to a more negative value, but the reverse voltage (X) remains the same. As the light intensity continues to increase, the I-V curve drops even further, as shown in the figure. In other words, the photo current increases with more light, though the reverse voltage is constant. However, the current cannot increase indefinitely because of the physical limitations of the photodiode. A point will be reached (saturation) at which the device will no longer respond to increases in light intensity.

An advantage of using photodiodes is that the photodiode reacts to changes in light intensity much more quickly than the photoconductor. This is because the charge carriers are accelerated by the depletion region's barrier field whereas the photoconductor's charge carriers move only by the slow thermal diffusion process. However, most conventional pn junction photodiodes are still not fast enough for most fiber optic work. This is because the depletion region in the photodiode is so small that a good portion of the light is absorbed by the semiconductor material outside of the depletion region.

4.11 DIODES, PHOTODIODES

4.11.2 Usual applications. Photodiode applications include card and tape readers, pattern and character recognition, shaft encoders, position sensors, and counters.

Both the PIN photodiode and the avalanche photodiode (APD) are very useful as photodetectors in fiber optics system applications.

4.11.2.1 Schmitt detector. In this particular circuit application, as shown in Figure 130, PIN photodiodes are superior to silicon npn phototransistors as signal detectors because when the frequency of the optical signal reached about 2 KHz, the output from the detector became very distorted whereas there was no signal distortion with the photodiode.

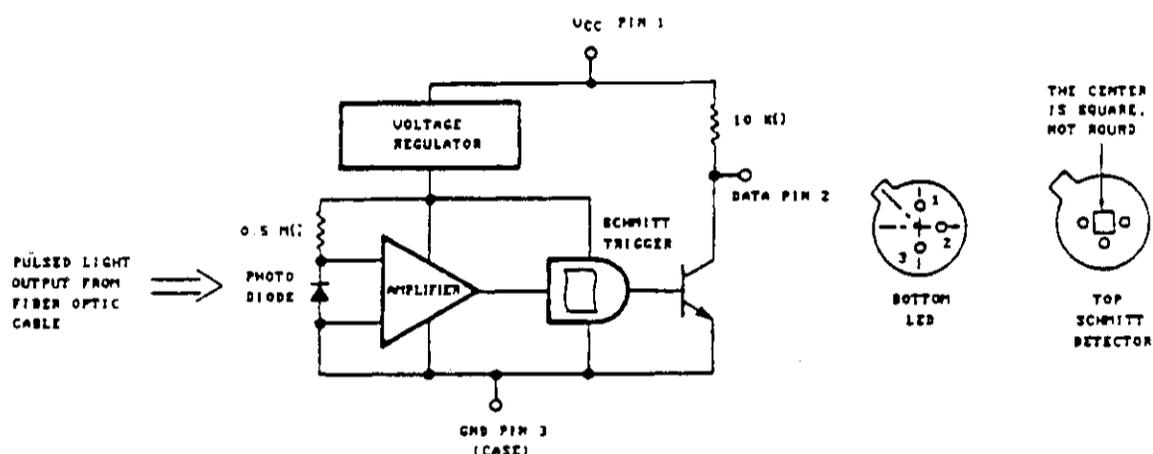


FIGURE 130. Schmitt detector circuit.

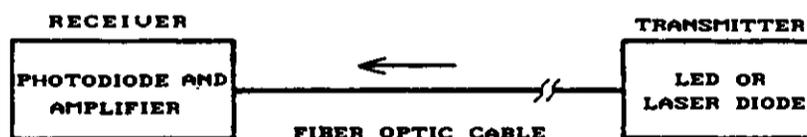
In addition, photodiodes, especially the PIN and APD types, are used extensively in fiber optic communication systems as detectors or as the key element in the receiver system as illustrated in the figures that follow.

It should be pointed out that there are three types of fiber optic communication systems; namely, simplex, half duplex, and full duplex, depending upon the direction in which their optical data may flow.

Reference to data direction applies to two terminals in a fiber optic system (transmitter and receiver). Between two terminals, data can travel one way, from one terminal to the other; two ways, from one terminal to the other and back again; and two-ways simultaneously, from one terminal to the other and back again at the same time. In other words, one-way, two-ways, and two-ways simultaneously are referenced to as simplex, half duplex, and full duplex, respectively.

4.11 DIODES, PHOTODIODES

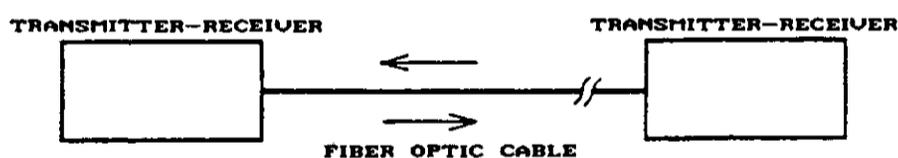
4.11.2.2 Simplex systems. Simplex systems are aptly named for they are the simplest kind of communication link possible. Figure 131 is a diagram of a typical simplex system. Note that the link consists of one transmitter connected to one receiver by one fiber. Data can travel in only one direction in this type of link, making it useful only for remote control or reception of telemetry. Unfortunately, this simple and inexpensive system has no means of feedback. That means that the sender has no way of knowing whether or not communication has been received correctly; therefore, there is no way to know when errors have occurred in transmission. Because of this inherent problem, simplex systems are used only in noncritical applications. However, the user may benefit from communications expense if these drawbacks are tolerable.



SIMPLEX FIBER OPTIC COMMUNICATION SYSTEM

FIGURE 131. Simplex fiber optics communication system.

4.11.2.3 Half duplex systems. Figure 132 is a diagram of a typical half duplex communication system. It consists of two transmitter-receiver pairs connected by a single fiber optic cable. In this system, communication can travel in both directions, but not at the same time.



HALF DUPLEX FIBER OPTIC COMMUNICATION SYSTEM

FIGURE 132. Half duplex fiber optic communication system.

To accomplish two-way communication on one line, the system must have an emitter and a photodetector coupled to each end of the fiber. This means having a combination LED/photodiode at each end of the system or, as shown in Figure 133, a way of splitting the transmission path at each end. Both of these arrangements are compromises in quality because (1) an LED/photodiode combination cannot work at peak efficiency as both a generator and detector of light, and (2) splitting the line results in smaller signal levels at the receiver. This means that the length of the line must be reduced to maintain the minimum acceptable signal level at the receiver.

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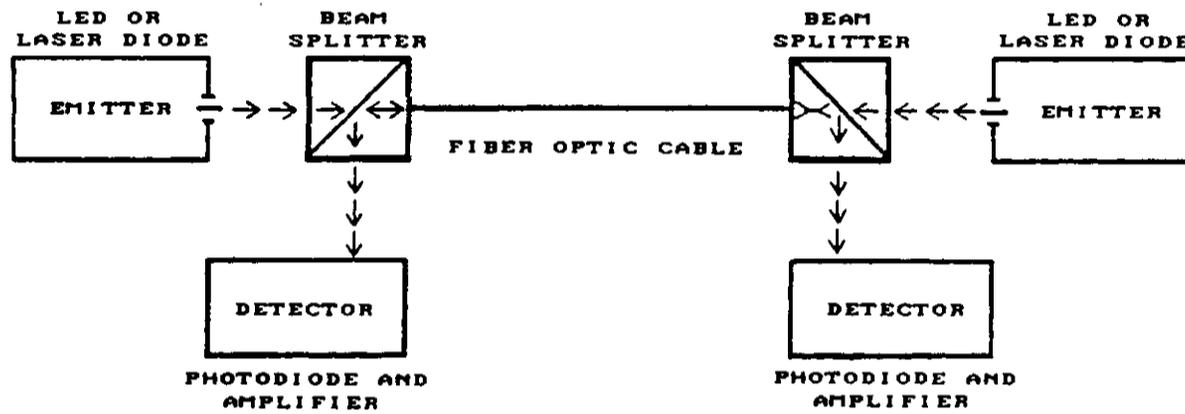


FIGURE 133. Half duplex system with split transmission path.

Special LED/photodiodes and beam-splitting connectors make half duplex systems more expensive and complicated than simplex systems. In terms of the number of lines used, half duplex systems are efficient because they only need one line. But in terms of the time spent waiting for the other party to stop sending so that you can send, the half duplex is inefficient. Also, because beam splitters reduce signal strength, half duplex systems are limited in the distance they can cover without a repeater. For these reasons, half duplex systems are also not widely used.

4.11.2.4 Full duplex systems. Figure 134 is a diagram of a typical full duplex system. It consists of two transmitters and two receivers connected by two fiber optic cables. Each cable carries communication in one direction only. In effect, a full duplex system is nothing more than two simplex systems side by side, operating in opposite directions.

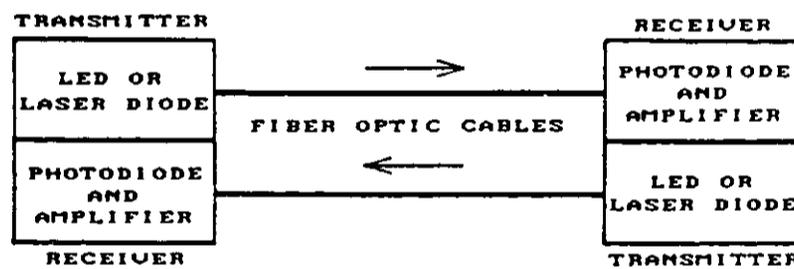


FIGURE 134. Full duplex fiber optic communication system.

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The disadvantage of this arrangement is that two fibers are not needed instead of one. However, the advantage of simultaneous communication often outweighs this disadvantage. Another advantage is that the system can be optimized for maximum length by using separate LEDs and photodetectors instead of LED/photodiode devices.

It is possible to operate a full duplex on one line by means of wavelength multiplexing. However, this technique is not currently used outside of the laboratory, although it may be used in future applications.

4.11.3 Physical construction.

4.11.3.1 Conventional pn junction photodiode and PIN photodiode. Figure 135 is a typical cross sectional drawing of a conventional photodiode. There have been many processes by which pn junctions have been formed. The technique that has become the most popular and is most used by industry is the planar process. In this process silicon is used as the starting semiconductor material and is followed by gas-phase doping of the solid silicon at high temperatures (1000 - 1200 °C). By diffusion, the dopant gases penetrate the solid surface of the silicon. Diffusions of n-type and p-type impurities may be made onto a previously doped region of an n-type or p-type substrate wafer by appropriate masking off one side of the wafer after lapping off the previously diffused doped layer. This serves to provide manufacturing versatility.

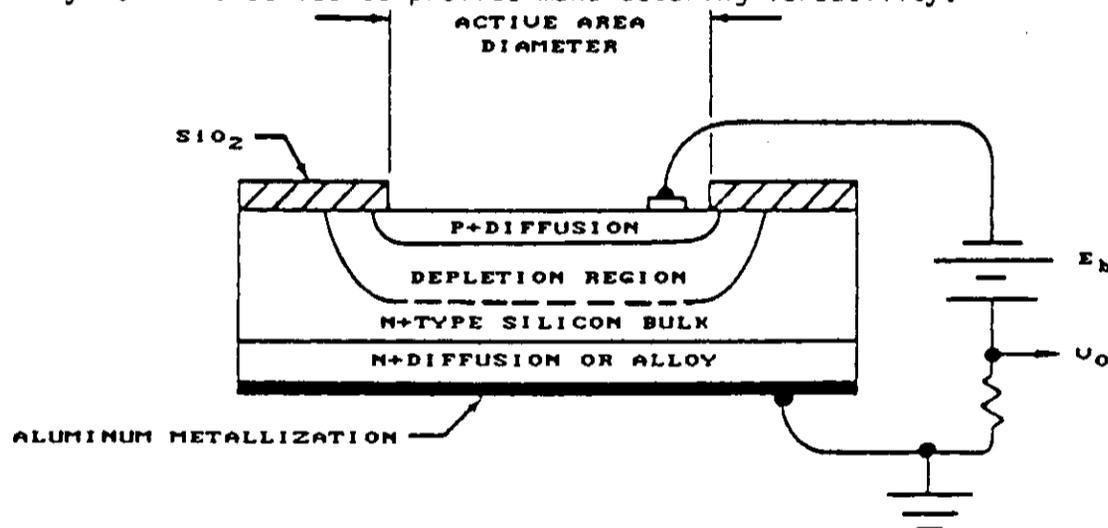


Figure 135. Cross section, planar diffused photodiode.

In the foregoing illustration, n-type bulk silicon is diffused on one side by n^+ dopant and on the opposite side by p^+ dopant. A depletion region between the n and p regions exists which is free of carriers while under appropriate reverse bias conditions. It is in this depletion area that the photon should be absorbed. The active area exposes the thin p^+ diffused region to the light beam. The light beam is in turn absorbed in the semiconductor material surface. When photons of the correct energy are absorbed into the semiconductor surface, electron-hole pairs are formed.

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The PIN photodiode is similar to the one described in Figure 135 with one important difference. This device has an intrinsic (I) layer between its p and n regions and is commonly referred to as a PIN photodiode. The intrinsic layer is undoped and has a bulk silicon resistivity of typically 300 to 500 Ω -cm compared to conventional photodiodes which can have bulk silicon resistivities in the 5 to 20 Ω -cm range. Therefore, the depletion region in this case will extend further into the "I" region than it would in a more heavily doped semiconductor. In other words, adding the "I" layer results in a much wider depletion region for a given reverse voltage near the breakdown value. The wider depletion area makes the PIN photodiode respond better to the lower light frequencies (longer wavelengths).

4.11.3.2 Schottky barrier photodiode. The Schottky barrier photodiode is illustrated in Figure 136. The Schottky photodiode (often called a surface diode) differs from the conventional photodiode in the method by which the p-type material is formed. In the conventional photodiode the pn junction is developed by a chemical diffusion process. In the Schottky photodiode, a metal-semiconductor barrier is formed over a thin silicon epitaxial layer (2 to 5 μ m) of several ohm-centimeters resistivity by evaporation of a thin gold metal film (10- to 20-nm thick).

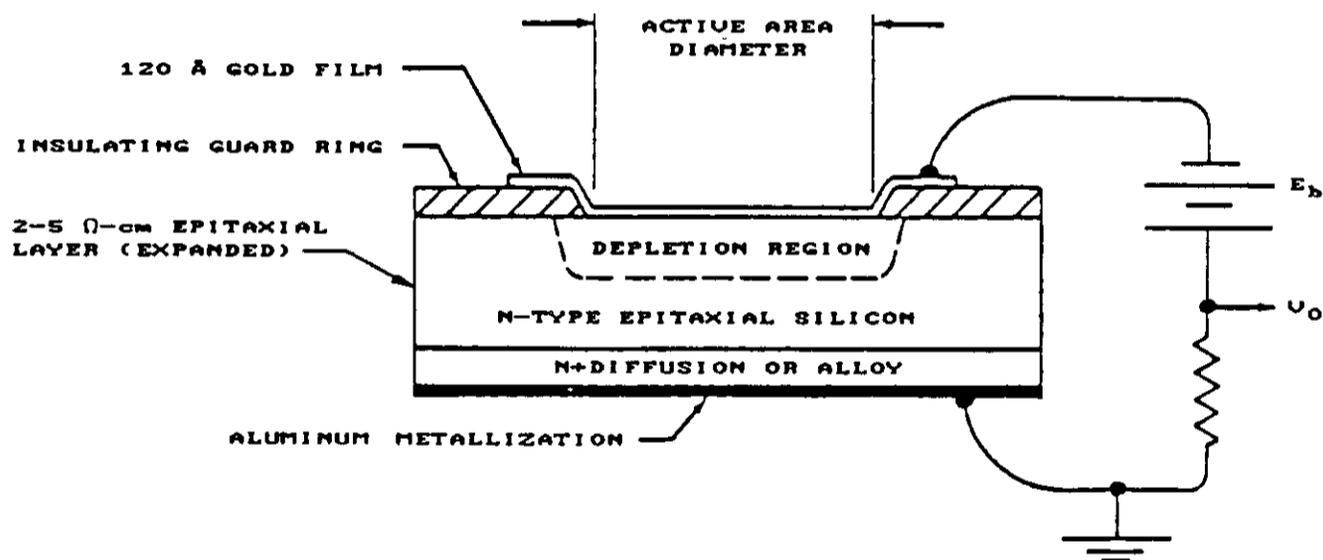
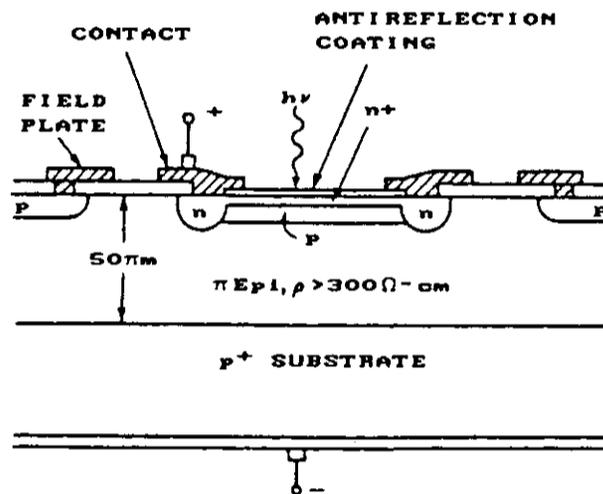


FIGURE 136. Cross section of Schottky barrier photodiode.

4.11.3.3 Avalanche photodiode. Figure 137 illustrates the general construction of an epitaxial silicon reach-through avalanche photodiode with n^+ -p- π - p^+ structure (π refers to the intrinsic silicon of about 300 to 500 Ω -cm). The diameter of the light-sensitive, high-gain region is typically 100 μ m.

4.11 DIODES, PHOTODIODES

FIGURE 137. Cross section, avalanche/photodiode.

In the foregoing figure, the antireflection coating for silicon devices is usually a thermally evaporated silicon monoxide film of the appropriate thickness and refractive index. Its purpose is to increase quantum efficiency and thereby increase the responsivity of the device near its peak value.

Typical packages commonly used for photodiodes are JEDEC TO-5, TO-18, and TO-46.

4.11.4 Military designation. Currently, there are no JANS or JANTXV military device designations on DESC's QPL list (April 1, 1986) for photodiodes. For high reliability applications such as space, the user can generate a specification control drawing (SCD) describing a manufacturer's generic type screened in accordance with MIL-S-19500 to JANS or JANTXV requirements as applicable. It is anticipated that with increasing usage of photodiodes and other optoelectronic devices in high reliability applications, QPL registrations will follow.

4.11.5 Electrical characteristics. Short, medium, and long wavelengths of photon power are absorbed at different depths within the pn junctions. The depth of penetration is dependent on the photon wavelength or energy as illustrated in Figure 138 for higher resistivity silicon ($>5 \Omega\text{-cm}$). Short wavelengths (higher photon energy), as expected, are absorbed near the surface. Long wavelengths (lower photon energy) may penetrate the entire die structure. To be most useful, the wavelengths should be absorbed in the depletion area.

The photon current is produced by electron-hole pairs being separated and drawn out in directions of more positive or negative sources whichever is the case. If the electron-hole pairs are generated outside the depletion area they will usually recombine and no photo current will be produced. The active p^+ diffused

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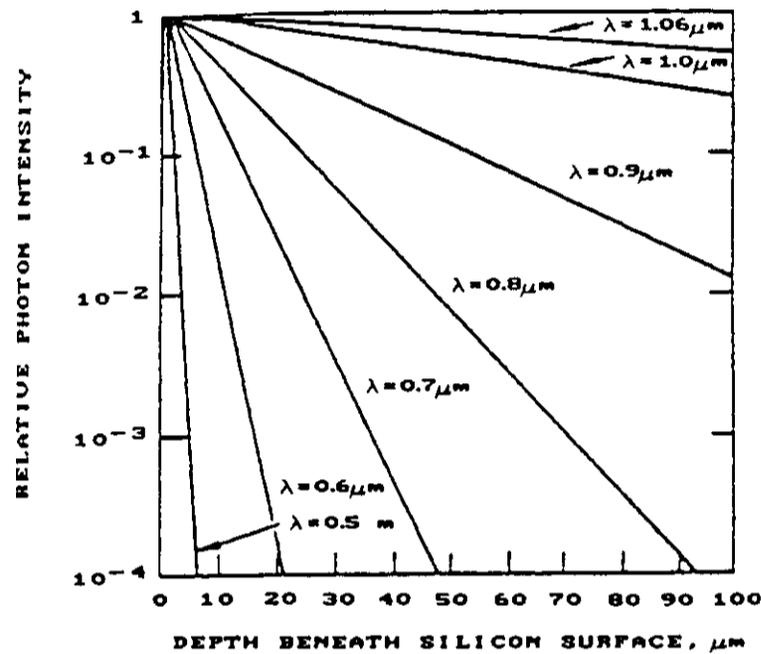


FIGURE 138. Decay of incident light intensity as a function of penetration depth into silicon substrate.

region should be extremely thin (less than $1 \mu\text{m}$) to ensure maximum penetration. As in other reverse-biased pn junction diodes, the depletion width can be made larger by increasing the reverse bias and thereby reducing the junction capacitance.

In a practical photodiode, 100 photons will create between 30 and 95 electron-hole pairs, thus giving the detector a quantum efficiency ranging from 30 to 95 percent. To achieve a high quantum efficiency, the depletion layer must be thick enough to permit a large fraction of the incident light to be absorbed. However, the thicker the depletion layer, the longer it takes for the photo-generated carriers to drift across the reverse-biased junction. Because the carrier drift time determines the response speed of the photodiode, a compromise has to be made between response speed and quantum efficiency.

For very low level light signals, all of the previous photodiodes considered fall short in their performance. Their responsivity prevents them from being used in electronic systems covering large distances (over a kilometer) because the signal attenuation at that distance becomes excessive. A group of photodiodes that can handle lower level signals is the avalanche-type (APD). With their internal amplification of light signals, they are well suited for long distance communication. These devices in general are more expensive than the corresponding PIN photodiodes.

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Table IX is a performance specification table and Figure 139 is a spectral response figure for a typical PIN photodiode. The response time for this photodiode is about 1 ns, which is much faster than the 30 μ s quoted earlier for the pn junction photodiode. Responsivity is about 0.5 A/W whereas dark current is about 1 nA at room temperature. In general, no figure for noise will be found in any vendor's performance specification. Instead, what will usually be given is the noise equivalent power (NEP), which is a measure of the light power required to produce a signal-to-noise ratio of one to one in the device. (Another name for NEP is sensitivity.)

Notice that the spectral response is not less than 40 percent of the maximum response for the wavelengths from 500 to 950 nm. This is typical for silicon photodevices.

Better response time than is possible with either a pn or PIN photodiode can be achieved through the use of the Schottky barrier photodiode.

The lower frequency photons contain less energy and tend to penetrate deeper into the diode's structure before producing electron-hole pairs and in many cases, do not produce pairs. The wider depletion region of the PIN photodiode increases the chance that pairs will be produced. The PIN photodiode is therefore more efficient over a wider range of light frequencies. The PIN device also has a lower internal capacitance due to the wide "I" region which acts like a wide dielectric between the p and n regions. This lower internal capacitance allows the device to respond faster to changes in light intensity. The wide depletion region also allows this device to provide a more linear change in reverse current for a given change in light intensity.

In the case illustrated, Figure 136, the active region has a very thin gold film that covers the metallized n-type or p-type silicon epitaxial layer. This film must be thin (200Å or 20 nm) to allow light penetration to the surface of the silicon epitaxial layer. The Schottky barrier photodiode has some advantages over the conventional photodiode. It operates well at wavelengths less than 500 nm and has a much simpler fabrication process. It should be pointed out that the Schottky photodiode does not operate well at high temperatures (greater than 70 °C) or with high light power (greater than 2 mW of radiant power). They have good response times, typically less than 25 ns.

These devices are recommended when high blue response (wavelengths less than 500 nm) or larger areas (greater than 1 cm²) are required.

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TABLE IX. Electrical characteristics of a PIN photodiode

Characteristic	Symbol	Min	Typ	Max	Unit
Dark current ^{1/} ($V_R = 20$ V, $R_L = 1.0$ Megohm) $T_A = 25$ °C $T_A = 100$ °C	I_D	- -	1.0 14	10 -	nA
Reverse breakdown voltage ($I_R = 10$ μ A)	$V_{(BR)R}$	100	200	-	V
Forward voltage ($I_F = 50$ mA)	V_F	-	-	1.1	V
Series resistance ($I_F = 50$ mA)	R_S	-	-	10	Ohms
Total capacitance ($V_R = 20$ V, $f = 1.0$ MHz)	C_T	-	-	4.0	pF
Responsivity (Figure 139)	R	0.4	0.5	-	μ A/ μ W
Response time ($V_R = 20$ V, $R_L = 50$ Ω)	t_{on} t_{off}	- -	1.0 1.0	- -	ns ns

^{1/} Measured under dark conditions. $H = 0$

For the avalanche photodiode shown in Figure 137, the guarding helps to ensure that the edge of the junction does not go into breakdown prematurely because of the high electric field strength at the edges. The guard ring is slightly doped to keep the electric field weak at the edge, forcing breakdown to occur in the bulk portion of the junction. This ensures a linear current reaction to the change in the light signal level.

The reach-through avalanche photodiode is composed of a high resistivity p-type material deposited as an epitaxial layer on a p⁺ (heavily doped p-type) substrate. A p-type diffusion or ion implant is then made in the high-resistivity layer (200 to 500 Ω -cm) followed by the diffusion of an n⁺ (heavily doped n-type) layer. For silicon, the dopants used to form these layers are normally boron and phosphorus, respectively. This configuration is referred to as a silicon p⁺- π -p-n⁺ "reach-through" structure.

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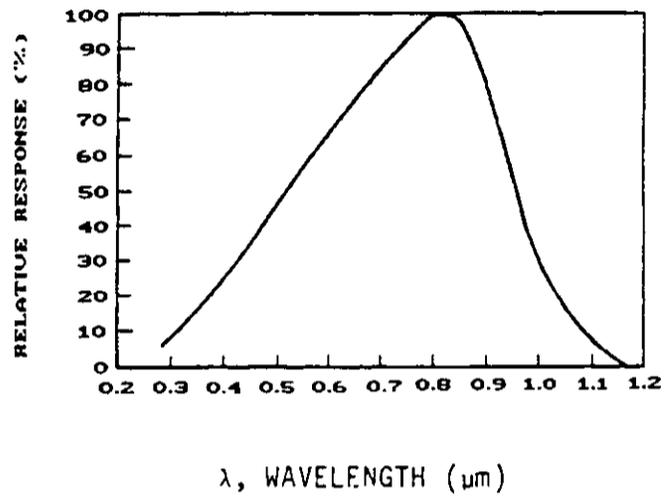


FIGURE 139. Spectral response.

The term reach-through arises from the photodiode operation. When a low reverse-bias voltage is applied, most of the potential drop is across the pn^+ junction. The depletion layer widens with increasing bias until a certain voltage is reached at which the peak electric field at the pn^+ junction is about 5 to 10 percent below that needed to cause avalanche breakdown. At this point, the depletion layer just "reaches through" to the nearly intrinsic region. The latter effect is illustrated in Figure 140.

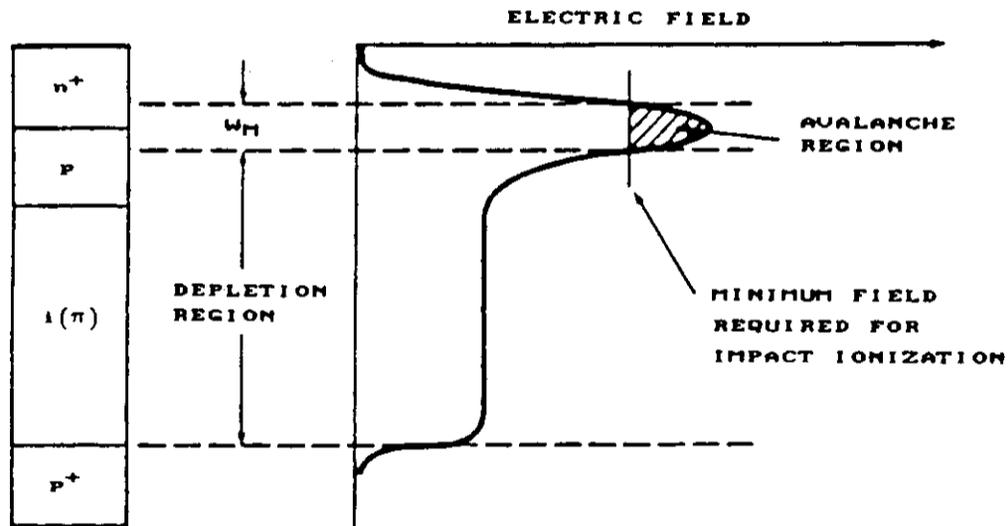


FIGURE 140. Reach-through avalanche photodiode structure and the electric fields in the depletion and multiplication regions.

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In normal usage the device is operated in the fully depleted mode. Light enters the device through the thin p^+ region and is absorbed in the π , intrinsic material, which acts as the collection region for the photogenerated carriers. Upon being absorbed the photon gives up its energy to create electron-hole pairs, which are then separated by the electric field in the π region. The photogenerated electrons drift-through the π region to the pn^+ junction where a high electric field exists (as shown in Figure 140). It is in this high-field region that carrier multiplication takes place.

In other words, electron-hole pairs generated by light signals generate many more electrons when the photodiode is in the avalanche state, so that a small amount of light generates a large change in photocurrent.

Table X gives typical performance values of an avalanche photodiode. The response time, listed as the rise time, is about 2 ns, which is not very much different from the PIN photodiode discussed earlier. Note that the responsivity in an avalanche photodiode can be as high as 75 A/W, which is many times greater than the responsivity of a PIN photodiode. Its noise equivalent power is high though, being 100 times greater than a PIN's NEP. At high-level optical signals, the shot noise component of the total noise makes the avalanche photodiode almost unusable.

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TABLE X. Electrical characteristics of an avalanche photodiode

Electrical Characteristics at $T_A = 22^\circ\text{C}$ At the dc Reverse Operating Voltage V_R Supplied with the Device	C30954E Light Spot Diameter 0.25 mm (0.01 in.)	Units
	Typ.	
Breakdown voltage, V_{BR}	375	V
Temperature coefficient of V_R for constant gain	2.2	V/ $^\circ\text{C}$
Gain	120	
Responsivity		
At 900 nm	75	A/W
At 1060 nm	36	A/W
At 1150 nm	5	A/W
Quantum efficiency		
At 900 nm	85	%
At 1060 nm	36	%
At 1150 nm	5	%
Total dark current, I_d	5×10^{-8}	A
Noise current i_n $f = 10$ kHz, $\Delta f = 1.0$ Hz	1×10^{-12}	A/Hz ^{1/2}
Capacitance, C_d	2	pF
Series resistance	-	Ω
Rise time, t_r $R_L = 50 \Omega$ $\lambda = 900$ nm 10% to 90% points	2	ns
Fall time $R_L = 50 \Omega$ $\lambda = 900$ nm 90% to 10% points	2	ns

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4.11.5.1 Device performance considerations. Photodetectors performance is based on five major criteria.

- a. Response time
- b. Responsivity and quantum efficiency
- c. Spectral response
- d. Noise
- e. Dark current.

In the following paragraphs, each of these criteria will be examined in more detail.

4.11.5.1.1 Response time and rise time. Response time is the speed at which a photodetector can change voltage or conductivity in response to a change in light intensity. Therefore, light pulses that are much shorter than a photo detector's response time will not be detected. Thus, it is important to choose a detector with a response time less than the pulse width of any signal that may be encountered in the system application.

Another name for response time is rise time. This is the time it takes the detector's output to reach the level that corresponds to the light intensity of the input signal. Figure 141 illustrates the idea of rise time.

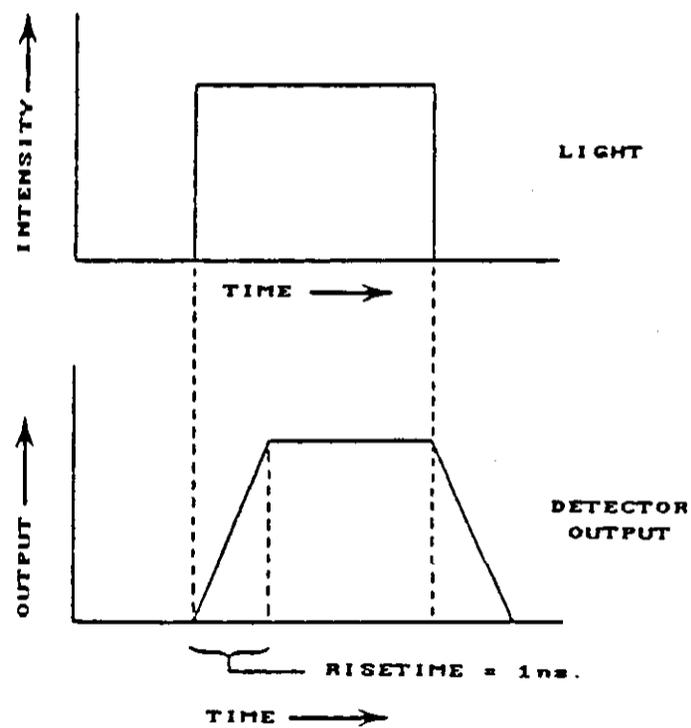


FIGURE 141. Rise time.

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The upper half of the figure represents the input light signal. The lower half of the figure represents the response of the detectors. Notice that upon reception of the light signal, the detector does not immediately rise to its full output until about 1 ns later which is a very good response time.

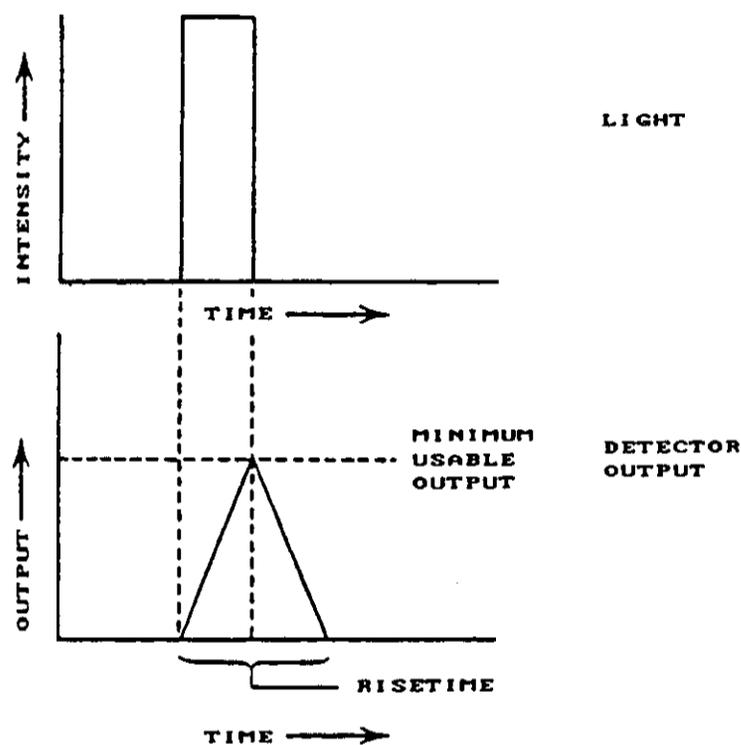


FIGURE 142. Insufficient rise time.

Figure 142 shows what happens when a narrow pulse is received by a very slow detector. Because the light pulse is narrower than the response time of the detector, the output of the detector never reaches its full value. If the output is too low, the receiver circuitry will not be able to distinguish it from noise. Therefore, the light signal pulse width and the detector response time must be compatible.

The response time is another way of measuring the amount of data a photo-detector can handle. If the response time is low, data pulses can be placed closer together and be made narrower. Because the data pulses can be made narrower in time, more pulses can be received per second than if the pulses were wider. Thus, a very short response time is a desirable feature in any photodetector.

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4.11.5.1.2 Responsivity and quantum efficiency. The performance of a photodiode can also be expressed in terms of its responsivity (R_A). The responsivity of a photodiode is simply a measure of how much output (reverse) current is obtained from a given light input. It is expressed as a ratio of output current (or photocurrent) in microamperes (μA) to the input radiant incidence which is measured in milliwatts per square centimeter (mW/cm^2) as shown below:

$$R_A = \frac{I_L}{P_0} = \frac{ng}{hv} + \frac{\mu A/cm^2}{mW/cm^2} \quad \frac{\mu A}{mW} = \frac{A}{W}$$

I_L is defined as the photocurrent generated when a given value of optical power (P_0) strikes the photodiode surface.

The responsivity of a photodiode also varies with the wavelength of the radiant energy striking the device surface and reaches a peak value when the quantum efficiency of the device is near its highest value. A typical photodiode will have a maximum responsivity of 1.4 μA per milliwatt per square centimeter. The quantum efficiency, η , is defined on page 4-120 with the terms q , h and v . In most photodiodes the quantum efficiency is independent of the power level falling on the detector at a given photon energy. Thus the responsivity is a linear function of the optical power. As shown above, the photocurrent I_L is directly proportional to the optical power (P_0) incident upon the photodetector, so that the responsivity R_A is constant at a given wavelength (for a given value of $h\nu$). Note, however, that the quantum efficiency is not a constant at all wavelengths because it varies according to the photon energy. Consequently, the responsivity is a function of the wavelength and of the photodiode semiconductor material because different materials have different band gap energies.

4.11.5.1.3 Spectral response. The foregoing graph shows the wavelength at which the detector is best suited to operate. At wavelengths away from the peak R_A value (vertical dashed lines), the amount of detectable signal falls off sharply, making communications more error-prone and inefficient. If a system is to be operated within a particular wavelength band, the spectral response error will reveal if the detector is suitable at those wavelengths. If the wavelengths have not been determined for the system, the spectral response curve will show what wavelengths are best for the detector.

Photodetectors, as can be seen in Figure 143, exhibit well defined, long wavelength thresholds (λ_c), the positions of which are determined by the magnitude of the band gap energy E_g or impurity activation-energy (E_A). The cutoff wavelength λ_c , corresponding to the given band gap energy is given by

$$\lambda_c = \frac{hc}{E_g} = \frac{1.24}{E_g(eV)}$$

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where the product of Planck's constant, h , and the velocity of light, c , is 1.24; E_g is expressed in electron-volts and λ_c in micrometers (microns).

As shown in Figure 143, the cutoff wavelength (λ_c) is about 1.1 μm for silicon, 1.88 μm for germanium and 1.65 μm for $\text{In}_{0.53}\text{Ga}_{0.47}\text{As}$. For longer wavelengths, as can be observed from the above equation, the photon energy is not sufficient to optically excite an electron from the valence band to the conduction band.

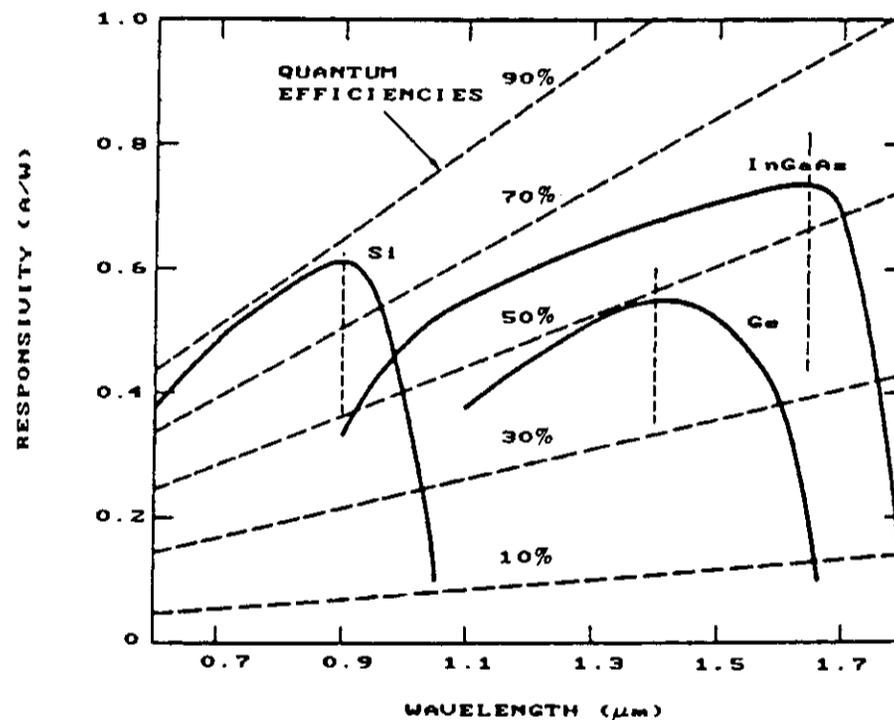


FIGURE 143. Spectral response curves (R_A vs λ) and quantum efficiency as a function of wavelength for PIN photodiodes for various semiconductor materials silicon, germanium and a III-V semiconductor alloy at 25 °C.

In Figure 144, the absorption coefficient α (measured in cm^{-1}) is plotted against the wavelength of incident radiation and the light penetration depth (μm) for various semiconductor materials used in photodiodes. At wavelengths longer than the band-edge wavelength (λ_c) (see Figure 144) the absorption decreases sharply. At the lower wavelength end, the photoresponse cuts off as a result of the very large α values at the shorter wavelengths. In this case the photons are absorbed very close to the photodetector surface where the recombination time of the generated electron-hole pairs is very short. The optically generated carriers thus recombine before they can be collected by the photodetector circuitry.

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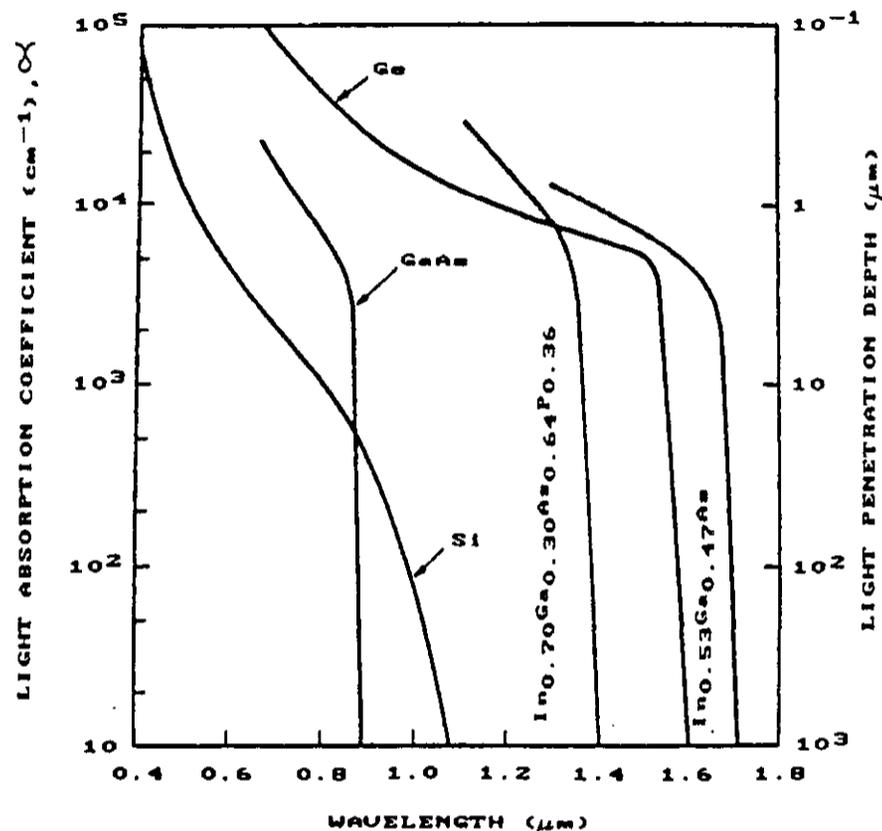


FIGURE 144. Absorption coefficient vs wavelength for Ge, Si, GaAs, InGaAsP, and InGaAs alloy at 25 °C.

A measurement similar to responsivity is quantum efficiency. Quantum efficiency tells what fraction of a single frequency light beam will be absorbed by electrons in a photodiode. In the ideal case, one electron-hole pair should be produced for each photon that strikes the diode surface, thus giving the ideal diode a quantum efficiency of 1 or 100 percent. However, the quantum efficiency of a typical photodiode will be lower than 100 percent and will vary with the wavelength of the incident radiant energy as shown in Figure 143.

The quantum efficiency, designated as η , is more quantitatively defined as the number of electron-hole carrier pairs (per unit area per unit time) per incident-photon (per unit area/unit time) of given energy $h\nu$ and is given by

$$\eta = \frac{\text{Number of electron-hole pairs generated}}{\text{Number of incident photons}} = \frac{(I_L/q)}{(P_0/h\nu)}$$

Here I_L is the average photocurrent generated by a steady-state average optical power (P_0) incident on the photodiode, q is the charge of an electron, and h is Planck's constant.

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4.11.5.1.4 Noise. Noise that is strong enough to mask a communications signal is called unwasted energy. In a photodetector, noise is made up of thermal noise and shot noise.

Thermal noise is caused by the thermal agitation of electrons in the detector. As the temperature rises electrons can absorb heat energy randomly, introducing a sporadic element into the detector's output. Usually this noise is so low (in the neighborhood of 2 nA) that it can be neglected, although it can be a problem in high temperature environments.

Thermal noise is not restricted to photodetectors. It can appear in other components as well. For instance, resistors in a receiver circuit can be a source of thermal noise. These extra thermal noise sources have to be considered when evaluating a system.

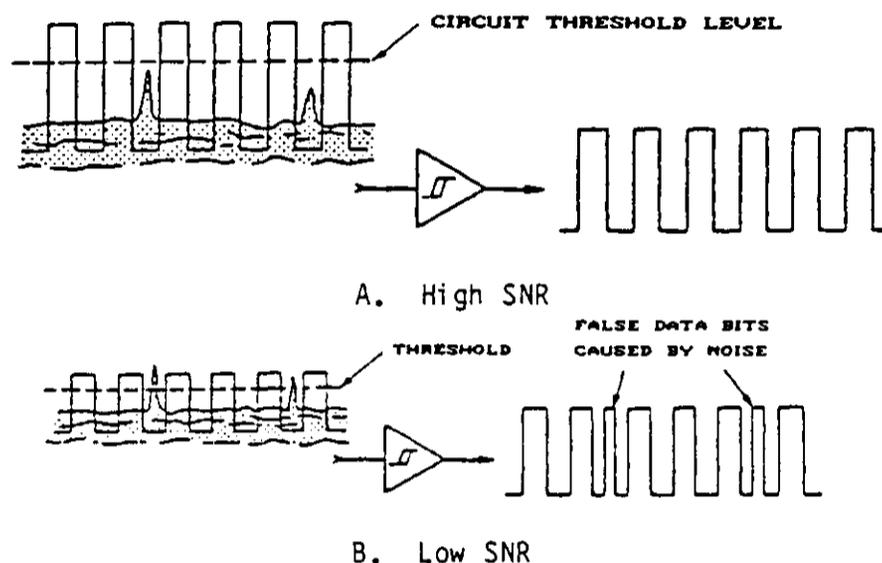
Shot noise is not a function of temperature, but of signal strength. As signal strength increases, that is, as the light intensity increases, the shot noise becomes greater. This is because the electrons are freed sporadically when light energy is present. Because of the numbers of electrons involved, the absorption appears to be even over time. But occasionally a large group of electrons are freed simultaneously, producing a burst of current, otherwise known as shot noise.

A measure of the effect of noise on a system is the signal-to-noise ratio (SNR). The SNR is the ratio of power of the detected signal and the underlying noise from all sources. It is commonly used as a measure of analog system performance.

Figure 145 illustrates the difference between a high and low SNR. The figure shows a signal and the background noise that is being applied to a pulse-restoring circuit. In Figure 145A, the SNR is high enough that the threshold detection level can be set well above the noise level. The output of the pulse-restoring circuit is then a faithful reproduction of the transmitted signal.

Figure 145B shows the same input signal greatly attenuated. To detect this signal, the threshold level of the pulse-restoring circuit must be set low. This makes it more likely that noise spikes will be detected and processed by the circuit, resulting in false data bits at the output. Therefore, to prevent loss of information, the detector must provide signals well above the noise level.

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FIGURE 145. The effect of SNR on received pulse data.

4.11.5.1.5 Dark current. Dark current is the current output of a photodetector when no light is present at the input of the detector. It is caused by leakage current through the detector, which is a result of the detector's reverse biasing. Because the leakage current increases with temperature, the dark current also increases with temperature. Excess dark current can also be caused by incorrect reverse biasing.

Dark current is also very high in the avalanche photodiode, being about 10 to 100 times greater than a PIN's dark current. Another disadvantage is a high reverse voltage needed to place the photodiode into its avalanche operating region. This reverse voltage is usually in the range of several hundred volts, which is much more than the voltage required by PIN photodiodes.

The avalanche photodiode is seldom used for short distance communications because of the cost of the device and its support circuitry. In very long distance work, though, its cost can be justified because of its high responsivity.

4.11.6 Environmental considerations. Typical environmental conditions and screens of photodiodes are similar to those given in paragraphs 4.1.6.9 Environmental considerations and 4.1.6.10 of Screening procedures in subsection 4.1 Diodes, general.

4.11.7 Reliability considerations. Because the failure mechanisms for photodiodes are similar to those of other diode types, paragraph 4.1.7 General reliability considerations in subsection 4.1 Diodes, general should be consulted for a discussion of diode failure mechanisms.

4.11 DIODES, PHOTODIODES

Because planar diffused photodiodes are usually fabricated from bulk silicon material having a range of 5 to 25,000 Ω -cm for the base region, semiconductor photodiodes with a base resistivity lying in the higher end of the resistivity scale would be much more sensitive to inversion layer formation at the silicon-silicon dioxide (thermal oxide) interface than conventional diodes.

Therefore, the incorporation of a guard ring structure (channel stopper) is necessary to minimize the surface contribution to the dc reverse leakage current in order to contract the magnitude of the dark current under reverse bias because the dark current is a combination of surface leakage current and bulk leakage current. The guard ring will also help prevent premature edge voltage breakdown failures, especially for APD devices which operate at a high-reverse bias near the breakdown voltage of the device.

Figure 146 shows the total dark current as a function of the active junction area with and without the guard ring. Note that large photodiodes do not have as significant a dark current reduction with the guard because a higher percentage of the dark current is due to bulk leakage. Under operating conditions where the guard ring and active area are biased at the same potential, the surface leakage is shunted around the load resistor and flows through the guard ring. In this manner, the much lower bulk leakage becomes the limiting source of shot noise current through the load resistor. Because the shot noise current of the detector varies directly as the square root of the leakage current, the total noise performance of the detector is greatly improved by the addition of the guard ring.

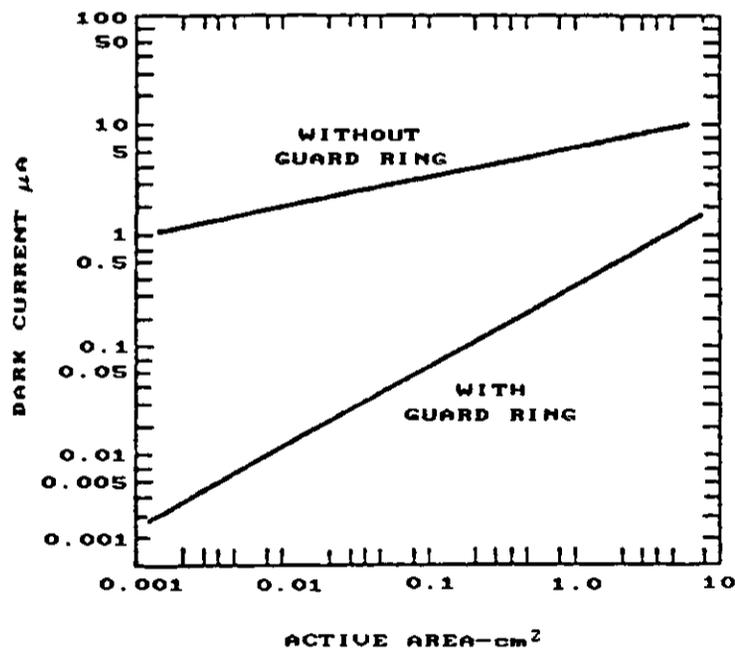
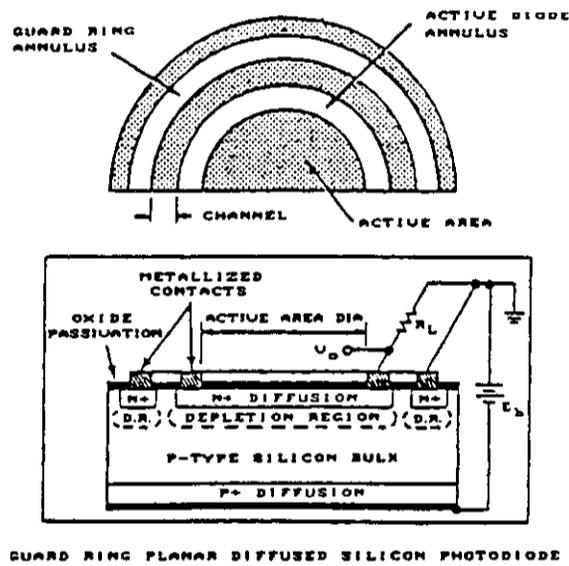


FIGURE 146. Dark current vs active junction area.

4.11 DIODES, PHOTODIODES

Surface leakage current is dependent on surface defects, cleanliness, bias voltage, and surface area. Bulk leakage current is dependent on the active area, the silicon resistivity and bias voltage. Both the surface leakage current and the bulk leakage current are temperature dependent. A good approximation for the temperature coefficient of dark current is that the dark current doubles for every 10 °C increase in operating temperature. Because dark current produces shot noise, it is evident that a reduction in dark current will improve the photo diode signal-to-noise ratio. The guard ring structure effectively accomplishes this by shunting surface leakage currents away from the load resistor as illustrated in Figure 147.



GUARD RING PLANAR DIFFUSED SILICON PHOTODIODE

FIGURE 147. Guard ring planar diffused silicon photodiode.

5.1 TRANSISTORS, GENERAL

5. TRANSISTORS

5.1 General.

5.1.1 Introduction. The transistor section is divided into five subsections: General; Low-Power Transistors which covers small-signal, switching, and chopper bipolar-transistors; High-Power Transistors which covers high-power bipolar-transistors; Field-Effect Transistors which covers low-power JFETs and MOSFETs devices; and Optocouplers. Radio frequency devices will be covered in Section 6.3, Microwave Transistors.

This general subsection will discuss the pertinent information that is common to all the subsections.

5.1.1.1 Applicable military specifications. There are many military documents that cover semiconductor devices. This section addresses itself to the primary documents.

<u>MIL-Spec</u>	<u>Title</u>
MIL-S-19500	General Specification for Semiconductor Devices
MIL-S-19500/---,	A group of military specifications used for procuring JAN, JANTX, JANTXV, and JANS transistors
MIL-STD-750	Test Methods for Semiconductor Devices
MIL-STD-975 (NASA)	NASA Standard Electrical, Electronics, and Electromechanical (EEE) Parts List
MIL-HDBK-217	Reliability Prediction of Electronic Equipment
DOD-HDBK-263	Electrostatic Discharge Control Handbook for Protection of Electrical and Electronic Parts, Assemblies and Equipment (Excluding Electrically Initiated Explosive Devices)

5.1.2 General definitions.

5.1.2.1 Abbreviations. Refer to MIL-S-19500 Appendix B for standard abbreviations used in this section. Only those not listed in MIL-S-19500 are listed herein.

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5.1.2.1.1 Chopper transistors.

IECS Emitter cutoff current
 hFE(INV) ... Static forward current transfer ratio (Inverted connection)
 r_{ec(on)} Small-signal emitter-collector on-state resistance
 V_{EC(ofs)} ... Offset voltage (Inverted connection)

5.1.2.2 Definitions. Definitions used in this handbook are the same as those found in MIL-S-19500 Appendix A.

5.1.2.3 Symbols. The following symbols are those generally accepted by electronic industries.

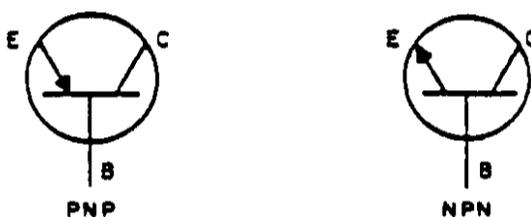
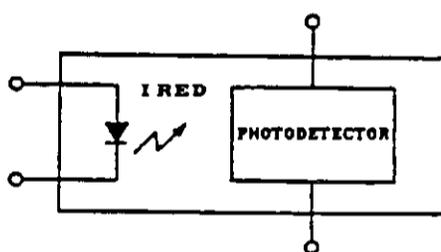
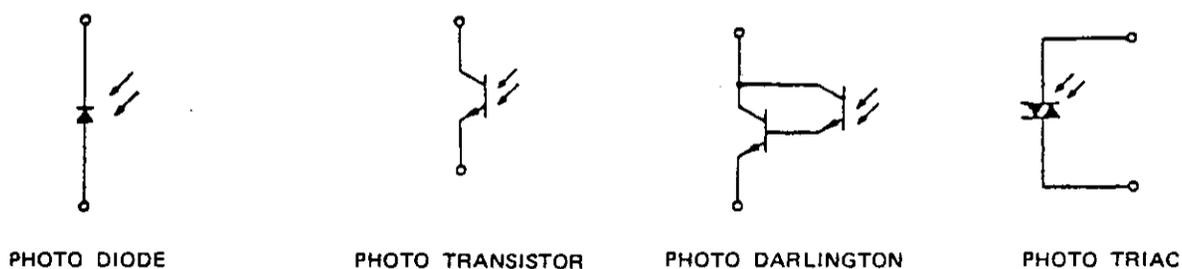


FIGURE 1. Bipolar transistors.



A. Basic optocoupler system



B. Symbol for photodetector

FIGURE 2. Optocouplers.

5.1 TRANSISTORS, GENERAL

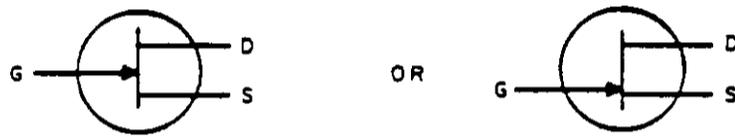


FIGURE 3. n-channel depletion mode JFET.



FIGURE 4. p-channel depletion-mode JFET.



n-Channel

p-Channel

FIGURE 5. Depletion-mode MOSFET.



n-Channel

p-Channel

FIGURE 6. Enhancement-mode MOSFET.

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5.1 TRANSISTORS, GENERAL

5.1.3 NASA standard parts. See General subsection 1.2 for a complete description of the NASA Standard Parts Program. In addition to this manual, the other principal element of this program is MIL-STD-975(NASA).

MIL-STD-975(NASA) is a standard parts list for NASA equipment with Section 13 containing a summary of standard transistors.

5.1.4 General device characteristics. The following is a brief description of the four basic types of transistors covered in this section.

5.1.4.1 Low power and chopper transistors. Low power transistors (<2 W) include npn and pnp devices that are used both in small-signal and switching applications. The chopper transistor is also a low power bipolar transistor with special parameters that are useful in chopper applications.

5.1.4.1.1 Small-signal. A transistor is said to be operating in the small-signal mode when the bias is such that the largest ac signal to be amplified is small compared to the dc bias current and voltage. Transistors used this way are normally biased at currents between 0.1 mA and 10 mA and voltage between 2 and 20 V.

5.1.4.1.2 Switching. The high value of off resistance and the low value of on resistance associated with a transistor makes the device as valuable as diodes for switching applications. The transistor has one important advantage over the diode. Its state is easily controlled using the base lead because a relatively small current in the base can control a large current in the collector where as the diode can only be switched by altering its bias. This switching gain makes the transistor a more versatile device. Switching transistors generally have a frequency operating range of greater than 500 MHz.

5.1.4.1.3 Choppers. A chopper is a low-power bipolar transistor intended primarily for use in an inverted operating mode in electronic chopper circuits. Its main characteristic is its low offset voltage $V_{EC}(ofs)$.

5.1.4.2 High-power transistors. This category covers devices that have a power dissipation range of two watts or greater. These devices are normally used in applications where the assumptions of linear operation are not valid and where variation in collector voltage and current are a significant fraction of the total allowable range of operation.

5.1.4.3 Field effect transistor (FET). A field-effect transistor consists essentially of a current-carrying channel formed of semiconductor material whose conductivity is controlled by an externally-applied voltage. The current is carried by one type of charge carrier only, with electrons in channels formed of n-type semiconductor material and holes in channels of p-type material; therefore, the field-effect transistor is sometimes called a unipolar transistor. This is to distinguish it from the bipolar transistor, which is a junction transistor the operation of which depends upon both types of charge carriers.

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There are two types of FETs: the junction FET (JFET) and the insulated gate FET (IGFET). The most commonly used insulated-gate FET is the metal oxide semiconductor FET (MOSFET).

MOSFETs can be structured to operate in both the depletion and enhancement-mode while JFETs operate only in the depletion-mode. MOSFETs are common both in the depletion and enhancement-mode for low-power devices. Also, both JFETs and MOSFETs are available as n- or p-channel devices.

5.1.4.4 Optocoupler. An optocoupler is a signal coupling device usually consisting of an infrared emitting-diode and a photosensitive semiconductor enclosed in a housing that is sealed against outside light. This device is characterized by nearly perfect input-output isolation due to the extremely low capacitance between the light source and the photodetector.

5.1.5 General parameter information. Electrical parameters will be discussed in each of the individual subsections.

5.1.6 General reliability considerations.

5.1.6.1 Failure modes and mechanisms. There are four failure modes for transistors: parametric degradation, shorts, opens, and mechanical defects. Tables I through IV describe the failure mechanisms for these failure modes, the methods of detecting the failure-causing defect before the parts are accepted by the user, and methods for reducing the probability of obtaining a part with such defects or of causing defects in good parts. It should be pointed out, however, that a very small percentage of parts will elude detection, and in-service failures will be encountered even with the most rigorous screening and process control methods.

5.1.6.1.1 Parameter degradation. Parameter degradation is the degradation of the electrical characteristics of a part. It is the most common failure mode and the most costly to detect. Although it usually does not result in a catastrophic failure of operating hardware, it can seriously degrade hardware performance if sufficient design margin is not allowed. Furthermore, it is usually indicative of a part whose expected life has been severely shortened. The mechanisms contributing to a parameter degradation are shown in Table I.

5.1.6.1.1.1 Contamination mechanisms. Contamination and corrosion are the most prevalent causes of parameter degradation and can be categorized by the process in which they are introduced: diffusion, oxide growth, washes, and gas ambient sealed in the can. Contaminated doping materials and ambient atmospheres during the diffusion processes result in wafers with a low yield of acceptable transistor die, usually as a result of improper resistivities or channel forming ions at the silicon-silicon dioxide interface. Bad dice are normally removed by the manufacturer through electrical testing.

5.1 TRANSISTORS, GENERAL

- a. Contamination detection methods. Elevated temperature bake-out for several hours in an inert atmosphere prior to capping is necessary to drive off surface contaminants. High temperature storage of sealed parts is effective in stimulating the effects of gas ambient and surface contaminants. The most effective method of detecting surface contaminants is through high temperature reverse bias (HTRB) testing. The temperature sufficient to activate the contaminating ion is usually 150 to 175 °C. Reverse bias of both junctions to achieve 10^6 V/cm field strengths is desired as it moves the ions to localized areas so that they can be detected on post HTRB leakage current measurements. Higher temperatures permit shorter test times but should be carefully considered to avoid starting other time dependent mechanisms, such as the Kirkendall effect. Burn-in at the rated junction temperature will aggravate the effects of bulk contaminants as well as corrosive contaminants. Large increases in leakage or degradation of gain will identify devices which are contaminated.
- b. Contamination minimization. Use of phosphosilicate glasses or silicon nitride passivations has been shown to be effective in gettering alkalis, thereby reducing their effects.

5.1.6.1.1.2 Bulk and oxide defect mechanisms. The next class of defects causing parameter degradation is somewhat related to contamination existing in the part. Contamination to some degree is inevitable so the more perfect the bulk and oxides, the less the effect of contaminants will be. Die faults and dislocations act as localization zones for mobilized gain failures.

Another phenomenon that is especially common to large area dice is inclusions of polycrystalline silicon, during the epitaxial growth, due to particulate contamination on the prepared silicon substrate surface. These propagate through the collector base junction and cause low breakdown voltages and high leakage currents.

Defective oxides having pin holes, cracks, misaligned oxide cuts, thin spots and other defects will contribute to several mechanisms. In bipolar transistors, mobile ions can migrate through discontinuous oxide to the silicon-silicon dioxide interface, depleting the base and thus degrading gain. In FETs and bipolar transistors, the discontinuities provide leakage paths from metal connections to active areas.

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TABLE I. Parametric degradation

Failure Mechanism	Description	Method of Detection	Method of Minimizing Defect
Surface contamination	<p>Ionic material on surface of Si or SiO₂ providing leakage paths (Cl, F1, etch)</p> <p>Alkali ions in or on oxides, creating channels (Na⁺, CA⁺⁺)</p> <p>Precipitated dopants on surface of silicon during oxide growth</p>	<p>High temperature reverse bias testing (HTRB) with temperature sufficient to mobilize contaminant and field concentrations sufficient to localize it, monitor leakage, and gain variations before and after test</p>	<p>Proper cleaning and etching sequences and recycling of solutions</p> <p>Practically unavoidable</p> <p>Keeping oxide growth temperature below diffusion temperature</p>
Contaminated gas ambient	<p>Noninert gases in the capping station</p> <p>Nonhermetically sealed cans</p> <p>Evaporated contaminants from die surface</p>	<p>Fine and gross leak seal tests to detect nonhermetic seal, sample residual gas analysis to detect lot problem with sealed ambient</p>	<p>Sealed capping station supplied with pure, dry, N₂ gas at pressure greater than 1 atm</p> <p>High temperature bake-out prior to capping</p>
Bulk defects	<p>Die cracks in active areas providing leakage paths</p> <p>Inclusions, dislocations, stacking faults, and other discontinuities providing localization zones for impurities</p> <p>Poor geometry design leading to areas of current crowding and hot spots</p>	<p>Power cycling will aggravate cracks; power aging at high temperature to localize impurities in fault areas, increase leakage and lower gain and breakdown voltage; close monitoring of gain and leakage before and after aging to detect parts with unstable characteristics</p>	<p>Visual inspection of each die, rejecting cracked ones</p> <p>Virtually impossible to eliminate all dislocations</p> <p>Careful die layouts and thermal analysis using thermal microprobe on an operating device to locate hot spots in design</p>
Oxide defects	<p>Cracks, holes, and thin spots providing leakage paths to exposed silicon</p>	<p>HTRB with temperature sufficient to mobilize contaminant and field concentrations sufficient to localize then monitor leakage and gain variations before and after test</p>	<p>Gross defects detectable with a visual inspection after oxide growth, high temperature growth desired to form dense, uniform oxides</p>
Die attachment	<p>Voids in eutectic which results in increased thermal impedance</p>	<p>X-ray normal to die looking for voids under die, precap visual inspection looking for voids around die</p>	<p>Proper temperature, scrubbing time, materials and cleanliness prevents voids</p>
Interconnect	<p>Resistive contact of metallization to silicon due to oxide growth in the window</p>	<p>High temperature storage will aggravate these phenomena, forward diode drop at each junction should be monitored, watching for slight increases (50 - 100 mV)</p>	<p>Complete oxide cuts and proper alloy procedure</p> <p>Temperature should be sufficient to form a Si-Al alloy</p> <p>Use monometallic interconnects, preferably gold wire, gold metallization</p>
Electrical stress	<p>Electrostatic discharges rupturing low leakage junctions</p> <p>Excess power and current</p>	<p>Identify all devices with 1 nA leakage current or less as susceptible and protect accordingly</p>	<p>Short all leads together until part is installed, handle in electrostatically neutral environment</p> <p>Current limits, transient suppressed power supplies and adequate oscillation suppression for dc operational</p>

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Other bulk defects which affect transistor operation, especially interdigitated power types, are misalignment of base and emitter diffusions or gradients in diffusion rates of the base and emitter that lead to irregular base widths and variations in emitter resistivity. These phenomena lead to localized secondary breakdown due to hot spot formation and punch-through under reverse bias. Die cracks which propagate into diffused areas reduce the breakdown voltage and increase the leakage current and ultimately produce shorted junctions.

- a. Bulk and oxide defect detection methods. Generally most defective dice are removed by the manufacturer prior to part assembly through wafer-level electrical tests. Visual die inspection will also eliminate dice with gross defects.

Thermal shock or power cycling will aggravate die cracks. HTRB and power burn-in at rated junction temperature with delta reject criteria on gain and leakage measurements will detect the majority of parts with time-dependent failures due to oxide and bulk defects.

- b. Bulk and oxide defect minimization. Discovery of most bulk defects during die fabrication is expensive and tedious usually involving high magnification microscopy. Bulk defects seem to be tolerated by industry provided that yields are high enough; however, since transistor dice with some bulk defects are unavoidable at the present state of semiconductor processing technology, they pose a life-limiting failure mechanism for a small population of parts.

Cleanliness and high temperature growth usually yield a very good grade oxide. A sample SEM inspection of the wafers will help detect oxide defects.

5.1.6.1.1.3 Die attach mechanisms. One of the most critical parameters of power devices and many small signal transistors is not electrical but thermal. Thermal resistance ($R\theta$) and junction-to-case ($R\theta_{JC}$), affects almost all other parameters because most parameters are dependent upon junction temperature. If this is too high, catastrophic part failure results. The primary interfaces affecting $R\theta_{JC}$ are the silicon-to-eutectic, eutectic-to-header, or insulator, and header- or insulator-to-stud. The most common defect affecting $R\theta_{JC}$ is voiding of the eutectic under the die leading to hot spots on the die and localized avalanching.

Voids can be caused by oxidized or contaminated silicon on the back of the die which precludes the formation of a good eutectic phase, intermediate phases formed in the eutectic die from diffusion or migration, inadequate gold-silicon preforms or platings, and poor die attach procedures resulting in the formation of undesirable intermetallics.

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- a. Die attach defect detection methods. X-ray of the header prior to capping, with the X-ray axis through the die surface, could be used to detect voids. Pre-cap visual inspections will detect voids around the edges of die but not under the die. Power aging at maximum junction temperature and high temperature reverse voltage tests will detect some of these failures. A nondestructive secondary breakdown test, increasing voltage to an acceptable threshold at elevated temperature, would eliminate many parts with hot spots on the die which contribute to premature secondary breakdown.
- b. Die attach defect minimization. Clean unoxidized surfaces and proper time-temperature controls for the attaching materials used are the only effective means for achieving proper die attachment. Gold-silicon eutectic materials afford the most reliable die solders, but sometimes even these may be unacceptable due to interaction with other part materials and instabilities with temperature and aging. Parts with soft solder die attachments should be carefully evaluated if power cycling requirements greater than 6000 cycles are required. Epoxies are not acceptable for use in high reliability applications.

5.1.6.1.1.4 Interconnect mechanisms. Interconnect flaws are another category of parameter degradation due to built-in defects. Interconnect wire degradations will be discussed later in the section dealing with opens, but their initial effect is an increase in contact resistance which increases the high current voltage drop in the forward direction.

Another phenomenon which increases contact resistance is the growth of silicon dioxide in the contact window; it is usually a result of incomplete etching of the silicon dioxide in the window.

- a. Interconnect defect detection methods: High temperature storage will aggravate both phenomena. High current forward voltage drop measurements, especially emitter-to-base, will usually detect parts with resistive contacts.
- b. Interconnect defect minimization. Proper cleaning of the base and emitter contact windows prior to metallization will help reduce interconnect defects.

5.1.6.1.1.5 Excess voltage, current, and power mechanisms. The last category of parameter degradation deals with the effects on structurally and electrically sound parts due to excess power, voltage, and current conditions. Whereas most overpower and current conditions result in shorts or opens, moderate excess power or current can lead to an increase in leakage and a decrease in gain as well as break-down due to formation of small channels or punch through sites in the junctions which are resistive in nature.

Degradation caused by overvoltage or current can be avoided through careful stress and thermal analysis of the part application and care in part and system level test conditioning.

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5.1.6.1.2 Shorts. Electrical shorts and their related failure mechanisms are shown in Table II. Most shorts are a result of poor workmanship by the part manufacturer and are easily detected by a vigorous pre-cap visual inspection. Parts which have permanent electrical shorts are rejected the first time they are electrically tested. Intermittent shorts usually occur during part test or installation and are very difficult to detect after the part is capped. These failures can be catastrophic to operating hardware if proper circuit redundancy is not provided.

5.1.6.1.2.1 Contamination mechanisms. The leading cause of shorts is conductive contamination internal or external to the part. Aluminum flakes that accumulate on probe points during the wafer level test that drop on the die across metallization paths, large fragments of silicon contamination that are introduced at the wafer level, excess eutectic that accumulate over the sides of the die, fragments of eutectic either attached or loose, and fragments and slivers of steel wool from tweezer cleaning, are common forms of contamination introduced during die attachment. Excess pigtailed and extra wires and balls from the wire cutting operation are types of conductive contaminants introduced during wire bonding. Improper header fabrication and plating results in flakes and slivers of Kovar or gold causing internal or external shorts. Improper storage and cleaning of die header assemblies and cans results in airborne contaminants, which are then sealed into the part. Weld slag balls can be introduced during the capping operation.

- a. Contamination detection methods. Bound particles can often be loosened with high-impact shock test. Most free particles can be detected with an electrically monitored vibration test, but this is usually expensive. Several part-users employ a technique referred to as particle impact noise detection (PIND) testing. Acoustic noise generated by particle impingement on the sides of the device can be monitored during random vibration coupled with a mechanical shock pulse. X-ray can detect large particles, two mils or greater, that are opaque to X-rays; this includes gold and iron.
- b. Contamination minimization. Conductive particles smaller than the smallest interconnect separation and nonconductive particles are not considered important. Ultrasonic cleaning of uncapped assemblies is one method of removing contaminants prior to capping. Glassivation over metallization is the best defense against conductive particles. A 100-percent pre-cap visual inspection would help weed out devices with contaminants.

TABLE II. Shorts

Failure Mechanism	Description	Method of Detection	Method of Minimizing Cause
Contamination	Aluminum flakes from probe points, silicon fragments, excess eutectic, excess pigtailed, extra wires, weld slag balls	X-ray Acoustic particle detection 100% precap visual	Ultrasonic cleaning Glassivation
Assembly defects	Improper die orientation, excess wire loops, bent posts, tall posts, misaligned bonds	100% precap visual X-ray Constant acceleration	Proper device design, use of proper assembly techniques, and implementation of in-process quality control check points
Die defects	Masking alignment, improper diffusion depth, lifting of undercutting of photoresist discontinuities, cracks, poor die attach, poor wire bonding, holes or thin spots in the oxide, metallization smears	100% precap visual High temp storage HTRB	Glassivation Guard rings
Excess voltage current and power mechanisms	Melted metallization, discolored or melted die vaporized bond wires	100% precap visual	--

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5.1.6.1.2.2 Assembly defect mechanisms. Assembly defects such as improper die orientation that forces interconnect wires to cross, connection of the wrong contact to the external pins, excess wire loops, down bonding where inadequate clearance is provided between the edge of the die and the wire, misalignment of wire on bond pad, and bent posts or those that are too high so that they touch the side of the can, can cause shorts.

- a. Assembly defect detection methods. A precap visual inspection eliminating parts with such defects is desired. X-ray will show crossed wires and improper post-to-cap clearance. Constant acceleration in all three axes will cause excess wire loops to touch other areas in the can. Constant acceleration in the Y1 direction will also tend to lift poor wire bonds off the bonding pads.
- b. Assembly defect minimization. Proper device design, the use of proper assembly techniques, and implementation of in-process quality control check points will greatly reduce assembly defects.

5.1.6.1.2.3 Die defect mechanisms: Die defects causing shorts can be placed in three categories: bulk, oxide, and metallization. The bulk and oxide defects which cause parameter degradation, discussed previously, often produce shorts. Masking misalignment and improper diffusion depths can result in shorted junctions. Lifting and undercutting of the photoresist or mask defects and misalignment can result in bridged junctions or metallization. Dislocations, inclusions, and other discontinuities in the junctions can also cause shorts.

Cracks in the die as a result of scribing or improper die-attach and wire bonding provide channeling paths along exposed silicon surfaces. Cracked dice can also result from gross mismatches in the temperature coefficient from header to die in power devices. Cracks, holes, and thin spots in the oxide under metal paths frequently result in shorts to the active areas under the metal for expanded contact devices (wire bond at the die is made on the oxide rather than at the oxide window). The contaminants on the die surface or buried in the oxide often produce inversion layers and channels sufficient to short junctions. Smears of metallization due to improper handling during the die-attach or wirebonding operation may also short active areas.

- a. Die defect detection methods. A good precap visual inspection will eliminate most gross die defects. High temperature storage and HTRB will aggravate oxide and bulk defects and mobilize contaminants to form shorting channels and inversion layers.
- b. Die defect minimization. Glassivation of dice has been successful in reducing smearing of metal. Thick eutectics or temperature coefficient matching phases such as molybdenum tabs are desired in power devices to eliminate thermally induced cracks. Channel stoppers are desired around the outside of the die to keep channels from propagating to the edge of the die in single diffused epitaxial devices, and guard rings

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or field plates around the base collector junction are often used to prevent channels in other geometries. Extension of base metallization over the base-collector junction line is a commonly used type of field plate.

5.1.6.1.2.4 Excess voltage, current, and power mechanisms. The last of the shorting mechanisms is electrical overstress. The most common is excess current causes aluminum from one metallization path to traverse across the surface of the silicon, underneath the oxide, to a metal path of different potential and forming what are known as white spears. Bulk breakdown due to localized thermal avalanche, punch through, and secondary breakdown is usually caused by excess temperature with bias applied. High frequency oscillations often occur during burn-in or circuit operation where degenerative feedback is not provided. Melted metallization, discolored or melted dice, and vaporized bond wires are all signs of misapplication of bias and signal or improper heat sinking.

5.1.6.1.3 Opens. The failure mechanisms producing opens are displayed in Table III. These mechanisms are frequently time-dependent and often require long periods to produce failures, and they can also be intermittent. These mechanisms pose a considerable question to the long-life performance of transistors. Open transistors are generally less catastrophic to hardware failure than shorted ones because they do not load power supplies, but they can still result in hardware failure due to loss of drive to succeeding stages. Redundancy is necessary where single-point transistor opens can cause catastrophic hardware failure.

5.1.6.1.3.1 Interconnect wire failure mechanisms. The most common cause of opens is interconnect wire failure. There are two types of metallization-interconnect wire systems used, gold wire-to-aluminum metallization and aluminum wire-to-aluminum metallizations. Each causes problems when not done properly.

The most common interconnect system in transistors is the gold wire-to-aluminum metallization interconnect. Insufficient time, temperature, or pressure during the formation of the bond results in weak or incomplete bonds.

A more significant mechanism is the Kirkendall effect, commonly called plague. Seemingly strong bonds grow weak with temperature and time due to intermetallic compound formation. Bond weakening is most severe in expanded contacts, but failures have also been observed in direct contacts.

The Kirkendall effect is the interdiffusion of two dissimilar metals at different rates producing voids at the metal interface. This phenomenon takes three forms, annular voids around the outer periphery of the ball, lateral voids along the entire surface of the ball, and depletion of the aluminum film.

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TABLE III. Opens

Failure Mechanism	Description	Method of Detection	Method of Minimizing Defects
Open interconnect wire at the die NOTE: The most common intermittent open-normal-open failure is due to die-wire bond failures	Gold-aluminum: Intermetallic formations (plaque) Kirkendall voids under ball Excess pressure resulting in nicked wire exiting ball Depletion or peeling of Al film around and under ball Aluminum-aluminum: Microcracking at heel of bond due to poor tooling and excess bonding pressure Both: Insufficient time, temperature pressure, or energy leading to incomplete bonds Insufficient bond pad area on die	100% bond pull (non-destructive to an acceptable threshold level) to detect weak or poorly formed bonds. Threshold for Au-Al 2 grams min pull, Al-Al 0.7 gram min. Low duty power cycling will aggravate microcracking in Al-Al wedge bonds. High temperature storage or cycling followed by mechanical shock will aggravate plaque and void failures	Use monometallic bonding system Oxygen in gas backfill seems to retard growth Visually inspect all bonds for improper ball dimensions and wire necking Proper alloying (sintering) of Al film prevents peeling Proper tooling design and controlled bonding parameters Control over bond parameters Bond pad area must be greater than the final surface area of the bond (i.e., metallization visible around bond)
Open interconnect wire at the post and along the wire	Insufficient weld area at the post Burned or incomplete welds Si inclusions in wire or greater than 1% content (Al wire) Twists, nicks, crimps, kinks, or scratches in wires which reduce the wire diameter Corrosion on or in the wire from drawing or cleaning operations which corrodes metallization or otherwise interferes with bonding	Visual inspection of wires and bonds will detect gross defects, 100% bond pull (nondestructive to an acceptable threshold) will detect weak wires and bonds, thermal shock will aggravate incomplete or improperly formed post bonds, high temperature storage followed by thermal cycling will anneal and break poor quality aluminum wire	Use double bond at the post Proper weld or bonding schedule Close metallurgical surveillance of wires used, 1% Mg wire appears to be stronger Workmanship Clean all wires before use
Die lifted from header	Voids in eutectic die mount Failures of ceramic insulation to support die under mechanical stress (for dielectrically insulated parts) Cracked die due to thermal mismatched or die and header (power devices)	Constant acceleration at high g levels will expose these failures	Workmanship Mount ceramics directly to header, do not support from leads Use thick eutectics or intermediate materials to provide some stress relief
Metallization	Photolithography defects (stained or dirty masks, lifted or underexposed photoresists, etc.) resulting in incomplete metallization Improper annealing (sintering) resulting in peeling or lifting metallization Scratches opening metal paths Scratches reducing cross-sectional area of metal resulting in migration failures (expanded contact and interdigitated geometries) Insufficient design width and thickness resulting in excess current densities and burned metal or migration	Electrical test will detect permanently open parts, visual inspection of die will detect gross metal defects, power burn-in at reasonably high current levels will accelerate migration failures NOTE: Electromigration of Al is only a problem on devices with expanded contacts and is discussed more in the microcircuit section	Clean masks, uniform photoresist deposition exposure and removal will help Sintering temperature should be reasonably high Glassivation over metal to protect it Glassivation over metal to protect it Maximum current densities should be 5×10^4 A/cm ² to minimize migration effects

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The next most widely used bonding system in transistors is aluminum wire, with 1 percent silicon, and aluminum metallization. The two failure mechanisms in this system are wire breakage due to silicon inclusion in the wire and micro-cracking at the heel of the bond due to necking of the bond by the bonding tool. Both failure mechanisms are time dependent and are a function of the thermal cycling, usually low rate, which causes mechanical movement and annealing of the wire.

Both gold:aluminum and aluminum:aluminum interconnect systems are prone to failure if the post bond does not provide sufficient wire-to-post surface contact. In addition, nicks and kinks in the wire, which reduce its designed diameter, could result in failure due to excessive current densities.

- a. Interconnect wire failures, detection methods. In both gold:aluminum and aluminum:aluminum interconnect systems a stringent 100 percent precap visual inspection of the die and post bonds and the wire will detect most common interconnect defects. A sample bond wire pull test could also be performed to assure that a good bonding schedule is being maintained by the manufacturer.
- b. Interconnect wire failures minimization. The majority of suppliers use gold:aluminum or aluminum:aluminum interconnect schemes. Double post bonds are recommended for either system to provide good adhesion to the post. Gold ball thermocompression bonds and ultrasonic aluminum bonds are some of the acceptable means of bonding the wire to the die. Wire quality and material content as well as metallization thickness and texture must be controlled by the manufacturer for reliable bonding. Bond location and adequate pad size also are important controlling factors in reliable interconnects.

5.1.6.1.3.2 Lifted die mechanisms. Excessive voids in the eutectic bond or undue mechanical stress can cause the die to lift off the header. In parts with dielectrically isolated collectors or gates, ceramic insulators which do not have adequate mechanical support to the header can break dice during mechanical stress and cause opens.

- a. Lifted die detection methods. Parts which have marginal die attachments can usually be detected through X-ray of the die header assembly, when looking for voids. Power pulse $V_{CE(SAT)}$ and $R_{\theta JC}$ measurements will also detect such defects.
- b. Lifted die minimization. Die and insulators should be firmly attached to the header rather than suspended from the leads. Ceramic temperature coefficients should be matched closely to silicon to preclude cracking due to thermal expansion.

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5.1.6.1.3.3 Metallization failure mechanisms. Metallization opens are usually a problem only in expanded contact devices since current is carried from the die to the wire bond by paths of metallization. Scratches due to improper die handling and missing metallization due to photolithography defects are commonly the sources of open metallization paths. Aluminum can also migrate from thin areas to create voids. This is a problem in rf and power devices where the metallization cross-sectional area and current requirements cause excessive current densities in the aluminum film. Improper alloying (sintering) of the aluminum film to the silicon and silicon dioxide surfaces results in peeling and lifting of the metallization. This also can be a problem in direct contact devices. Another mechanism is failure of the metallization to make contact with silicon due to an incomplete etch of the silicon dioxide at the window or growth of silicon dioxide at the interface of the silicon and aluminum. Molybdenum gold metallization system failures are sometimes caused by excessive undercutting of the molybdenum during etching or inadequate alloying of the molybdenum.

- a. Metallization failure detection methods. A precap visual inspection will detect open or degraded metallization fingers in uncapped parts. Electrical testing will reveal those sealed parts having open metallization paths. Metal voiding due to migration can be discovered by a forward burn-in at rated current. Several integrated circuit users require a sample SEM inspection of each wafer to determine adequate metal coverage over oxide cuts.
- b. Metallization failure minimization. Use of phosphosilicate glasses retards metal migration, except in gold metallization systems, and also protects the metal surfaces from scratches during die handling. High temperature storage will stimulate the growth of oxide at the contact window. The use of gold metallization significantly reduces metal migration effects because of the relative immobility of gold; however, while the gold is immobile, the moly underneath is not.

Derating maximum current limits to keep metallization current densities below 5×10^4 A/cm² will considerably reduce electromigration effects.

5.1.6.1.3.4 Excessive voltage, current, and power. Usually, overpower conditions will melt interconnect wires or vaporize metallization films due to excessive current density. Shorts in the die also cause open interconnects under normal power conditions.

5.1.6.1.4 Mechanical degradation. The last group of failure mechanisms is shown in Table IV and deals with failures that prevent proper installation of the part. While these failures are normally discovered during and just after part installation, they cause significant delay and occasionally go undetected until a considerable investment has been made in the using hardware. They are most easily detected prior to installation by an external visual inspection.

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TABLE IV. Mechanical defects

Failure Mechanism	Description	Method of Detection	Method of Minimizing Defects
Unsolderable or unweldable leads	Finger greases or acids on leads Exposure of leads to corrosive solvents Exposure of part to O ₂ and H ₂ O in high temperature environments Poor plating Improper metal content Improper dimensions Bent, broken, nicked, scratched, or twisted leads due to improper handling or test equipment	External visual inspection, dimensional inspection, solderability/weldability tests	Use of tweezers and gloves when handling parts Control of cleaning processes and storage of parts Use of dry N ₂ environments during high temperature storage QC inspection by part manufacturer of plating integrity QC inspection by part manufacturer of lead metals QC inspection by part manufacturer QC inspection by part manufacturer
Degraded seals	Cracked glass due to mishandling Bubbles or voids in glass Improper can-to-header seal ring dimensions Contaminated can-to-header seal ring Improper can-to-header weld schedule	External visual inspection; hermetic seal tests, fine and gross, with maximum leak rate of 5×10^{-8} Atm/cc/sec	100% inspection of headers upon receipt by part manufacturer 100% inspection of headers upon receipt by part manufacturer 100% inspection of headers upon receipt by part manufacturer 100% inspection of headers upon receipt and proper cleaning Regular certification of weld schedules and welders by part manufacturer
Degraded part package	Exposure to corrosive solvent or gases Finger acids and greases Improper dimensions and material content Deformation of part can or header due to excessive mechanical stress Galled or stripped threads and broken studs for stud-mounted packages Poor finish and flatness on bottom of stud surface precluding proper thermal interface Improper insulator resulting in poor mechanical support or high thermal impedance for collector or gate insulated devices	External visual inspection, dimensional inspection External visual inspection, thermal impedance measurement base-to-headersink on installed parts Thermal shock/high impact shock with electrical measurements prior to and post shock thermal impedance measurement junction-to-case	Store and test parts in dry inert environment Handle parts with tweezers or gloves QC inspection of headers and cans by manufacturer Control of mechanical shock and acceleration fixtures and handling and packaging Use of properly derated stud torque values by part user, burn-in power parts in free air using free air ratings to achieve max temperature QC inspection of headers and lot test of thermal impedance case to heat sink by part manufacturer Use ceramic insulators which are solid (unperforated or slotted) that are eutectically mounted to part header rather than supported by lead parts
Mismarked or unmarked	Wrong part numbers placed on part Unintelligible marking smears, blobs Tags applied to leads Part marking inks/paints soluble in cleaning solutions	External visual inspection Marking permanency test	Workmanship Workmanship No tags applied to leads, stamp part ID on package Use of indelible inks/paints

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5.1.6.1.4.1 Degraded leads and packages. Corrosion of leads and cans results from any one of a number of poor practices. Handling of parts without hard protection and storage of parts in corrosive or oxidizing environments degrades leads preventing sound solder joints when the part is installed. Oxidized surfaces on the can where heat flow is required increase thermal impedance. Poor plating also contributes to solderability problems and reduces thermal conductivity. Improper material content and use of improper dimensions result in poor welds when contact resistance welding is used. Contamination can result in weakened joints. Scratched, bent, broken, and twisted leads make the part more difficult to install properly.

5.1.6.1.4.2 Studs. Galled or stripped threads, irregular or hollow surfaces on base of stud, and improper attachment of ceramic insulators, for dielectrically isolated devices, result in failures of stud mounted devices. The latter two usually cause a marked increase in thermal impedance of the device resulting in overheating and die failure. Damaged threads usually result in broken studs when proper torque is applied or in insufficient torque due to added mechanical resistance.

5.1.6.1.4.3 Seals. Cracks and bubbles around the external leads result in loss of hermeticity. The effects of a contaminated ambient have been discussed previously.

5.1.6.1.4.4 Part marking. Soluble inks, smeared or smudged marking, or absence of marking may cause the loss of traceability of screened qualified parts. The use of insoluble inks is mandatory. All part marking is to be stamped on the can. Do not use tags on leads or cans which may outgas into the hardware in which the parts are used or which would otherwise degrade the parts. An external visual inspection of all parts will eliminate such defects.

5.1.6.2 Transistor failure rate models. Failure rate models for these devices can be found in MIL-HDBK-217.

5.1.6.3 Environmental considerations. All MIL-STD-975 transistor devices require a vigorous screening and conformance testing per MIL-STD-19500 and its applicable slash sheet. These tests are performed to assure that these devices can withstand certain levels of environmental conditions; for example, pressure, vibration, moisture, temperature cycling, mechanical stress, and, if required, radiation. The user of these devices should take precautions so that these levels are not exceeded in the design of a system. Nevertheless, the following are some of the important environmental considerations.

5.1.6.3.1 Electrostatic discharge. One of the environmental factors which is not screenable is electrostatic discharge (ESD); therefore, the user must take special precautions when handling ESD sensitive devices; refer to DOD-HDBK-263 for guidance.

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5.1.6.3.2 Radiation considerations. To insure that a transistor functions properly in space applications, the design engineer must consider the effects of radiation exposure. A designer of radiation resistant systems must know how radiation affects the circuits and components of the system. When semiconductor devices are exposed to radiation environments, changes occur in their rated electrical parameters. The magnitude of the changes is a function of such things as the type of radiation (neutron, proton, electron, gamma or heavy ions) and the duration of exposure.

Transistors exposed to radiation may display changes in the following electrical characteristics:

- a. Breakdown voltage decreases
- b. V_{CE} (sat) increases
- c. Decrease in h_{FE}
- d. Leakage current increases
- e. Decrease in $V_{GS(th)}$ N-channel MOSFET
- f. Increase in $V_{GS(th)}$ P-channel MOSFET
- g. $V_{DS(on)}$ increases

Therefore, when designing with transistors, the design engineer must consider the fact that these parameters will change. The degree of parametric alteration and the long term effects are dependent upon the device technology and the manufacturing process. These vary from vendor to vendor and device to device. Power MOSFETs in particular are very susceptible to total dose induced radiation damage, and non-radiation hardened devices may fail at levels as low as 2 Krads (Si). Also, recent testing has shown that these devices exhibit catastrophic burn-out failure when exposed to energetic heavy ions. Since the cosmic rays present in natural space environment do contain some heavy ion flux, it is recommended that, for critical applications the specifics of each device type should be discussed with a radiation expert and/or manufacturer. For a more detailed discussion of radiation and radiation effects, see the section in microcircuits, paragraph 7.1.7.7.

5.1.6.4 Application considerations.

5.1.6.4.1 Thermal considerations. Temperature is one of the most critical reliability factors. For example, if the failure rate for a device is 0.01 failures/ 10^6 hours at a junction temperature of 75 °C, at 150 °C the failure rate would be 600 times that value, assuming a surface-inversion failure mechanism with an activation energy of 1.0 eV, which is a common failure mechanism for silicon transistors. Therefore, precautions must be taken to maintain a device junction temperature at the design operating level throughout its useful life. Some of the design precautions that can be taken are as follows:

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- a. Design for the lowest feasible operating junction temperature.
- b. Heat sink design and device mount down are crucial. For instance, a reduction in $R_{\theta JA}$ of 1.0 C/W for a power device dissipating 25 W would reduce the operating junction temperature by 25 °C.
- c. Assure that the device mount down surface is flat or slightly convex to provide proper seating.
- d. The device case temperature should be measured under actual operating ambient conditions to assure that the design operating junction temperature is not being exceeded.

$$T_J = T_C + R_{\theta JC} * P_d$$

- e. Allow sufficient latitude in circuit design to accommodate some changes in device parameters with time.

5.1.6.4.2 Electrical considerations. To assure proper device operation throughout the lifetime of a system the following design criteria should be considered:

- a. Allow for aging of electrical parameters.
- b. Protect critical devices from transients by the use of snubbers, clamps, or other protective means.
- c. Keep in mind device parameter variation with temperature when designing a circuit.
- d. Beware of the unspecified parameter. It is possible that a JANTXV2N3714 from one vendor will function well in a circuit whereas the same device from another vendor will not, even though the device meets all specified parameter limits. When possible, devices from all possible sources should be tested under actual operating conditions.

5.1.6.4.3 Mechanical considerations. A mechanical reliability problem for hermetically sealed devices is handling. Bend leads can break the nonhermetic seal and can also break or loosen bond wires. Careless handling can cause particles to be released inside the package and that causes problems; therefore, care must be taken when handling hermetically sealed devices.

5.1.6.5 Screening. Both environmental and electrical screening are performed in accordance with the MIL-S-19500 slash sheet on all MIL-STD-975 devices. Users of these devices should become familiar with the test requirements to assure that they are not exceeded in the design of a system.

5.1.6.6 Derating factors. Derating factors for these semiconductor devices are in MIL-STD-975, and the user should refer to this document for derating factor guidelines.

5.2 TRANSISTORS, LOW-POWER

5.2 Low-power.

5.2.1 Introduction. In recent years, integrated circuits have rapidly replaced many of the low-power functional circuits in which low-power discrete transistors were used. This is especially true in the case of digital and linear applications. The user should also refer to the microcircuit section of this handbook concerning these types of applications.

This low-power section will cover small-signal switching, and chopper transistors.

5.2.1.1 Small-signal. The transistor is a nonlinear active device. Figure 7 illustrates that, although slightly nonlinear throughout its range, the transistor's nonlinearities become very pronounced at the very low and high current and voltage levels below point A and above point B. If an ac signal is applied to the base of a transistor without dc bias, conduction would take place only during one half-cycle of the applied signal, and the amplified signal would be highly distorted. To avoid this problem, a dc bias operating point "OP" is chosen (see Figure 7). This bias moves the transistor's operation to the more linear portion of its characteristics. There the linearity, although not perfect, is acceptable and results in amplification with low signal-distortion.

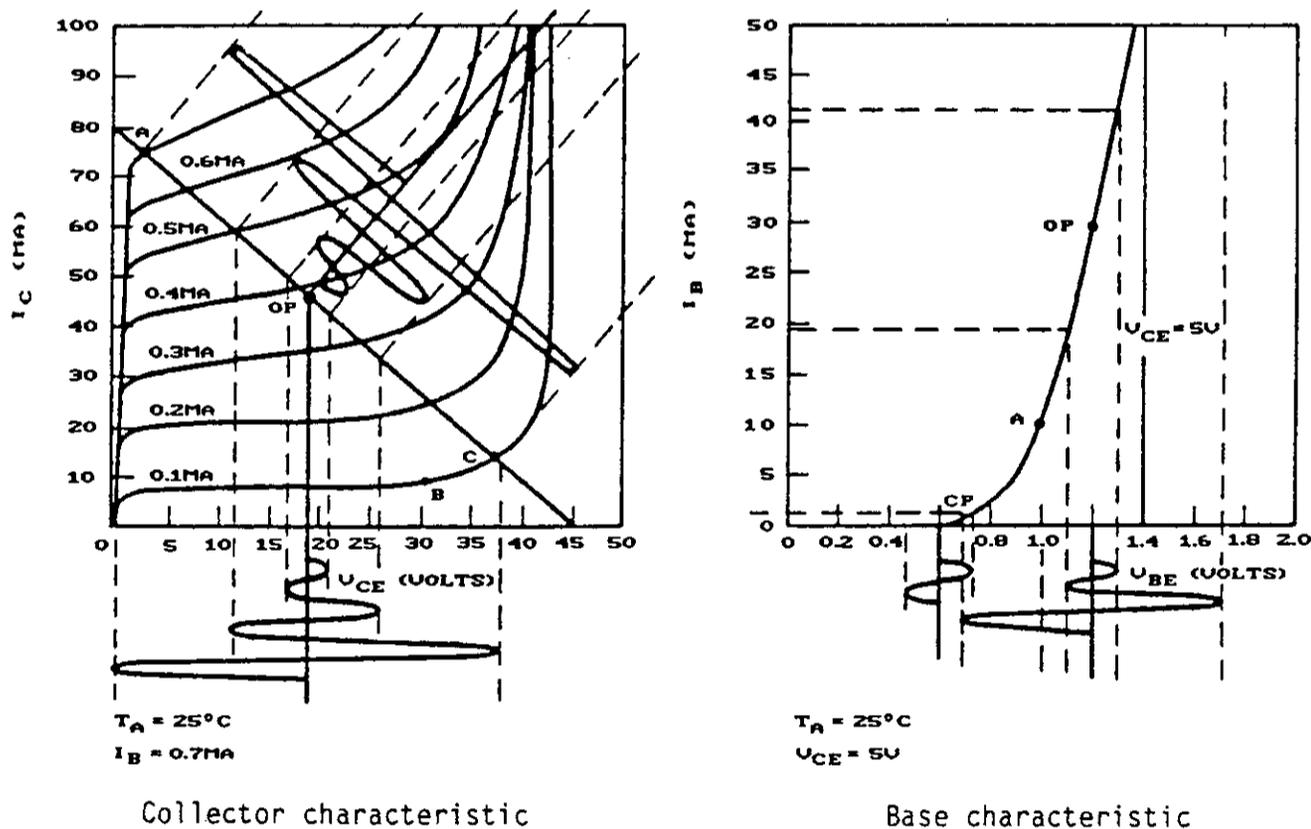


FIGURE 7. Typical V-I characteristics of a silicon planar transistor.

5.2 TRANSISTORS, LOW-POWER

The application of a dc bias is still not sufficient. A transistor could be biased exactly in the middle of its linear range and be operated at such large signal swings (see Figure 7) that the signal would encroach upon the nonlinear area and results in increased distortion. This is quite common in Class A audio output or driver stages.

In a great number of transistor applications, normal operating signal levels are small. Examples of such applications are the rf and most amplifier stages of radar, radio, and television receivers. Even after detection, as in audio or servo preamplifiers, the signal can be at a low-level.

In low-level stages, signal swings run from less than 1 μ V to about 10 mV under normal operating conditions (for which these stages are generally designed). Therefore, it is important to analyze the transistor under conditions where the largest ac signal to be amplified is small compared with the dc bias current and voltage. The transistor is then said to be operating in the small-signal mode. Transistors used in this way are normally biased at currents between 0.1 and 10 mA and voltages between 2 and 10 V. An insufficient bias can cause distortion whereas an excessive bias exhibits unnecessarily increased power dissipation and a higher noise figure (the latter is primarily important in input stages). If the bias is sufficiently increased to make the stage operate in the high voltage nonlinear region, distortion will increase.

5.2.1.2 Switching. The most common usage for transistors is as a direct analog of the mechanical switch. That is, the transistor is made to have a high resistance (off-state) and a low resistance (on-state).

Figure 8 shows the on-state and off-state conditions. Point A occurs at a low current I_0 and relatively high voltage implying a high resistance or an off condition. In a properly designed circuit, I_0 will approach I_{CBO} . This current can be extremely small in a silicon planar transistor. Point B occurs at a relatively low voltage and high current point implying low resistance or an on condition.

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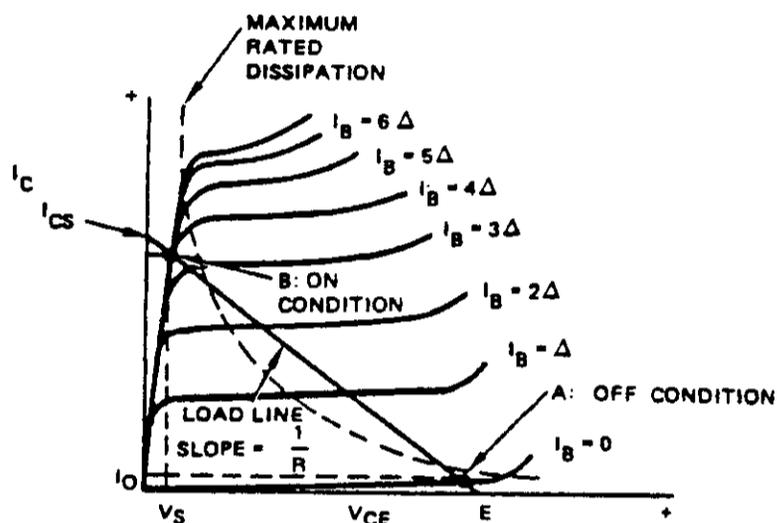


FIGURE 8. Generalized voltage-current characteristics for an npn transistors.

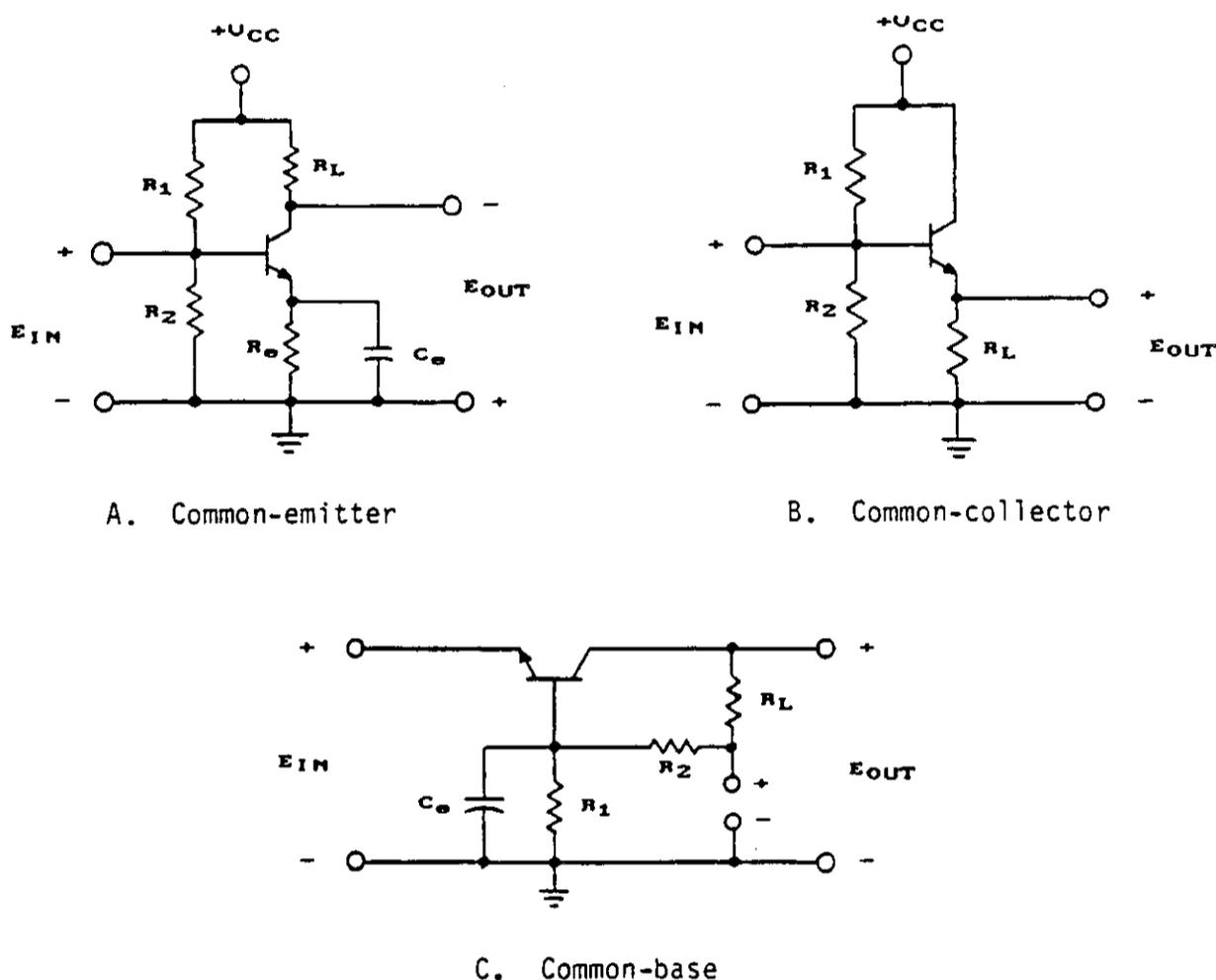
5.2.1.3 Choppers. A bipolar chopper transistor has similar characteristics (in the standard bias configuration) as a low power bipolar transistor; however, in the inverted configuration (where the collector is connected as the emitter and the emitter as the collector) the bipolar chopper transistor is designed to operate with a higher gain, a higher BV_{ECO} , and a lower V_{ec} . These characteristics make the chopper transistor useful in both ac and dc chopper applications.

5.2.2 Usual applications. There are many applications in which the small-signal, switching, and chopper transistors can be used but only the basic applications are covered in this section.

5.2.2.1 Basic amplifiers. There are three basic types of biasing configurations. These are the common-emitter (CE), common-collector (CC) (also known as an emitter follower), and the common base (CB).

The basic common-emitter configuration, shown in Figure 9A, is the most versatile and is capable of giving both voltage and current gain. Its input and output impedance vary the least with variations in the input and output load impedance, and its input signal is inverted at the output.

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FIGURE 9. Basic biasing configurations.

The basic circuit of a common-collector is shown in Figure 9B. It has a current gain that approaches the CE configuration, but has a voltage gain less than one. Its main characteristic is its high input impedance. It has wide applications as a buffer stage between a high and low impedance source. There is no signal phase inversion when using this configuration.

The basic common-base circuit configuration is shown in Figure 9C. This configuration does not provide current gain but it provides voltage gain that approaches the CE configuration. Its low input impedance makes it useful for matching low impedance sources with high impedance sources. Its high cutoff frequency characteristic (f_{hfb}) also makes it useful for some high frequency applications. There is no signal phase inversion when using this configuration.

In the following paragraphs, a few of the basic amplifier circuits will be discussed using a simplified approach.

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5.2.2.1.1 Single-stage amplifier. Figure 10 shows a typical single-stage amplifier using a 2N2222A npn transistor. With the resistance values shown, the bias conditions on the transistor are 1 mA collector current and 6 V from collector to emitter. At frequencies at which C₂ provides good bypassing, the input resistance is given by the formula

$$R_{in} = (1 + h_{fe}) h_{ib}$$

For the 2N2222A, at a design center of 1 mA, the input resistance would be 51 times 20, or about 1000 Ω .

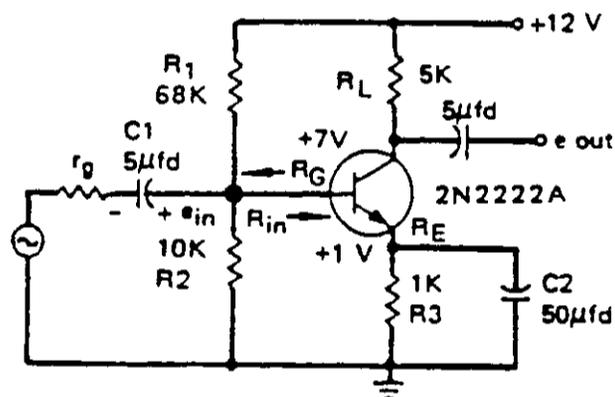


FIGURE 10. Single-stage amplifier.

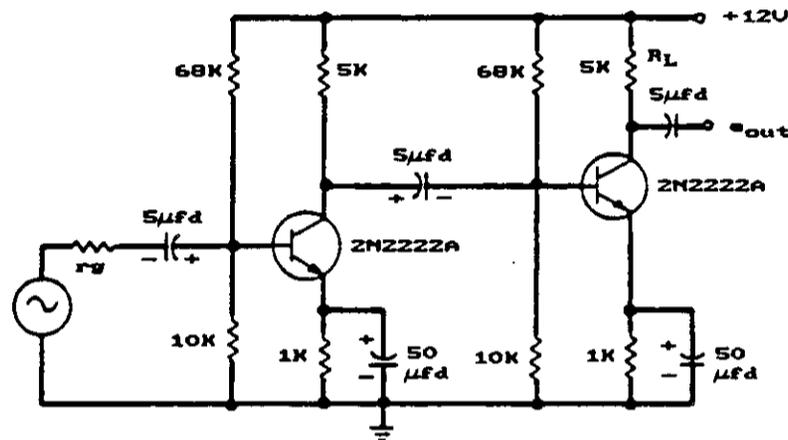
The ac voltage gain e_{out}/e_{in} is approximately equal to R_L/h_{ib} . For the circuit shown, this would be $5000/20$, or 250 (48 dB). The low frequency gain will drop 3 dB when the reactance of C₂ equals the parallel impedance of R₃ and R_E where

$$R_E \approx \frac{R_G + h_{ib}}{h_{fe} + 1}$$

and R_G is the parallel impedance of the bias network and generator r_g. Whereas the low frequency gain loss is mostly circuit dependent, the high frequency gain loss can be due to transistor characteristics.

5.2.2.1.2 Two-stage RC-coupled audio amplifier. The circuit of a two-stage RC-coupled amplifier is shown in Figure 11. The input impedance is the same as the single-stage amplifier and would be approximately 1000 Ω .

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FIGURE 11. Two-stage RC-coupled audio amplifier.

The load resistance for the first stage is now the input impedance of the second stage in parallel with 5 K Ω . The approximate voltage gain for the two-stage circuit is given by the formula:

$$A_v = h_{fe} \frac{R_L}{h_{ib}}$$

5.2.2.1.3 Class B push-pull output stages. In the majority of applications, the output power is specified; therefore, a design will usually begin at this point. The circuit of a typical push-pull Class B output stage is shown in Figure 12.

5.2 TRANSISTORS, LOW-POWER

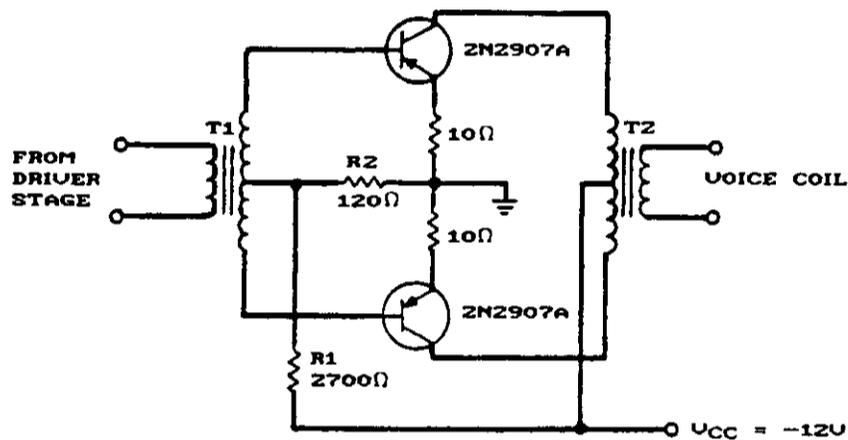


FIGURE 12. Class B push-pull output stage.

The voltage divider, R_1 and R_2 gives a slight forward bias (about 0.52 V) on the transistors to prevent crossover distortion. The 10- Ω resistors in the emitter leads stabilize the transistors to prevent thermal runaway. Typical collector characteristics with a load line are shown below in Figure 13.

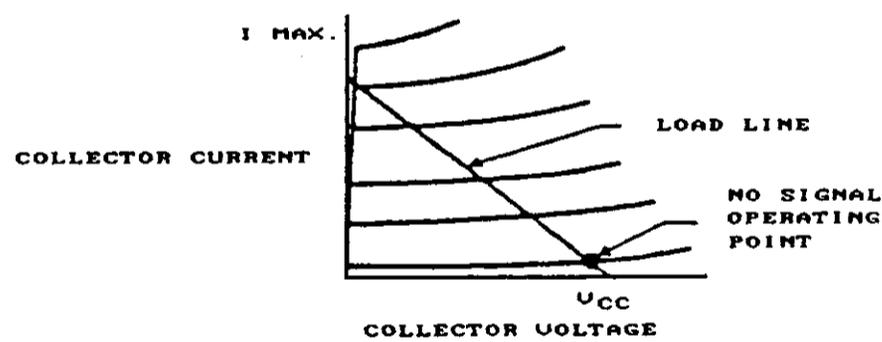


FIGURE 13. Typical collector characteristics and load line.

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The maximum ac output power, without clipping and using a push-pull stage, is given by:

$$P_o = \frac{I_{\max} V_{CE}}{2}$$

where V_{CE} = collector to emitter voltage with no input signal.

Moreover, the load resistance R_L is equal to:

$$R_L = \frac{V_{CE}}{I_{\max}}$$

And since the collector-to-collector impedance is four times the load resistance per collector, the output power (P_o) can be given by:

$$P_o = \frac{2V_{CE}^2}{R_{C-C}} \quad (1A)$$

Where R_{C-C} = collector-to-collector load resistance.

Thus, for a specified output power and collector voltage, the collector-to-collector load resistance can be determined. For an output power in the order of 50 mW to 850 mW, the load impedance is so low that it is essentially a short circuit compared with the output impedance of the transistors. Thus, unlike small signal amplifiers, no attempt is made to match the output impedance of transistors in power output stages. The power gain is given by:

$$\text{Power gain} = \frac{P_{\text{out}}}{P_{\text{in}}} = \frac{I_o^2 R_L}{I_{\text{in}}^2 R_{\text{in}}}$$

Since I_o/I_{in} is equal to the current gain (β) for small load resistance, the power gain formula can be written as follows:

$$\text{Power gain} = \frac{\beta^2 R_{C-C}}{R_{B-B}} \quad (1B)$$

where

R_{C-C} = collector-to-collector load resistance

R_{B-B} = base-to-base input resistance

β = grounded emitter current gain.

Since the load resistance is determined by the required maximum undistorted output power, the power gain can be written in terms of the maximum output power by combining equations (1A) and (1B) to give

$$\text{Power gain} = \frac{2\beta^2 V_{CE}^2}{R_{B-B} P_o}$$

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5.2.2.1.4 Class A output stages. The Class A output stage is biased as shown on the collector characteristics in Figure 14.

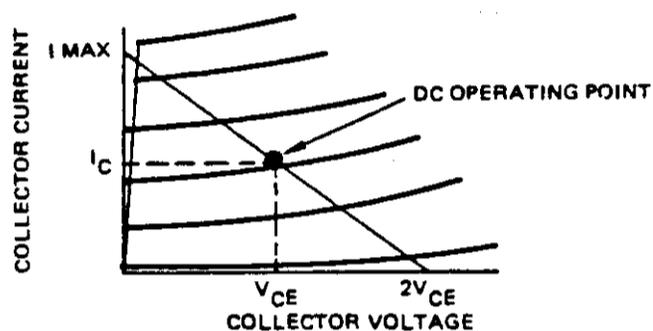


FIGURE 14. The dc operating point of class A audio amplifier.

The dc operating point is chosen so that the output signal can swing equally in the positive and negative direction. The maximum output power without clipping is equal to the following:

$$P_o = \frac{V_{CE} I_C}{2}$$

The load resistance is then given by:

$$R_L = \frac{V_{CE}}{I_C}$$

By combining these two equations, the load resistance can be expressed in terms of the collector voltage and power output is given by:

$$R_L = \frac{V_{CE}^2}{2P_o}$$

For output powers of 20 mW and above, the load resistance is very small compared with the transistor output impedance, and the current gain of the transistor is essentially the short circuit current gain β . Thus for a Class A output stage, the power gain is given by:

$$\text{Power gain} = \frac{\beta^2 R_L}{R_{in}} = \frac{\beta^2 V_{CE}^2}{2R_{in}P_o}$$

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Four other common approaches are used for analytical determination of the stability of the system. These approaches (from direct examination of the system differential equation solutions) are: Routh's criterion, Nyquist's criterion, Bode's attenuation and the phase shift method, or a combination of these. In nearly all oscillators, the amplifier that we have labeled A has a voltage gain and current gain greater than unity (and consequently, a power gain greater than unity), and the amplifier that we have labeled B has gain less than unity. An active device, such as a transistor, furnishes the gain, and the resistors, capacitors, and inductors provide the loss and phase shift to insure the proper polarity and amplitude of the feedback.

The following paragraphs discuss the various types of oscillators and their applications.

5.2.2.2.1 Phase shift oscillators. Figure 16A depicts a simple versatile transistor amplifier. Its current gain is stabilized against transistor variation. The means used to stabilize both operating point and small signal current gain also allow it to be used over a collector voltage range of 2 to 24 V.

When this simple amplifier is combined with the phase shift network in Figure 16B, oscillation will occur at a frequency where there is a 360-degree total phase shift; 180 degrees of this 360 degrees is furnished by the grounded emitter amplifier, and 180 degrees is furnished by the high pass network. Figure 16C connects both together and provides a 5 K Ω pot for frequency adjustment. This pot adjusts frequency from about 200 to 400 Hz. Both h_{ie} and h_{ob} enter as terms in the expression for frequency, but the unusually low impedances chosen provide excellent temperature and voltage stability. Output is derived across the collector and is approximately equal to VCC. Frequency of oscillation is as follows:

$$f \approx \frac{1}{2\pi \sqrt{6k^2 C^2 + 4 R R_L C^2}}$$

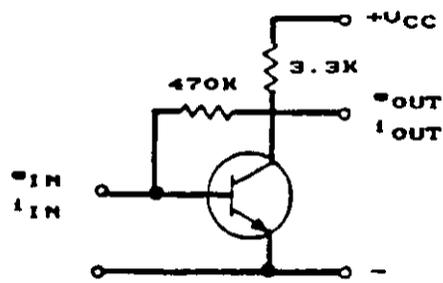
(This equation assumes $R > 10 h_{ie}$ and $1/h_{oe} > 10 R_L$)

The h_{fe} for sustained oscillation is as follows:

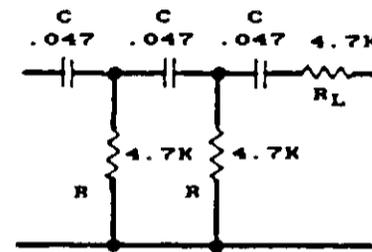
$$h_{fe} \approx 22 + \frac{30R}{R_L} + \frac{4R_L}{R_L}$$

5.2.2.2.2 Parallel-T oscillators. Figure 17 shows another RC phase shift oscillator using a parallel-T network. In this case, the simple amplifier is supplemented by an emitter follower to eliminate h_{ie} loading variations. Frequency stability of 0.2 percent is possible over the temperature range of -55 °C to +80 °C. In both phase shift oscillators, the effect of h_{oe} variations is swamped by the low (3.3 K Ω) collector load.

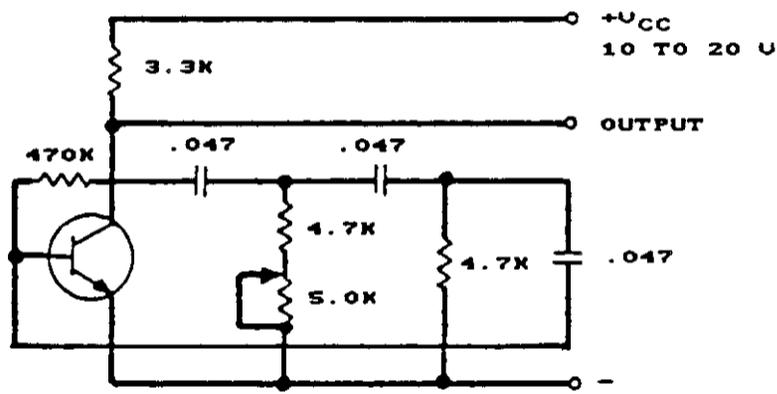
5.2 TRANSISTORS, LOW-POWER



A. Basic amplifier



B. Phase shift network



C. Phase shift oscillator with frequency adjust

FIGURE 16. Phase shift oscillator.

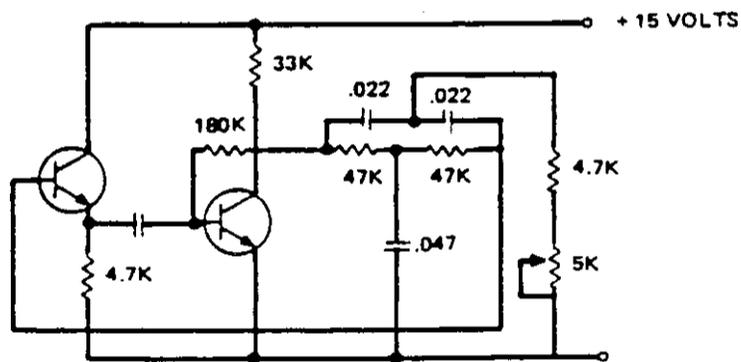


FIGURE 17. High stability parallel-T oscillator.

5.2 TRANSISTORS, LOW-POWER

5.2.2.2.3 Resonant feedback oscillators. These are among the most prevalent and useful oscillators. For their resonators, they use either an inductance capacitance combination or a crystal electromechanical resonator. They are characterized by circuit simplicity, good power efficiency, and good stability. Figure 18A demonstrates how ac coupling between input and output of a transistor amplifier is accomplished. This circuit is classed as a tuned collector oscillator. L_1 and C , in the collector of transistor Q , form a parallel resonant circuit.

Furthermore, the mutual coupling between L_1 and L_2 (called a tickler winding) provides an input signal current whose direction and magnitude are set by the physical arrangement of L_1 relative to L_2 , and by the direction and magnitude of the current through L_1 .

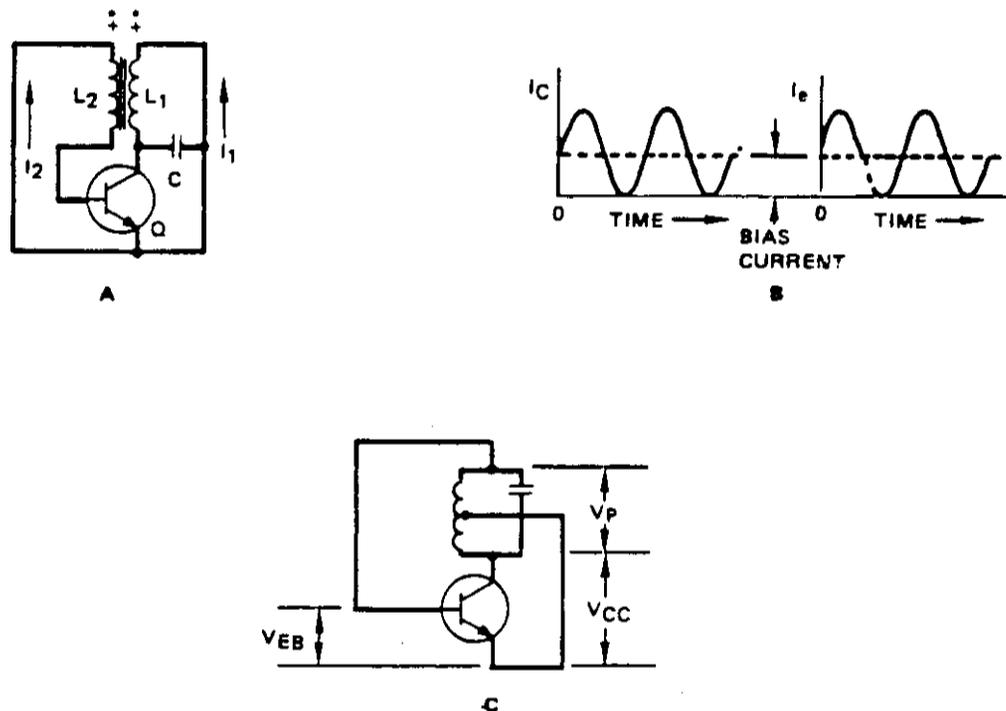


FIGURE 18. Resonant feedback oscillator.

The dots shown on the coils indicate winding start and phase coincidence so that an increase in collector current causes an increase in base current. This will provide the regenerative or positive feedback required to initiate oscillation. The core placed in the transformer is shown to indicate unity coupling between L_1 and L_2 for oscillator operation. The collector current and base current waveforms of Figure 18B show an offset which comes from a bias current which is not shown in this simple ac circuit. This offset current insures that

5.2 TRANSISTORS, LOW-POWER

the oscillator will start. Reversal of the increasing or decreasing base current is a limited cycle and comes about because the current gain of a transistor rolls off at both very high and very low collector currents. The exact points in I_C where reversal occurs are determined by loop gain and losses, and will always lie between transistor cutoff and saturation. The period of oscillation is very nearly set by the familiar relation between L_2 and C .

$$f_r = \frac{1}{\sqrt{2\pi\sqrt{L_2 C}}}$$

This assumes that the core coupling L_1 and L_2 is operated over a reasonably linear portion of its B-H characteristic. A more nearly exact expression, including the mutual inductance M , is

$$f_r = \frac{1}{2\pi\sqrt{C(L_2 + L_1 + 2M) - (L_2 L_1 - M)^2 h_{ob}/h_{ib}}}$$

In Figure 18C, an auto transformer can be substituted for the two winding transformers, and the emitter rather than the base may be allowed to float. This preserves the proper feedback polarity and is the basic circuit of the Hartley oscillator. Furthermore, it is a grounded-base oscillator and stability criteria are appropriately expressed in terms of grounded-base hybrid parameters. Analysis of this type of oscillator most frequently concerns itself with limit conditions. First, and fundamental, is the ability of the oscillator to start and continue oscillation. For this purpose small signal hybrid parameters may be used to establish the power gain of the circuit or the equivalent current or voltage gains. The Barkhausen criteria for loop unity gain forms the most convenient analytical approach. In terms of the mutual inductance M and the inductances $L_1 + L_2$, oscillation requires:

$$h_{fb} = \frac{L_2 + M}{L_1 + M}$$

In power applications, device ratings become important limit parameters to examine. The first point to consider is the power dissipation (P_d) of the device compared with the input power (P_i) needed in an attached load. For example, to furnish 1 W of output power (P_o) into a stipulated load, and by an additional winding or other means, provide a correct reflected load to the collector of the transistor, we can relate the P_d of the device to the desired P_o using the following equation:

$$\eta = \frac{P_o}{P_i} \times 100\% = \frac{P_o}{P_o + P_d} \times 100\% \quad (\text{neglecting circuit loss})$$

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where

R_L = required load V_{CC} = collector supply voltage
 P_o = output power
 P_i = input power
 P_D = power dissipated in the transistor
 η = efficiency as a percent

Because in most cases, oscillator efficiency will be greater than the theoretical Class A efficiency of 50 percent (relationship is in the biasing requirements needed to start and sustain oscillations); therefore, in our example, we can use 50 percent as the worst case efficiency for calculating the required P_D , $0.5 = 1/1 + P_D$ where $P_D = 1.5 - 0.5 = 1$ W.

The next concern is that of collector voltage swing. After selecting a supply voltage (V_{CC}), the maximum swing will be determined by the value of the load resistor (R_L).

where

$$R_L = \frac{V_{CC}^2}{2 P_o}$$

If $R_L = \infty$, then the worst case occurs, and the oscillator tank voltage swing is the largest. The peak voltage (V_p) appearing across the tank circuit is:

$$V_p = \frac{V_{CC}}{R_L} Q\omega L$$

$$Q = \frac{\omega L}{R_L} \text{ (assuming an unloaded } Q \text{ of greater than } 10)$$

then

$$V_p = \frac{V_{CC}\omega^2 L^2}{R_L^2} \text{ and the peak stress across the transistor, then becomes}$$

$$V_{CE} = V_{CC} \frac{(1 + \omega^2 L^2)}{R_L^2}$$

The peak stress across the emitter base junction is related to V_{CE} through the transformer turns ratio. Therefore:

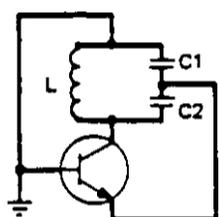
$$V_{EB} = \frac{V_{CC} \omega^2 L^2 N}{R_L^2} \text{ where } N = \text{turns ratio}$$

5.2 TRANSISTORS, LOW-POWER

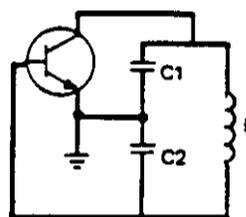
If the capacitor used to resonate the collector tuned circuit is split and used to form the feedback divider network, a Colpitts oscillator results. A simplified ac circuit for this configuration is shown in Figure 19A. Both C_1 and C_2 together set the effective capacity against which L resonates.

Figure 19B is identical to 19A except that the emitter rather than the base is shown at ground. One may relate nearly any LC resonant oscillator to the Hartley or Colpitts types, even though, at very high frequencies part of the circuit capacitive may be hidden as transistor capacitance.

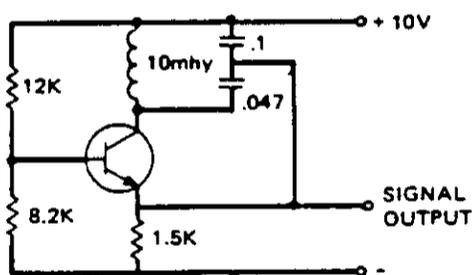
Figure 19C illustrates a practical 10 kHz Colpitts oscillator having a temperature drift rate of 0.035%/°C. This is the total drift rate and is determined by the temperature rate of incremental permeability of the coil core material.



A. Common-base Colpitts oscillator



B. Common Emitter Colpitts oscillator



C. 10 kHz Colpitts oscillator

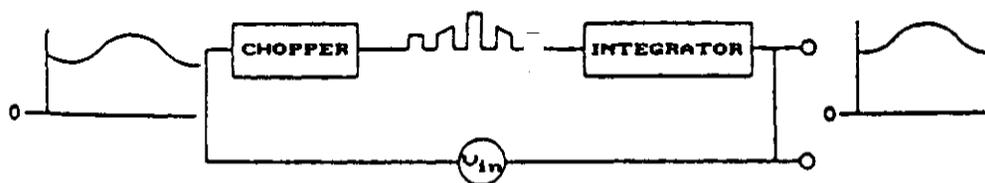
FIGURE 19. Colpitts oscillator.

5.2 TRANSISTORS, LOW-POWER

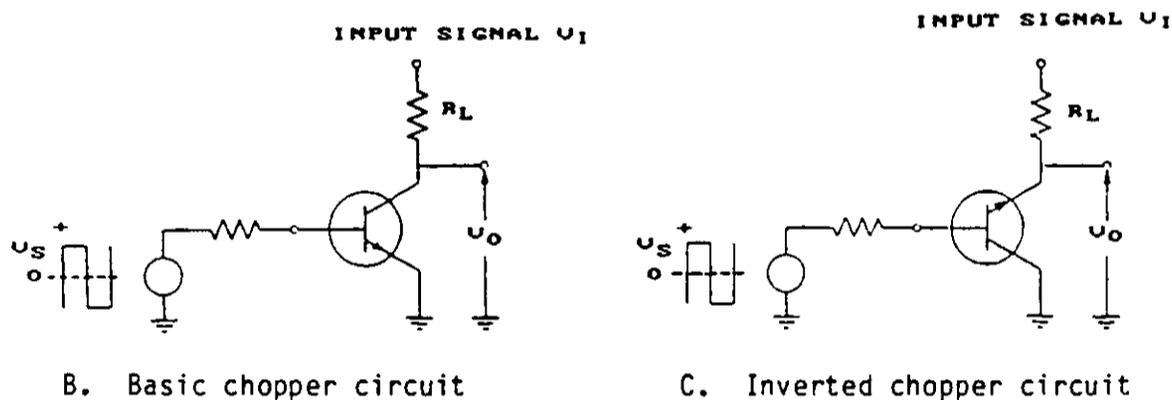
For the purpose of stability analysis, the Thevanin equivalent of the emitter resistor and the bias divider, together with transistor h_{jb} and the loaded voltage divider, are lumped to form the feedback loop.

If very high frequency stability is desired, the frequency determining network should be buffered from the amplifier. This provision cushions the frequency determining network from the inevitable changes induced from electrical environmental variation, but it also demands higher losses in coupling networks and therefore a higher amplifier gain to satisfy stability criteria. The alternative to this is a lower loss frequency determining network. The lower loss network is an alternative way of saying that a "high Q" is required where Q is defined as the energy stored per radian of angular period divided by the energy lost per radian of angular period. This Q definition is the usual figure of merit associated with resonant circuits but is equally applicable to many other networks having a transcendental solution.

5.2.2.3 Chopper. A chopper circuit can be used to convert a slowly varying dc signal to a fixed frequency ac signal, but with the amplitude variations that match the amplitude variations of the original signal. The basic circuit is shown in Figure 20.



A. Block diagram showing action of a chopper circuit



B. Basic chopper circuit

C. Inverted chopper circuit

FIGURE 20. Basic chopper circuit.

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In this process a very small dc signal, after being converted to ac, can be amplified by a RC-coupled amplifier which eliminates the problem of thermal generated current (I_{CBO}) becoming mixed with the signal current. However, at these very small signal levels, a problem known as voltage offset causes an error in the output of a standard type bipolar transistor. This offset voltage can be greatly reduced by the use of a bipolar chopper transistor that is especially designed so that, when operated in the inverse mode (as shown in Figure 20C), the offset voltage, $V_{EC(ofs)}$, is in the microvolt range. Moreover, other inverse parameters such as I_{ECS} , $h_{FE(INV)}$, and $r_{ec(on)}$ are also specified so as to aid in the circuit design.

5.2.2.4 Switching applications. A proper choice of the I_C/I_B (force gain) ratio can minimize switching time for a given transistor. Some circuit techniques are available that further improve switching speed. Two common techniques for this purpose, a collector-catcher circuit or the use of a speed-up capacitor in the base.

A simplified collector-catcher circuit is shown in Figure 21. Improvement in switching speed is obtained because the transistor is not allowed to go into saturation. Storage time, therefore, is drastically reduced. With no input pulse applied to the circuit, the transistor is initially biased off by $-V_{BB}$. A positive pulse of voltage turns the transistor on and the collector voltage begins to drop from $+V_{CC}$ toward $+V_R$. If V_R is greater than the sum of the voltage drop across CR and V_{CE} just out of saturation, then the collector is clamped at some value of voltage which maintains the transistor out of saturation.

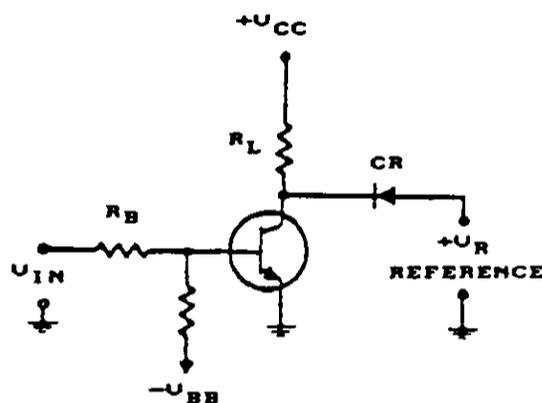


FIGURE 21. Collector-catcher circuit.

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The difficulty with this circuit is that the maximum I_C is essentially beta-dependent and because beta varies with temperature, this circuit is very unstable. Burnout of the diode or transistor is possible.

More practical circuits are shown in Figures 22 and 23. In these circuits, the voltage V_B and the diode keep the transistor out of saturation. However, the feedback arrangement tends to keep I_C constant by automatically varying the base drive. For example, if beta increases, I_C tends to rise and the collector voltage begins to decrease. But if the base drive is decreased, the circuit is returned close to its original operating condition. The only disadvantage of this circuit is the requirements of an isolated power supply V_B .

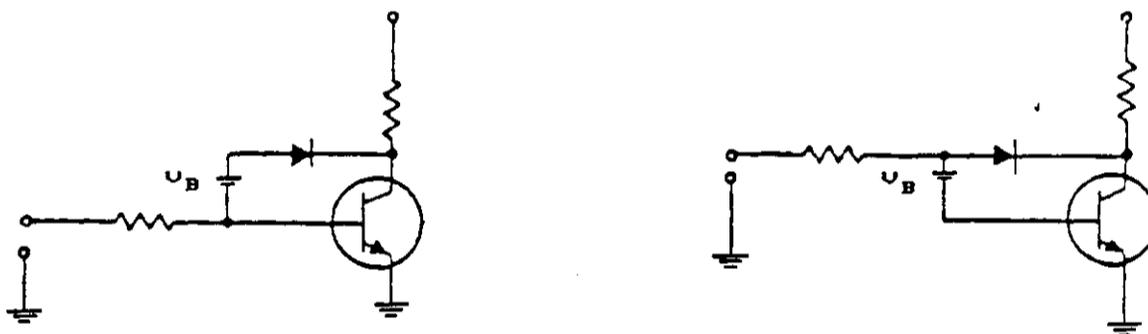


FIGURE 22. Collector-catcher circuits in which feedback is used.

A practical circuit (also known as a Baker clamp) is shown in Figure 23. Operation is similar to that described above. The drop across CR_2 acts as the V_B supply. The diode pair CR_1 constitutes the feedback arrangement. The function of CR_3 is to keep the transistor turned off under low V_{BE} voltage to reduce delay time. C acts as a speed-up capacitor. Use of the proper speed-up capacitor effectively increases turn-on drive and turn-off drive without supplying large amounts of on base current. As a result, faster rise times and faster fall times are achieved without the penalty of long storage time.

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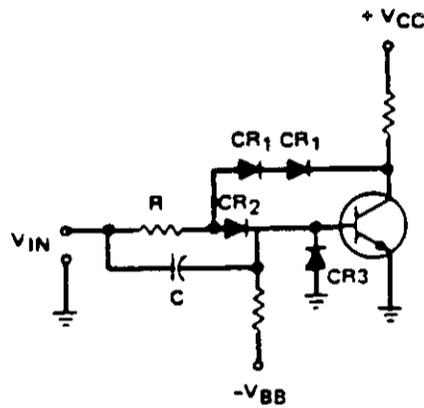


FIGURE 23. Practical collector-catcher circuit.

An example of a circuit that uses a speed-up capacitor is shown in Figure 23, and waveforms for this circuit are shown in Figure 25.

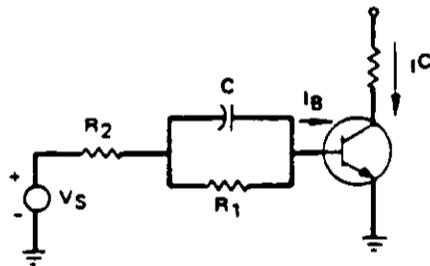


FIGURE 24. Circuit with speed-up capacitor.

The optimum value of C for fastest response can be found experimentally. If V_S is large compared with V_{BE} and R_2 can be neglected, the charge stored on the capacitor while the transistor is on is V_C . This charge should equal the stored base charge for best response. Practical values for R_2 , V_S , and V_{BE} will modify this relation.

5.2 TRANSISTORS, LOW-POWER

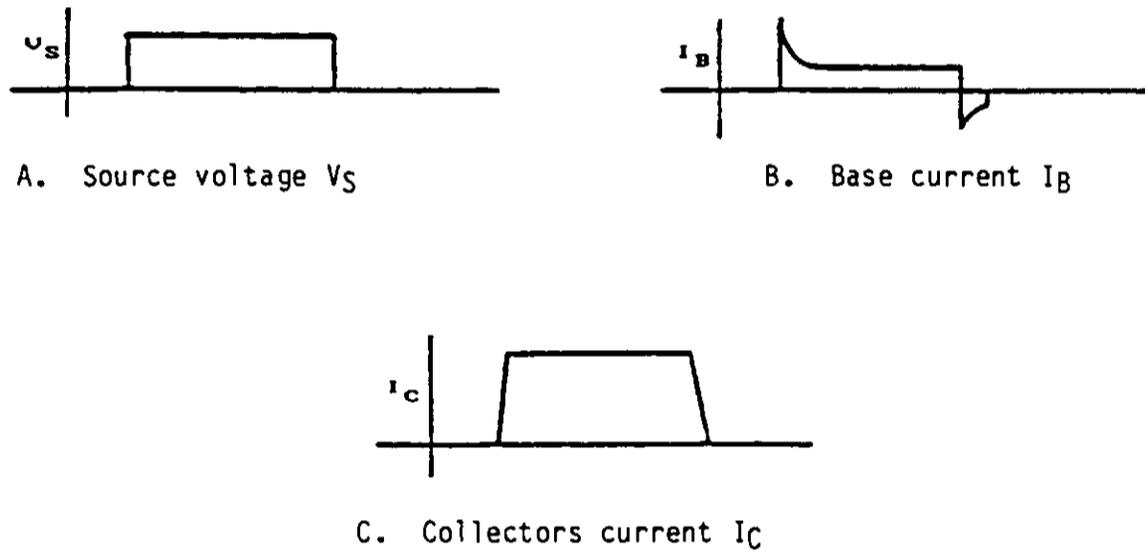


FIGURE 25. Waveforms for circuit shown in Figure 24.

5.2.3 Physical construction.

5.2.3.1 Die structures. The two basic structures of low-power and chopper transistors are the mesa (single and double) and planar types. These structures are shown in Figure 26.

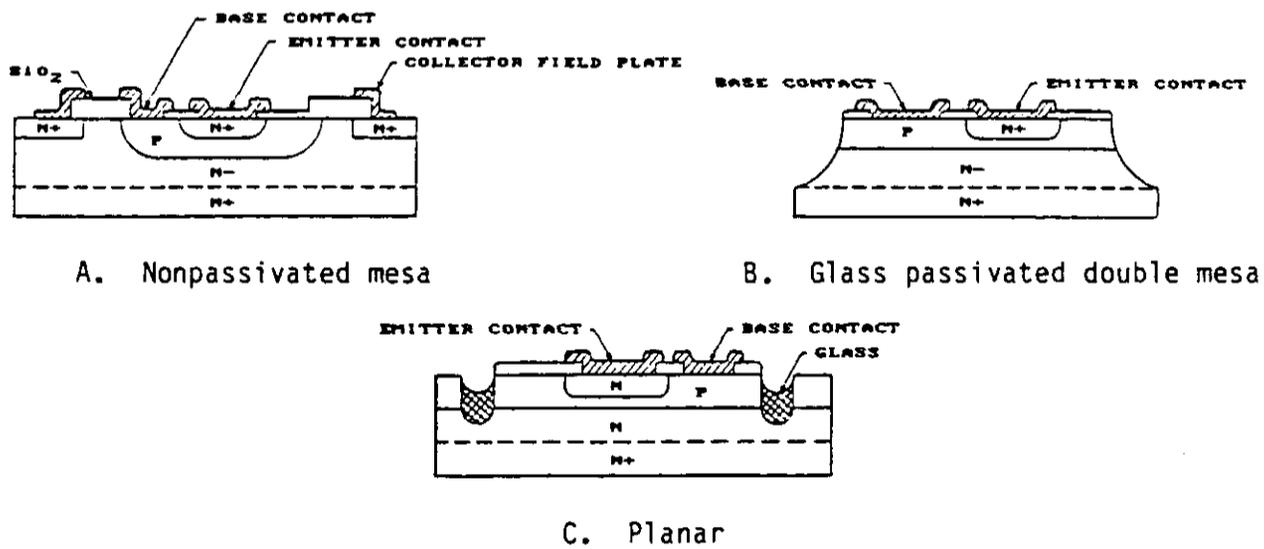


FIGURE 26. Basic die structures.

5.2 TRANSISTORS, LOW-POWER

Planar devices offer the lower leakage current because both junctions are always protected by silicon dioxide or some other passivant. In a mesa structure, the base-collector junction might not always be protected by a primary passivant. If this is the case, some type of conformal coating may be used during the assembly operation as a secondary passivant.

There are three basic die processes that are used to produce the mesa and planar structures. These are the double diffused (epi-collector), the triple diffused, and the epi-base process. By using a combination of these processes and die structures, several different transistor die-types can be produced (with the exception of an epi-base planar). It is, therefore, possible that three JANTXV2N5667 devices, each from three different manufacturers, could contain three different die types. These three devices, although they may have passed all JANTXV2N5667 electrical parameters and other requirements, could each possess different characteristics. It is these differences that could lead to a problem known as the unspecified parameter; for example, while one manufacturer's JANTXV device functions well in a circuit, the same type device from another manufacturer may not.

5.2.3.2 Package construction. Standard packages used for these devices are the T0-5, 18, 39, 46, 78, and 91. Figure 27 shows the basic construction for a T0-5.

Die bonds can be either soft solder or gold eutectic and both are reliable.

Wire bonding can be done with either gold or aluminum wire. The bonding techniques used can be thermocompression or ultrasonic weld to the die, and ultrasonic, thermocompression, or electrical weld to the leads. These bonds are reliable if proper process controls are implemented.

Sealing of the package is done in an inert atmosphere using electrical-weld techniques. It is possible that weld splashes might occur which form loose particles inside the package; therefore, MIL-STD-975 devices require that a PIND test and an X-ray be done to detect any defective devices.

5.2 TRANSISTORS, LOW-POWER

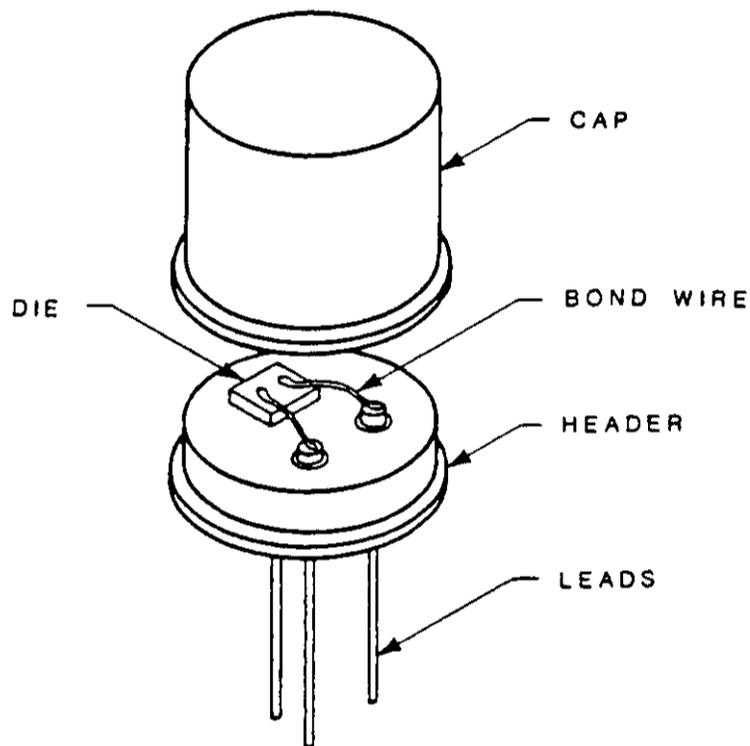
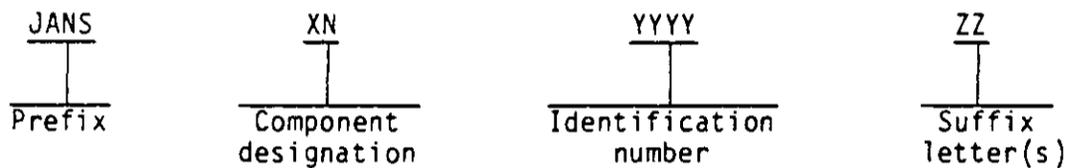


FIGURE 27. Typical TO-5 transistor construction.

5.2.4 Military designation. The military designation for transistors is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-15900 for details.

The component designation is 2N for transistors.

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The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices, suffix M, longer or shorter terminal leads (suffix L or S), or any other letter to indicate a modified version.

5.2.5 Electrical characteristics. The electrical parameters of interest may be separated into static and transient groups. This is somewhat arbitrary because the same parameters may be in both groups.

5.2.5.1 Static parameters.

5.2.5.1.1 Collector cutoff current, I_{CB0} . I_{CB0} is defined as the dc collector current when the collector-base junction is reverse biased and the emitter is open. Its value is determined by the voltage applied as shown in Figure 28 and the temperature at which it is measured as is indicated in Figure 29. I_{CB0} essentially varies exponentially with temperature and above the "knee" of the voltage curve, and it tends to follow an exponential variation with voltage.

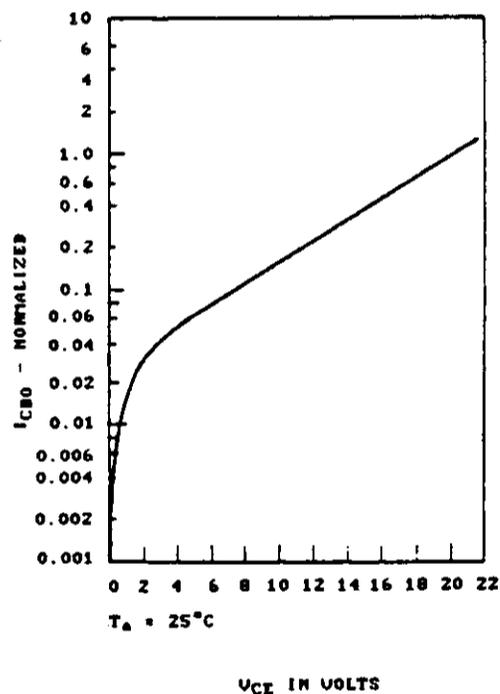


FIGURE 28. Behavior of I_{CB0} with voltage.

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To eliminate the need to take voltage variation into account each time a circuit is designed, the I_{CB0} is usually specified at a voltage near the maximum rating of the transistor. It is then assumed that I_{CB0} is constant for voltages lower than this value. The temperature which determines the I_{CB0} value of a device is the junction temperature, not the ambient. If the basic measuring circuit is studied, it is seen that the power dissipated in the transistor is the product of I_{CB0} and V . Since the I_{CB0} and the power are very small, the junction temperature is essentially that of the ambient temperature. Many manufacturers label the leakage current versus temperature curves with the ambient rather than junction temperature.

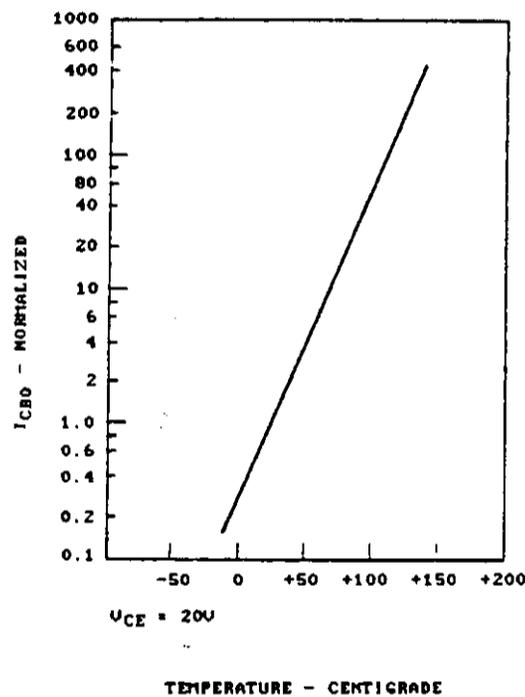
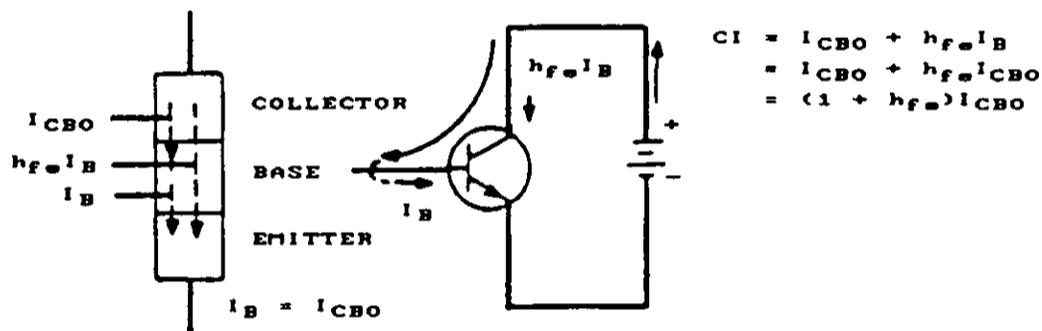


FIGURE 29. Behavior of I_{CB0} with temperature.

5.2.5.1.2 Collector cutoff current, I_{CE0} , I_{CER} . I_{CE0} is defined as the dc collector current when the collector-emitter junction is reverse biased and the base is open-circuited. I_{CE0} is important in the design of transistor circuits because it determines how close to a true off-state condition can be obtained. For example, the circuit in Figure 30 shows the base lead open rather than the emitter. Leakage current I_B (essentially I_{CB0}) flows across the reverse biased collector-base junction as before, but this current cannot return to the voltage source unless it flows across the base-emitter junction. Because polarities are such that this junction tends to be forward biased and this leakage current is essentially indistinguishable from a base-current supplied externally, the transistor will amplify the current to produce an additional current, $(h_{fe} \times I_{CB0})$ in the collector. The net result is that there is a total collector current (I_{CE0}) that is equal to $(1 + h_{fe}) \times I_{CB0}$.

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FIGURE 30. Effect of transistor gain on leakage current.

If a finite resistance is placed between the base and the emitter (R_{BE}) as shown in Figure 31, some of the I_{CBO} current will be shunted through this resistor. This shunted portion of the leakage current would not be amplified and therefore a collector current (I_{CER}) will flow. The I_{CER} would be much less than I_{CEO} , and as R_{BE} approached a zero value, I_{CER} would approach I_{CBO} . Conversely, if R_{BE} approaches an infinite value (usually 1 K Ω or more) I_{CER} would approach I_{CEO} .

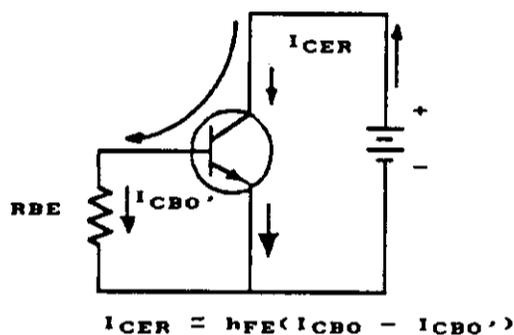
5.2.5.1.3 Emitter cutoff current (I_{EBO}). If the base-to-emitter of a transistor is reverse biased, there will be a leakage current (I_{EBO}) similar in every way to I_{CBO} except that it flows from emitter to base. Thus, to reverse bias a transistor, it is necessary to allow for I_{CBO} and I_{EBO} to flow out of the base lead. When I_{EBO} is not specified, it is usually assumed to be equal to I_{CBO} .

5.2.5.1.4 Current gain (h_{FE}). The direct current gain is of great interest in the design of transistor circuits and is defined as the following:

$$h_{FE} = \frac{I_C}{I_B}$$

where I_B and I_C are the absolute value of the base current and collector current respectively. The more commonly used parameter, h_{fe} , is essentially the ratio of a change in collector current for a small change in base current.

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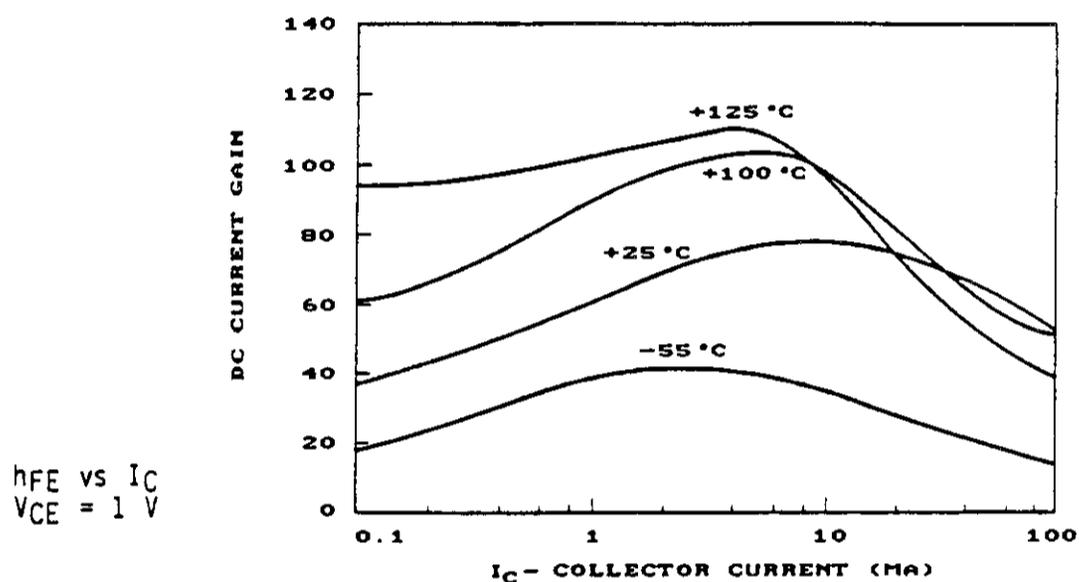
FIGURE 31. Effect of base-emitter resistor on leakage.

The value of h_{FE} is usually measured at a voltage between collector and emitter; this value is close to the saturation voltage as this represents a minimum value. Normally, h_{FE} is not a very strong function of collector-emitter voltage outside of saturation. It is, however, a rather strong function of junction temperature and of collector current. Figure 32 is a set of typical curve for h_{FE} as a function of I_C . Each curve is associated with a different temperature.

The most important feature in Figure 32 is that over most of the current range, the gain decreases as the temperature decreases. This rule cannot be applied indiscriminately because the reverse begins to be true beyond 10 mA and 100°C. A second feature is that the gain has a definite maximum which may be quite broad at room temperature or rather sharp at 125°C. The collector current at which this maximum occurs is a function of the junction temperature. It follows that when selecting an operating point, the temperature range over which the circuit will be expected to operate should be considered.

It sometimes happens that a decreasing gain with increasing temperature is desirable. Magnetic cores, for example, often require less drive at high temperatures than at low. Generally, however, this characteristic cannot be controlled sufficiently well to be useful.

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FIGURE 32. Variation of hFE with temperature and current.

5.2.5.1.5 Collector saturation voltage [$V_{CE(sat)}$]. The collector saturation voltage is the parameter that effectively limits how closely the transistor approximates a closed switch. Figure 33 and 34 shows how this parameter varies with temperature, current ratio and collector current.

Figure 33 shows that when the temperature is held constant while the circuit current ratio is increased (or I_B decreased), the saturation voltage changes linearly.

In Figure 34 the curves indicate that temperature is not a particularly strong influence and that the saturation voltage increases with increasing temperature; however, this depends very much on the device being used. In some devices the saturation voltage is almost completely independent of temperature, while in others the temperature coefficient can be negative overall or part of the temperature range.

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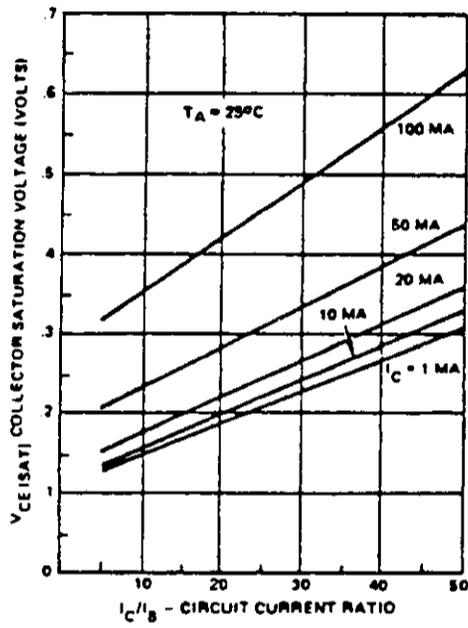


FIGURE 33. Variations in $V_{CE(sat)}$ with force gain and I_C .

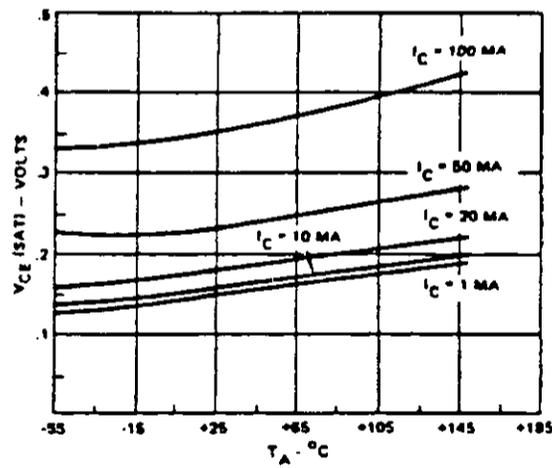


FIGURE 34. Variations in $V_{CE(sat)}$ with temperature and I_C .

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5.2.5.1.6 Base-emitter saturation voltage, $V_{BE(sat)}$. Two sets of $V_{BE(sat)}$ curves similar to the curves for collector saturation are shown in Figures 35 and 36. The characteristic feature is that the slopes of these curves are opposite to those shown in Figures 33 and 34. The temperature coefficient is negative and varies little over the entire range. Nominal values are about 2 mV change per degree centigrade.

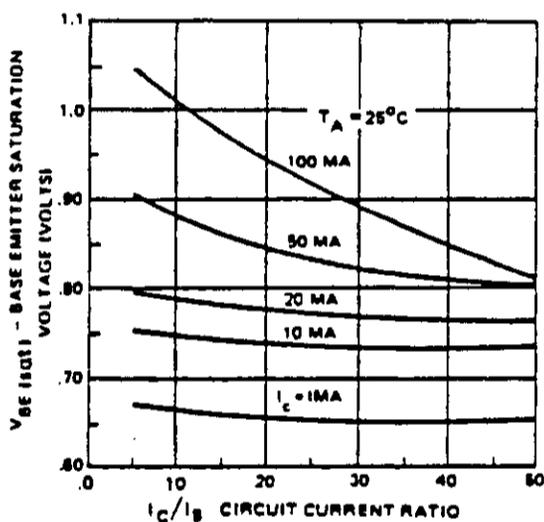


FIGURE 35. Variations in $V_{BE(sat)}$ with force gain and I_C .

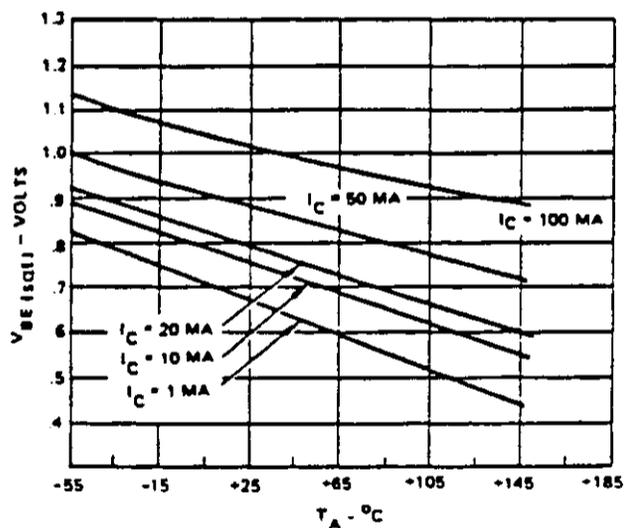


FIGURE 36. Variations in $V_{BE(sat)}$ with temperature and I_C .

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5.2.5.2 Transient parameters of switching transistors. The factors which influence the transient response of the transistor are basically associated with the diffusion time of the carriers across the base region, the effect of capacitances due to the collector-base and base-emitter junctions, the associated parasitic capacitances between leads and from case to leads, and the operating conditions of the circuit. In predicting transient response, it is convenient to think of the turn-on delay time (t_d), the current rise time (t_r), the storage or turn-off delay time (t_s), and current fall time (t_f) as dependent variables whose values depends upon the operating conditions of the device as well as the capacitances and diffusion parameters. It follows that the calculations of the time intervals (t_d , t_r , t_s , and t_f) can be very complex.

Before considering the time intervals of a transistor and the effects of the collector-base and base-emitter capacitances and resistances on these intervals, the presence of these reactive components within the transistor is discussed.

There are two types of capacitances associated with any semiconductor junction: transition capacitance (C_T) and diffusion capacitance (C_D).

C_T is due to the high electric field in the depletion region caused by the voltage across the barrier. Hence C_T is voltage dependent. C_D is due to the current flowing through the depletion region. Hence C_D is current-dependent and therefore, the total junction capacitance is the sum of C_T and C_D .

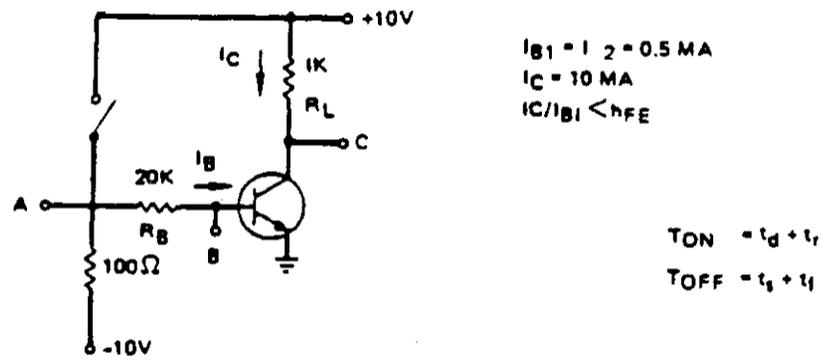
The collector capacitance (C_C) is made up primarily of the transition capacitance, (C_{TC} or C_{obo}), because the diffusion capacitance is small in a reverse biased junction. On the other hand, the emitter-base capacitance can be either diffusion capacitance (C_{DE}) or transition capacitance (C_{ibo}) if reversed biased.

The active portion of the base region of a transistor is not equipotential but exhibits an ohmic resistance to the flow of base current. This parasitic resistance is called the base-spreading resistance (r_b).

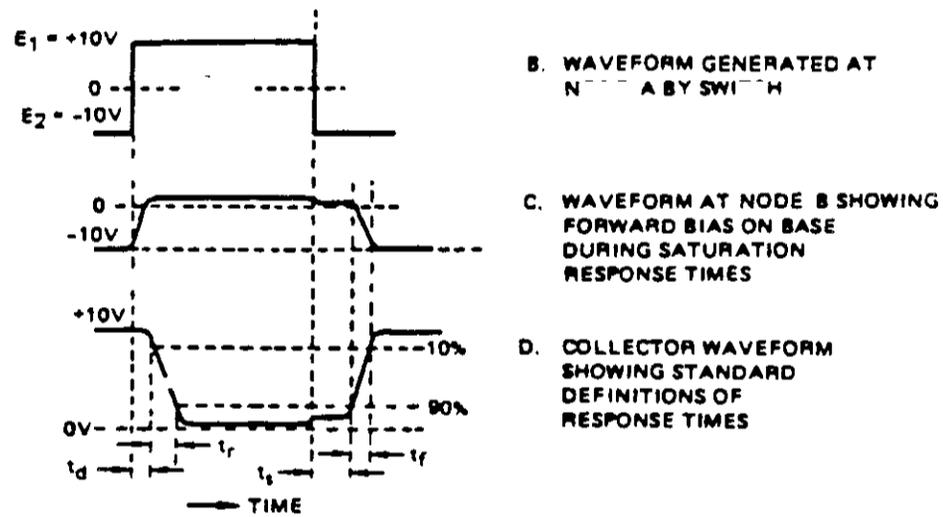
This base resistance is not purely resistive but takes on a distributed form (transmission line) in many transistor structures. However, to avoid complications, it will be assumed that r_b is resistive. As will be seen later, r_b is a very objectionable parameter because it contributes to the deterioration of transistor performance in many ways.

The time intervals (t_d , t_r , t_s , and t_f) mentioned above are defined in Figure 37. These definitions are commonly accepted for measurements and require no further explanation beyond that the 10 and 90 percent points of the collector waveform are taken as the points at which measurements are to be made. The collector waveform is the voltage from collector to emitter. For most calculations, we shall use collector current rather than collector voltage as the reference thereby avoiding some difficulty with what is meant by rise-time (t_r). In Figure 37D, the voltage is falling during t_r , because the current is increasing (or rising) during this interval.

5.2 TRANSISTORS, LOW-POWER



A. Typical circuit



B. Waveforms

FIGURE 37. Transient response.

5.2 TRANSISTORS, LOW-POWER

5.2.5.2.1 Charge-control theory. To better comprehend switching parameters, it is necessary to have an understanding of the transistor as a charge-controlled device. A thorough understanding will also permit accurate comparisons among the switching abilities of different transistors as well as providing data to calculate switching speeds in a variety of circuits. The charge required to perform various switching functions can be readily deduced by examining the common hybrid-equivalent circuit of Figure 38, which is a simple but fairly accurate representation of a transistor.

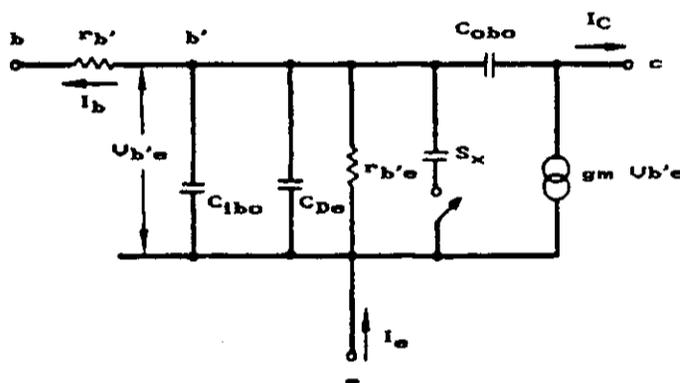


FIGURE 38. Hybrid- π transistor equivalent circuit.

The capacitances C_{1b0} and C_{0b0} are the transition capacitances of the emitter-base and collector-base junctions respectively plus stray capacitance associated with the transistor package. Transition capacitance is somewhat analogous to a parallel plate capacitor in that charges of opposite polarity are separated by a narrow region depleted of charge. The emitter and collector transition capacitances are always present, and their capacitances increase somewhat as the reverse junction voltage decreases.

The diffusion capacitance C_{De} represents the charge involved in the flow of carriers across the base region, this is somewhat analogous to water flow in a pipe. The carrier flow is by diffusion; hence the name "diffusion capacitance." Because a transistor has a finite base width, the flow of carriers through it constitutes a base charge. The charge in the base will, therefore, be proportional to the emitter current, (I_e). Because a barely perceptible change in base-emitter voltage can change the emitter current of a transistor over wide ranges, a capacitor can be used to represent the base charge if it has a value proportional to emitter current. Therefore, the capacitance C_{De} has a value proportional to emitter current; C_{De} also depends somewhat upon collector voltage since the base width narrows as collector voltage increases.

5.2 TRANSISTORS, LOW-POWER

The remaining capacitance, S_x , and the switch represents the storage phenomenon. The switch closes and connects S_x to the circuit only when the transistor is saturated. The behavior of the store charge is nonlinear, being dependent upon values of the currents I_{b1} and I_c and upon temperature.

5.2.5.2.2 Delay time, t_d . The switching process, as the transistor is turned on from an off condition and then off from an on position, can be explained in terms of the capacitance. When the transistor is off, a reverse-bias voltage is present on both junctions, and only the capacitances C_{ibo} and C_{obo} are effective. The voltage on the base must be changed to forward-bias the base-emitter junction so that the transistor will turn on. The time required for the input base current to charge C_{ibo} and C_{obo} to a voltage level that forces the emitter to inject current is the delay time, t_d . The total change in charge on C_{ibo} and C_{obo} during the delay time interval is termed the off-bias charge, Q_{ob} .

5.2.5.2.3 Rise time, T_r . The rise-time interval commences as the emitter begins to inject current, causing the diffusion capacitance to enter into the picture. The base voltage must now change from its value at the threshold of injection to its final value. This voltage change requires change in charge upon C_{ibo} and C_{De} . In addition, the voltage across C_{obo} changes as the flow of collector current causes the collector voltage to drop from the cutoff level to the saturation voltage level. Thus, a charge is required by C_{obo} .

The time taken for the input current to supply the charge required by C_{ibo} , C_{De} , and C_{obo} is the rise-time interval. The total charge required by these capacitances to turn on a transistor is termed the active region charge, Q_A .

5.2.5.2.4 Storage time, t_s . To keep the transistor in the saturation region, a base current (I_{b1}) larger than that required to just maintain I_c is supplied. This condition results in an excess charge Q_x to be stored in the transistor. The storage excess charge is represented on the transistor equivalent circuit of Figure 38 by the inclusion of S_x and the switch.

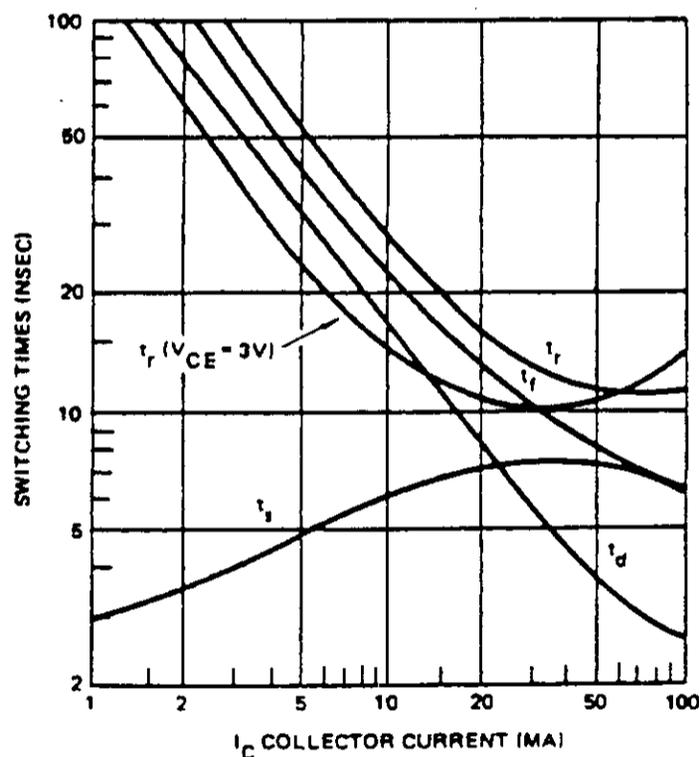
Before the transistor can be turned off, the excess charge stored on S_x must be removed. It can be removed either through an external path or by internal recombination. Since the internal recombination time is long, faster switching is achieved by supplying a reverse-bias voltage to the transistor input circuit. The reverse bias permits a reverse current (I_{b2}) to flow and considerably shortens the time to deplete S_x of charge. The time for the excess charge to leave S_x is the storage time (t_s).

5.2.5.2.5 Fall time, t_f . At the conclusion of storage time, the fall-time interval commences. The charge which must be removed is the same as the rise time; that is, C_{ibo} , C_{De} , and C_{obo} must be charged from the voltages of the on condition to the voltages at the threshold of conduction. Often, however, some excess carriers stored in remote regions of the transistor did not have time to leave during the storage time interval and exit during fall time, producing a "tail" on the fall-time waveform.

5.2 TRANSISTORS, LOW-POWER

The total charge which appears in the base circuit during turn-off is called the total control charge (Q_T). It represents the charge stored on all transistor capacitances and the store during the period of time the transistor was held on. It is normally the largest amount of charge involved in the switching process and, therefore, must be known as a function of all circuit variables (i.e., I_C , I_B , V_{CC} , and T_J) in order to design a circuit for any set of conditions.

5.2.5.2.6 Collector current effects on transient response. As illustrated in Figure 39, the diffusion capacitance charge becomes appreciable for this transistor at currents above 10 to 20 mA, depending upon collector voltage. At low currents, the rise time when $V_{CC} = 3$ V is faster than the rise time when $V_{CC} = 10$ V because the charge required by C_{ob} is less. However, at high currents, rise time at 10 V is faster than rise time at 3 V. This behavior occurs because at high currents, rise time is determined primarily by the diffusion capacitance, whose average value is smaller when switching from high collector voltages.



$B_F = 10$
 $V_{CE} = 10$ V

FIGURE 39. Typical switching times.

5.2 TRANSISTORS, LOW-POWER

Fall time is less than rise time at all current levels because of the effect of recombination. Simply stated, the recombination current is subtracted from the turn-on current I_{b1} and added to the turn-off current I_{b2} . Therefore, the current available to charge Q_A during rise time is less than the current available to discharge Q_A during fall time.

Storage time behavior is very complex and is the most difficult transistor parameter to predict. Because of the space limitation in this manual, storage time will not be discussed. The t_s curve of Figure 39 is merely intended to serve as a guide.

5.2.5.3 Transistor frequency limitations.

5.2.5.3.1 Gain-bandwidth product (f_t). When operated at low frequencies in the common emitter configuration, the transistor exhibits a short-circuit gain ($R_L \ll r_{OUT}$) of h_{fe0} . This value may vary in small-signal transistors from a low of twenty to a high of several hundred. As the signal frequency is increased, the magnitude of h_{fe} decreases and its phase shift increases. This is because carriers when they are injected into the base-emitter junction will take a certain time to cross into the collector region. Figure 40 shows that there are three time constants limiting the speed of the injected carriers: the emitter time-constant $r_e C_e$, the collector time-constant $r_e C_c$ (the collector is shorted to the emitter), and the base transit-time τ_B .

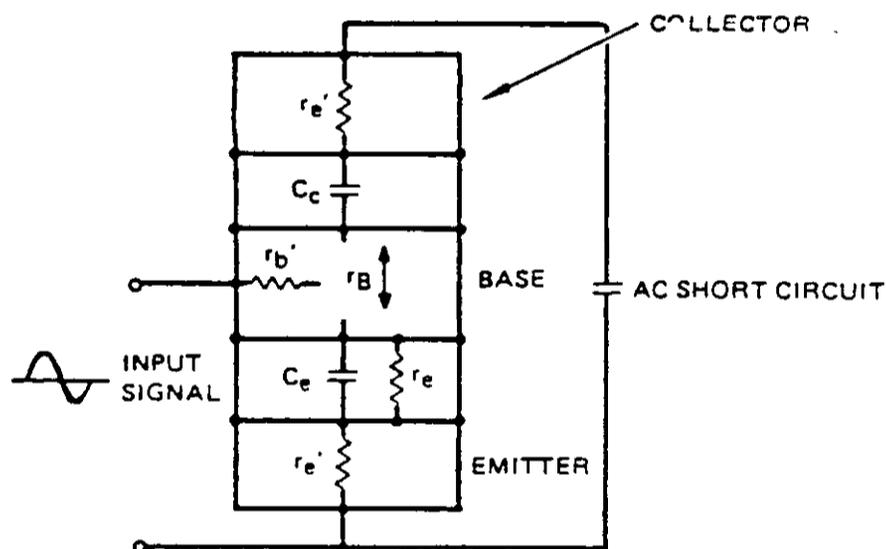


FIGURE 40. Various time-constants limiting the gain-bandwidth product of the transistor.

5.2 TRANSISTORS, LOW-POWER

The sum of these three constants is defined as the gain-band-width product f_t , because it is the frequency at which h_{fe} falls to unity; it is given in the following equation:

$$f_t = \frac{1}{2\pi[\tau_B + r_e (C_e + C_c)]}$$

Even if C_e ($C_{b'e}$) and C_c (C_{TC}) were made very small, the base transit-time would still limit the frequency response of the transistor. Therefore, high gain-bandwidth (f_t) transistors must be designed with extremely thin base regions or an accelerating field added into the base region. A well designed high-frequency transistor might have

$$r_e C_{b'e} \text{ of } 25 \times 1 \times 10^{-12} = 0.025 \text{ ns}$$

$$r_e C_{TC} \text{ of } 25 \times 0.4 \times 10^{-12} = 0.010 \text{ ns}$$

$$\tau_B \text{ of } 125 \times 10^{-12} = 0.125 \text{ ns}$$

Total time for carriers to reach collector = 0.160 ns and the equation is as follows:

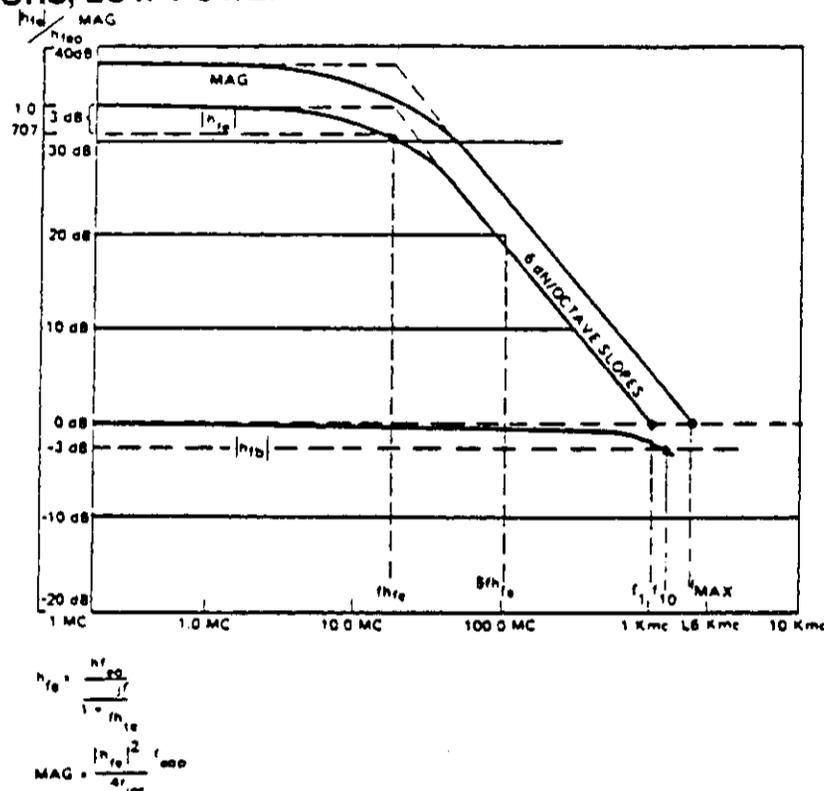
$$\text{Hence } f_t = \frac{1}{2\pi TC_{\text{total}}} = \frac{1}{628 \times 160 \times 10^{-12}} \approx 1.0 \text{ GHz}$$

A large collector bulk resistance will add a fourth time-constant of $r_c' C_{TC}$, which, if $r_c' = 100 \Omega$ and $C_{TC} = 0.5 \text{ pF}$ will give the carriers another delay of $100 \times 0.5 \times 10^{-12} = 0.05 \text{ ns}$, reducing f_t to 760 MHz. Therefore, a good high frequency transistor should also exhibit low extrinsic collector series resistance (r_c'). In epitaxial transistors, r_c' is small, and this fourth time-constant can be made negligibly small.

Figure 41 plots h_{fe} , h_{fb} , and maximum available power gain (MAG) versus frequency for a typical UHF transistor.

The first deduction that can be made is that there is an exact relationship between f_t and h_{fe0} ($h_{fe0} \times f_{h_{fe}} = f_t$). Second, at a frequency five times this beta cutoff frequency ($f_{h_{fe}}$) a dB/octave slope has been reached. Along this -6 dB/octave slope, the product of h_{fe} and its corresponding frequency is a constant, which is defined as the gain-bandwidth product (f_t).

5.2 TRANSISTORS, LOW-POWER

FIGURE 41. MAG, $|h_{fe}|$, $|h_{fb}|$ vs frequency of typical UHF transistor.

5.2.5.3.2 Alpha and beta cutoff frequencies. As previously stated, the beta-cutoff frequency ($f_{h_{fe}}$) can be used as an aid to locate the proximity of the 6 dB/octave slope. Actually, modern transistor circuit analysis has done away with the formerly much used "alpha cutoff frequency" ($f_{h_{fb}}$) and "beta cutoff frequency" ($f_{h_{fe}}$). A small-signal transistor is primarily frequency-limited by its emitter, base, and collector time-constants, and may not be usable at $f_{h_{fb}}$ (due to feedback, there may not even be an $f_{h_{fb}}$). Therefore, these two frequencies will simply be defined below.

- Alpha cutoff frequency is the frequency at which the common base current gain α , falls to 0.707 of its low frequency value. In modern transistors $f_{h_{fb}}$ is usually somewhat above f_t .
- Beta cutoff frequency is the frequency at which the common emitter current gain β , more recently identified as h_{fe} , falls to 0.707 of its low frequency value (h_{fe0}).

5.2.5.3.3 Maximum frequency of oscillation. Maximum frequency of oscillation (f_{max}) is the frequency at which the maximum available unilateralized power gain (MAG) falls to unity. Considering the equation for MAG in Figure 41, it can be seen that as $|h_{fe}|$ drops to unity there still is a power gain given by $r_{oep}/4r_{ies}$ (impedance ratio to output to input impedance). Hence f_{max} will be generally higher than f_t . The f_{max} is sometimes also referred to as the (power gain)^{1/2} (bandwidth) product; stated as $\sqrt{PG} \times BW$ which gives a 6 dB PG/octave slope, but should not be confused with f_t .

5.2 TRANSISTORS, LOW-POWER

5.2.6 Environmental considerations. Refer to paragraph 5.1.6.3 Environmental considerations.

5.2.7 Reliability considerations.

5.2.7.1 Failure mechanisms. The failure mechanisms for low power transistors are discussed in detail with the stresses that cause these failures, under paragraph 5.1.6 Reliability considerations.

5.2.7.2 Derating. Refer to MIL-STD-975 for device derating guidelines.

5.3 TRANSISTORS, HIGH POWER

5.3 High-power.

5.3.1 Introduction. High-power transistors are used for controlling large amounts of power in various circuits by means of relatively low-power control signals. This category covers transistors that have a power dissipation of 2 W or more.

For high-power transistors, the design engineer will make frequent use of the common-emitter collector family of curves illustrated in Figure 42. This plot of the collector current versus collector-emitter voltage at various constant base current levels is divided into three regions; the limitations concerning these regions are also shown. The following describes the operation regions of all npn transistors (the polarities are reversed for a pnp transistor).

- The active region is the region where V_{BE} is positive but is less than V_{CE} . Moreover, the collector current is a function of base current and consists of two areas: that of continuous operation and that of pulsed operation. These two operating areas are limited only by the maximum current and maximum voltage ratings of the device. In the pulsed operation area, pulse duration is also a limitation.
- The saturation region is characterized by a positive V_{BE} greater than V_{CE} and is limited by the maximum base and collector currents only.
- The cut off region, in which V_{BE} is negative, is limited only by the manufacturer's maximum voltage rating.

It is possible for a transistor to operate in any or all of these regions, depending on its mode of operation.

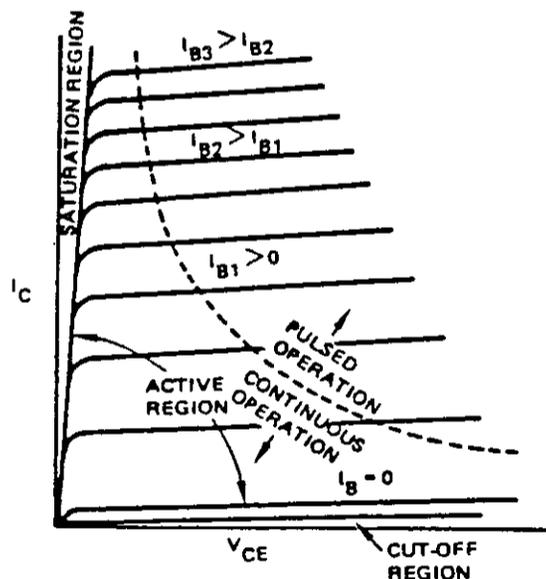


FIGURE 42. Collector family of curves for common emitter connection.

5.3 TRANSISTORS, HIGH POWER

5.3.1.1 Operating modes. High-power transistors are devices for controlling large amounts of power in various circuits by means of relatively low-power signals. The operating mode of the transistor will depend upon the type of control desired. For example, the device may be used as a linear amplifier where the output signal should be a faithful reproduction of a lower power input signal. On the other hand, the device may be used as a highly efficient switch that simply opens and closes a circuit. The various modes of operation will be considered separately and related to the collector family of curves.

5.3.1.1.1 Class A. The Class A mode of operation is one in which collector current flows at all times and the load line lies within the active region. For example, the audio amplifier of Figure 43 operates in the Class A mode. The transistor is provided with a source of dc bias, I_{BQ} , by voltage V_{BB} and resistor R_B . The quiescent operating point Q is determined by the intersection of the dc load line (resistance of the transformer primary winding) and the collector curve corresponding to the quiescent base bias I_{BQ} . If an ac signal of base current is superimposed upon the quiescent bias current, such that I_B varies between the limits of I_{Bmax} and I_{Bmin} , the instantaneous operating point will traverse the ac load line as shown. Although for simplicity a resistive load is shown, in practice reactive loads are frequently encountered in which the load line is an ellipse rather than a straight line.

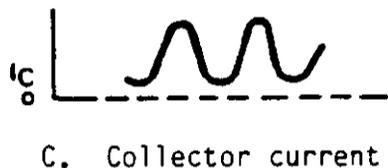
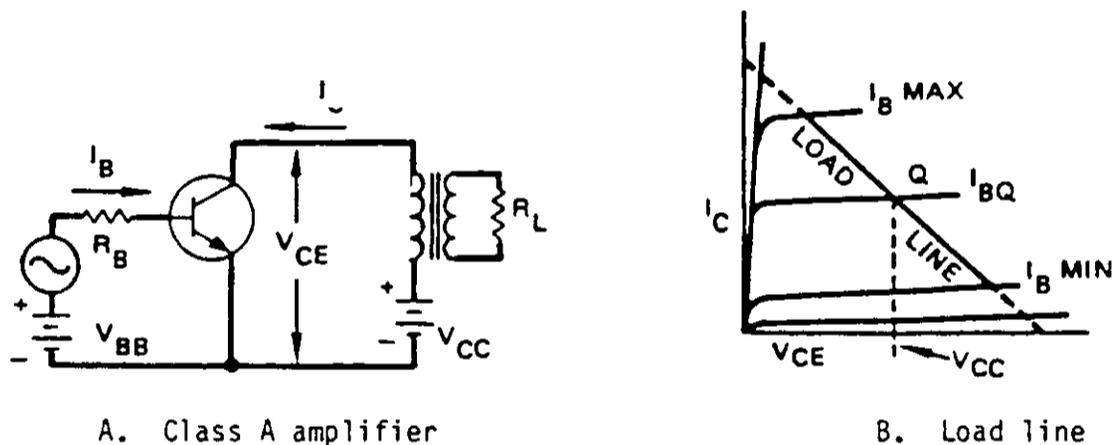


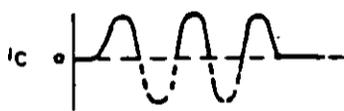
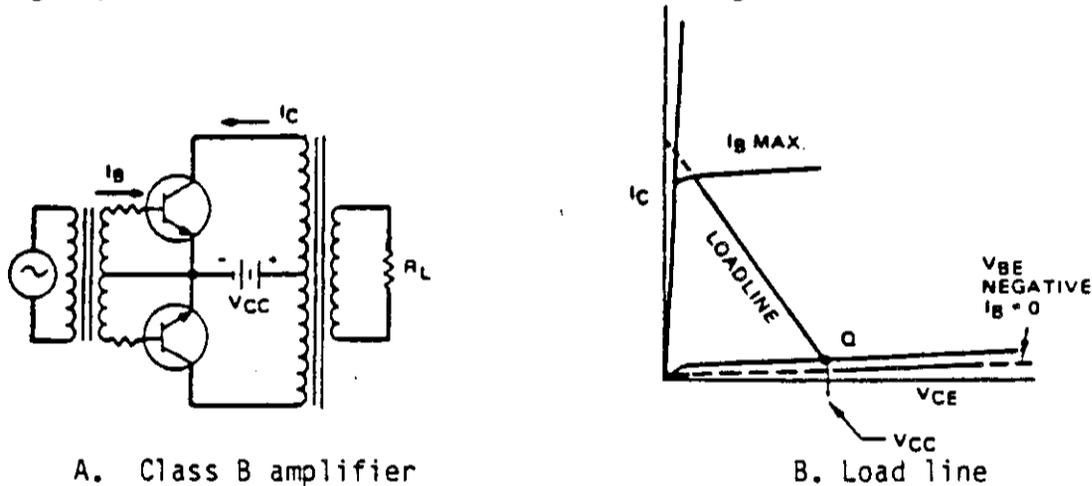
FIGURE 43. Example of class A mode.

5.3 TRANSISTORS, HIGH POWER

The Class A mode is not limited to audio amplifiers but includes all applications in which the collector current flows continuously.

5.3.1.1.2 Class B. The Class B mode is one in which the quiescent base bias is approximately equal to zero, and the corresponding collector current is also equal to zero. When the ac signal is applied, collector current flows for approximately one half of each cycle during the period when V_{BE} is positive.

During the alternate half-cycles V_{BE} is negative, and the collector current is cut off. One common application of the Class B mode is in push-pull audio amplifiers as shown in Figure 44. Note that part of the load line is in the active region, and the remainder is in the cutoff region.

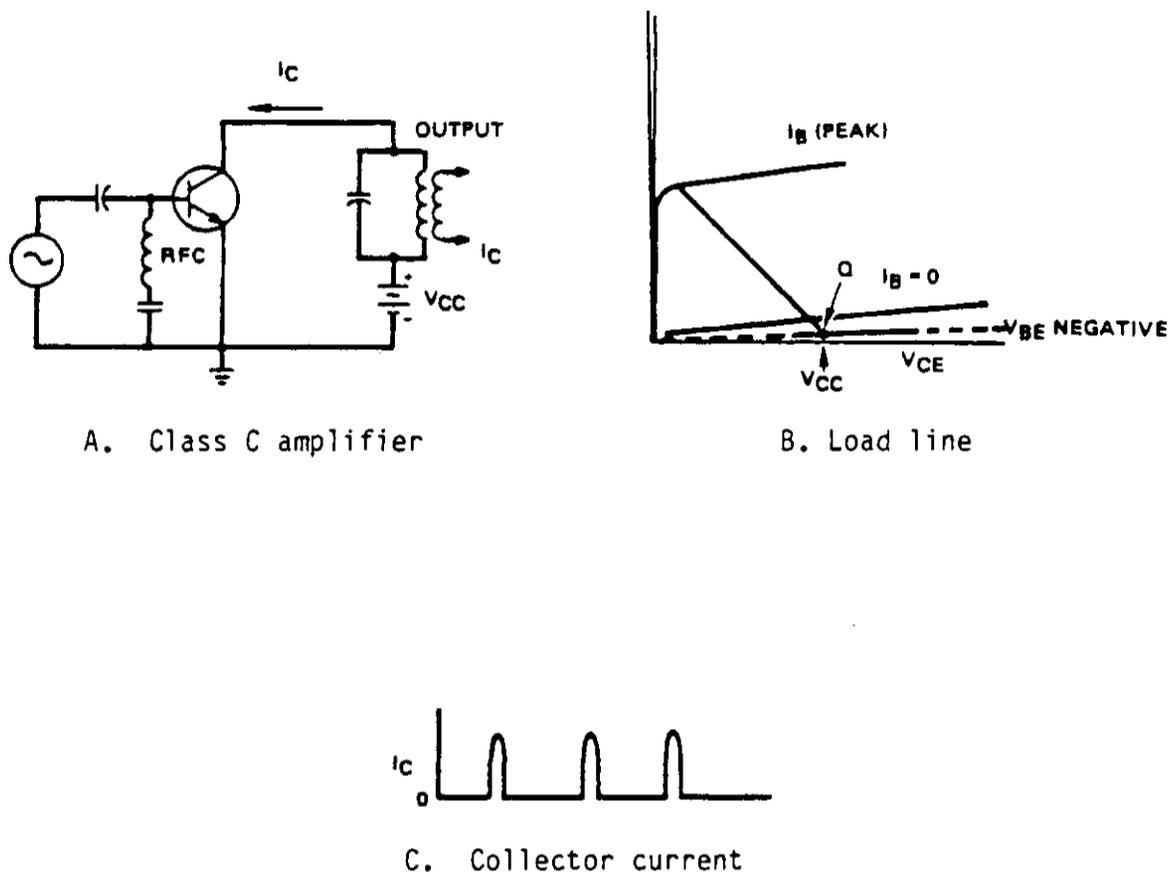


C. Collector current

FIGURE 44. Example of Class B mode.

5.3.1.1.3 Class C. The Class C mode is one in which the quiescent base-emitter bias is negative, and the corresponding collector current is completely cut off. When an ac base drive signal is applied, collector current flows for considerably less than one half of each cycle. A common application for the Class C mode is in rf power amplifiers as shown in Figure 45.

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FIGURE 45. Example of Class C mode.

5.3.1.1.4 Class D. The Class D mode is the term used to designate the operation of the transistor as a switch. For this type of operation it is necessary to consider three states: (1) The off state when the transistor is not conducting, (2) the on state when it is conducting, and (3) the "transition" state when it is changing from off to on or vice versa. In the off state, the base-emitter is reverse-biased (V_{BE} negative) so that the operating point is in the cut-off region. In the on state, the base current is raised to a level somewhat in excess of the value needed to maintain the collector current at the on value. This condition is sometimes described as overdriving the base. The operating point is in the saturation region. The value of V_{BE} under this condition is greater than that of V_{CE} ; in other words, both the emitter-base junction and the collector-base junction are forward-biased and both the off and on states are low-power dissipation conditions for the transistor because in the former case the current is very low, and in the latter the voltage is low.

5.3 TRANSISTORS, HIGH POWER

In the Class D mode, the transition from off to on or vice versa is made very quickly by applying a positive or negative step of drive to the base. Therefore, the operating point does not remain in the transition state, which is in the active region, longer than a few microseconds. A simple switching circuit is shown in Figure 47.

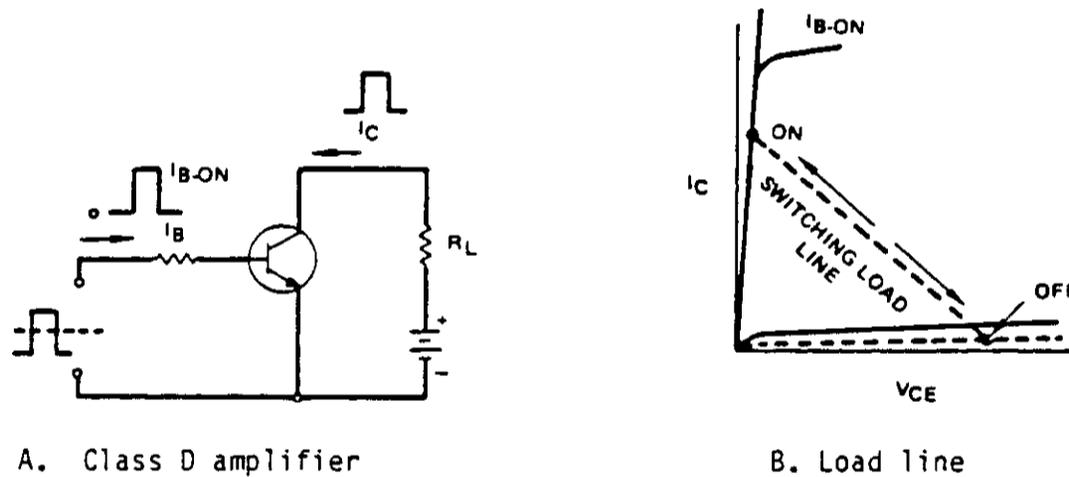


FIGURE 46. Example of Class D mode.

5.3.2 Usual applications.

5.3.2.1 Transistor inverters. Transistor inverters have become an important means for converting dc to ac power in a wide variety of applications. They power complicated electronic systems of military, commercial, and space applications and are widely used as airborne power sources. In such applications, the transistor inverter offers advantages of high efficiency, absence of moving parts, low weight, small size, and high reliability.

5.3 TRANSISTORS, HIGH POWER

A typical inverter circuit is shown in Figure 47 and is comprised of two power transistors, a transformer with a square-loop core, and a source of dc voltage. The circuit operates as follows: assume transistor Q_1 is conducting and transistor Q_2 is not conducting. The supply voltage, V_{CC} , is connected to primary winding 2, which causes an opposing emf to be induced in winding 2 and in all other windings on the core. Feedback windings 1 and 4 are connected in the proper phase so as to bias bases of Q_1 and Q_2 in the off state. This condition will persist until the square-loop core material saturates. Magnetizing current in winding 2 then suddenly increases until the base drive of Q_1 is no longer sufficient to keep the transistor in saturation. Voltage across Q_1 increases and voltage across winding 2 decreases. The resulting induced voltages in windings 1 and 4 turn Q_1 off and Q_2 on. The supply voltage now is connected across winding 3, and the process repeats until the core saturates in the other direction, and another reversal occurs, and so on. The resulting output voltage is a square wave with the frequency

$$f = \frac{V}{4NAB_m} \times 10^8$$

where

- V = applied voltage (volts)
- N = number of turns in winding 2 or 3
- A = core cross-sectional area (cm^2)
- B_m = saturation flux density (gauss)

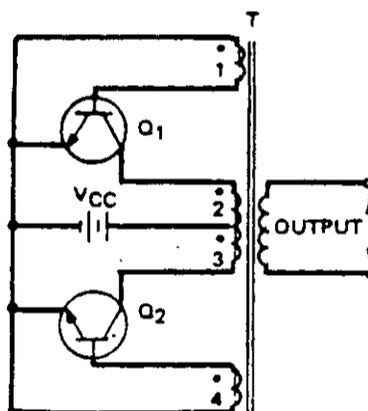


FIGURE 47. Basic inverter circuit.

5.3 TRANSISTORS, HIGH POWER

Typical voltage and current waveforms for one transformer inverter operation are shown in Figure 48. It can be seen from the collector-to-emitter voltage waveforms that each device is subjected in the off condition to a voltage approximately twice the supply voltage plus any induced voltage that may occur in the circuit due to such things as leakage inductance. Also significant is the fact that the same maximum collector current, i_p , is required for switching action whether this current is primarily reflected load current, as in Figure 48B, or totally magnetization current, as in Figure 48C. This will obviously limit efficiency at low output loads.

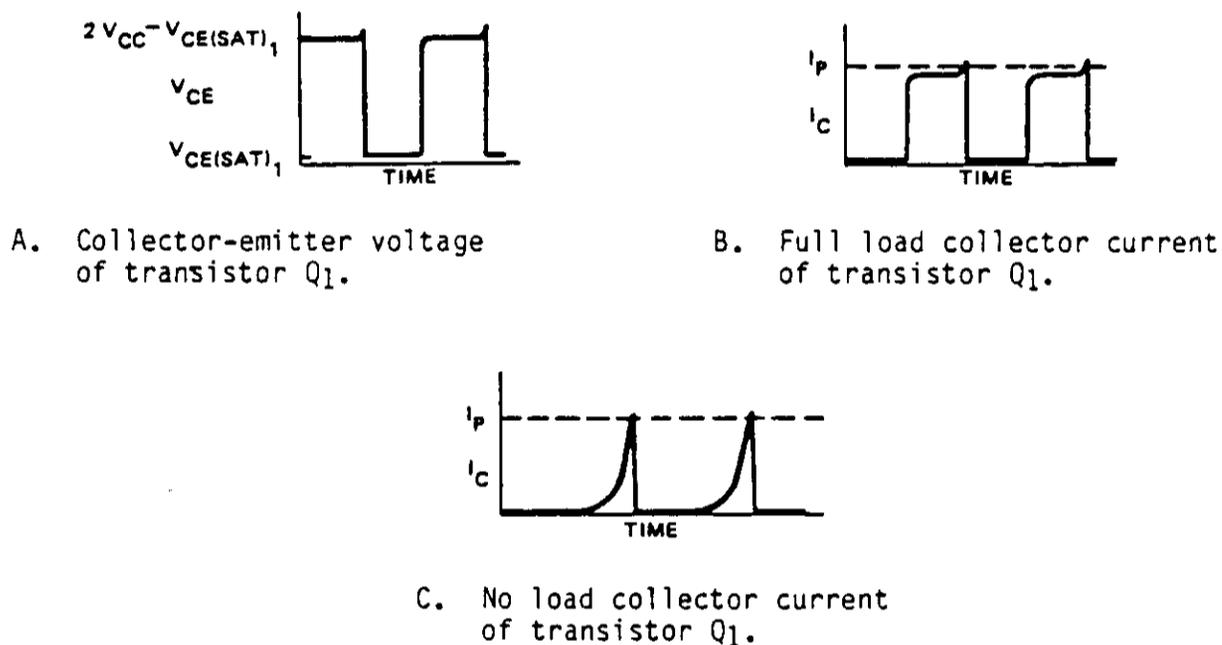


FIGURE 48. Waveforms of basic inverter circuit.

Although the common-emitter connection is most useful, a common collector is sometimes used (when chassis ground polarity permits) because the transistors may be mounted without insulation on a chassis, giving better thermal contact.

At high-frequency and high-output power, the transformer requirements for saturated operation and efficiency present a difficult problem in the basic single-transformer inverter.

5.3.2.2 Saturable reactor controlled inverters. For power levels greater than approximately 25 W, several improvements can be made in the basic circuit of Figure 47. These improvements will provide a more economical circuit, improved transistor operation, and greater circuit flexibility. The modified circuit shown in Figure 49 is a conventional push-pull transformer circuit with feedback derived from additional windings on the transformer as in the previous circuit. A parallel feedback circuit with an additional transformer winding (N_C), voltage divider (P), and saturable reactor (SR) has been added. This modified circuit will oscillate and provide a square wave output voltage at

5.3 TRANSISTORS, HIGH POWER

a frequency determined by the saturable core characteristics and the voltage applied to the winding on the core. The power output transformer is not driven into saturation; therefore, conventional lower-cost core materials may be used. Transistor operation is improved because the peak current, which the transistor must switch, is determined primarily by the load and not by the magnetizing current of a saturated output transformer.

Because the saturable reactor is required to pass only very short pulses of power, the saturation current is much smaller than that for the saturated power transformer of the basic inverter. With only one function to perform, the saturable reactor design can be improved to provide a faster transition from high to low impedance and, thereby, reduce the transistor switching time. The reduced current and shorter switching times significantly lower the transistor dissipation particularly at higher frequencies.

Greater design and circuit flexibility is achieved by permitting the operating frequency of the circuit to be controlled rather than fixed by the supply voltage as in the circuit of Figure 47. With the voltage divider (P), the voltage across the reactor can be varied to produce a variable output frequency.

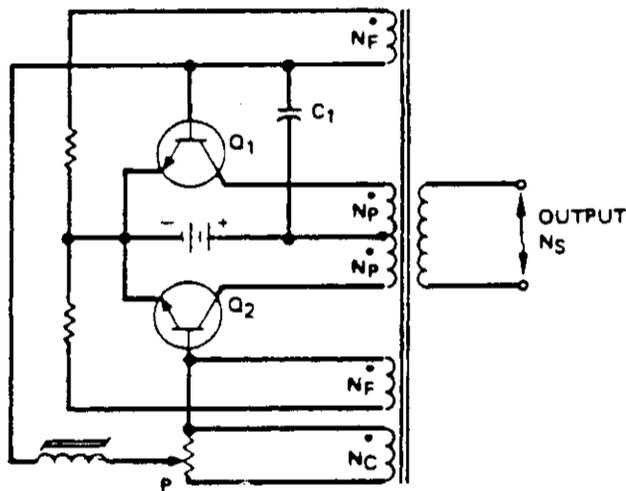


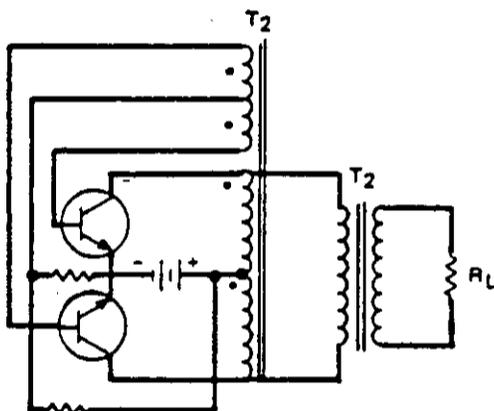
FIGURE 49. Saturable reactor-controlled inverter.

The design of this type of circuit is relatively simple. The output transformer must be selected to support the highest input voltage at the lowest desired frequency without saturating. This precaution will insure that the oscillation is always under the control of the frequency locking reactor circuit. The voltage supplied by the resistive divider to the saturable reactor should be at least equal to twice the voltage supplied by the feedback windings. The circuit will operate with lower voltages applied to the reactor. However, more reliable operation is achieved at the higher voltage. In calculating the turns required on the reactor, the total voltage appearing across the output terminals of the reactor as well as the voltage supplied by the resistive divider network must be considered.

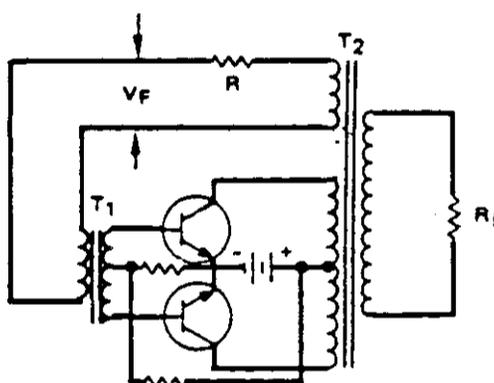
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Tests of the improved circuit indicate that the transistor switching time can be reduced to less than one-half that obtained with the basic circuit and that the dissipation occurring during the switching transient can be reduced to about one-fourth the value obtained with a saturated output transformer. The described circuit has been constructed in power ratings up to 500 W. Variable frequency converters have been built to operate over frequency ranges of 10 Hz to 1 KHz and up to frequencies of 20 kHz. These inverters are used to supply power to motors, electronic equipment, and a variety of other loads.

5.3.2.3 Two-transformer inverters. Other methods of obtaining high output power are the two-transformer inverters, shown in Figure 50. In these circuits, only the small driver transformers (T_1) saturate. This significantly reduces the magnetizing currents which the transistors must switch in the basic single-transformer inverter. The use of normal core material in the nonsaturating output transformer reduces transformer cost and increases efficiency.



A. Two Transformer inverter



B. Two transformer inverter with frequency control

FIGURE 50. Two-transformer inverter.

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Frequency control may be accomplished as shown in Figure 50B where voltage (V_F), is regulated to provide constant frequency or varied to provide variable frequency. The circuits of Figure 51 are recommended to decrease transistor switching time and thereby reduce collector dissipation.

Fast switching is especially important at higher frequencies. Figure 51A shows capacitors in parallel with the base drive resistor of each transistor. This tends to promote faster switching by producing a spike of current limited only by the base-emitter junction impedance during turn-on. This capacitor also provides a low impedance path for the reverse junction current during turn-off. The capacitor discharges through the base drive resistor, which it shunts after the base-emitter junction is reverse biased. The capacitance must be greatly increased to provide any measurable improvement.

The circuit in Figure 51B uses cross-coupled capacitors to aid turn-on and turn-off. When Q_1 is on and Q_2 off, capacitor C_2 is charged to approximately $2 V_{CC}$ through the conducting base-emitter of Q_1 . When Q_2 starts to turn-on, C_2 can discharge through Q_2 and apply reverse bias to the emitter-base junction of Q_1 to turn that transistor off. C_1 is aiding the turn-on of Q_2 as it charges to the off voltage level of Q_1 .

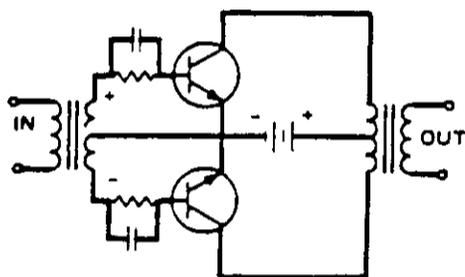
5.3.2.4 Audio amplifiers. The most important requirement of an audio power amplifier is to provide power gain over a wide band of frequencies with minimum distortion. This requirement is met by operating the transistor in the active region. The choice of circuit and operating mode will depend on the application.

In the Class A mode, the transistor is biased to some quiescent operating point with no signal applied. The ac signal then swings the operating point on either side of the quiescent point so that ideally the output collector current is a precise amplified reproduction of the input base current (see Figure 43 of this section). Because the quiescent operating point must allow the collector current to swing both positively and negatively, it is usually near the midpoint of the load line. Consequently, for the transformer-coupled Class A amplifier, the quiescent power dissipation in the transistor is equal to the input power as shown below.

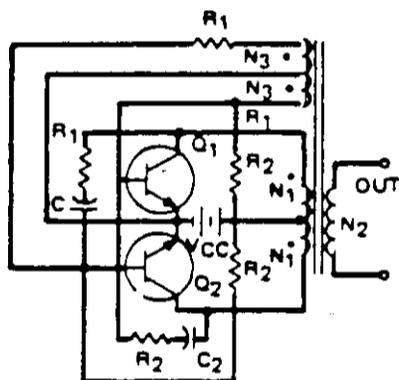
$$P_{IN} = V_{CC} \times I_Q$$

where V_{CC} is the supply voltage and I_Q is the quiescent current. Since the operating point swings both positively and negatively with the signal, the average input power remains constant in the Class A amplifier. The maximum transistor dissipation occurs at the zero signal or quiescent condition, and the maximum ideal efficiency, which occurs at maximum signal, is 50 percent. Because of the relatively high quiescent power dissipation and low efficiency, Class A amplifiers are usually limited to low-power levels.

5.3 TRANSISTORS, HIGH POWER



A. Inverter with speed-up capacitors



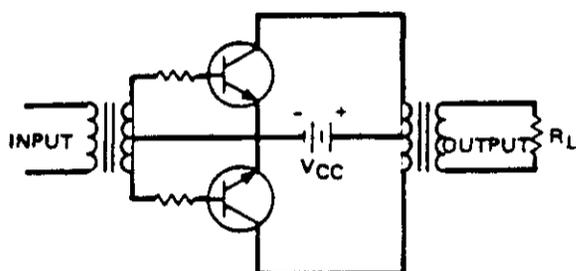
B. Inverter with cross-coupled feedback-capacitor

FIGURE 51. Inverter speed-up circuits.

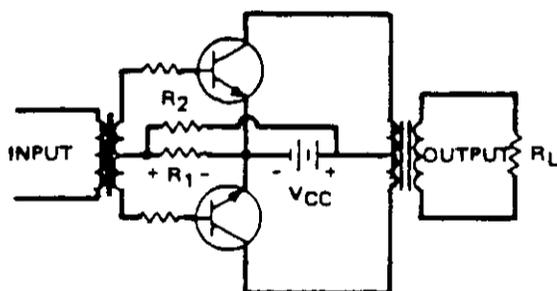
In the Class B mode, the quiescent point bias is zero, so that the zero signal power dissipation in the transistor is zero. The ac signal swings the operating point alternately into the active region and into the cutoff region. Since conduction occurs for only 180 degrees of the ac cycle, it is necessary to operate the transistors in pairs in a push-pull circuit as shown in Figure 52. The two transistors are driven from a split-phase source (a centertapped transformer), and they conduct alternately.

In Class B operation, the maximum ideal efficiency is 78 percent. The maximum power output obtainable for Class B is five times the dissipation rating of the individual transistors. In contrast, the power obtainable in the Class A mode is only one-half the dissipation rating.

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FIGURE 52. Class B amplifier.

The Class AB mode is often used for applications in which low distortion is a prime consideration. This mode is intermediate between Class A and Class B. It is necessary to overcome the so-called crossover distortion which results from the nonlinearity of the transistor input characteristics at low-current levels. Figure 53 shows how the Class B circuit of Figure 52 may be modified to accomplish this purpose. A resistor network, comprised of R_1 and R_2 , is used to bias both transistors slightly into the active region for the quiescent condition. Thus, the conduction angle for each transistor is made slightly greater than 180 degrees, and there is a slight amount of overlapping near zero of the signal ac wave when both transistors are conducting. The quiescent point is chosen so that the base current is beyond the nonlinear knee region of the input characteristic. The transistor dissipation is still minimum at zero signal although it is no longer zero as with Class B. Also, the maximum efficiency is reduced somewhat from the Class B theoretical value. The Class C and Class D modes are not capable of linear operation.

FIGURE 53. Class AB amplifier.

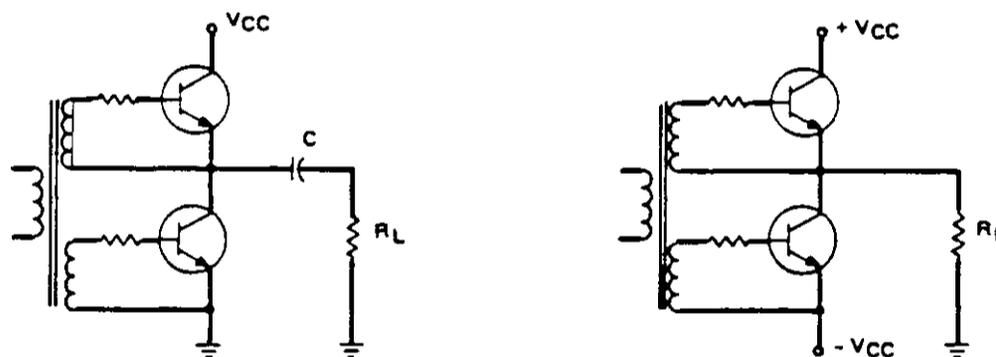
The following discussion is limited to the push-pull Class B (or AB) amplifier, which is the most useful type for high-power applications.

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A number of circuit configurations are available to the amplifier designer. The most familiar is the transformer-coupled circuit illustrated in Figures 52 and 53. This circuit is commonly used in applications where high-power levels and narrow-to-moderate bandwidths are required such as public address systems, sonar, and servo amplifiers. The output transformer allows great flexibility in matching the amplifier to a wide range of load impedances.

The phase shifts in transformers at the high and the low frequency ends of the amplifier bandwidth frequently lead to instability when the transformer is included in an inverse feedback loop. These problems have led to the development of a number of "transformerless" power amplifier circuits. One of these is the series transistor circuit shown in Figure 54. Two methods are shown for connecting the load. In Figure 54A, a coupling capacitor is used. In Figure 54B, the load is directly connected, which necessitates a dual power supply to set the input end of the load at ground potential. One disadvantage in this circuit is that it still requires 180-degree phase inversion for the base drive. This may be provided by a transformer with its attendant phase shift problems as shown in Figure 54, or by a split-load phase inverter as shown in Figure 55. The latter solution also may present problems such as difficulty of impedance matching a need for very large coupling capacitors.

These difficulties may be eliminated by the complementary circuit as shown in Figure 56. This circuit avoids the need for a transformer or a split-phase input. Advantage is taken of the availability of pnp as well as npn devices. The amplifier operates as a true push-pull amplifier because one transistor is driven on while the second is driven off and vice versa. The complementary circuit also allows the design of a direct-coupled amplifier.



A. Capacitor coupled load

B. Dual power supply

FIGURE 54. Series output amplifier.

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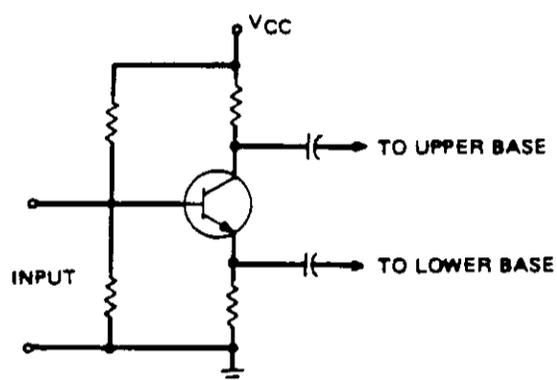


FIGURE 55. Split-load phase inverter.

A variation of the complementary circuit is the quasi-complementary circuit shown in Figure 57. This circuit allows the use of identical npn output transistors. The driver transistors give additional current gain, thus reducing the previous stage driving requirement.

The upper npn-npn pair of transistors is connected in the Darlington configuration and behaves as a high-gain npn device. The lower pnp-npn pair works as a high-gain pnp device. Analyzed in this way the circuit is seen to be the equivalent of that shown in Figure 56.

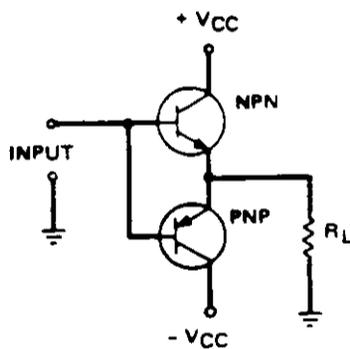


FIGURE 56. Complementary circuit.

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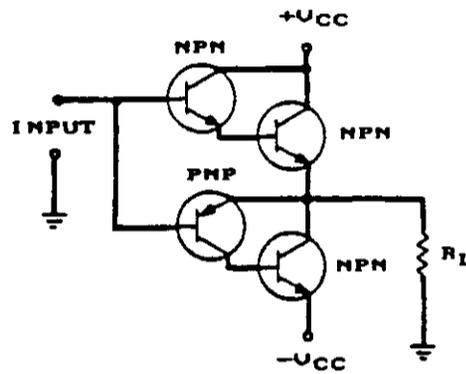
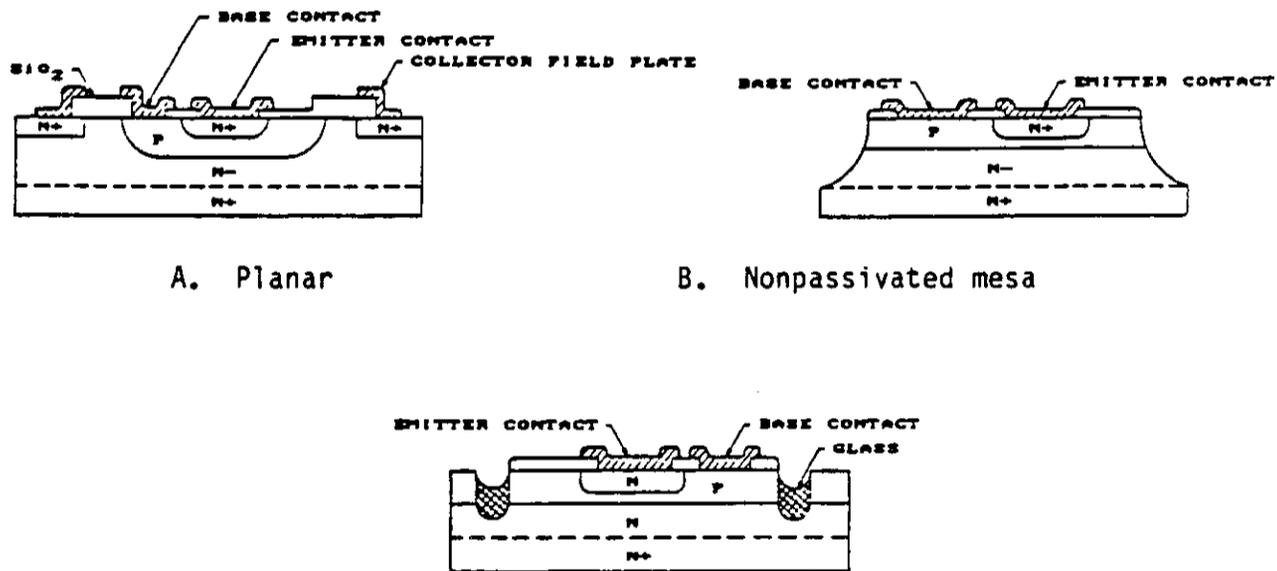


FIGURE 57. Quasi-complementary circuit.

5.3.3 Physical construction.

5.3.3.1 Die construction. Two basic types of structures are used in die fabrication for high-power transistors. These are the mesa (single and double) and planar structures. These structures are shown in Figure 58.



C. Glass-passivated double mesa

FIGURE 58. Basic die structures.

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The planar structure offers the lowest leakage current because both the E-B and C-B junctions are always protected by ultra clean silicon dioxide or some other type of primary passivant. For spaceflight and critical aerospace applications semiconductor devices containing a conformal coating should not be used. Conformal coatings may have inherent problems when exposed to radiation and temperature extremes, such as outgassing, and physical and/or chemical alterations.

Moreover, three basic die processes are used to produce these two types of structures. These are the double diffused (epi-collector), triple diffused, and epi-base processes. It is, therefore, possible by using a combination of these processes and structures, to produce a variety of transistor types (with the exception of an epi-base planar). Consequently, three manufacturers supplying the same JANTXV or JANS device could provide devices containing dice with different structures and process combinations. Although these devices may pass all electrical requirements, they might possess different electrical characteristics. Such a situation can lead to a problem known as the unspecified parameter, that is, whereas one manufacturer's JANTXV devices functions well in an application, the same type from another manufacturer might not.

5.3.3.2 Packaging. Some of the standard packages used in the packaging of these devices are the T0-3, T0-66, T0-111 and MT-53. Figure 59 shows the basic construction of a T0-3 package.

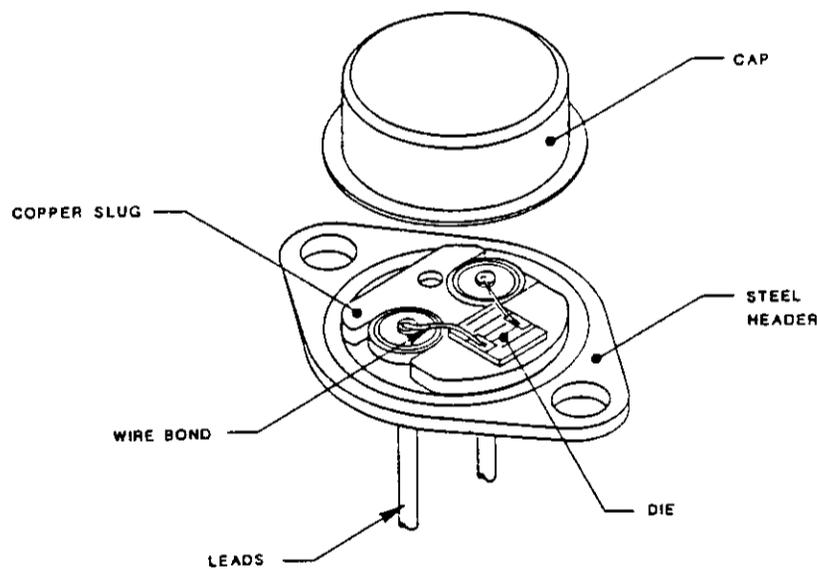


FIGURE 59. Typical T0-3 transistor construction.

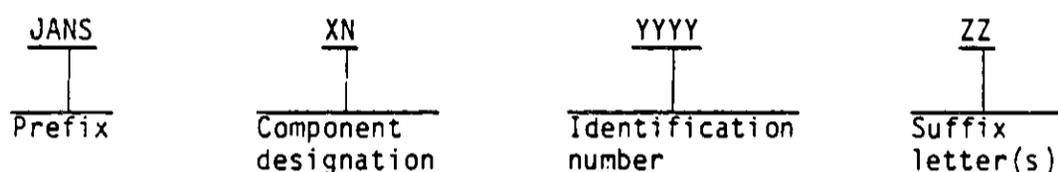
Die bonding methods used can be either soft solder or gold eutectic with moly interface. Both methods have been found to provide reliable bonds. However, if power cycling capability greater than 6000 cycles (at T_c deltas of 100 °C) are required, gold eutectic/molybdenum would be the best choice.

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Although there are various methods of making a die-to-lead interconnections the most popular wirebonding technique is ultrasonic bonding to the die and ultrasonic or electrical weld to the lead, with aluminum as the interconnecting wire material.

Sealing of the package is done in an inert atmosphere using electrical weld techniques. Due to this welding technique, it is possible that weld splashes might occur causing particles to be released inside the package. Nonetheless, all MIL-STD-975 devices require that a PIND and X-ray test be performed to remove any defective devices.

5.3.4 Military designation. The military designation for transistor is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 2N for transistors.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices, suffix M, shorter or longer terminal leads, suffix S or L, or any other letter to indicate a modified version.

5.3.5 Electrical characteristics. A knowledge of the capabilities and limitations of power transistors is essential for the economical and reliable application of these devices. Much of the required information can be obtained from the manufacturer's technical data sheets. However, the circuit applications engineer must have a clear understanding of the meanings of the various ratings and characteristics, how they were derived by the manufacturer and how they are related to his application in order to make optimal use of such information.

Some of the basic electrical characteristics have already been discussed in paragraph 5.2.5 in the low-power subsection; these are also applicable to high-power devices. The following paragraphs will emphasize some of the electrical characteristics of high-power devices and discuss thermal considerations.

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5.3.5.1 Breakdown and leakage characteristics. I_{CBO} is the dc current that flows when the collector-base junction is reverse-biased by a specified voltage. It is temperature dependent; therefore, the temperature must be specified as one of the test conditions. A typical collector blocking characteristic is shown in Figure 60. As the voltage is increased, a value is finally reached beyond which the current increases very rapidly for small increments of voltage. This is the avalanche region. The breakdown voltage $[V_{(BR)CBO}]$ is that voltage in or near the avalanche region.

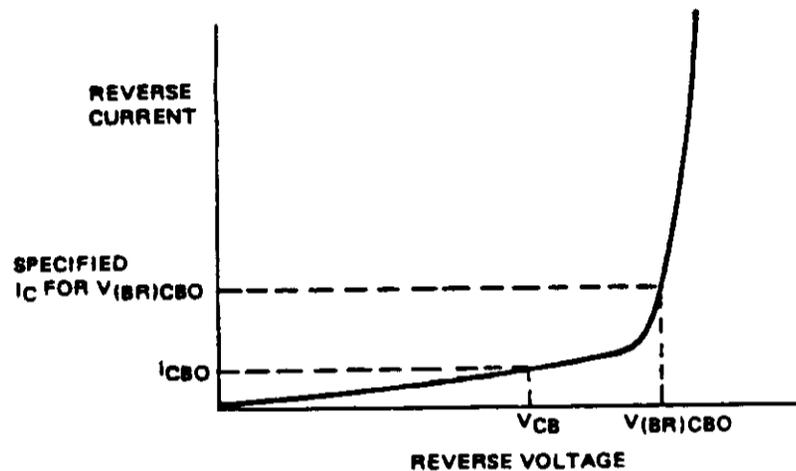


FIGURE 60. Collector junction blocking characteristics.

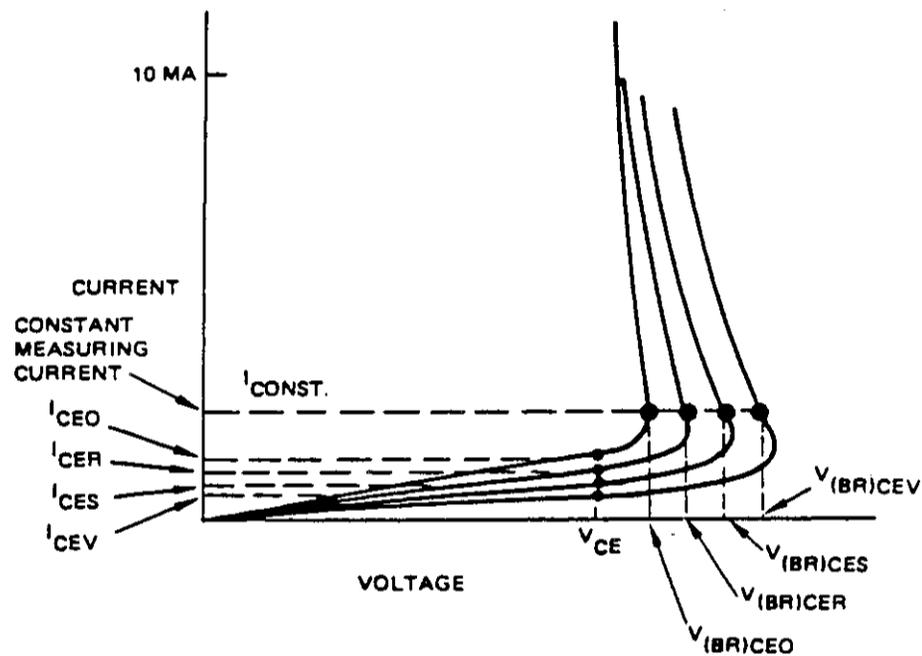
I_{EBO} is the dc current which flows when the emitter-base junction is reverse-biased by a specified voltage. The same comments and shape of characteristic curves apply to I_{EBO} as to I_{CBO} . However, in high-power transistors the emitter breakdown voltage $[V_{(BR)EBO}]$ is usually designed to be much less than $V_{(BR)CBO}$.

I_{CEV} , I_{CES} , I_{CER} , and I_{CEO} are the dc currents that flow when the collector and emitter terminals are connected to a specified voltage with polarity such that the collector-base junction is reverse-biased and the base terminal is returned to the emitter terminal through a circuit designated by the third letter in the subscript as follows:

- V -- reverse base-emitter bias
- S -- short circuit
- R -- resistance
- O -- open circuit.

The common emitter blocking currents are also temperature-dependent, particularly I_{CEO} . Figure 61 compares the relative leakage current levels for these four parameters. Also shown are the corresponding breakdown voltages.

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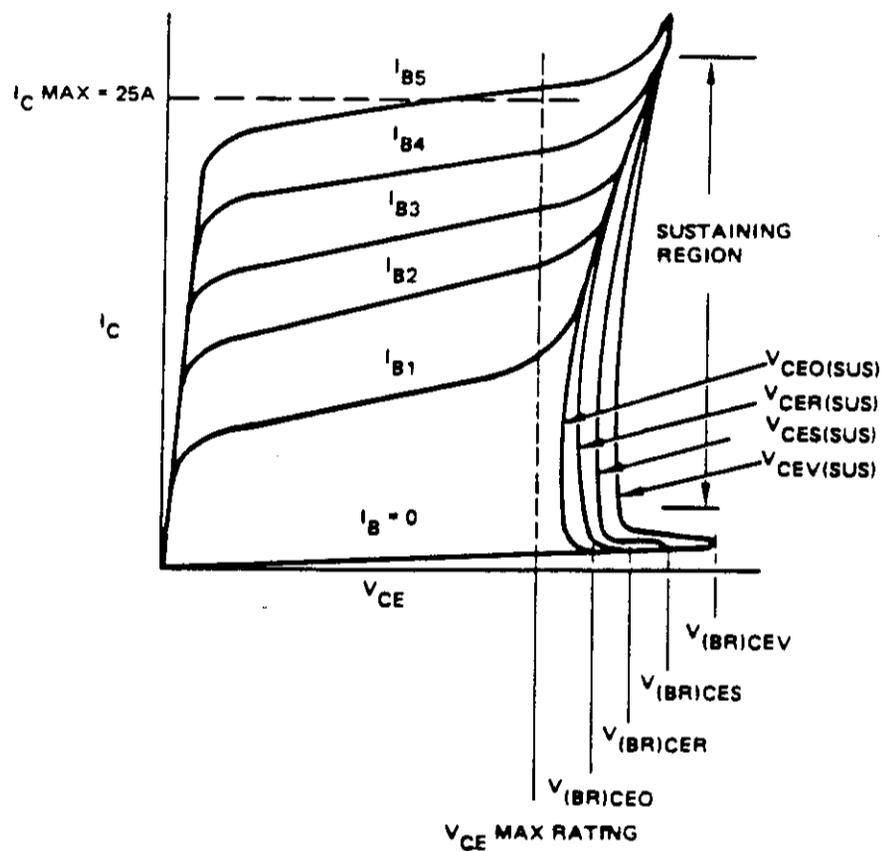
FIGURE 61. Common emitter blocking characteristics.

5.3.5.2 Sustaining voltages. Because of the negative resistance region, the breakdown voltages described above are still inadequate to specify completely the voltage capability of a transistor. Whereas the breakdown voltages are usually specified at a current of several milliamperes, the applications engineer may well be interested in the voltage limitations of full rated current.

The complete collector family of curves for a typical 30 ampere transistor is shown in Figure 62. It should be noted that the current scale of this family is greatly compressed as compared to that of Figure 61. Also, the shape of the blocking portion of the curves is somewhat exaggerated for the sake of clarity.

As shown in Figure 62, the V_{CE0} , V_{CER} , V_{CES} , and V_{CEV} curves, after passing through the breakdown points ($V_{(BR)CE0}$, etc.) and the negative resistance region, finally reach minimum values. This portion of the curves is known as the sustaining region because the voltage remains relatively constant over a wide range of collector currents. Also shown in Figure 62 are the curves for various constant values of base drive, I_{B1} , I_{B2} , etc., which constitute the

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FIGURE 62. Transistor collector family of curves.

active region of the collector family. All of the forward-bias base curves finally bend over and become asymptotic to the open-base sustaining voltage curve.

The symbol for designating the sustaining voltage level is formed by adding (sus) to the suffix; e.g., $V_{CE0(sus)}$, although the nomenclature BV_{CE0} has been used in some literature. The sustaining voltages are specified at a current near the minimum voltage portion of the curves. For most high-power transistors, this current is in the range of several hundred milliamperes. The open base sustaining voltage [$V_{CE0(sus)}$] is one of the most important parameters

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for a power transistor because it is the absolute upper voltage limit of the collector family in the active area.

5.3.5.3 DC current gain. The static value of the forward current transfer ratio (dc current gain in the common emitter configuration) is one of the most important parameters for transistors. Its value is required in low frequency applications involving large current swings and also in high frequency applications to establish biasing conditions. The common emitter dc current gain is defined as the ratio of dc collector current to dc base current and is represented by the symbol h_{FE} .

$$h_{FE} = \frac{I_C}{I_B}$$

The value of h_{FE} is not constant, but varies with collector voltage V_{CE} , collector current I_C , and junction temperature T_J . It is, therefore, necessary to specify V_{CE} , I_C , and T_J as test conditions in order to measure h_{FE} .

V_{CE} is usually specified at a value sufficiently above the saturation voltage so that the test operating point is beyond the region of maximum curvature shown in Figure 63.

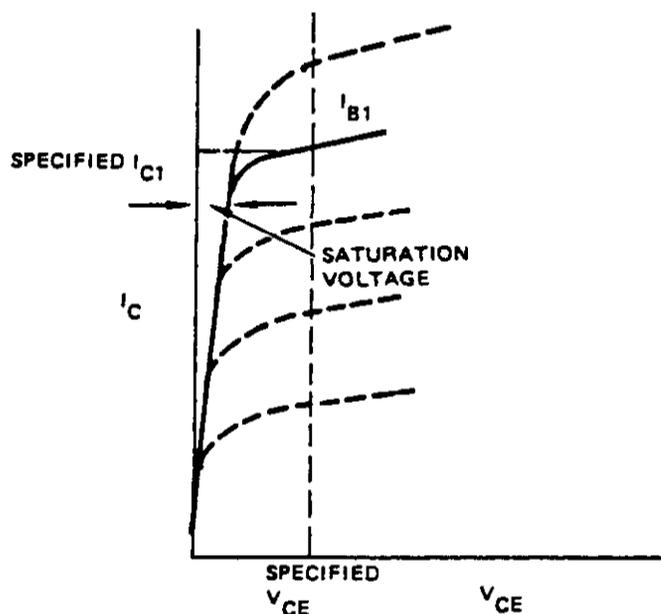
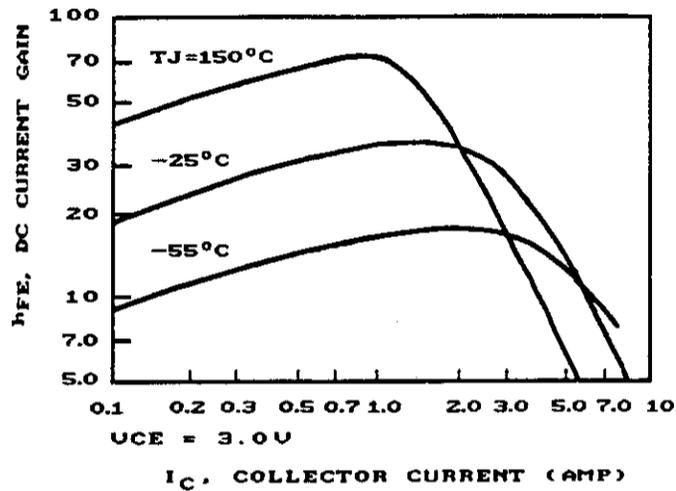


FIGURE 63. Test conditions for h_{FE} .

A typical curve for h_{FE} versus I_C at constant V_{CE} and three temperatures is given for high-power transistors in Figure 64.

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$T_J = 150^\circ\text{C}$.

FIGURE 64. DC gain versus collector current.

5.3.5.4 AC current gain. The small-signal, short-circuit, forward current transfer ratio or ac current gain in the common emitter configuration is required in amplifier applications. It is defined as the ratio of ac collector current to ac base current and is represented by the symbol h_{fe} .

$$h_{fe} = \frac{di_c}{di_b} = \frac{I_c}{I_b}$$

at $v_{ce} = \text{constant}$.

If h_{fe} is plotted against frequency, the curve shown in Figure 65 results.

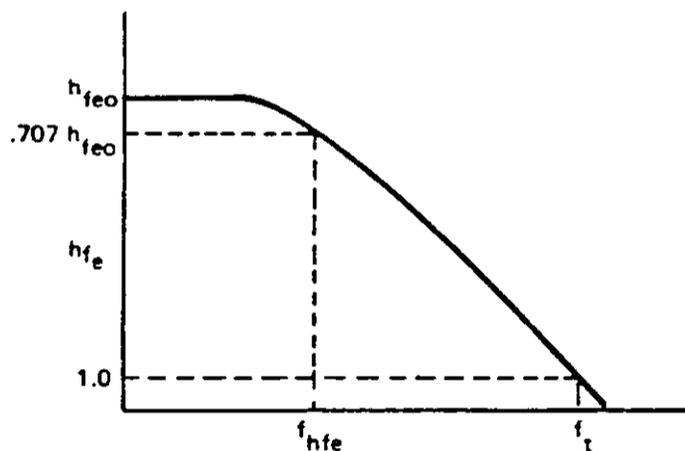


FIGURE 65. h_{fe} versus frequency.

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A number of important terms appearing on this curve should be noted:

- a. h_{fe0} is the low-frequency value of h_{fe} , usually measured at 1 KHz.
- b. $f_{h_{fe}}$ (or β cutoff) is the frequency at which the ac gain has fallen to $0.707 h_{fe0}$.
- c. f_t is the frequency at which the ac gain has fallen to 1.0.

At frequencies greater than 10 MHz, it is necessary to use high-frequency techniques to obtain ac gain measurements.

5.3.5.5 Thermal considerations.

5.3.5.5.1 Secondary breakdown. A power circuit design cannot be considered reliable unless the circuit has been checked to insure that the transistors will not undergo secondary breakdown. Operating within the power-temperature ratings and insuring against thermal runaway will not alone guarantee circuit reliability.

Figure 66 shows a sketch of the voltage-current characteristics of a transistor operating in the reverse breakdown mode. At low collector currents, the voltage across the device exceeds the open base breakdown rating. The peak of the curve and the negative resistance region is called either the first or normal breakdown and results from avalanche action in the transistor. However, as current in the avalanche mode is increased to higher values, a critical current (I_m) is reached at which the voltage across the device drops to a very low level. This behavior is called secondary breakdown.

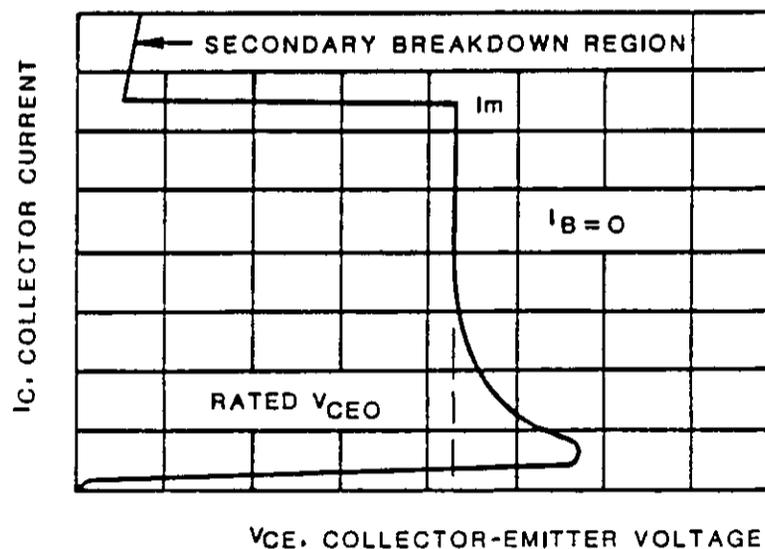


FIGURE 66. Manifestation of secondary breakdown in a transistor.

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Transistors need not be operating in avalanche breakdown in order to encounter secondary breakdown. Figure 67 shows a family of collector curves and the locus of critical or trigger currents at which the transistor enters secondary breakdown. As collector voltage is increased, maximum voltage occurs at lower currents and becomes extremely low as the emitter-base junction becomes reverse-biased. It has also been found that the amount of time a power pulse is applied at a particular operating point also determines whether or not secondary breakdown will occur. The observed behavior is a result of hot spots forming in the device as a result of nonuniform current density.

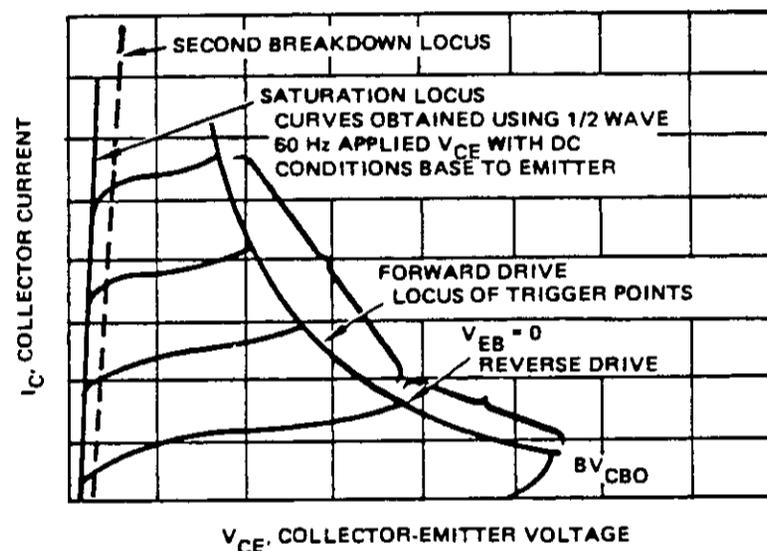


FIGURE 67. Locus of secondary breakdown trigger points.

Secondary breakdown in transistors is triggered when the temperature of some part of the material achieves an extremely high value. It causes material to go into intrinsic conduction and become a very low resistance material. If the current is removed or held to very low values while the transistor is in secondary breakdown, no particular harm occurs and the semiconductor will act normally when it cools.

However, when the transistor enters secondary breakdown, a collector-emitter short results. The reason for this is that it is difficult to prevent the junction temperature from exceeding the melting point of the material. High pin-point temperatures in the transistor are produced because certain biasing conditions cause current to concentrate into a few very small areas. These small areas sustain the full collector current which normally, under ideal conditions, would be distributed evenly along the entire semiconductor active region. If the temperature at one of these hot spots becomes sufficiently high, secondary breakdown will occur. The resulting decrease in voltage across the device normally will result in an increase of current through the transistor because of the action of the other circuit elements. One small area must now

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sustain even more current than before, so the temperature can become extremely high. Extreme temperatures can rapidly change the characteristics of the device or melt the material. If enough melting occurs in the region of the high current path from collector to emitter, upon cooling a low resistance path or short circuit from collector to emitter will exist.

If it were possible to remove the current within a few microseconds after the onset of secondary breakdown, no particular damage to the transistor might result. Actually, under far less ideal conditions it has been possible to observe devices being switched in and out of secondary breakdown.

The current at which secondary breakdown occurs decreases markedly with increases in collector-emitter voltage when operating in the normal active region mode. The curves of Figure 68 illustrate this behavior and are typical for most types of semiconductors.

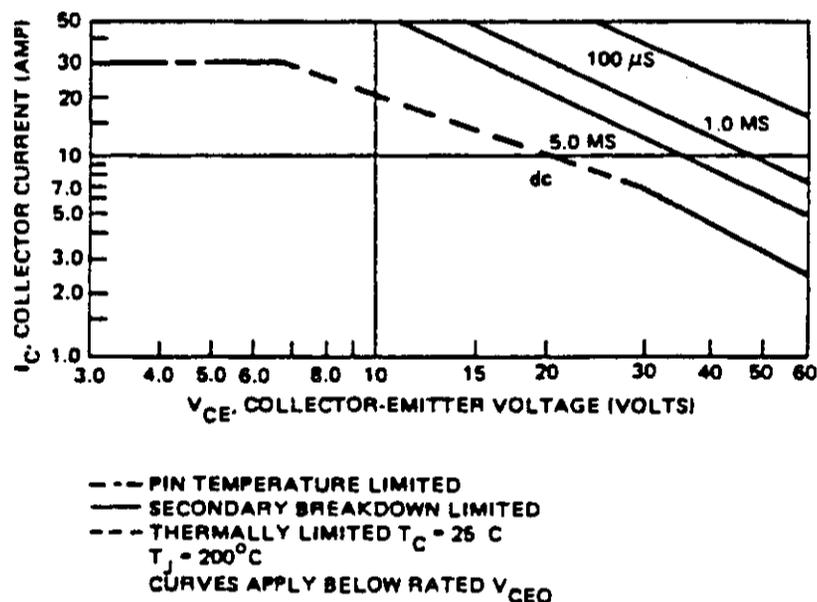


FIGURE 68. Example of an active region safe operating area.

The reason for collector voltage being an important variable is that as collector voltage is increased, the base width is reduced, thereby accentuating current crowding effects because the higher electric field caused by the shorter path reduces the current spreading or fan-out. Therefore, transistors with narrow base widths (i.e., higher f_t) will encounter secondary breakdown at lower power levels than lower frequency devices, other conditions being equal.

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Because of the secondary breakdown problem, many vendors of power transistors provide safe operating area (SOA) charts as shown in Figure 68. The solid lines show secondary breakdown limitations whereas the dotted lines represent thermal limitations. For this transistor family, dc currents above 30 A cause excessive emitter pin temperatures (on low-level devices, I_C is limited by the bonding wire at low values of V_{CE}); therefore, operation above 30 A dc is not recommended. Above 6.5 V, the allowable dc current is junction temperature limited. If the case temperature is 25 °C, power dissipation must not exceed 200 W. The dc curve also shows that the dc power level must be lowered from the 200 W level as voltage increases above 30 V, if secondary breakdown is to be avoided. At case temperatures higher than 25°C, thermal resistance information must be taken into account to derate the power dissipation limit to keep the junction temperature (T_J) below its limits, $T_{J(max)}$. The secondary breakdown limitation curve is valid when $T_J < T_{J(max)}$. Therefore, the curve is not derated with increases in case temperature, because the curve is dependent on junction temperature only.

The various pulse curves given show no thermal limitation for currents to 50 A. The transistor power is limited solely by secondary breakdown. (At higher case temperatures the transient, thermal resistance must be used to determine if operation is within $T_{J(max)}$.) The given pulse curves are used if the duty cycle is 10 percent or less. For higher duty cycles, the curves will gradually degrade to the dc curve.

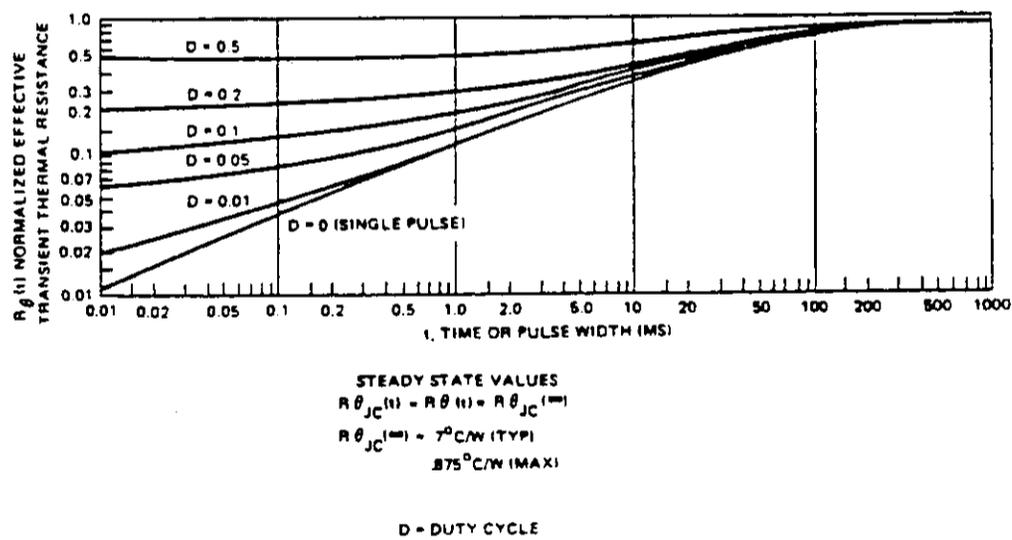


FIGURE 69. Example of thermal response data.

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It is common industry practice to rate power devices at a case temperature of 25 °C, even though it is very unlikely that the case would ever be held at 25 °C in a practical operating situation. It is instructive to find the effect on the safe area curve for this particular transistor if the case were at 125 °C.

In Figure 69, which shows the thermal response of a typical power transistor, the normalized effective thermal resistance, $R_{\theta}(t)$, can be read for a single pulse ($D=0$) as 1.0 for dc, 0.23 for 5 ms, 0.13 for 1 ms, and 0.034 for 0.1 ms. The effective thermal resistance, $R_{\theta JC}(t)$, is found by multiplying these normalized values for the steady state value of 0.875 °C/W. The allowable power is found from

$$P_D = \frac{T_J - T_C}{R_{\theta JC}(t)}$$

where $T_J - T_C = 75^\circ\text{C}$ in this example.

The power levels allowed are, respectively, 86, 370, 660, and 2500 W. These values can be used to construct a safe area curve based upon $T_C = 125^\circ\text{C}$ as shown in Figure 70.

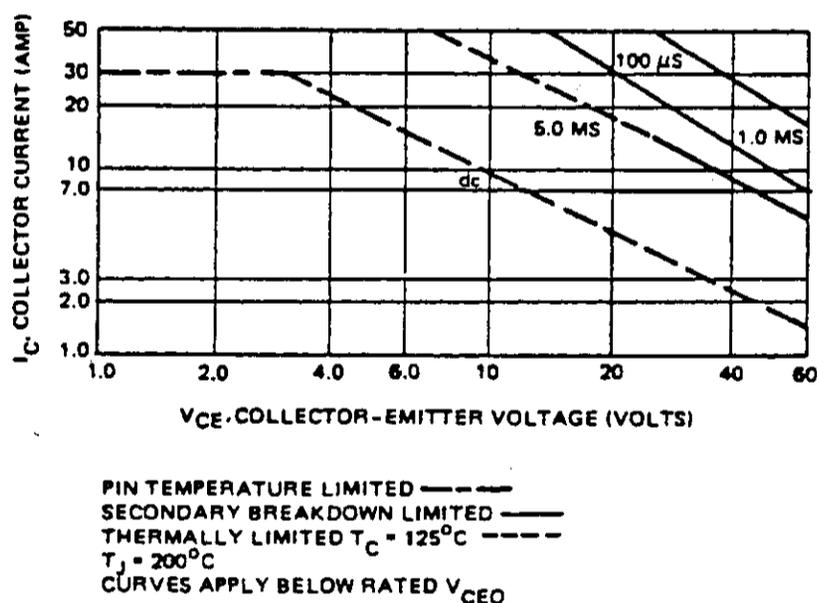


FIGURE 70. Constructed safe operating area data at $T_C = 125^\circ\text{C}$ for transistor of Figures 67 and 68.

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The secondary breakdown curves are unchanged. They are already based upon $T_J = T_{J(max)}$. The dc operation is entirely thermally limited and thermal and thermal limitations appear on the lower voltage ranges of the longer-pulse times. However, at 100 μ s no thermal limitations appear below 50 A. Such behavior is typical of most transistors; i.e., dc power becomes thermally limited at fairly low case temperatures, whereas under short pulse operation power dissipation is limited by secondary breakdown even at fairly high case temperatures.

The curves indicate the limits of power dissipation which may be sustained for a time interval as shown by the designated line; i.e., a 20-V, 30-A power pulse may be sustained for 1 ms without encountering secondary breakdown. In switching applications, the load line may traverse along one of the limit lines for the time indicated.

Figure 71 shows some of the circuits and load lines that might be encountered. Note that load lines can vary considerably in different applications because of the effects of transistor switching speed, circuit capacitance and inductance values, strays due to wiring, and transformer leakage inductance.

The first five circuits have typical inductive load lines. The most demanding is the transformer-coupled audio amplifier (see Figure 71A). Under normal operation, the load line is resistive and as long as it falls below the dc safe area line, no failures should occur. However, should the load be opened with signal applied, a reactive, nearly rectangular load line could result. It is important that the drive be restricted to keep the collector current within bounds and the whole load line below the dc safe operating area or failure could result.

The solenoid drivers (Figures 71B and C) cause relatively few problems. Peak currents are restrained by R_L and switch off is fast. The collector current remains fairly constant until the clamping level of the diode is reached, then the current transfers to the diode. The simplest and fairly reliable way to check for safe areas is to obtain the collector current waveform and make an equivalent rectangular model having the same peak amplitude and a width adjusted to have the same energy as the actual pulse. Assume collector voltage is constant at the clamp level (V_K)--it will generally rise to this level quickly--and note if this condition is within the safe area. For example, suppose the voltage rises from A to B in 1 μ s and it takes 10 μ s for the current to linearly decay from point B to point C. A 6 μ s rectangular pulse of amplitude I_{max} would contain the same energy. Locate the point on the safe area at a current I_{max} and a voltage of V_K (point B) and note how much time is allowed there. If it is greater than the equivalent rectangular pulse, the transistor is safe. Generally, at times below 10 μ s, all transistors can withstand their rated current and voltage simultaneously, so that this mode of operation seldom results in transistor failure.

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The power inverter (Figure 71D) and the ignition system (Figure 71E) can be treated in the same way as the relay drivers. The fact that no clamp exists in Figure 71D makes the circuit more sensitive to transients, but the critical point is still at the high power corner and the waveform modeling approach just described will work.

The switching motor control (Figure 71F) and the switching regulator (Figure 71G) differ from the previous circuits in that a large power transient occurs on both the switch-off interval (because of the inductor) and switch-on intervals (because of the diode recovery current); therefore, both intervals need to be checked. The same modeling principle applies, but both forward bias SOA (FBSOA) and reverse bias SOA (RBSOA) curves have to be scrutinized.

Should both voltage and current change at about the same speed, the foregoing approach may be too conservative. For a more exact analysis, it is necessary to plot the load line and note the time between certain intervals. A comparison to the SOA curve can then be made. Note that the pulse curves indicate that the transistor operating point can spend a given time at any one point on the curve or may travel along the curve for the same given time.

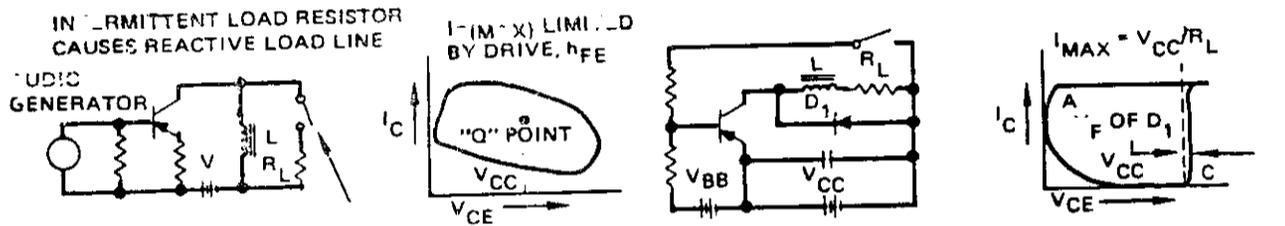
5.3.5.5.2 Power transistor cooling. Transistors with power ratings greater than 1 W are usually provided with a large, flat surface that can be clamped against a metal exchanger. The purpose of the heat exchanger is to transfer the heat to a larger surface from which it can then be dispersed by a cooling medium. The heat must pass through several thermal impedances as it flows from its source at the collector junction to the final cooling medium. These "thermal impedances" may be modeled as a series circuit, as shown in Figure 72.

The heat flowing through the thermal impedances will produce a succession of temperature drops, just as current flowing through electrical resistance will produce voltage drops. The value of the thermal impedance ($R_{\theta JC}$) between the collector junction and case is a function of the thermal design of the transistor. It is affected by the thermal conductivity of the various solders that bond the parts together. The $R_{\theta JC}$ values for transistors are given on their data sheets.

The thermal characteristics of the transistor are also specified indirectly on the data sheets in terms of thermal derating curves. These derating curves are often preferred over simple $R_{\theta JC}$ values because they include thermal transient effects and secondary breakdown limitations.

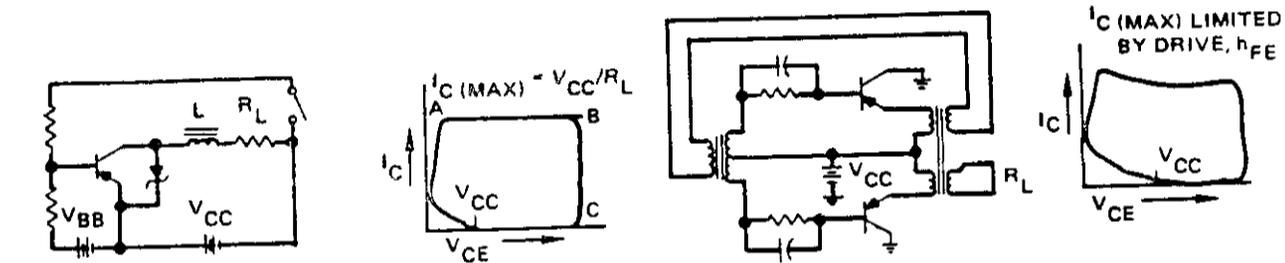
Although a maximum junction temperature is given in data sheets, this parameter cannot be measured directly. Consequently, most of the values and curves for transistor thermal analysis are based upon the case temperature (T_C) which can be measured by conventional thermocouple probes.

5.3 TRANSISTORS, HIGH POWER



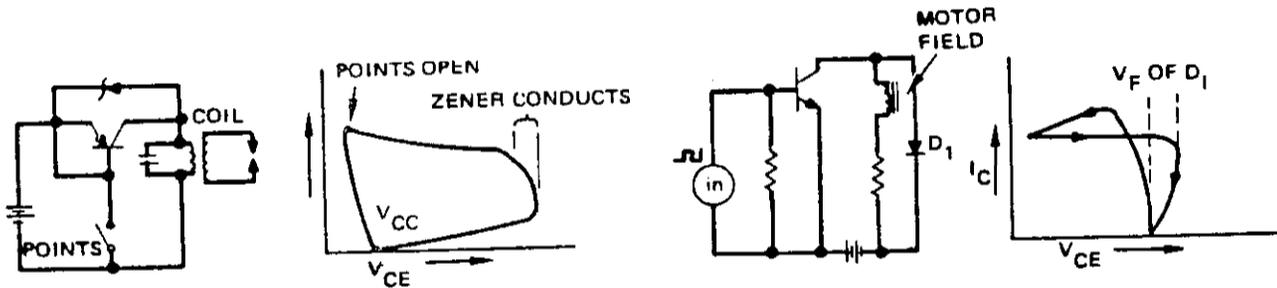
A. Class A audio amplifier

B. Solenoid driver with rectifier diode to suppress inductive kick



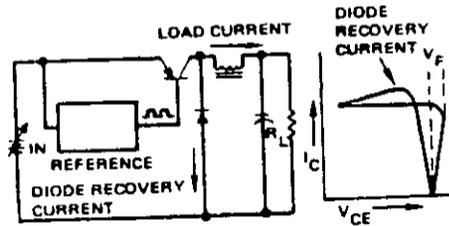
C. Solenoid driver with zener diode suppression

D. Power inverter



E. Ignition system

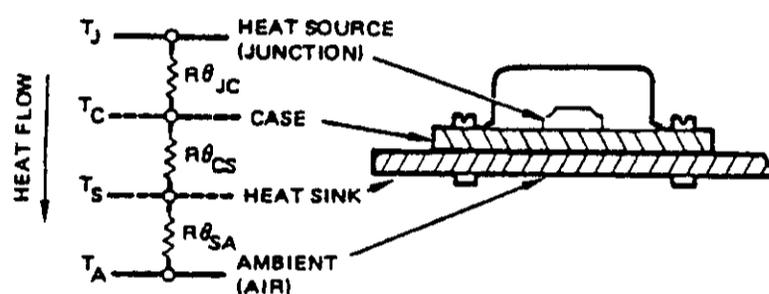
F. Switching type motor control



G. Switching type series regular

FIGURE 71. Examples of circuits.

5.3 TRANSISTORS, HIGH POWER

FIGURE 72. Thermal circuit.

Because of the importance of the case temperature in practical design problems, it is useful to note how T_C should be measured. The case temperature reference point for stud-mounted transistors is in the plane of the mounting surface, directly below the center of the collector junction. Access to this reference point is obtained by drilling an axial hole in the threaded stud as shown in Figure 73.

For case styles in which the mounting surface is clamped to the sink by bolts through a flange, such as the T0-3 case, the above procedure must be modified. The practical approach is to choose the nearest reference point to the ideal location. For the T0-3, the thermocouple should be placed into a very small hole drilled into the top surface of the mounting flange, at the end of the case closest to the transistor die (see Figure 74).

5.3.5.5.3 Mounting practices. The next thermal impedance in the heat removal circuit to be considered is $R_{\theta CS}$, the small but unavoidable thermal impedance at the interface between the transistor case and the heat sink (see Figure 72). In order to obtain the maximum rating from a power transistor, certain recommended heat sink mounting practices should be observed.

Regardless of the type of heat sink employed, the objective is to obtain the best possible thermal contact between the flat mounting surface of the transistor case and the sink. In order to avoid high thermal drop, a large mating surface area should be provided. Holes should be drilled or punched clean with no burr or ridges.

Careless deburring can seriously increase the thermal resistance and result in transistor overheating.

5.3 TRANSISTORS, HIGH POWER

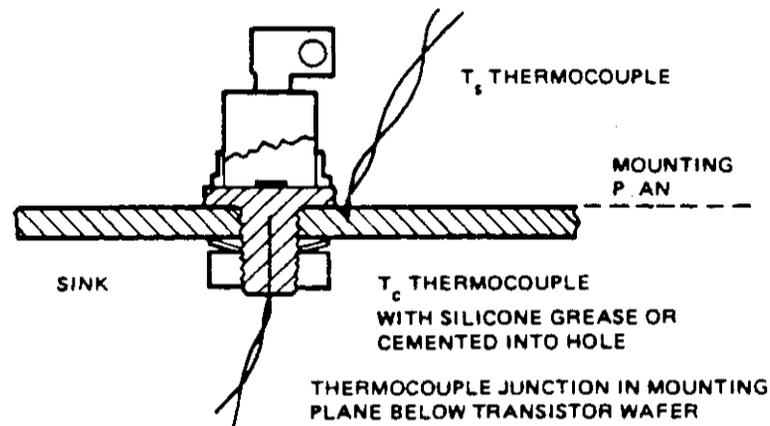


FIGURE 73. Thermocouple placement for stud-mounted case.

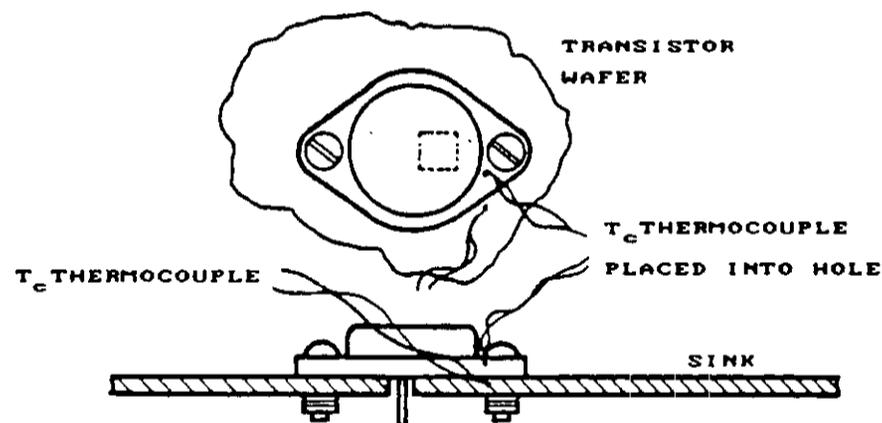


FIGURE 74. Thermocouple placement for TO-3 case.

However, even with the above precautions the surface will not be perfectly smooth. The two surfaces at the interface are actually in contact at only a limited number of points, with the remainder of the interface area consisting of a thin gap. Because air has very poor thermal conductivity, a marked reduction in $R_{\theta CS}$ can be obtained by filling the gap with an approved heat sink compound (HSC). Table V gives typical values of $R_{\theta CS}$ for various types of power transistors, both with and without HSC on the mounting interface. Care should be exercised to exclude foreign particles, particularly when using HSC, since this material can easily capture grit and dust during mounting.

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The nuts used on transistor studs should be carefully torqued down. Insufficient torque can lead to overheating, whereas excessive torque can strip threads, elongate studs and damage the transistor collector junction. It is important to remember that the studs are made from relatively soft copper to obtain high thermal conductivity. They cannot withstand the same torque as comparable-sized steel bolts. Torque wrenches should always be used when mounting power transistors and the specification limits for both maximum and minimum torque carefully observed. The effect of torque on the value of $R_{\theta CS}$ is illustrated in Figure 75.

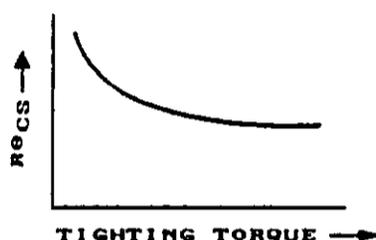
Recommended contact force should be maintained throughout the operating temperature range and allowance made for the stress relaxation of components with time. The practice of screwing the transistor into a tapped hole should be avoided because temperature cycling can produce "thermal ratcheting" and gradual loosening of the device.

It is sometimes necessary to electrically insulate the transistor case from the sink by means of an insulating washer. Unfortunately, this washer contributes appreciably to the value of $R_{\theta CS}$, the magnitude of this contribution being shown in Table V. Therefore, it is preferable to mount the transistor directly on the sink and insulate the sink from the rest of the circuit wherever possible.

TABLE V. Typical values of $R_{\theta CS}$

Stud Threads	Thermal Impedance, °C/Watt			
	$R_{\theta JC}$	$R_{\theta CS}$ (Dry)	$R_{\theta CS}$ (HSC)	$R_{\theta CS}$ (Insulator-HSC)
1/4-28	0.75	0.7	0.3	1.1
5/16-24	0.7	0.5	0.2	0.7
5/16-24	0.5	0.5	0.2	0.7
1/2-20	0.45	0.3	0.15	0.5
T0-3	1.5	0.5	0.2	0.7

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FIGURE 75. Relation between tightening torque and thermal impedance.

5.3.5.5.4 Heat sinks. The heat sink-to-ambient thermal impedance, $R_{\theta SA}$, is a function of the heat sink design. Copper and aluminum have the highest thermal conductivity (of the common metals) and are therefore usually preferred as heat sink materials.

An effective heat sink may take many forms. The simplest form is merely a flat, square plate of aluminum or copper 1/32- to 1/4-inch thick. This is mounted in the vertical plane to obtain the maximum "chimney effect." The term chimney effect, of course, relates to the natural convection type drafts set up as the air next to the surface of the heat sink is heated and rises, pulling more air past the heat sink surface. But even in these natural convection arrangements, it is possible to transfer heat away from the sink by radiation. This secondary transfer of heat by radiation can contribute appreciably to the total heat transfer, if the emissivity of the metal heat sink surface is increased by painting, oxidizing or anodizing (also see paragraph 5.3.5.5.5 Heat dissipation by radiation). Table VI shows some typical characteristics of simple, square plate heat sinks.

TABLE VI. Thermal characteristics of square plates

Sink Description	$R_{\theta SA}$, °C per Watt
5" x 5" x 1/8" Bright aluminum	4.3
5" x 5" x 1/8" Painted aluminum	2.5
7" x 7" x 1/8" Bright aluminum	2.7
7" x 7" x 1/8" Painted aluminum	1.6

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A further reduction in $R\theta_{SA}$ is obtained by using forced convection cooling. Figure 76 shows how increasing air velocity can markedly reduce the thermal impedance.

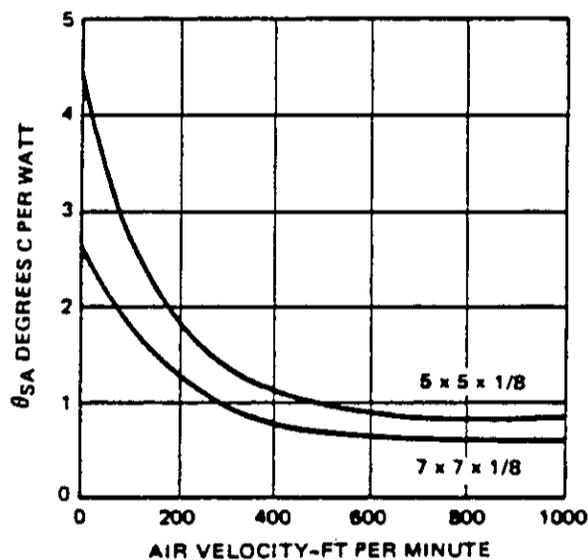


FIGURE 76. Sink-to-ambient thermal impedances for 1/8-inch thick square aluminum plates.

For applications where the thermal requirements are more critical, the simple flat plate sink may not be adequate. Die case and extruded aluminum sinks or assemblies of copper fins are available from many heat sink suppliers. They provide more efficient thermal paths and allow considerably more cooling capacity in a given space. The $R\theta_{SA}$ values for these improved heat sinks range from 3°C per watt to 0.3°C per watt for natural convection and as low as 0.1°C per watt for forced convection. Figure 77 shows how $R\theta_{SA}$ is related to overall volume for typical commercially available heat sinks. Additional information may be obtained from the heat sink manufacturers.

5.3 TRANSISTORS, HIGH POWER

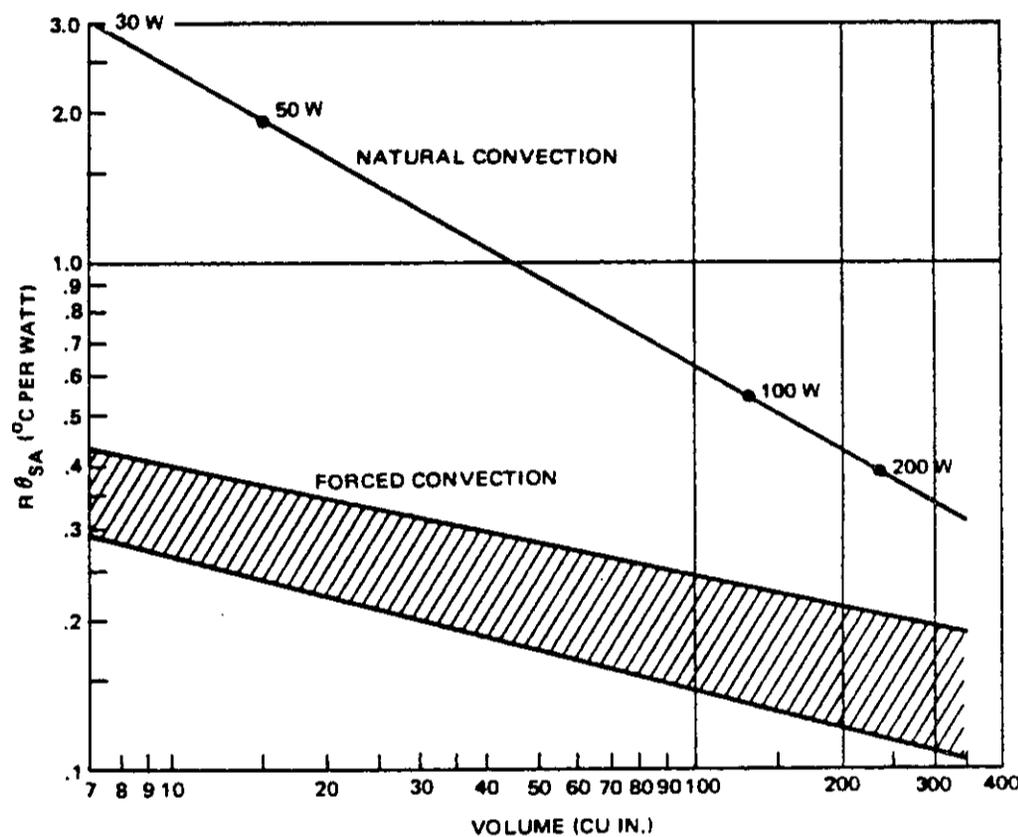


FIGURE 77. Thermal impedance, θ_{SA} versus volume (overall) of commercial heat sinks.

5.3.5.5.5 Heat dissipation by radiation. Radiation, the only mode of heat transfer that can take place across a vacuum, involves the transfer of photons from a hotter body, the emitter, to a cooler body, the absorber. An example is the heat one feels when standing next to a fireplace.

Opaque bodies absorb part and reflect the rest of the radiant energy which falls on the surface. The amount absorbed and reflected depends on the surface characteristics of the body, such as color and finish. Perfectly black bodies absorb all the radiant energy whereas perfectly shiny bodies reflect all energy. The radiation characteristics of a surface are characterized by a dimensionless quantity called the emissivity. A perfect absorber and emitter has an emissivity of unity whereas a perfect reflector has an emissivity of zero. Real bodies have emissivities between zero and one.

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The rate of heat transferred by radiation is low when the difference in temperatures between the emitting and absorbing bodies is small, say 20 °C or less, and the temperatures of the bodies are close to normal ambient temperature (25 °C). As shown in Figure 78, the thermal resistance decreases rapidly as the temperature difference increases, because the heat transfer rate is proportional to the quantity $T_E^4 - T_A^4$

where

T_E = the absolute temperature of the emitter
 T_A = the absolute temperature of the absorber

The absolute temperature is measured relative to absolute zero; in the metric system, it is measured K.

Guidelines for minimizing the thermal resistance to radiation are:

- a. High emissivity values for absorber and emitter
- b. Good view of the absorber by the emitter
- c. Larger absorber and emitter areas.

These guidelines are more important in high altitude (>70,000 ft) and a space environment where heat dissipation by natural or force convection is almost nonexistent.

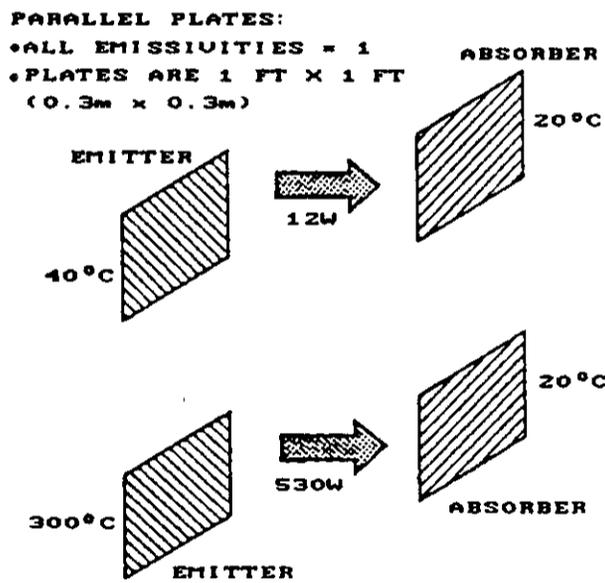


FIGURE 78. Rate of radiation heat transfer as a function of emitter and absorber temperatures.

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5.3.6 Environmental considerations. Refer to paragraph 5.1.6.2 Environmental considerations.

5.3.7 Reliability considerations. Refer to paragraph 3.1.6 Reliability consideration, for a complete list of failure mechanisms and application considerations.

5.4 TRANSISTORS, FIELD EFFECT

5.4 Field effect transistors (FET).

5.4.1 Introduction. The field effect transistor is a semiconductor device in which the flow of charge carriers is controlled by a charge on the gate electrode. Unlike conventional transistors which are bipolar devices (i.e., performance depends on the interaction of two types of carriers, holes, and electrons, field-effect transistors are unipolar devices (i.e., operation is basically a function of only one type of charge carrier, holes in p-channel devices and electrons in n-channel devices).

There are two basic types of FET devices, the junction FET (JFET) and the metal-oxide-semiconductor FET (MOSFET). They can be structured to operate in either the depletion-mode or enhancement-mode; however, enhancement-mode JFETs are almost nonexistent. Both JFETs and MOSFETs are available as either p- or n-channel devices.

5.4.1.1 JFET operation. Before considering the operation of a JFET, a review of pn junction behavior will be valuable. Figure 79 illustrates a basic silicon pn junction.

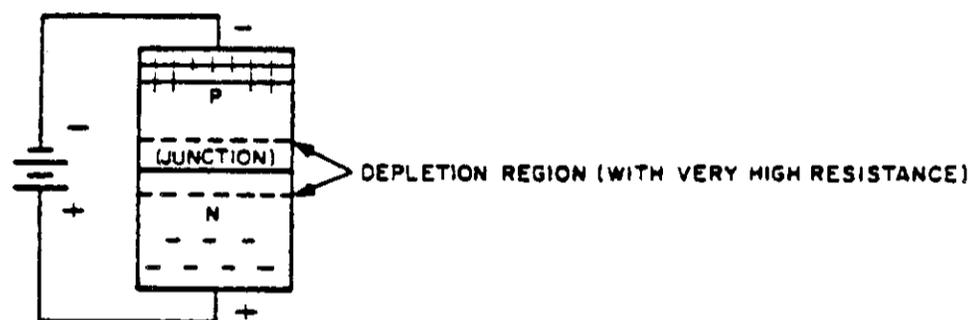
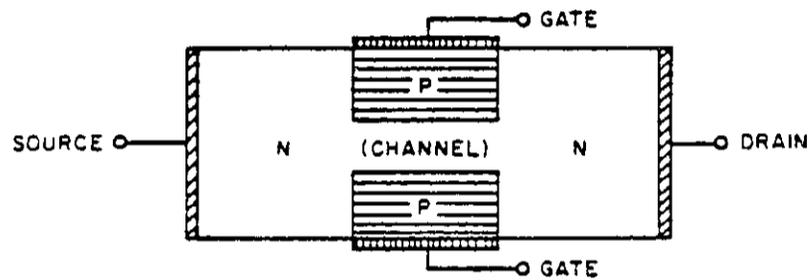


FIGURE 79. Basic silicon pn junction.

With the voltage connected as shown, the junction is reverse biased because the positive and negative carriers (holes and electrons) are attracted away from the junction. The empty area created at the junction is the key to JFET operation. This area is called the depletion region and contains neither holes nor free electrons in sufficient quantity for easy current flow; therefore, resistance is very high. Only a small leakage current flows.

The JFET can best be described as a bar of n-type silicon with two p-type regions diffused into it as shown in Figure 80.

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FIGURE 80. Physical configuration of a n-channel JFET.

The n-region between the two p-regions is called the channel. The source connection is so named because this is the source of the carriers which enter the device. The opposite connection is called the drain because this is where the carriers flow out. The gate terminals (which are normally connected together to one lead of the device or one external circuit connection) control the turn-on and turn-off.

In the configuration shown in Figure 81A, both junctions are reverse biased and a depletion region is created at the pn junction. If these regions extend across the channel and meet, a pinch-off condition is developed.

If a small positive voltage is applied on the drain as shown in Figure 81B, the two depletion regions are widened even more because of the reverse-bias voltage.

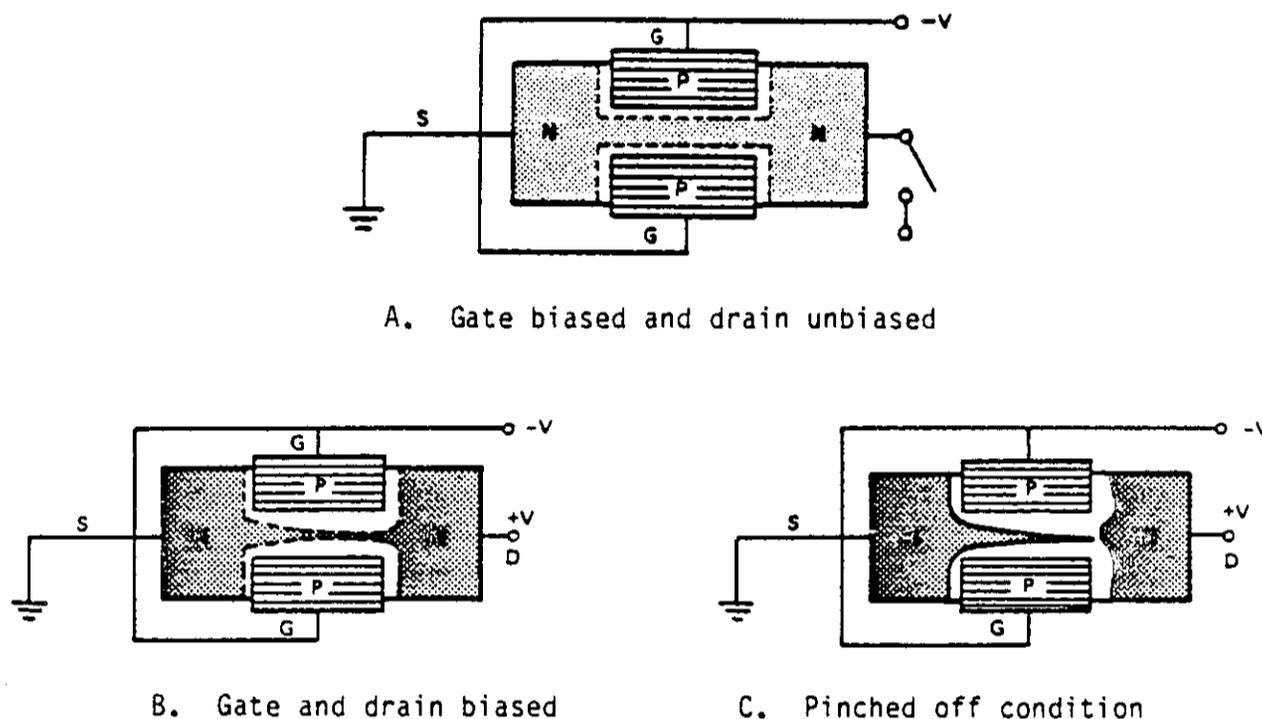
As reverse bias voltage is increased, the width of these depletion regions increases. If sufficient voltage is applied, the two depletion regions will extend completely through the channel and meet at one point, as shown in Figure 81C. In this condition, the channel for drain-source current (I_D) flow is pinched off.

The voltage value at which this condition occurs is a parameter normally specified on JFET data sheets as the gate-source cutoff voltage ($V_{GS(off)}$). Raising the voltage above $V_{GS(off)}$ will result in no appreciable decrease in I_D .

It can be seen from the above that a JFET transistor is a voltage-control device which can perform the complete switching function.

The operation of a p-channel JFET is the same, except that the polarities are reversed.

5.4 TRANSISTORS, FIELD EFFECT

FIGURE 81. Operation of an n-channel JFET.

5.4.1.2 MOSFET basic operation. The following paragraphs discuss the operation for both an enhancement-mode and a depletion-mode MOSFET device.

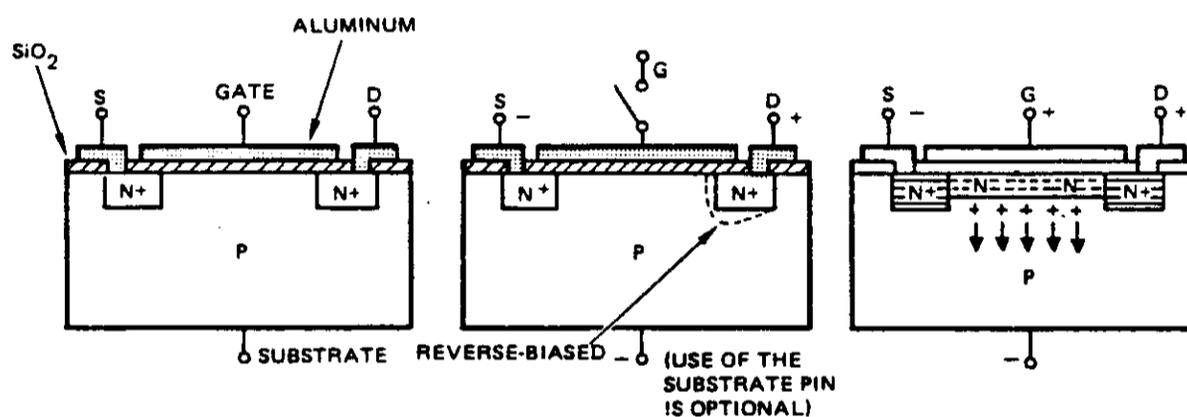
5.4.1.2.1 Enhancement-mode MOSFET. Figure 82A illustrates an unbiased enhancement-mode device.

An aluminum plate (gate) and the p-type substrate form equivalent plates of a capacitor, with the silicon dioxide acting as a dielectric. The other aluminum areas are used as bonding pads for the source and drain connections. If voltage is applied between the source and drain, only leakage current flows, because the pn junction will be reverse biased (gate open), as shown in Figure 82B.

In order to make the device conduct, a positive charge is applied to the gate. Because unlike charges attract, the available free electrons in the substrate are pulled up to the area under the silicon dioxide, as illustrated in Figure 82C. This concentration of electrons causes the normally p-type region to become n-type in a layer on the device between source and drain. This induced channel allows electrons to flow from source to drain. The gate-source voltage (V_{GS}) at which I_D barely starts to flow is called V_{GS} threshold (V_{GSth}). This configuration of a MOSFET, which uses the induced channel principle, is called an enhancement-mode MOSFET because the current flow is enhanced by the application of gate voltage.

5.4 TRANSISTORS, FIELD EFFECT

The operation of a p-channel enhancement-mode device is the same except that polarities are reversed.



A. Unbiased B. Drain biased C. Gate and drain biased

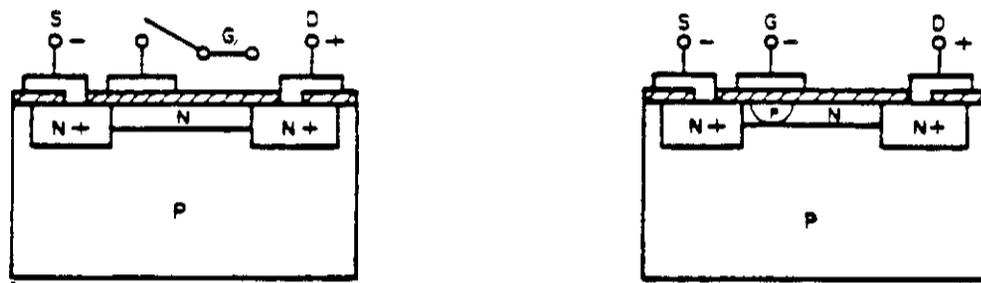
FIGURE 82. Operation of an n-channel enhancement-mode MOSFET.

5.4.1.2.2 Depletion-mode MOSFET. In this structure, a lightly doped n-channel is diffused into a p-material, as illustrated in Figure 82A. As a result, current will flow from source to drain with no voltage on the gate.

However, if a negative voltage is applied to the gate, the free electrons will be repelled out of the diffused channel, as illustrated in Figure 83B, creating a current-limiting condition similar to the no-gate-voltage condition of the enhancement-mode MOSFET described previously. The V_{GS} required to cutoff I_D is also called $V_{GS(off)}$ as in a JFET device. In fact, JFET and MOSFETs have very similar characteristics, because they are both depletion type devices.

Again the operation of a p-channel depletion-mode device is the same except that polarities are reversed.

5.4 TRANSISTORS, FIELD EFFECT



A. Drain biased

B. Gate and drain biased

FIGURE 83. Operation of an n-channel depletion-mode MOSFET.

5.4.2 Usual applications. There are three basic single-stage amplifier configurations for MOSFETs: common-source, common-drain, and common-gate. Each provides certain advantages in particular applications.

5.4.2.1 Common-source. The common-source arrangement shown in Figure 84 is most frequently used. This configuration provides a high input impedance, medium to high output impedance, and voltage gain greater than unity. The input signal is applied between gate and source, and the output signal is taken between drain and source. The voltage gain without feedback, A , for the common-source circuit may be determined as follows:

$$A = \frac{g_{fs} r_{os} R_L}{r_{os} + R_L}$$

where: g_{fs} is the gate-to-drain forward transconductance of the transistor

r_{os} is the common-source output resistance

R_L is the effective load resistance

The addition of an unbypassed source resistor to the circuit of Figure 84 produces negative voltage feedback proportional to the output current.

The voltage gain with feedback, A' , for a common-source circuit is given by

$$A' = \frac{g_{fs} r_{os} R_L}{r_{os} + (g_{fs} r_{os} + 1) R_s + R_L}$$

where R_s is the total unbypassed source resistance in series with the source terminal.

5.4 TRANSISTORS, FIELD EFFECT

The common-source output impedance with feedback, Z_0 , is increased by the unby-passed source resistor as follows:

$$Z_0 = r_{os} + (g_{fs} r_{os} + 1) R_s$$

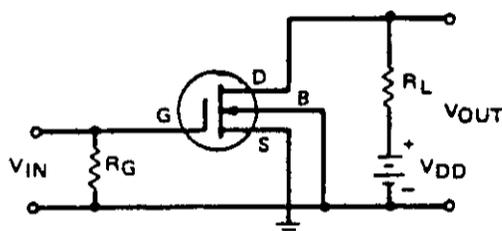


FIGURE 84. Basic common-source circuit for a MOSFET.

5.4.2.2 Common-drain. The common-drain arrangement, shown in Figure 85, is frequently referred to as a source-follower. In this configuration, the input impedance is higher than in the common-source configuration, the output impedance is low, there is no polarity reversal between input and output, the voltage gain is always less than unity, and distortion is low. The common-drain is used in applications which require reduced input-circuit capacitance, downward impedance transformation, or increased input-signal-handling capability.

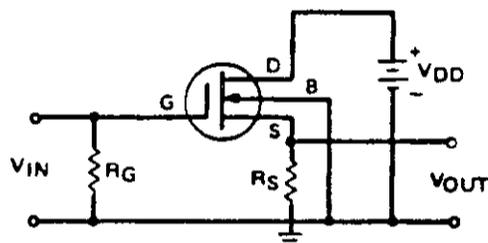


FIGURE 85. Basic common-drain circuit for a MOSFET.

5.4 TRANSISTORS, FIELD EFFECT

The gain A' of the common-drain circuit is given by

$$A' = \frac{R_s}{\frac{\mu + 1}{\mu} R_s + \frac{1}{g_{fs}}}$$

The amplification factor (μ) of a MOSFET is usually much greater than unity. The equation for the gain of a common-drain can be simplified as follows:

$$A' = \frac{g_{fs} R_s}{1 + g_{fs} R_s}$$

For example, if the gate-to-drain forward transconductance, g_{fs} , is 2000 microsiemens [2×10^{-3} siemen], and the unbypassed source resistance, R_s , is 500 ohms, the resulting stage gain, A' , would be 0.5. If the same source resistance is used with a transistor having a transconductance of 10,000 microsiemens [1×10^{-2} siemen], the stage gain would increase to 0.83.

If the resistor, R_g , were returned to ground, as shown in Figure 85, the input resistance, R_1 , of the source follower would equal R_g . However, if R_g were returned to the source terminal, the effective input resistance, R_1' , would be given by:

$$R_1' = \frac{R_g}{1 - A'}$$

where A' is the voltage amplification of the stage with feedback.

If the load resistance and the effective input capacitance, C_1' , of the common-drain are reduced by the inherent voltage feedback, the following equation holds:

$$C_1' = c_{gd} + (1 - A') c_{gs}$$

where c_{gd} and c_{gs} are the intrinsic gate-to-drain and gate-to-source capacitances, respectively.

The effective output resistance, R_0' , of the common-drain stage is given by the following equation:

$$R_0' = \frac{r_{os} R_s}{(g_{fs} r_{os} + 1) R_s + r_{os}}$$

where r_{os} is the transistor common-source output resistance
 g_{fs} is the gate-to-drain forward transconductance
 r_{os} is the common-source output resistance
 R_s is the source resistance, and r_0' is the effective output resistance.

5.4 TRANSISTORS, FIELD EFFECT

The common-drain output capacitance, C_0' , may be expressed as follows:

$$C_0' = c_{ds} + c_{gs} \left(\frac{1 - A'}{A'} \right)$$

where c_{ds} and c_{gs} are the intrinsic drain-to-source and gate-to-source capacitances, respectively, of the MOS transistor.

5.4.2.3 Common-gate. The common-gate circuit shown in Figure 86 is used to transform a low input impedance to a high output impedance. The input impedance of this configuration has approximately the same value as the output impedance of the source-follower circuit. The common-gate circuit is also a desirable configuration for high-frequency applications, because its relatively low voltage gain makes neutralization unnecessary in most cases. The common-gate voltage gain, A , is given by

$$A = \frac{(g_{fs} r_{os} + 1) R_L}{(g_{fs} r_{os} + 1) R_G + r_{os} + R_L}$$

where R_g is the resistance of the input-signal source.

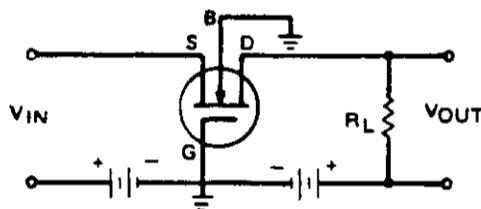


FIGURE 86. Basis common-gate circuit for a MOSFET.

5.4.3 Physical construction.

5.4.3.1 Die structure. Figure 87 illustrates the basic JFET structure, and Figure 88 shows the basic enhancement-mode MOSFET.

5.4 TRANSISTORS, FIELD EFFECT

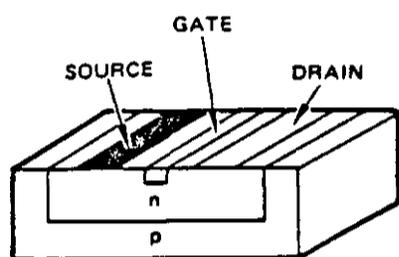


FIGURE 87. Structure of an n-channel JFET.

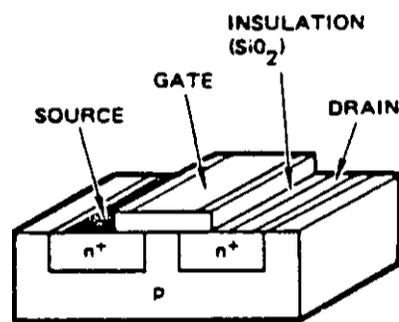


FIGURE 88. Structure of an n-channel enhancement-mode MOSFET.

5.4.3.2 Packaging. Some of the standard packages used for FET devices are the T0-5, T0-18, T0-39, and T0-72. Figure 89 shows the basic construction for a T0-5 device.

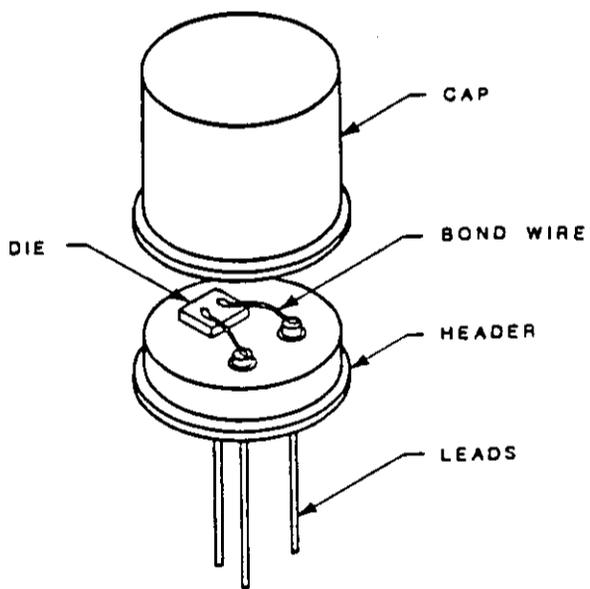


FIGURE 89. Typical T0-5 FET construction.

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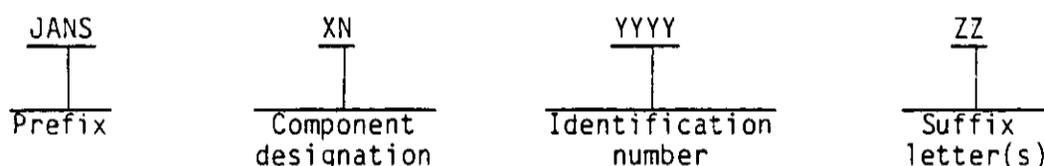
5.4 TRANSISTORS, FIELD EFFECT

Die bonding of these devices can be with either soft solder or gold eutectic. Both methods have been found to be reliable.

Wire bonding is done by using either aluminum or gold wires. The wire bonding methods most frequently used are thermocompression or ultrasonic bonding.

Sealing of the package is performed in an inert atmosphere by using electrical welding techniques; consequently, it is possible that weld splashes may occur and cause particles to be released inside the package. However, all MIL-STD-975 devices require 100 percent PIND testing and X-raying which would detect any defects.

5.4.4 Military designation. The military designation for transistors is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975.

The component designation is 2N for transistors.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices, suffix M, shorter or longer terminal leads, suffix S or L, or any other letter to indicate a modified version.

Radiation hardness assurance (RHA) designation code is placed after the prefix. See MIL-S-19500 for details.

5.4.5 Electrical characteristics. In both types of FETs, the operation is similar. Both rely on changing the charge on the gate electrode and the amount of charge in the conductive channel to change the drain current (see Figure 90).

The current-voltage relation of the FET is nearly the square law; it is

$$I_D = I_{DSS} \left[1 - \frac{V_{GS}}{V_{GS(off)}} \right]^2$$

5.4 TRANSISTORS, FIELD EFFECT

where

V_{GS} is the gate voltage

$V_{GS(off)}$ is the value of V_{GS} necessary to reduce the drain current to zero

I_{DSS} is the zero-gate-voltage drain current (saturation drain current)

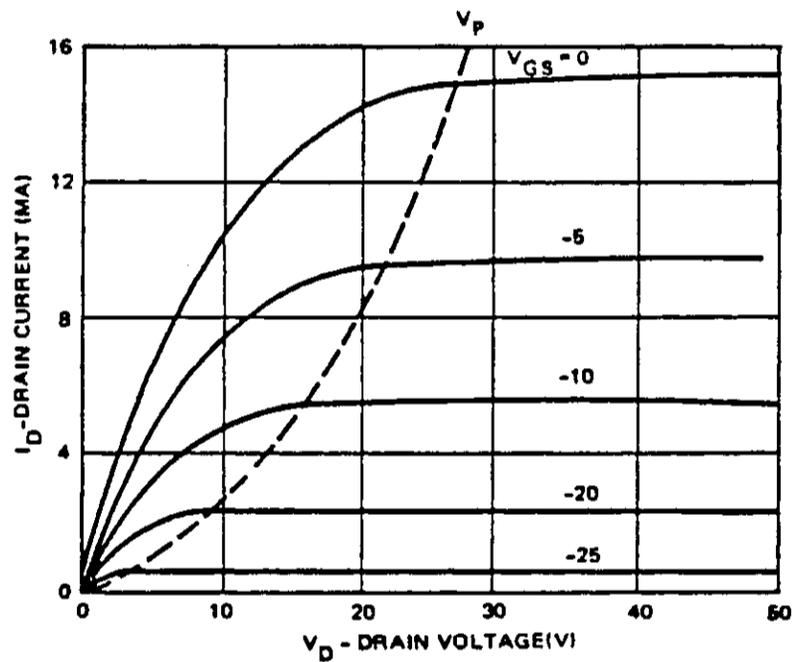


FIGURE 90. Typical output characteristics of a JFET.

The transconductance of the FET varies with gate voltage according to the relation:

$$\frac{\gamma I_D}{\gamma V_G} = g_m = g_{m0} \left[1 - \frac{V_{GS}}{V_{GS(off)}} \right]^{n-1}$$

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The basic current-voltage relationship for a depletion-mode MOSFET operating in the common-source configuration is shown in Figure 91. At low drain-to-source potentials and with the gate returned to source ($V_{GS} = 0$), the resistance of the channel varies linearly with voltage, as illustrated in region A-B. As the drain current increases beyond point B, the voltage drop in the channel produces a progressively greater voltage difference between the gate and points in the channel which are closer to the drain. As the potential difference between gate and channel increases, the channel is depleted of carriers (becomes constricted), therefore the drain current increases at a much slower rate with further increase in drain-to-source voltage, as shown in region B-C. An additional increase in drain-to-source voltage beyond point C produces no change in drain current until point D is reached. This condition leads to the description of region B-D as the "pinch-off" region. Beyond point D, the transistor enters the "breakdown" region and the drain current may increase dramatically. The upper curve in Figure 91 also applies to enhancement-mode MOSFETs, provided that the gate voltage V_{GS} is large enough to produce channel conduction.

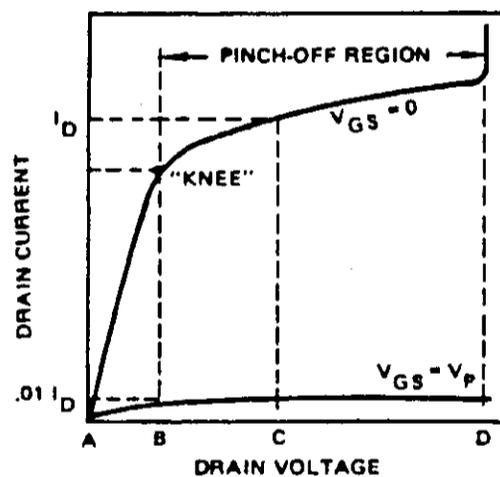


FIGURE 91. Basic current-voltage relationship for a depletion-mode MOSFET.

A MOSFET channel may achieve self pinch-off by the intrinsic IR drop alone, by a combination of intrinsic IR drop and an external voltage applied to the gate, or by an external gate voltage alone, which has the same magnitude as the self pinch-off IR drop V_p . In any case, channel pinch-off occurs when the sum of the intrinsic IR drop and the extrinsic gate voltage reaches V_p . The pinch-off voltage V_p is usually defined as the gate cutoff voltage $V_{GS(off)}$ that reduces the drain current between 0.1 and 1 percent of its zero-gate-voltage value at a specified drain-to-source voltage (this corresponds to the "knee" voltage, point B in Figure 91, of the zero-gate-voltage output characteristics).

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The pinch-off region between points B and D in Figure 91 is where MOS FET transistors are especially useful as high impedance voltage amplifiers. In the ohmic region between points A and B, the linear variation in channel resistance makes the device useful in voltage-controlled resistor applications, such as the chopper unit at the input of some dc amplifiers.

Typical output-characteristic curves for n-channel MOSFETs are shown in Figure 92. (For p-channel MOSFET, the polarity of the voltage and current are reversed.) In the pinch-off region, the dynamic output resistance, r_{os} , of the transistor may be approximated from the slope of the output-characteristic curve at any given set of conditions.

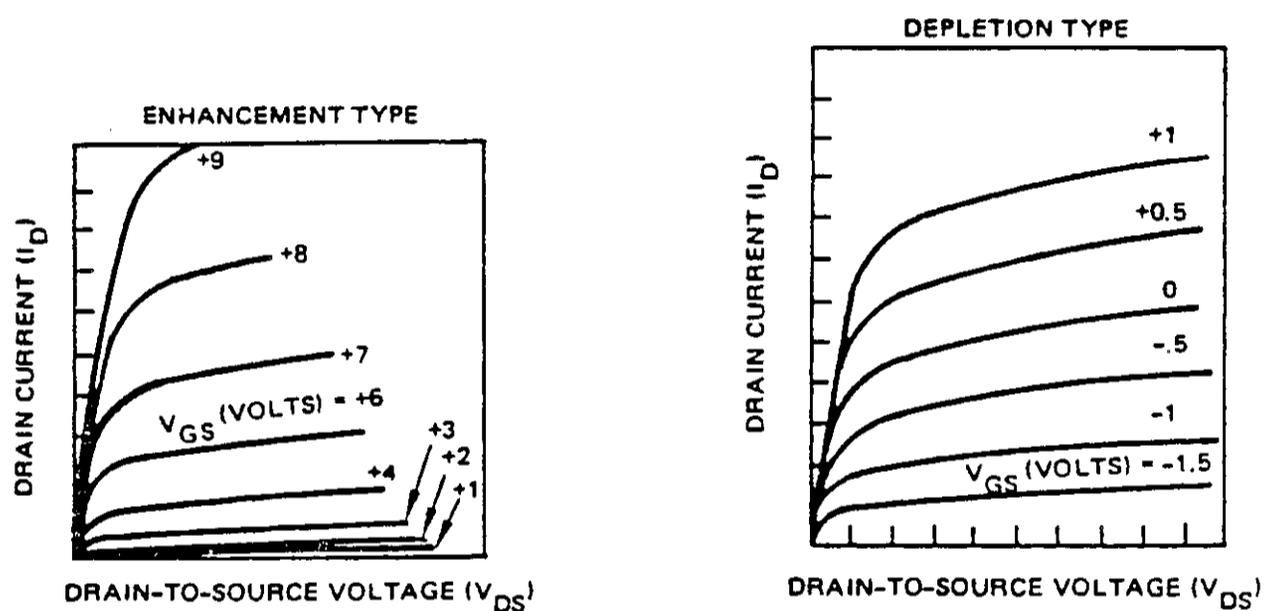


FIGURE 92. Typical output characteristics curves for a n-channel MOSFET.

Typical transfer characteristics for n-channel MOSFETs are shown in Figure 93 (polarities would be reversed for p-channel devices). The threshold voltage shown in Figure 93 is an important parameter for enhancement mode MOSFETs, because it provides a desirable region of noise immunity for switching applications.

5.4 TRANSISTORS, FIELD EFFECT

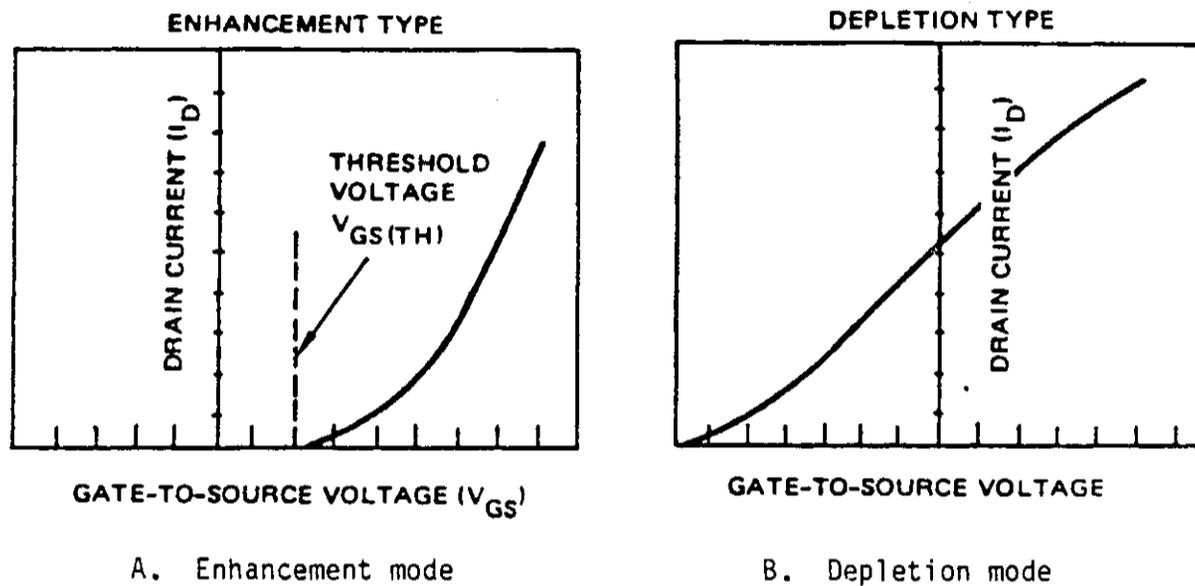


FIGURE 93. Typical transfer characteristics for a n-channel MOSFET.

5.4.6 Environmental considerations. Refer to paragraph 5.1.6.3, Environmental considerations in the General section.

5.4.7 Reliability considerations. The failure modes seen in FETs are also commonly seen in bipolar transistors. These failure modes are discussed in paragraph 5.1.6, "Reliability considerations" in the General section.

An additional defect may arise if caution is not exercised when applying voltage to the MOSFET device. Performance of MOS transistors depends on the relative perfection of the insulating layer between the control electrode (gate) and the active channel. If this layer is punctured by an inadvertent application of excess voltage to the external gate connection, the damage is irreversible. If the damaged area is relatively small, the additional leakage may not be noticed in most applications. However, greater damage may degrade the device to the leakage levels associated with JFET transistors. Therefore, it is very important that appropriate precautions be taken to ensure that the MOSFET gate-voltage ratings are not exceeded.

5.5 TRANSISTORS, OPTOCOUPPLERS

5.5 Optocouplers.

5.5.1 Introduction. Optically coupled isolators, also called optocouplers, are used to isolate electrical systems from each other in an electronic circuit. Optocouplers allow very good circuit control with a high degree of electrical isolation between the input and output. These isolators are ideally suited for eliminating problems such as ground loop isolation, common mode noise rejection, and electromagnetic interference. These devices replace mechanical relays and pulse transformers. Optocouplers provide electrical isolation of potentially dangerous voltages in the equipment.

An optically coupled isolator consists of a light source coupled through an optically transparent insulation to a light detector and is housed in a light-excluding package. The light source may be an incandescent or neon lamp, or a light-emitting diode (LED). The transparent insulation may be air, glass or plastic. The detector may be a photoconductor, photodiode, phototransistor, photo silicon controlled rectifier (SCR), photo Darlington, or an integrated combination photodiode/amplifier. The discussion will be limited here to optocouplers having infrared-emitting diodes (IRED) coupled to a solid state photo-detector. Diagrams of some basic optically coupled isolators are shown in Figure 94.

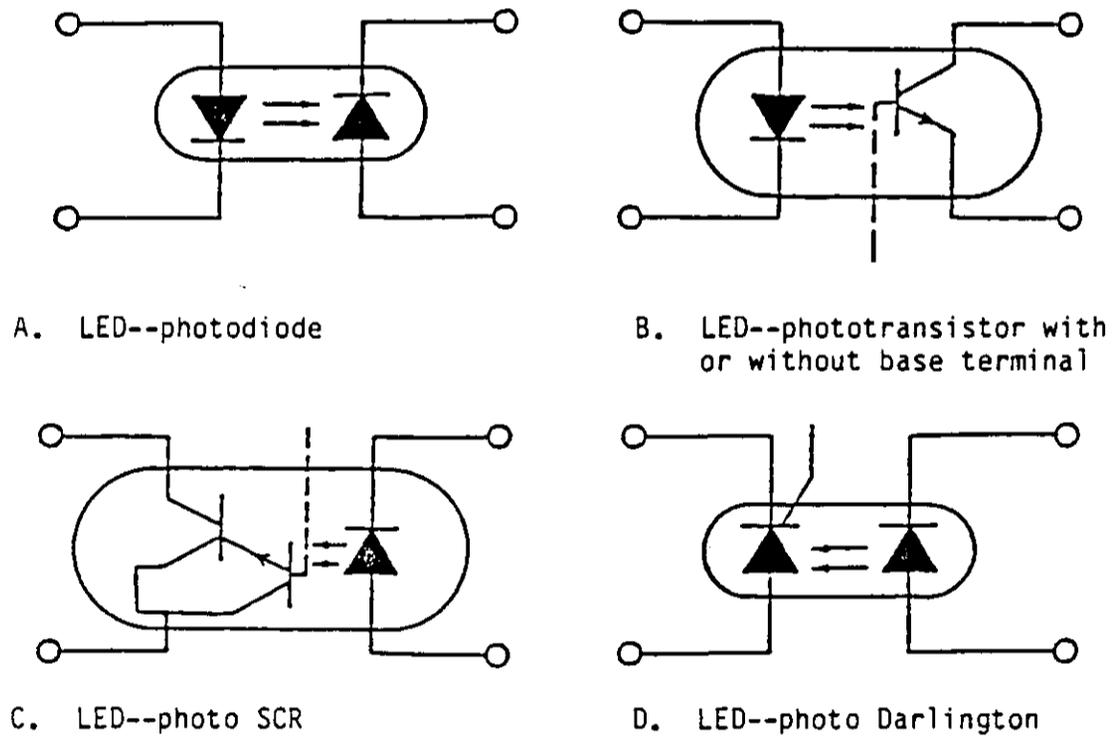


FIGURE 94. Basic types of optically coupled isolators.

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Because optocouplers have no external optical properties, they are specified as electrical devices. Two functional characteristics that define an optically coupled isolator are how well they transfer information from input to output and how efficiently they maintain electrical isolation from input to output. Some important optocoupler parameters that determine these two characteristics are current transfer ratio (CTR), isolation resistance, isolation capacitance, and isolation voltage.

5.5.1.1 Current transfer ratio. The current transfer ratio, which is a measure of the optocoupler efficiency, is the ratio of the output current over the input current and is generally expressed as a percentage. It is dependent upon the radiative efficiency of the IRED, as well as the area, sensitivity, and amplifying gain of the detector.

5.5.1.2 Isolation resistance. This is the dc resistance between the input and the output of the optocoupler. Typically, the isolation resistance is very high and has a value of about 100 to 1000 G Ω (10^{11} to 10^{12} Ω).

5.5.1.3 Isolation capacitance. Since the optocoupler consists of two semiconductor materials (emitter and photodetector) separated by a transparent insulating material, this results in an isolation capacitance between the input and output. Typical values of isolation capacitance range from 0.5 to 3.0 pF.

5.5.1.4 Isolation voltage. One of the primary functions of an optically coupled isolator is to provide electrical isolation between the input and output circuits. The isolation rating refers to the maximum voltage difference that can be safely applied across the device without danger of insulation breakdown.

5.5.2 Usual applications. Optically coupled isolators are used to transmit information between electrically isolated electronic circuits. Typically, this isolation has been provided by relays, isolation transformers, line drivers, and receivers. However, if an optically coupled isolator is utilized, it will provide certain unique advantages over its mechanical counterparts.

- a. Smaller size
- b. Wide operating range
- c. High gain
- d. Faster operating speeds
- e. High-voltage electrical isolation
- f. Compatibility with circuits.

Some typical circuits and systems applications of optically coupled isolators follow:

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5.5.2.1 Line receivers. An optically coupled isolator can provide line isolation between two systems coupled by a transmission line. Figure 95 shows a typical interface system using TTL integrated circuitry coupled by a twisted pair line.

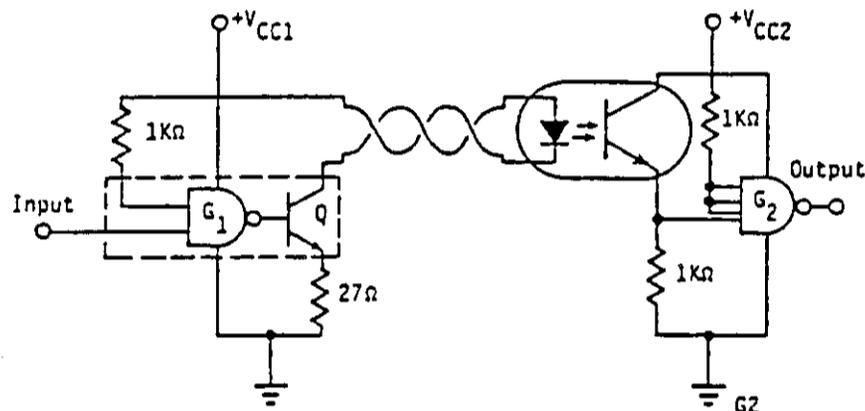


FIGURE 95. Typical interface system with TTL-integrated circuitry, twisted line pair, and optical coupler.

Gate G_1 and transistor Q constitute the input stage which drives the transmission line and emitter of the optically coupled isolator. At the receiving end of the line, the phototransistor is coupled to a fast switching gate (G_2) for fast pulse generation. In this system the optically coupled isolator provides isolation for both noise generated by electromagnetic interference and high voltage differences between the input and output.

An isolation standard transformer or relay can also accomplish this, but it would not be as fast as an optically coupled isolator. A standard line driver and receiver combination can eliminate noise and increase speed, but cannot provide isolation for high voltage differences between input and output. Digital input units containing optical isolators protect the microcomputer from standard as well as over voltage conditions.

5.5.2.2 Solid state relays. Optically coupled isolators can eliminate undesirable relay noise and spikes in relay circuits by providing a high degree of isolation between the input and output. A block diagram of a typical solid-state relay using an optically coupled isolator is shown in Figure 96.

The control stage consists of discrete transistors or integrated circuits whereas the output stage consists of high-power switching devices.

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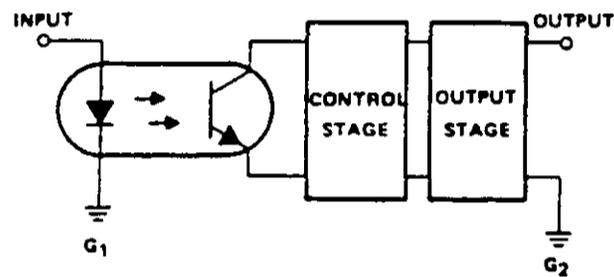


FIGURE 96. Typical solid state relay using an optically coupled isolator.

5.5.2.3 Digital logic interface. A very useful application of optically coupled isolators is interfacing between digital systems. An optically coupled isolator has output currents compatible with TTL inputs. It is beneficial to use this device when high voltage differences exist between systems. Other important applications of an optically coupled isolator are those involving 54/74 TTL and similar digital integrated circuit families. There is a wide variety of standard TTLs with different logic levels or logic conditions. A general interface circuit using an optocoupler is shown in Figure 97.

When the output of a logic circuit 1 (Gate 1) is low (V_{OL1}), the output of the optically coupled isolator is also low (V_{OL2}). Since V_{OL2} is the input to logic circuit 2 (Gate 2), it must be less than the maximum required low input voltage (V_{IL2}), to hold logic circuit 2 (Gate 2) in a stable state as shown below.

$$V_{OL2} (\text{isolator}) < V_{IL2} (\text{max})$$

Optically coupled isolator specifications should be evaluated to be level compatible with the TTL logic. The R_1 and R_2 values can be calculated using the following equations:

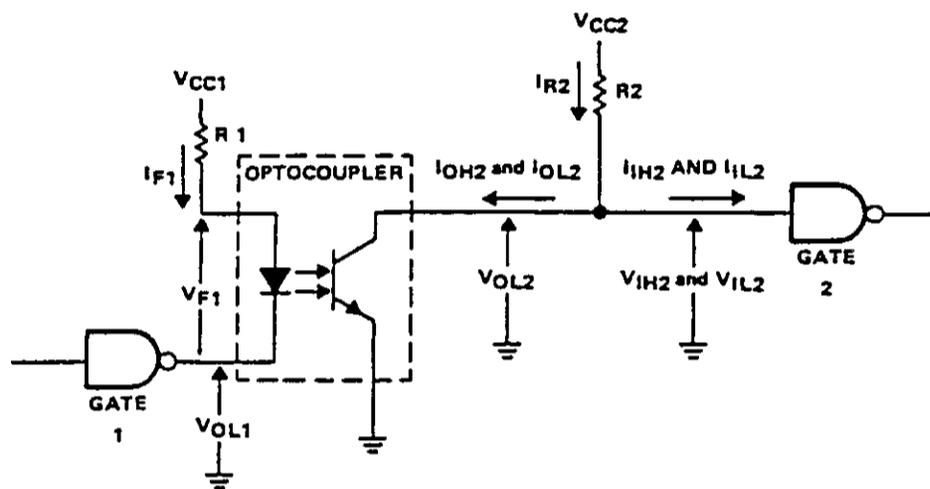
$$R_1 = \frac{V_{CC1} - V_{F1}(\text{typ}) - V_{OL1}(\text{typ})}{I_{F1}(\text{typ})}$$

$$R_2 (\text{min}) = \frac{V_{CC2}(\text{max}) - V_{OL2}(\text{max})}{I_{OH2}(\text{max}) + I_{IH2}(\text{max})}$$

$$R_2 (\text{max}) = \frac{V_{CC2}(\text{min}) - V_{IH2}(\text{min})}{I_{OH2}(\text{max}) + I_{IH2}(\text{max})}$$

R_2 is selected between the limits of $R_2(\text{min})$ and $R_2(\text{max})$.

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NOTE: V_{OL2} = low-level output voltage of coupler when coupler is on
 V_{IL2} = low-level input voltage specified for Gate 2.

FIGURE 97. Optocoupler interface circuit.

5.5.2.4 Pulse amplifiers. Pulse amplification, as well as isolation, can be achieved by using an optically coupled isolator with a pulse amplifier. This circuit is shown in Figure 98, which illustrates an optical isolator used with a UA741 operational amplifier to amplify the pulse appearing at the anode of the IRED. The gain of this circuit is controlled by the feedback resistor R_F .

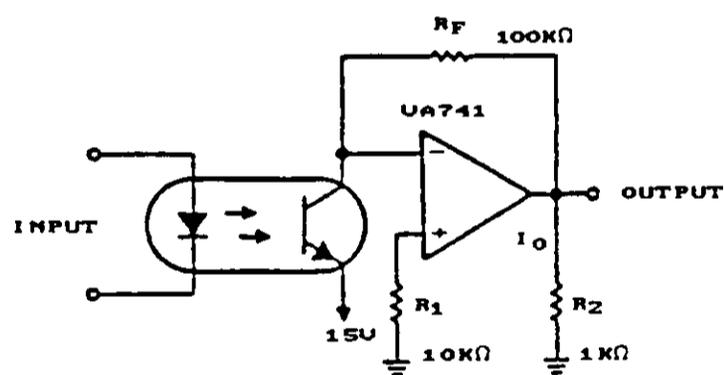


FIGURE 98. Isolated pulse amplifier using optically coupled isolator and operational amplifier.

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5.5.3 Physical construction, packaging configurations, and device combinations. An optically coupled isolator contains both an IRED and a photodetector in the same package with a coupling medium of infrared-transmitting glass or a silicone rubber material. Three major packages are widely used:

- a. Transistor can packages
- b. Ceramic DIP packages.

The construction of each is described below.

5.5.3.1 Transistor package. One effective construction approach is shown in Figure 99. The phototransistor and the LED are separated by a piece of glass, which serves as a dielectric and light pipe. The close proximity of about 5 mils of the two active elements allows a high CTR, about 200 percent, and an isolation voltage of about 1.5 KV.

Exact positioning of the elements or of the case is not required because the phototransistor is designed to have a base area larger than the LED. This gives a bigger target for the LED during bonding and makes its placement less critical.

The package is assembled by attaching the phototransistor to a transistor-type header. A thin layer of clear silicone resin is applied and then covered by a thin piece of glass. The partial assembly is baked for about an hour to cure the epoxy resin. Another thin layer of clear silicone epoxy resin is applied to the glass and the LED is positioned in it. The assembly, with LED, is again baked for curing. At this time, the wire bonding operation takes place. Once the wire bonding is completed, the entire assembly is coated with the clear resin and cured. The last step is to attach the cover to the header. The overall case outline and internal connections are shown in Figure 100.

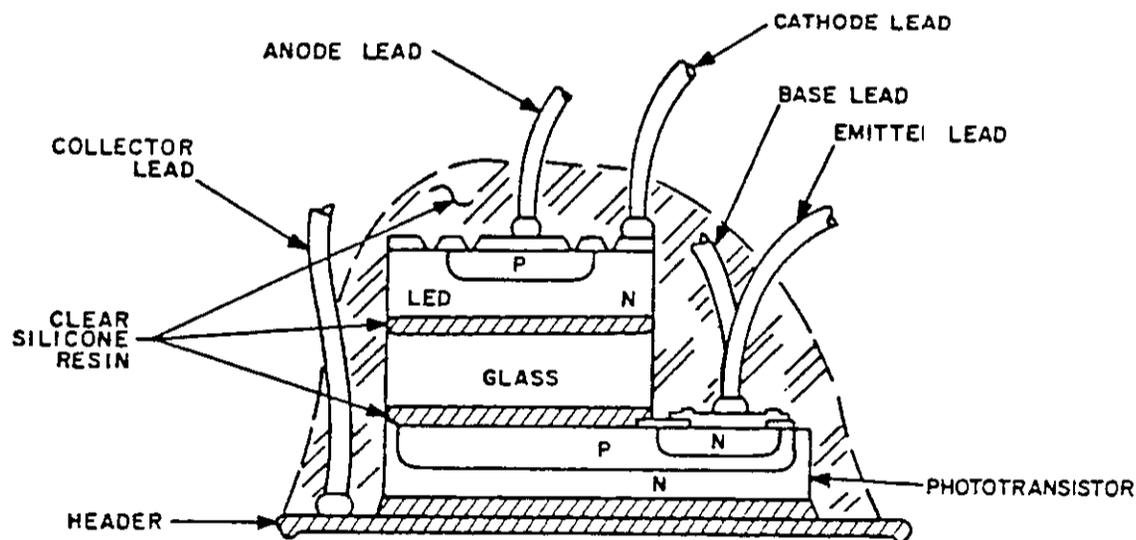
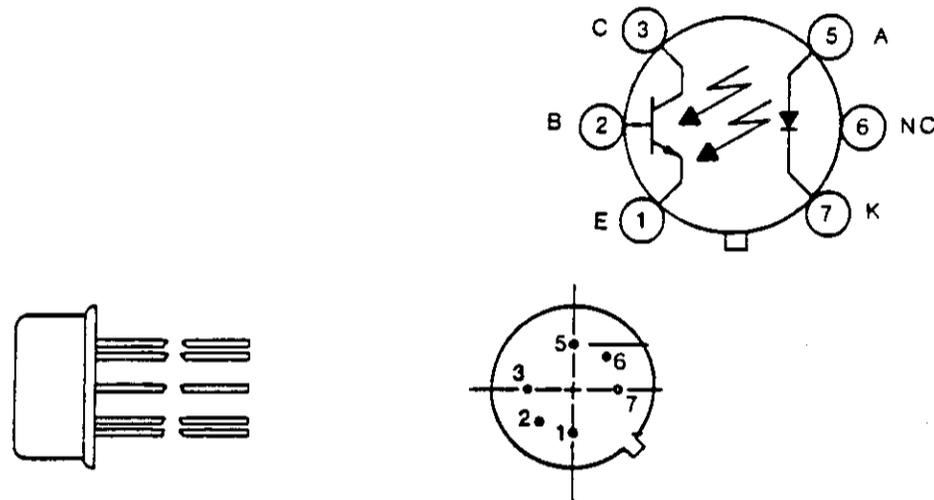


FIGURE 99. A typical phototransistor package.
SEE NOTE PAGE 5-118

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FIGURE 100. A typical transistor case outline.

5.5.3.2. High reliability ceramic package. Some of the high reliability optocouplers are packaged in a 16-pin ceramic package and hermetically sealed. Figure 101 shows the mechanical construction of a 16-pin hermetically sealed, optically coupled isolator.

Silicon detectors are die-attached and wire bonded inside a 16-pin ceramic package. GaAsp emitters (LEDs) are die-attached and wire bonded to a separate ceramic insert. Solder preforms are applied between the ceramic package and the insert and soldered in place. The assembly is then potted with transparent insulating material to improve the optical coupling and electrical insulation between the emitter and detector. Finally, a metal lid is attached to the package to insure a hermetic seal. The finished optically coupled isolator can withstand storage temperatures from -65 to $+150$ °C, operating temperature from -55 to $+125$ °C and 98 percent relative humidity at 65 °C without failure. Some manufacturers of optocouplers use an organic polymer (black-pigmented silicon rubber) to surround the light channels to isolate the channels in their multichannel devices to reduce cross talk. However, this imposes a limitation on constant acceleration testing to 5000 g, which might cause lateral movement of the pigment and reduce the current transfer ratio.

Note:

For those applications which are of a critical nature, opto devices containing opaque or transparent conformal materials should not be used. These coatings may have inherent problems when exposed to radiation and temperature extremes, such as outgassing and physical and/or chemical changes. Only hermetically sealed devices should be used.

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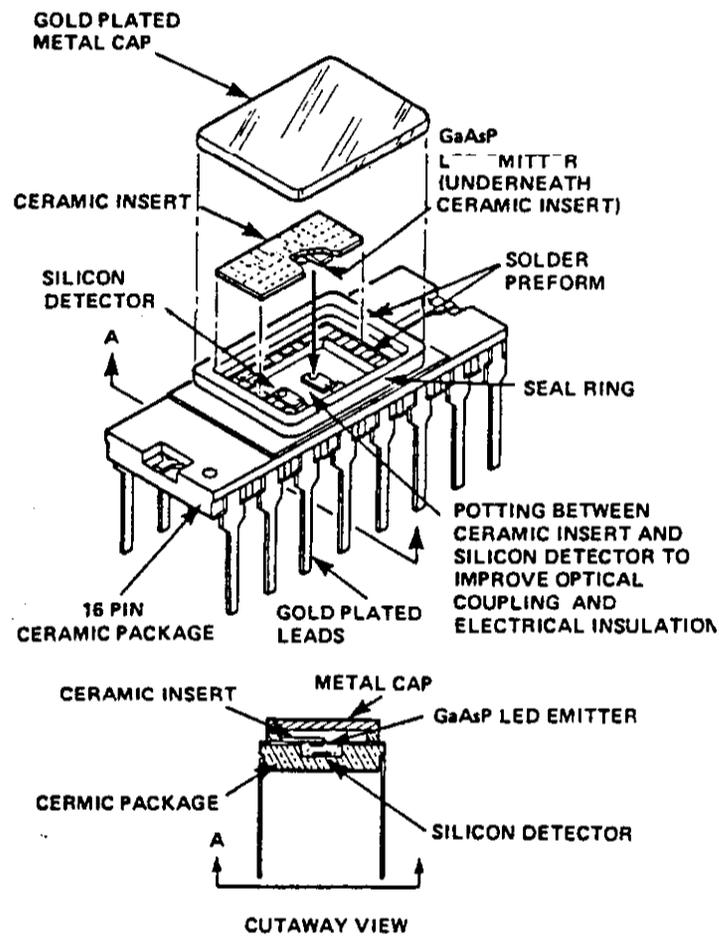
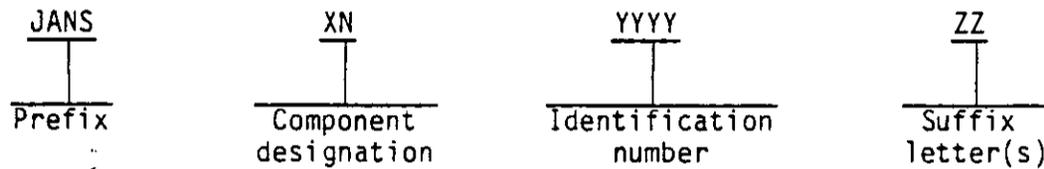


FIGURE 101. Mechanical construction of an optically coupled isolator in a hermetically sealed 16-pin ceramic package.

5.5.4 Military designation. The military designation for optocouplers is formulated as follows:



MIL-HDBK-978-B (NASA)

5.5 TRANSISTORS, OPTOCOUPERS

The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JAN TXV or JAN S in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is 4N and 6N for optocouplers.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix M), longer or shorter polarity (suffix L or S), or any other letter to indicate a modified version.

5.5.5 Electrical characteristics. One of the most widely used general purpose optically coupled isolators uses a gallium arsenide light-emitting diode as the input and a silicon phototransistor as the output. It has a useful current transfer ratio (CTR) and is reasonably fast. Photo-Darlington outputs have higher CTR, but are slower. These two types are useful, but the collector-emitter breakdown voltage is limited to 25 to 50 V. By using a photo silicon-controlled rectifier, reverse voltages up to 400 V may be obtained. The speed of the photo SCR is between that of a phototransistor and a photo-Darlington. The fastest optocouplers use photodiodes in conjunction with a transistor amplifier or a logic gate to supply the output of the optically coupled isolator. Photodiodes are not used alone because they have very low CTR of about 0.1 percent.

Since the LED/phototransistor optocouplers are most widely used, these devices will be discussed in detail and a comparison of optocouplers using different output devices will be provided. The LED/phototransistor optocoupler consists of a GaAs LED and a silicon phototransistor.

5.5.5.1 LED characteristics. For most applications, the basic LED parameters I_F and V_F are sufficient to define the input. Figure 102 shows these forward characteristics, providing the necessary information to design an LED drive circuit.

Another important LED characteristic is the luminous intensity or the power output (P_O). At very low values of I_F , no radiative current predominates and, hence, very little or no light is emitted. As I_F increases, the radiative current begins to flow and increases faster than the non-radiative portion. This relationship allows a fairly linear relationship between I_F and luminous intensity for a range of forward currents. Eventually, the nonradiative current rate increases faster and the luminous intensity levels off as the radiative current virtually stops increasing.

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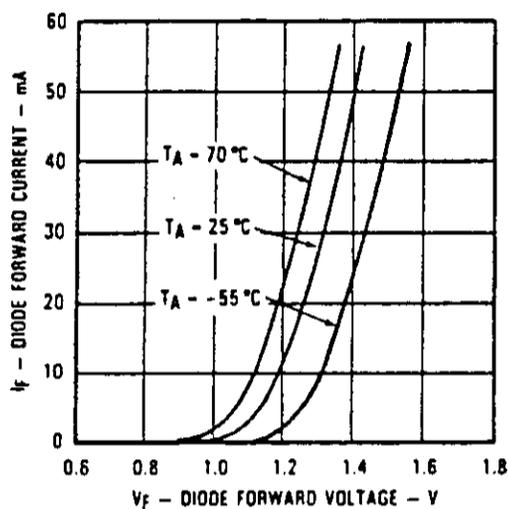


FIGURE 102. Diode forward current vs forward voltage.

Another important factor to consider is that the output of an LED decreases with time. Lifetime is often expressed as the time required for the output to fall to 50 percent of its original value. The typical lifetime may be 10,000 to 100,000 hours of operation. The LED degradation mechanisms will be discussed in paragraph 5.5.7, Reliability considerations.

Typical electrical characteristics of LEDs used in optically coupled isolators are forward voltage (V_F) at $I_F = 10$ mAdc, ranging from 0.8 to 1.5 V, and maximum reverse leakage current of about 100 μA at $V_R = 2.0$ Vdc).

Although the specifications do not call out any optical characteristics, the LED and the photodetector must have matching spectral characteristics to optimize the operation of the optically coupled isolators. Typically, when forward current (I_F) is passed through a GaAs LED, the optocoupler emits infrared radiation that peaks at a wavelength around 900 nanometers. The sensitivity of the photodetector should peak at the same wavelength for maximum signal coupling. The detector material (which is silicon) also peaks at this particular wavelength.

5.5.5.2 Phototransistor characteristics. The output of the optically coupled isolator is typically a phototransistor. The radiant energy from the LED falls on the base surface of the phototransistor. Phototransistors are designed to have a very large base region and, therefore, a very large collector base area with a small emitter. The incident energy, in the form of photons, creates electron-hole pairs in the base region and causes current to flow in the circuit.

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In most common test circuits the base is left open, the emitter is grounded, and a positive voltage is applied to the collector. However, the base lead is left available for unique circuit applications.

Typical phototransistor characteristics are listed in Table IX.

5.5.5.3 Coupled characteristics. Because the output stage of an optically coupled isolator is a phototransistor, the on-state parameters to be specified are $V_{CE(sat)}$, $I_C(on)$, and $I_B(on)$. Additional parameters that fully characterize the optically coupled isolator are current transfer ratio (CTR), isolation voltage, isolation resistance, and isolation capacitance. These additional characteristics are listed in Table VII.

TABLE VII. Typical isolator characteristics

Device Type	Current Transfer Ratio (CTR) (%)	Isolation Voltage V_{iso} (V)	Isolation Resistance (Ohms)	19500 Slash Sheet
4N22	25	1000	10"	486
4N23	60	1000	10"	486
4N24	100	1000	10"	486
4N47	50	1000	10"	548
4N48	100	1000	10"	548
4N49	200	1000	10"	548

NOTE: This table is not to be used for part selection; instead, use MIL-STD-975.

5.5.5.4 Switching characteristics. Switching characteristics are important for circuit applications and are usually specified by rise time (t_r) and fall time (t_f).

TABLE VIII. Switching characteristics

Device Type	Maximum Rise Time t_r (μs)	Maximum Fall time t_f (μs)	Comments
4N22, 4N23, 4N24	20	20	
4N47, 4N48, 4N49	25	25	Phototransistor mode
4N47, 4N48, 4N49	3	3	Photodiode mode

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5.5.5.5 Optically coupled isolator characteristics. The current transfer ratio (CTR) and isolation voltage are the characteristics unique to an optically coupled isolator. The characteristics of 4N22, 4N23, 4N24, 4N47, and 4N48 are discussed in paragraph 5.5.5.3. All these devices are included in the qualified parts list of MIL-S-19500/486 and /548.

However, various types of optocouplers that meet special application requirements are available. Comparisons of the characteristics of various isolators are listed in Table IX.

TABLE IX. Comparison of different optically coupled isolators

Output Device <u>1/</u>	CTR	Speed	CISO, RISO	Voltage Isolation	Package
Photo-transistor	10% to 100% min	1 - 10 μ s	1 - 3 pF; 10 - 100 G Ω	1 - 5 kV	DIP, metal can
Photo-Darlington	100% to 500% min	50 - 200 μ s	1 - 3 pF; 10 - 1000 G Ω	1 - 5 kV	DIP, metal can
Photo SCR	Approx. 20 mA trigger current	2 - 20 μ s	1 - 3 pF; 10 - 1000 G Ω	1 - 2.5 kV	DIP
Photodiode and transistor amplifier	7% - 400% min	0.5 - 60 μ s	--	1 - 2.5 kV	DIP
Photodiode and logic gate	400% - 600% min	t _{plh} , t _{pfl} 50 - 100 ns	--	1 - 2.5 kV	DIP

1/ All the devices use LEDs for the input.

5.5.6 Environmental considerations. Typical environmental conditions and screens that the optically coupled isolators are capable of withstanding are not substantially different from those given in paragraph 5.1.6.3 of Screening procedure in subsection 5.1, Transistors, general. However, caution should be observed when selecting test conditions for constant acceleration and vibration, so that the dielectric resin material is not affected.

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5.5.7 Reliability considerations.

5.5.7.1 Failure modes. The optically coupled isolator differs from standard semiconductor devices in that it contains a GaAs/GaAsp LED chip and a silicon npn phototransistor chip, coupled with a light transmission medium. The major failure modes are CTR (current transfer ratio) degradation and the isolation voltage breakdown. The CTR degradation occurs due to light-emitting diode and phototransistor degradation. The isolation voltage breakdown is related to the light transmission medium and package-related defects.

5.5.7.2 Failure mechanisms.

5.5.7.2.1 Light-emitting diode (optocoupler input). Most current transfer degradation is due the reduction in the efficiency of the light-emitting diode within the optocoupler. The LED current consists of two components, a diffusion current and a space-charge recombination current.

$$I_F (V_F) = A_e \frac{qV_F}{KT} + B_e \frac{qV_F}{2KT}$$

where

- A is the diffusion current component
- B is the recombination current component
- q is the electron charge
- K is the Boltzman constant
- T is temperature in K
- V_F is the forward voltage

The diffusion current component is the radiative or light-emitting current and the recombination current is nonradiative. As the LED ages during use due to an increase in the value of B, the recombination current increases, then the radiative current will decrease for a fixed total LED current, causing a reduction in the light output of the LED. The specific reasons for the increase in the recombination current are not fully understood.

Dark-line defects are the dominant physical degradation mechanism of a light-emitting diode. Material or crystal imperfections act as growth sites for dislocation lines under forward bias conditions. The dislocation lines appear as dark regions giving rise to the term "dark line defects." As the dark line grows the light output progressively decreases.

Increasing the device temperature or operating current also accelerates the degradation rate. The rate depends on the device structure, assembly of the LED chip into the package and package thermal resistance. Impurities in the chip due to process contamination, exposed junction and metallization processes can also cause LED degradation.

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A typical CTR degradation with time is represented in Figure 103.

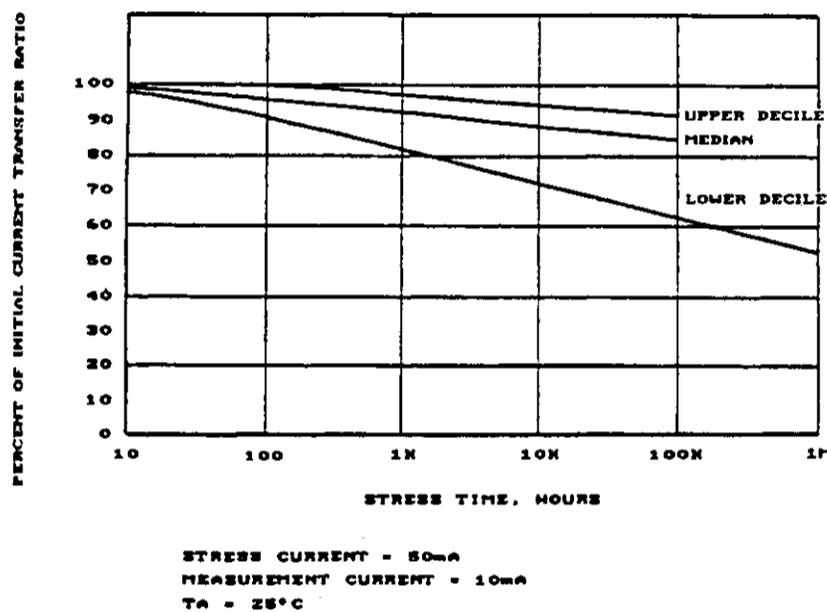


FIGURE 103. Current transfer ratio vs time.

In summary, the LED degradation depends on the following factors:

- a. Quality of starting materials (wafer with few crystal dislocations and processed by the liquid phase epitaxial method).
- b. Clean processing
- c. LED structure or design
- d. Metallization process
- e. Operating temperature
- f. Operating current.

5.5.7.2.2 Detector degradation. Although the detector has less influence than the light emitting diode, detector stability is very important. In a photo-transistor the failure modes are h_{FE} instability, increase in off state leakage current, and drop in breakdown voltages. The leakage instability is caused by the formation of the surface inversion channel in the base region of the

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phototransistor. The effect of this surface inversion can be remedied by subjecting these devices to a thermal anneal without bias. The surface inversion channel is considered to originate from ionic contamination in the silicon polymer material that covers the phototransistor. The large base area of the phototransistor is very sensitive to such contamination. The leakage current (I_{CEO}) on a phototransistor increases rapidly when subjected to high temperature reverse bias (collector-base), when such contamination is present.

Suitable design and processing of the phototransistor with appropriate passivation can be used to produce a stable and reliable detector.

5.5.7.2.3 Mechanical failure mechanisms. The failure modes are generally opens or shorts. Assembly-related failure mechanisms are:

- a. Broken bond wire at dielectric interface
- b. Lifted bond wire off the chip, due to a thermal expansion mismatch between the chip and the resin covering it
- c. Improper bonding; bond wire shorting to the edge of the chip or package lead
- d. Contamination carried with the package by the clear resin on the light pipe between the emitter and the detector. Optically coupled isolators are especially susceptible to this contamination. The effects of such contamination were previously discussed under detector degradation.
- e. Dielectric instability. Optically coupled isolators transmit signals from input to output, while maintaining a high degree of isolation. Human safety and equipment protection are often critically dependent upon dielectric stability of the insulator under severe field conditions.

5.5.7.3 Derating. The largest single source of failure of these devices was proved to be operating above allowable levels of thermal and electrical stress. Accordingly, it is imperative that derating of parts be invoked to enhance the reliability of military systems. Users should refer to MIL-STD-975 for derating factor guidelines. The life of the LED in the optically coupled isolator can be extended by pulsed operation at low drive currents by a factor equal to the inverse of the duty cycle. High temperature operation is the most destructive stress a semiconductor can encounter. It will result in early end of life, electrical parameter drift, and general degradation of the device's electrical and mechanical characteristics.

MIL-HDBK-978-B (NASA)

6.1 MICROWAVE DEVICES, GENERAL

6. MICROWAVE DEVICES

6.1 General.

6.1.1 Introduction. This section contains information on the various types of microwave components used in microwave communications and radar. Microwave components can be divided into three categories: active, passive, and hardware. The components range from dielectrically tuned transistor oscillators to hybrid power dividers and rectangular waveguides. The wide range of parts make general descriptions difficult, but there are some parameters and concepts that apply to all microwave devices. General microwave diode and transistor electrical parameters and reliability considerations are covered by the diode and transistor general sections. Microwave parts do not appear in NASA standard parts lists with the exception of some rf transistors and Schottky diodes. This constraint requires designers of high-reliability microwave systems to choose parts that meet all system requirements and are manufactured using high quality tests and controls. In most cases, this will require additional tests, process controls, and selected qualification tests because the products do not come from qualified suppliers. Some components are not mentioned in this section because they cannot be adequately treated in a handbook of this type. Examples are: assemblies such as antennas; advanced components such as HEMTs, GaAs MMICs, millimeter wave components, and subsystems; complex tubes, antenna feeds, and electromechanical switches.

Applicable military specifications are listed below.

<u>Mil Spec</u>	<u>Title</u>
MIL-A-3933	Attenuators, Fixed, General Specification for
MIL-A-22641	Adapters, Coaxial to Waveguide, General Specification for
MIL-A-24215	Attenuators, Variable (Coaxial and Waveguide), General Specification for
MIL-C-3928	Switches, (Coaxial), Radio Frequency Transmission Line, General Specification for
MIL-C-15370	Couplers, Directional (Coaxial Line or Waveguide), General Specification for
MIL-C-28790	Circulators, Radio Frequency, General Specification for
MIL-D-3954	Dummy Load, Electrical, Waveguide
MIL-D-39030	Dummy Loads, Electrical, Coaxial and Stripline

MIL-HDBK-978-B (NASA)

6.1 MICROWAVE DEVICES, GENERAL

<u>Mil Spec</u>	<u>Title</u>
MIL-F-3922	Flanges, Waveguide, General Purpose, General Specification for
MIL-F-39000	Flanges, Waveguide, Ridge, General Specification for
MIL-I-28791	Isolators, Radio Frequency, General Specification for
MIL-L-3890	Lines, Radio Frequency Transmission (Coaxial, Air Dielectric)
MIL-M-28837	Mixers, Stages, Radio Frequency, General Specification for
MIL-P-23971	Power Dividers/Power Combiners, and Power Divider/Combiners, General Specification for
MIL-S-55041	Switches, Waveguide, General Specification for
MIL-W-23068	Waveguides, Rigid, Circular
MIL-W-23351	Waveguides, Single Ridge and Double Ridge, General Specification for
MIL-STD-1352	Attenuators, Fixed and Variable, Selection of
MIL-STD-1358	Waveguides, Rectangular, Ridge and Circular, Selection of
MIL-STD-1637	Dummy Loads, Electrical, Waveguide, Coaxial, and Stripline, Selection of

6.1.2 General definitions. This list is presented as an aid for the interpretation and understanding of the specific microwave sections.

Attenuation. The reduction in amplitude or power of an electromagnetic wave as it propagates down a lossy transmission line.

Characteristics impedance, Z. The square root of the ratio of the series impedance to the shunt admittance of the equivalent circuit of a transmission line.

MIL-HDBK-978-B (NASA)

6.1 MICROWAVE DEVICES, GENERAL

Insertion loss. A measure of the amount of loss a signal will suffer as it propagates through a component. In general, the measured value contains both the reflected loss and transmission loss.

Isolation. A measure of the reduction of power from the power at the input that will be expected at an isolated port of a component. In the case of a switch, isolation is the difference (in decibels) of power at the load with the switch off as compared to the switch on.

Reflection coefficient, Γ or ρ . The ratio of the voltages of the reflected signal to the incident signal when a signal is applied to the impedance terminating a transmission line. In general, the terminating impedance will not match the transmission line impedance and there will be reflected energy.

Voltage standing wave ratio (VSWR). The ratio of the maximum voltage to the minimum voltage of the standing wave produced by a mismatched termination on a transmission line.

6.1.3 NASA standard parts. Microwave parts do not appear in MIL-STD-975 with the exception of two Schottky diodes and several rf transistors.

6.1.4 General device characteristics. Device characteristics, such as materials, processes, package, and electrical characteristics, are covered in the individual microwave component subsections.

6.1.5 General parameter information. Details of electrical parameters for microwave diodes and transistors appear in this section as well as in the diode and transistor sections of this handbook. Details of electrical parameters for the remaining microwave components are presented in the section for the particular device. General electrical parameters include the following.

- a. VSWR. VSWR is usually calculated from the return loss of a device. The return loss is the relation between the power returning along the line from a mismatched load to the power incident to a load. Thus, a return loss of 10 dB means that the reverse power traveling in a line from a mismatched load is 10 dB below the reference power incident on that mismatched load. For example, the Standing Wave Ratio (SWR) with a certain mismatched load is 2.0, which is equal to a return loss of about 9.5 dB (i.e., the reflected power is 9.5 dB below the incident power).

That quantity for the same SWR can be confirmed as follows: The voltage vector of the backward-flowing power is 0.333 (if incident voltage were 1.0). Power returning down the line is then $(0.333)^2$ or 0.111. Note that 0.111 is approximately -9.5 dB. Mathematically, return loss is equal to $-20 \log_{10} \rho$ (where ρ is the voltage reflection coefficient), and VSWR is $\frac{1+|\rho|}{1-|\rho|}$

6.1 MICROWAVE DEVICES, GENERAL

- b. Insertion loss. Expressed in decibels is $10 \log \frac{P_{out}}{P_{in}}$

Careful calibration of a test circuit can reduce the contribution of reflection loss to this number. The result will be the approximate transmission loss of a line with finite attenuation.

- c. Isolation. Isolation is calculated the same as insertion loss in the case where the isolation of the output port from the input port is desired. The isolation of one output port from another output port is the difference in decibels of the power out of one of the first ports from the second port with power into the input. If more than one signal is present, the isolation of one particular signal power from the total power output may be calculated.

6.1.6 General reliability considerations. The general reliability considerations for transistors and diodes also apply to microwave transistors and diodes. Unique considerations for specific types are presented in this section. Many passive parts dissipate very little power and may be used at full rating over the full temperature of operation. Exceptions are loads, attenuators, and ferrite devices. Active components, such as switches, attenuators, and amplifiers, should be used in applications with proper consideration for derating and reliability.

Three major factors contribute to microwave component reliability.

- a. Good basic device design and good mechanical design of packages and transmission medium
- b. A good manufacturing process
- c. Quality and reliability.

Only when all three factors are optimized will component reliability be at its maximum.

6.2 MICROWAVE DEVICES, DIODES

6.2 Microwave diodes.

6.2.1 Introduction. The only microwave diode that is included in MIL-STD-975 is one type of Schottky diode; however, microwave diodes are included in this handbook to provide a technical understanding.

Microwave diodes can be divided into three general groups which are determined by their general application. These applications are generation, detection, and control of microwave energy. Even though microwave devices utilized in these applications are similar in construction to other diodes, microwave diodes are sufficiently different to warrant a special discussion of their radio frequency (rf) properties.

Discussion of microwave diodes is facilitated by a review of the terms describing device characteristics. Following are several important parameters.

Conversion loss. The ratio of the available rf input power to the available intermediate frequency (IF) output power under the specified conditions.

Figure of merit. The measure of excellence of a video crystal in a video receiver.

Impedance at intermediate frequency (IF). The impedance presented at the output terminals of the mixer when the device is driven by the local oscillator under the specified conditions.

Load impedance. The input impedance of the diode load circuit at the converted frequency of the signal used for the measurement of conversion loss or overall noise figure.

Minority carrier lifetime. The measure of the time it takes to switch a microwave diode from totally on to totally off.

Mixer ratio frequency impedance. The impedance measured at the local oscillator terminals of a mixer under specified conditions.

Negative resistance. The dynamic slope of the current versus voltage (IV) curve when the microwave diode is tested in a suitable high frequency fixture.

Output noise ratio or noise temperature ratio. The ratio of the available noise-power output of the diode at the IF when driven by a local oscillator under the specified test conditions, to that of a resistor at standard temperature (293 ±5 K).

Overall noise figure. The ratio of the available signal-to-noise power ratio at the input to the signal-to-noise power ratio at the output of a network. This value is usually expressed in decibels.

6.2 MICROWAVE DEVICES, DIODES

Series resistance. A measure of the sum of the loss elements in the internal structure of the microwave diode such as lead and contact resistance as well as bulk semiconductor resistance.

Tangential sensitivity. Generally referred to as the signal power below a 1 mW reference level necessary to produce an output pulse whose amplitude is high enough to raise the noise change by an amount equal to the average noise level.

Threshold voltage. The minimum dc voltage applied to a Gunn diode to achieve threshold current. In a typical case, $V_{TH} = 0.33$ V operating.

Threshold current. The maximum dc current applied to a Gunn diode. Typically, $I_{TH} = 1.5$ I operating.

Transition time. A measure of the time for a microwave diode to switch from a reverse conduction state to a totally off state.

Video impedance. The impedance at the specified frequency presented at the output terminals of a semiconductor diode under the specified conditions.

6.2.2 Usual applications. As noted previously, microwave diodes can be divided into three general groups as determined by their application. The first group provides generation of microwave power. The rf frequency multiplication is best accomplished through use of the varactor device, whereas the bulk effect, avalanche, and tunnel diodes best provide power generation for amplifier applications. Noise power and oscillator power generation both use the avalanche diode. The second general group provides microwave power control. Attenuator, limiter, and switching applications are effectively controlled by the PIN diode, whereas tuning is best controlled by the varactor diode. Phase shifting may incorporate either type. The third general group covers the receiving or detecting of microwave power. The function of detection and mixing in receiving rf energy is best accomplished by the Schottky barrier diode, point contact, tunnel diode, or back diode. The rectification of microwave power is accomplished by use of the Schottky barrier diode.

6.2.2.1 Generation of rf power. Generation of rf power by frequency multiplication is often accomplished with the step recovery diode (SRD). The SRD is a type of varactor that provides efficient and versatile frequency multiplier performance. Its advantage is its charge storage and recovery characteristics. The SRD stores charges while biased in the forward direction by either part of a sinusoidal wave or by a steady bias current. It is turned off by either the negative portion of the sinusoid or by a negative pulse. This device is designed to conduct for a period of time in the reverse direction as minority carriers stored near the junction are depleted. When the stored minority carriers are depleted, an abrupt step in current that is rich in high-order harmonics occurs.

For discrete frequency multiplication the voltage pulse should be resonated (e.g., in a quarter wavelength transmission line) to generate an exponentially

6.2 MICROWAVE DEVICES, DIODES

decaying sine wave. To generate a comb spectrum, the voltage pulse is terminated in a resistive load (see Figure 1).

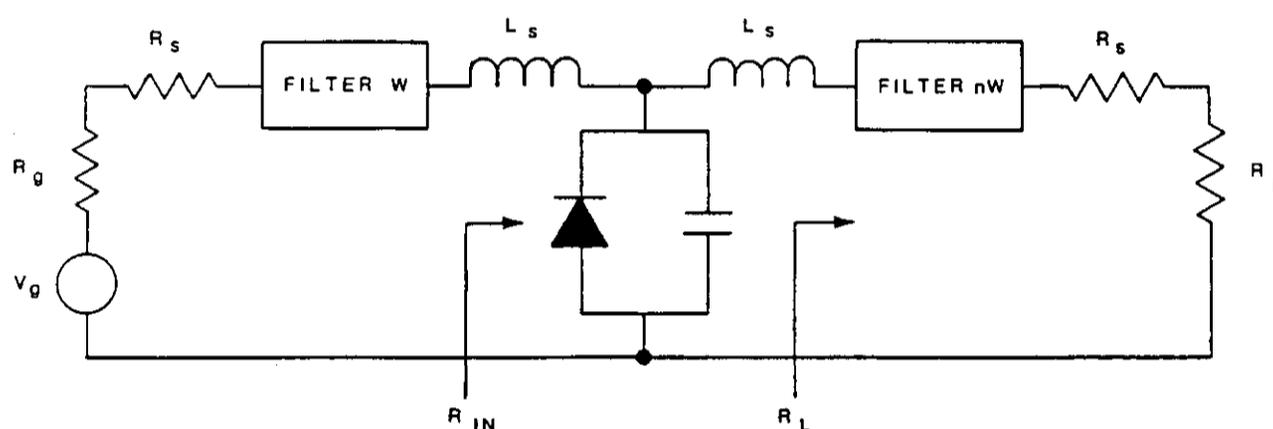


FIGURE 1. SRD multiplier circuit.

The ideal storage diode should have the following properties:

- a. Infinite capacitance in the forward direction (zero incremental voltage drop)
- b. Infinitely fast switching from reverse storage conduction to cutoff
- c. Zero capacitance in reverse direction
- d. Long minority carrier lifetime
- e. No excessive parasitic elements; a small amount of series resistance is tolerable and a small amount of series inductance may be desirable to reduce the resistance-capacitance transition time.

A second application of the varactor for rf power generation is in parametric amplifiers and up converters in which the varactor becomes the active element that produces a negative resistance. The term parametric is derived from the achievement of amplification, oscillation, and up conversion or harmonic multiplication by choosing a lossless parameter in the system. Parametric amplifiers, when cooled to cryogenic temperatures, offer low-noise performance for low-noise receiver applications.

In the reflection parametric amplifier circuit a high-power, high-frequency pump signal is reactively mixed with the incoming low-frequency signal. The resulting lower sideband is supported in a resonant circuit whereas the upper sideband

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signal is terminated in a high impedance. The lower sideband frequency mixes again with the pump signal providing an additional component to the low-frequency signal. This double reactive mixing process introduces a 180-degree phase shift between the incoming and outgoing signal frequency which results in the appearance of a negative resistance. The power is drawn from the pump source (see Figure 2).

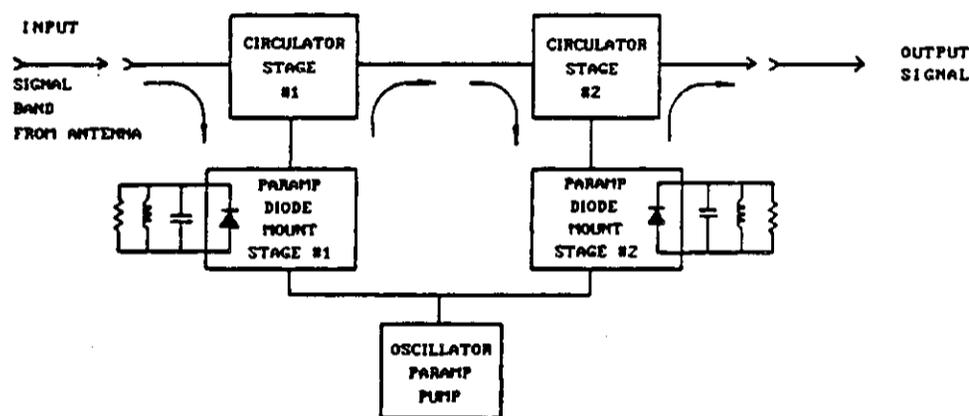


FIGURE 2. Reflection parametric amplifier.

Bulk-effect microwave devices (Gunn diodes) differ considerably from conventional pn junctions in that there is no discrete junction area. The bulk semiconductor material is usually gallium arsenide whereas conventional diodes are silicon. The n-type bulk material, when dc-biased, creates a charged region that travels from the negative terminal through the bulk material. In a microwave resonant circuit this charged region produces microwave power generation in oscillator or amplifier applications.

The major microwave applications of the Gunn device have been in low-power oscillators in the 6 to 20 GHz range with average output power to about 200 mW. Gunn devices have been used in applications exceeding 90 GHz. This device has about a 10-dB less noise figure and requires a lower bias voltage than the impact ionization avalanche time (IMPATT) diode. However, the Gunn device generates less power than the IMPATT.

The peak power output of a Gunn diode occurs at the transit time frequency which is inversely proportional to the length of the device and directly proportional to the electron drift velocity. This is in turn, proportional to the operating voltage. The power output of a Gunn oscillator increases rapidly beyond a threshold voltage until optimum transit time is obtained, then decreases with further

6.2 MICROWAVE DEVICES, DIODES

increased voltage. If possible, the diode should operate slightly below the power maximum (1 V) and never over it. Diodes become noisy when operated above the power-peak voltage.

Thus, the important characteristics of the device are the IV threshold voltage, turn-on voltage, peak power voltage, and turn-off voltage. Figure 3 shows the characteristic shape of the diode current versus voltage and microwave power versus voltage.

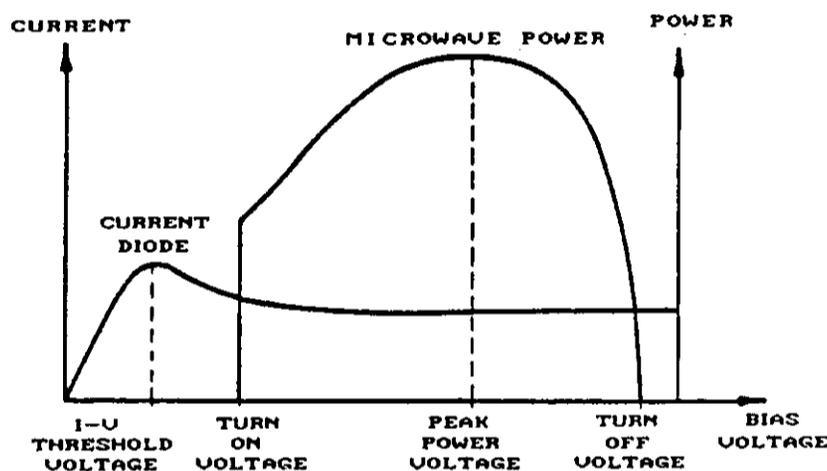


FIGURE 3. Gunn oscillator IV characteristic curve and power-voltage curve.

The frequency over which the device exhibits negative resistance will determine whether or not the device will oscillate in a given cavity. A reduced height cavity is often used to combat low temperature turn-on problems. Figure 4 shows a Gunn oscillator circuit with varactor diode tuning.

A Read diode or IMPATT diode amplifier has the ability to produce more microwave power than any other semiconductor device at frequencies as high as 300 GHz. The IMPATT diode utilizes conventional pn, PIN, or Schottky junctions under external, reverse bias. The diode couples energy into a microwave field by providing charge carriers from avalanche breakdown of an n^+p region at one end of the diode to a drift region at precisely the time the applied ac field is starting to decrease. The clump of holes obtained from the avalanche region

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begin traveling through the diode intrinsic, or drift, region under the force of the dc field. At this point the holes are traveling against the direction of the ac field; therefore, they give up energy to the ac field. Traveling through the drift region opposite the applied ac voltage produces a current with the opposite polarity. This implies a negative resistance has been developed. The drift region is made just long enough for the charge clumps to reach the end of the drift region as the ac voltage again starts going positive.

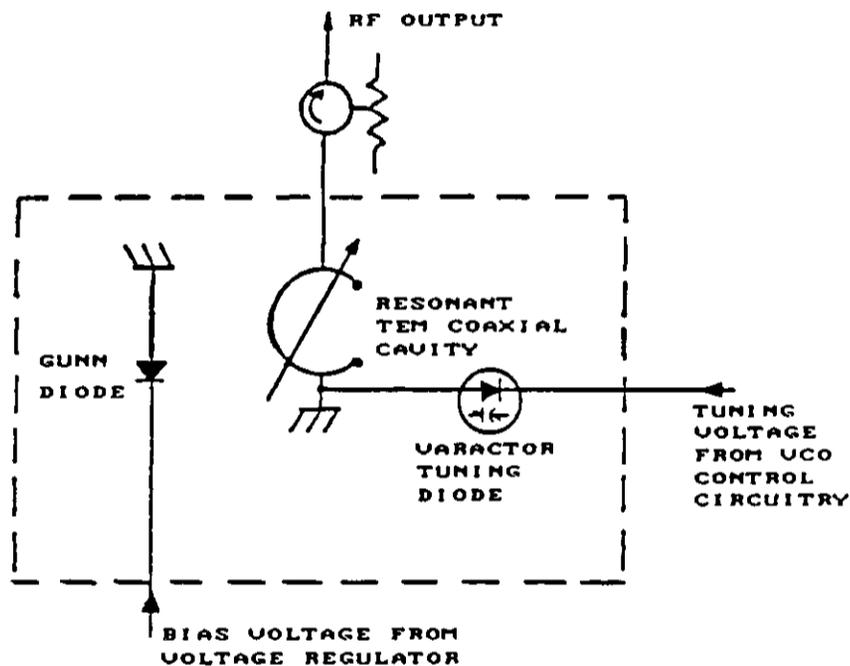


FIGURE 4. Gunn oscillator circuit with varactor diode tuning.

IMPATT diodes are often combined in such a fashion that the total output power is approximately equal to the sum of the powers of the individual diodes. Combiners can be 2-way or N-way. The 2-way combiners, such as rat-race couplers, hybrid couplers, or 2-way Wilkinson combiners, can be combined to form an N-way combiner where $N = 2^n$ and n is an integer. This method is not efficient for $N > 4$. The N-way combiner combines diodes in a single structure. This structure may be either a nonresonant or a resonant type. In the nonresonant type, the power sources are isolated from one another so power combining can only occur if it is excited by an external signal. In the resonant type, the power sources are coupled together so the circuit can be used as an oscillator without the use of an external locking signal. Figure 5 shows one type of resonant combiner oscillator.

6.2 MICROWAVE DEVICES, DIODES

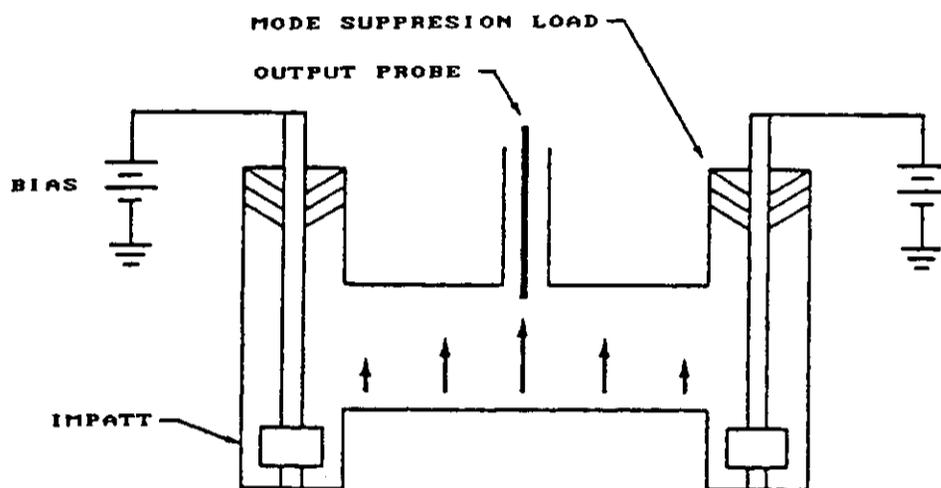


FIGURE 5. IMPATT combiner oscillator.

The trapped plasma avalanche triggered transit (TRAPATT) diode provides microwave energy from approximately 0.5 to 4 GHz, with efficiency approaching 50 percent. However, the TRAPATT diode requires a complicated circuit and gives a higher noise level than the IMPATT diode.

Tunnel diodes and back diodes have normal diode junctions except the pn-doping level is much higher than conventional devices. Electrically, tunnel diodes provide negative resistance at microwave frequencies and are useful as low-noise amplifiers due to their inherent low-noise figures. Back diodes differ from tunnel diodes in that they allow little current flow when the diode is reverse biased. Tunnel and back diodes can be used in detector and mixer applications.

6.2.2.2 Microwave power control. PIN diodes utilize an undoped semiconductor layer referred to as an intrinsic layer, located between the p- and n-doped layers. This intrinsic region provides low-junction capacitance that has little or no change due to reverse bias. Under forward biasing conditions, the intrinsic layer becomes a conductor. When the PIN device is placed in a microwave transmission application, the device has the effect of a shunt attenuator. In limiter applications, the device acts as a lossy switch.

The switching action results from variations of the I-layer resistance with applied dc signal. At zero or reverse bias, R_i is high and the diode acts as a fairly high capacitance at microwave frequencies. In the forward biased state, the I-layer resistance R_i is lowered due to conductivity modulation.

Conductivity modulation can be produced either by sufficiently high-level microwave power or by dc bias.

6.2 MICROWAVE DEVICES, DIODES

The breakdown voltage is approximately proportional to the I-layer width. The intrinsic resistance is proportional to the square of the I-layer width and inversely proportional to the minority carrier lifetime and dc bias current

$$R_{RF} = \frac{W^2}{\tau \cdot I_{DC}} \quad (\text{as shown in Figure 6}).$$

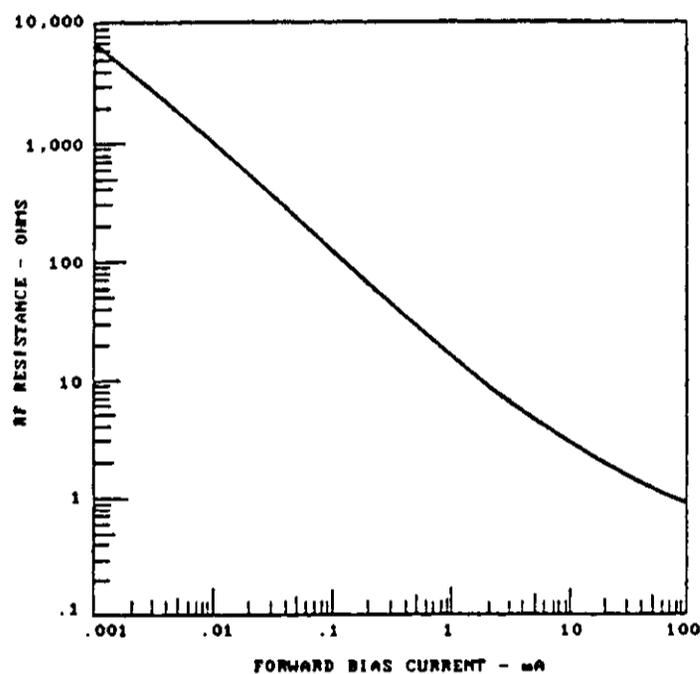


FIGURE 6. Typical PIN diode rf resistance vs forward bias current.

The PIN diode may be used in series or shunt configurations. The key electrical parameters are the insertion loss, isolation, and bandwidth of the particular application.

Four simple attenuator circuits are shown in Figure 7. The forward biasing is not shown. The first attenuator is the simplest and cheapest implementation. The second circuit has the highest isolation whereas the "TEE" and "Pi" circuits have the broadest bandwidth. A diode in shunt (not shown) could be implemented in waveguide. If n shunt diodes are spaced at quarter wavelength intervals, the overall attenuation can be increased by more than n times that of a single diode.

A series PIN switch and a shunt PIN switch are shown in Figures 8 and 9 respectively. When used as a switch, the residual attenuation that exists in the PIN diode when the switch is on is called insertion loss. The attenuation provided

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when the switch is off is called isolation. If the diode is assumed to be a pure resistance at rf, the attenuation for each circuit is a function of the ratio of the circuit resistance to the diode resistance. As the bias on the diode is varied the load resistance as seen by the source also varies. Reflection from the mismatch is the primary source of attenuation, but some of the energy is also transmitted or dissipated in the PIN diode.

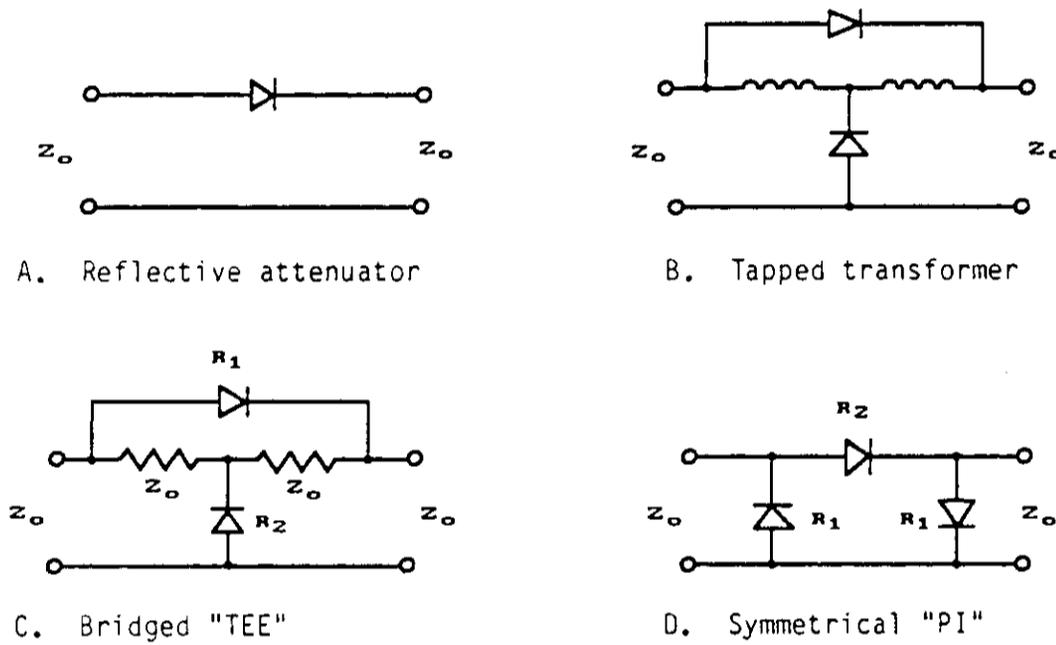


FIGURE 7. PIN diode attenuator circuits.

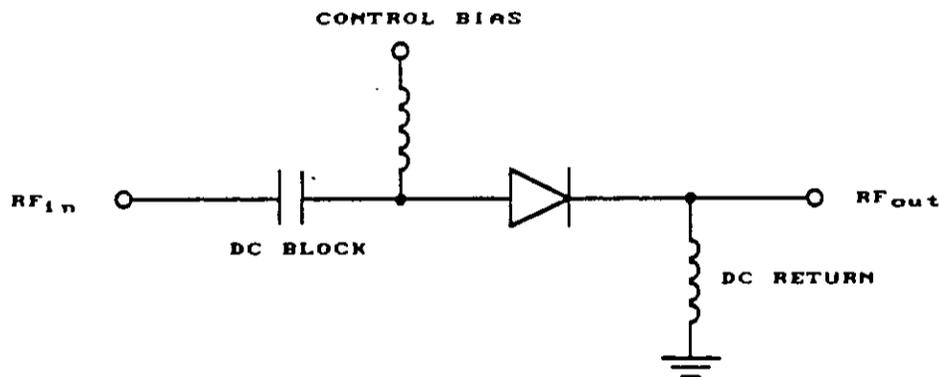
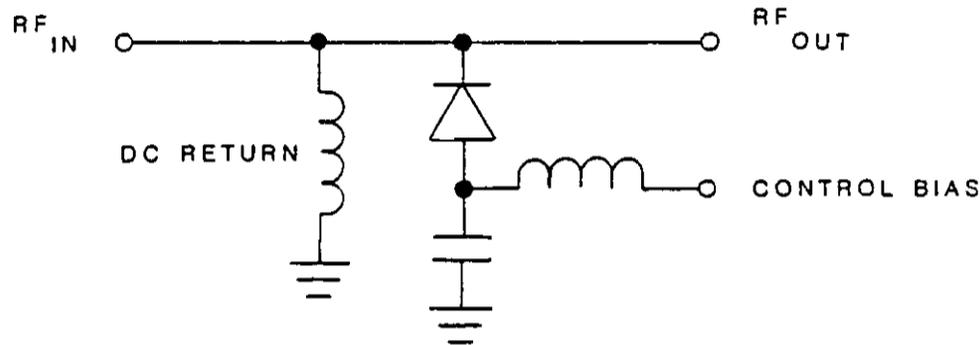


FIGURE 8. Series PIN reflective switch.

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FIGURE 9. Shunt PIN reflective switch.

A packaged part, or even a wire-bonded chip component, has stray parasitic elements that must be tuned out at microwave frequencies by the addition of external reactances. Such circuits are generally referred to as resonant switches in which the PIN diode is used essentially to switch the circuit parameters from a parallel resonant condition to a series resonant condition. The high and low impedances produced by the parallel and series resonant circuits, respectively, constitute the on and off states of the switch.

A basic limiter utilizes the decrease in PIN diode resistance when incident rf power increases to reflect some of that power and maintain a nearly constant output power. The limiting action only comes into play when the input power reaches a threshold value. After that, the output power will remain constant for increasing input power until the diode resistance is reduced to its smallest value. The output will begin to increase if any additional power is input.

Figure 10 shows the output power versus the input power for a shunt-mounted 50 ohm limiter diode.

The power rating, breakdown voltage, intrinsic resistance, and speed of the diode determine the limiter characteristics. The turn-on characteristics of the PIN diode in the limiter can be improved by using a detector diode to add dc current to the PIN diode junction at the onset of rf power.

Varactor diodes have a modified pn junction constructed to maximize the decrease in junction capacitance with an increase in reverse voltage. Consequently, they are used as the voltage-controlled, tunable impedance in voltage-controlled oscillators (VCOs) and phase shifters.

Figure 11 shows an equivalent circuit of a VCO. The active device and cavity could be the Gunn diode oscillator discussed previously, a field effect transistor (FET), or crystal oscillator. Important parameters are capacitance versus voltage, capacitance ratio (a measure of the tuning range), and the Q (a measure of the efficiency).

6.2 MICROWAVE DEVICES, DIODES

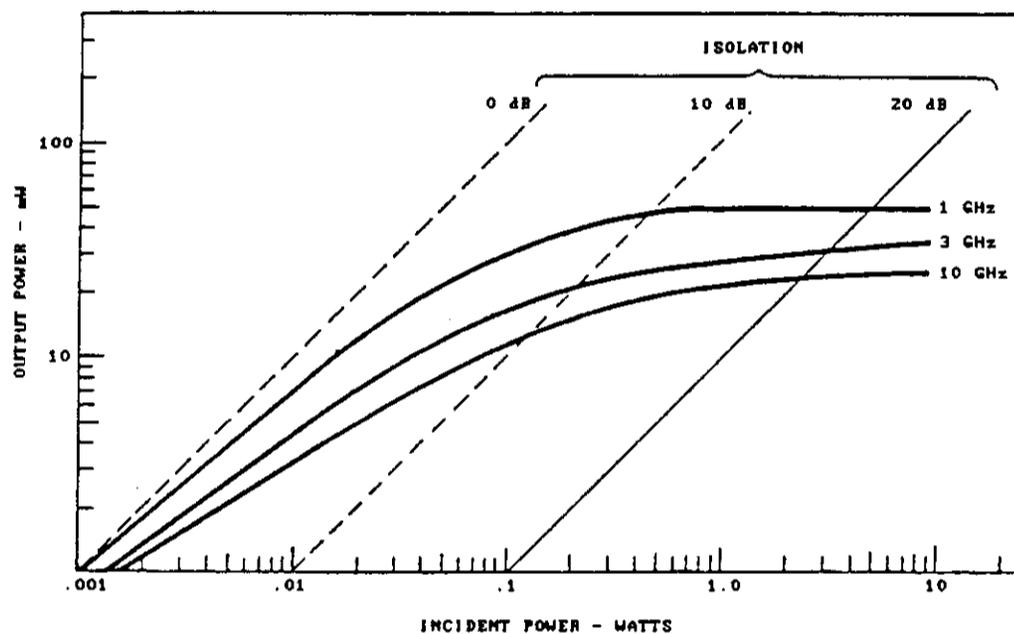


FIGURE 10. Output power vs input power for a 50-ohm limiter diode.

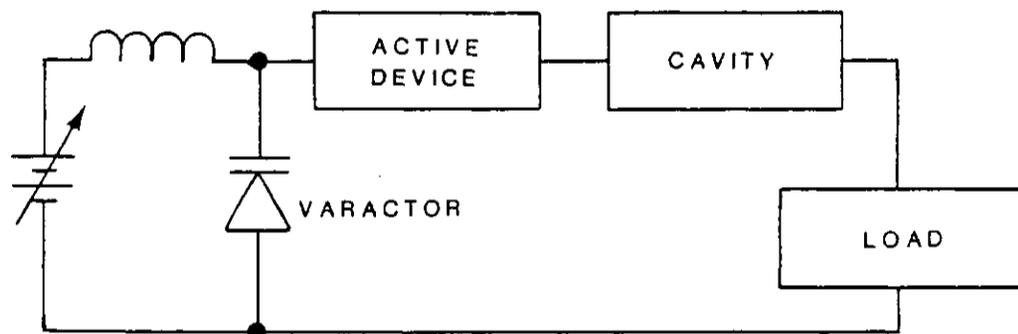


FIGURE 11. Equivalent circuit of a VCO.

6.2 MICROWAVE DEVICES, DIODES

6.2.2.3 Microwave detection. Schottky-barrier diodes differ from the normal silicon pn construction in that a metal layer deposited on the silicon forms the junction. As a result of this special layer, the device has a small junction capacitance and a nonlinear voltage-current relationship. This type of diode is useful as a microwave detector because of its low noise properties and efficient rectifying properties.

The point-contact diode is similar to the Schottky barrier diode except that the anode contact is accomplished by using a tungsten whisker to form a pressure contact to the silicon. A more detailed discussion of this oldest of the microwave semiconductors is provided in the last portion of this section.

Tunnel diodes have the lowest output resistance and lowest 1f noise characteristics due to the high doping levels of the back diode semiconductor wafer. This results in a faster video rise time and larger bandwidth. The tunneling region of the IV characteristic, where the detector operates under small signal conditions, is relatively independent of temperature. Figure 12 shows a zero-biased detector circuit.

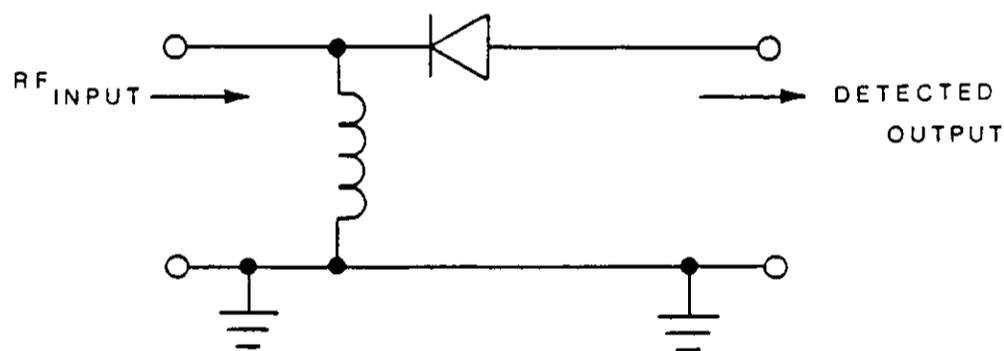


FIGURE 12. Detector circuit.

Schottky barrier diodes are more reliable when operated with a low impedance load. High-resistive loads are acceptable if the diode is in the square law region. A 50Ω load is best for fast pulse shape preservation. The input match and R-C time constant of the input circuitry will also affect the shape of a fast changing pulse. High-power detector loads should be less than $1,000\Omega$ with 300 to 500Ω giving maximum power transfer.

6.2.3 Physical construction. To accomplish the various microwave functions previously discussed, microwave diodes are constructed in a different manner from

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the normal pn junction and associated packaging. The following brief outline of the comparative internal junction construction (Figure 13) is presented as an aid to the discussion of electrical characteristics in paragraph 6.2.5 of this section. Details on beam lead devices are given in section 3.1, Transistors, general.

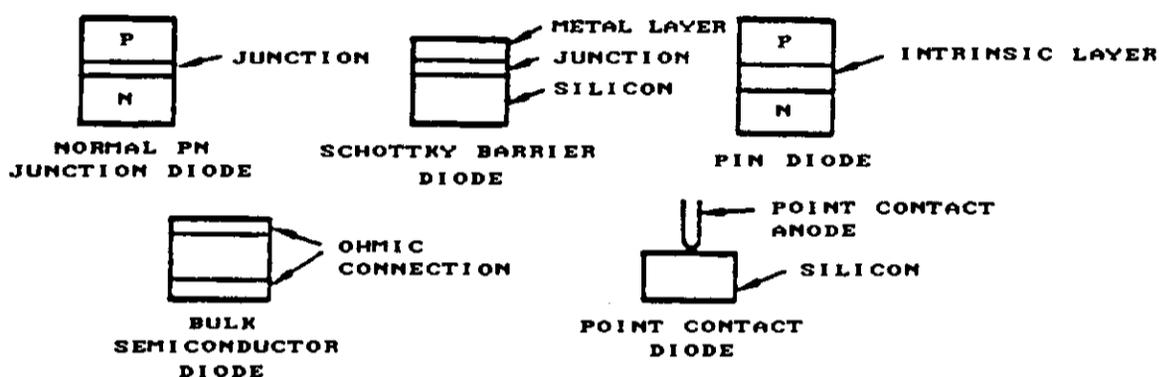
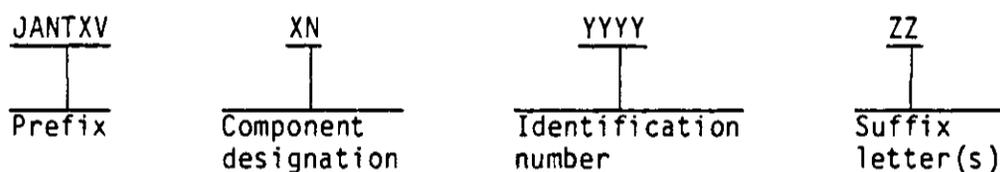


FIGURE 13. Microwave diode construction.

External packaging of microwave diodes varies according to the application. Figure 14 illustrates a few of the commonly used types.

6.2.4 Military designation. The military designation for microwave diodes is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is IN for microwave diodes.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describe matched devices (suffix M), reverse polarity (suffix R), or any other letter to indicate a modified version.

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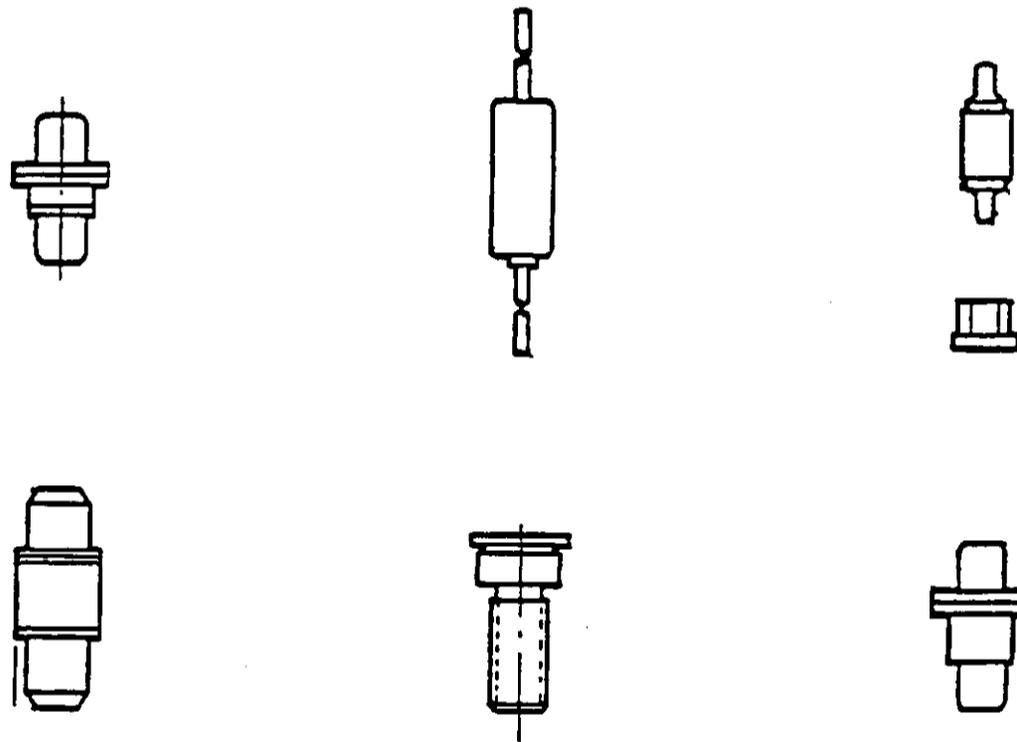


FIGURE 14. Microwave diode packages.

6.2.5 Electrical characteristics. Table I presents some typical detector and mixer diode electrical characteristics for JAN-type rf and microwave diodes. An examination of this oldest class of microwave diodes points out their applications.

6.2.5.1 Detection and mixing based on the dc EI curve. The curve for an ideal diode is shown in Figure 15. In this instance the forward current is infinite for the smallest conceivable amount of forward voltage. The reverse current is zero for any value of reverse voltage. The ideal diode would make a perfect detector or mixer. As a low-level detector operating at close to zero volts, the extreme nonlinearity of the characteristic curve over the range of some small plus-to-minus voltage would lead to perfect or lossless detection. As a mixer the dV/dI at some operating point above zero volts would once again result in perfect or lossless mixing. Because conditions are seldom ideal, consideration will be given to more practical cases.

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TABLE I. Electrical ratings

Device Type No.	Test Freq. (MHz)	Operate Freq. (MHz)	Noise Figure (dB)	Conversion Loss (dB)	Output Noise Ratio	VSWR Max.	IF Impedance (ohms)		Burn-out (Ergs)
							Min.	Max.	
1N830A	100	100	---						
1N25	1000	---	---	---	---	---	---	---	375
1N82A	1000	---	14.5	8	2.5	1.3	100	400	---
1N21WE	3060	---	7.0	---	---	---	---	---	---
1N21WEM	3060	---	7.0	5.5	1.5	1.3	350	450	5
1N21WEMR	3060	---	7.0	5.5	1.5	1.3	350	450	5
1N32	3295	1550-	---	5.5	1.5	1.3	350	450	5
		3060	---	---	---	---	---	---	375
1N31	9375	5200-	---						
		9375	---	---	---	---	---	---	175
1N23WE	9375	---	7.5						
1N23WEM	9375	---	7.5	6.0	1.4	1.3	335	465	2
1N23WEMR	9375	---	7.5	6.0	1.4	1.3	335	465	2
1N78	16000	---	---	6.0	1.4	1.3	335	465	2
1N78C	16000	---	9.5	7.5	2.5	---	325	625	1
1N26	23984	---	---	6.0	1.9	1.5	400	565	1
1N26B	23984	---	1.0	8.5	2.5	---	300	600	.1
1N53	34860	---	---	7.5	2.0	1.5	400	600	.1
				8.5	2.5	1.6	400	800	---

NOTE: These parts are not included in MIL-STD-975. This table is not intended to be a part selection list.

6.2 MICROWAVE DEVICES, DIODES

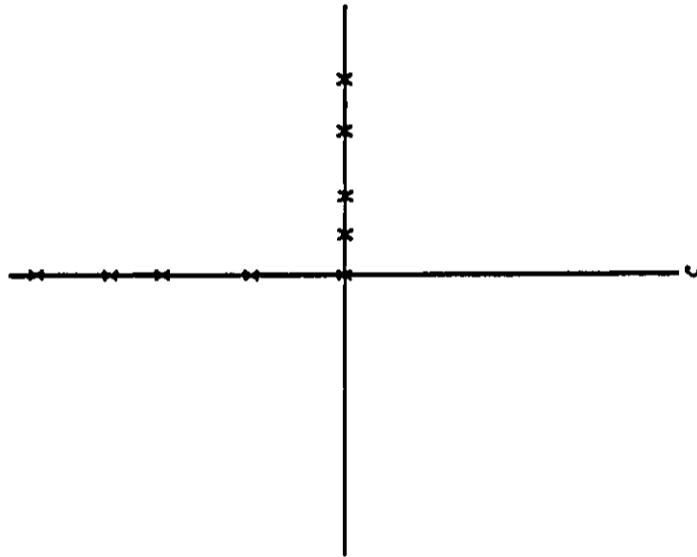
FIGURE 15. An ideal diode dc characteristic curve.

Figure 16 is a plot of the EI curve of the point-contact diode. Concerning detection, forward conduction starts at zero for the point-contact diode, and the ordinate, in this case, is 0-10 μA . This scale has been chosen because detection is being discussed. For the purpose of this discussion, the reverse current is assumed to be zero over the range of the signal input.

If a small rf voltage (40 mV P-P) is applied, it results in a detected output as shown in Figure 17.

As can be seen, the point-contact diode is detecting efficiently due to the fact that there is conduction over the positive half cycle of the rf input.

The mixer diode can be treated in somewhat the same manner. Mixers typically use Schottky barrier diodes or hot-carrier diodes. Figure 18 is a plot of the EI curve of a Schottky barrier diode with the ordinate, in this case, being from 0-10 mA. This scale has been chosen because the mixer diode is now being discussed and conduction in the range of milliamperes is important, not microamperes, as was the case for the detector. Since the local oscillator supplies the bias level, we see that the Schottky barrier diode will operate as an efficient mixer, for the purpose of this discussion, with a local oscillator bias of about 500 mV peak-to-peak and a signal of 40 mV peak-to-peak. It should be pointed out, however, that dc bias will also enable the Schottky barrier diode to operate at lower drive levels.

6.2 MICROWAVE DEVICES, DIODES

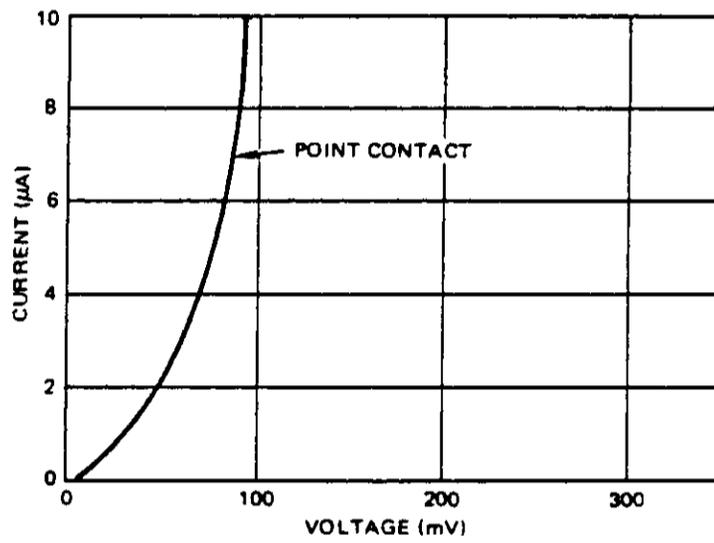


FIGURE 16. Forward direction dc characteristic EI curve.

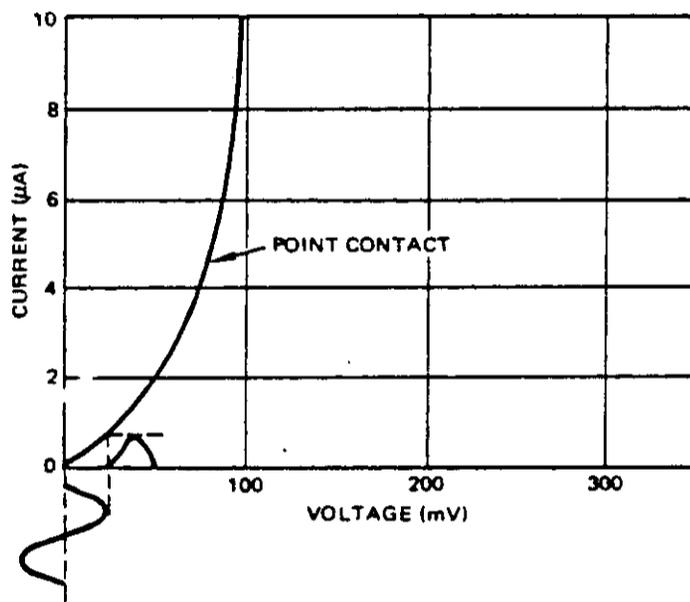
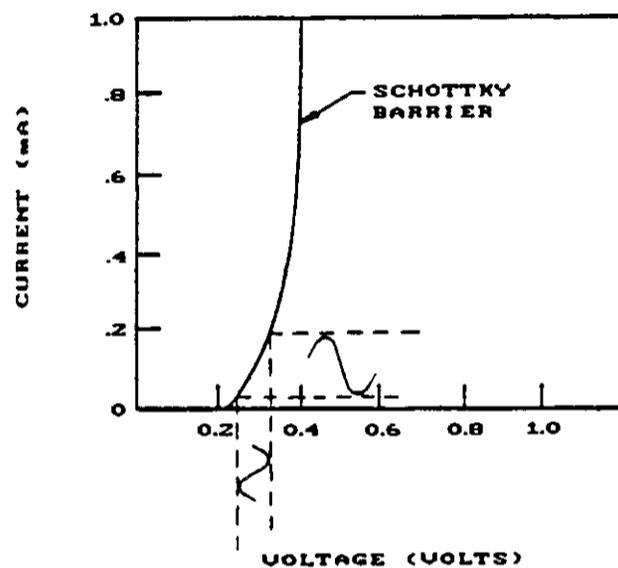


FIGURE 17. Video detection based on the dc characteristic curve without dc bias.

6.2 MICROWAVE DEVICES, DIODES

FIGURE 18. Mixing based on the dc characteristic curve.

Shown in the next several figures are the actual forward and reverse characteristics of point-contact diodes. Comparisons are made between typical S, X, and Ka-band types, and typical diode characteristics versus temperature.

Figure 19 shows the forward characteristic for the point contact diodes in the S-band, X-band, and Ka-band. The curves are similar.

Figure 20 is a plot of the reverse characteristic of the point contact diode. Point contact diodes have very low reverse voltage drops for any given reverse current. Although not generally specified, point contact diodes typically have 3-V peak-inverse voltage (piv) at 100 μ A.

Figure 21 shows the forward characteristic of a point-contact diode versus temperature. At currents of 1 mA or more, the change versus temperature is much less than at currents of 100 mA or less. This is to be expected, because internal heating due to the I^2R drop (forward conduction) tends to swamp out any differences due to ambient temperature. This is not so at lower current values.

Figure 22 shows the reverse characteristics of the point-contact diode versus temperature. As can be seen, the change in piv at 150 $^{\circ}$ C is extreme, becoming one-third of its room temperature value for this diode.

6.2 MICROWAVE DEVICES, DIODES

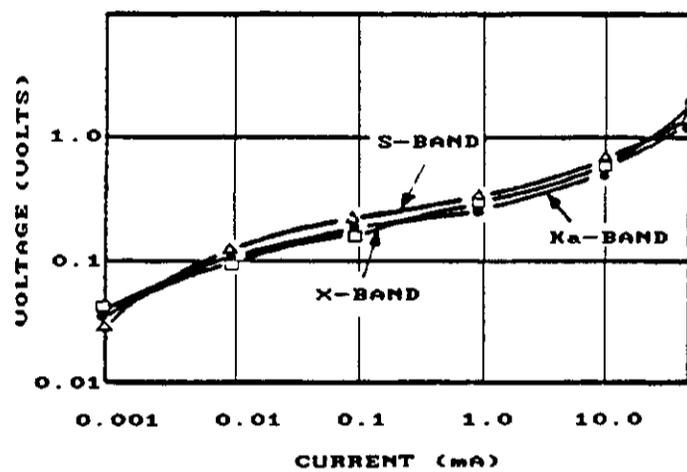


FIGURE 19. Typical forward dc characteristics at 25 °C for point-contact diodes.

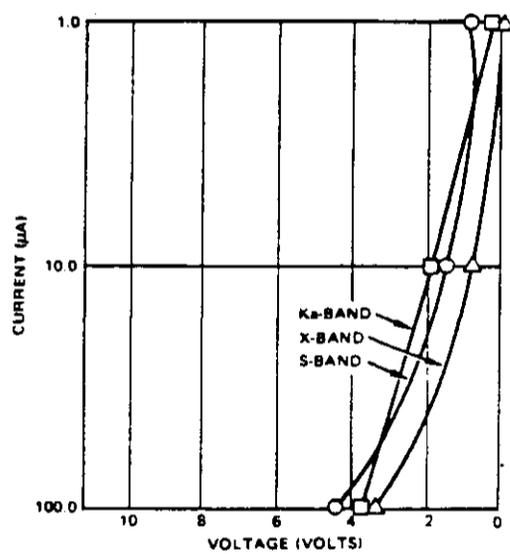


FIGURE 20. Typical reverse dc characteristics at 25 °C for point-contact diodes.

6.2 MICROWAVE DEVICES, DIODES

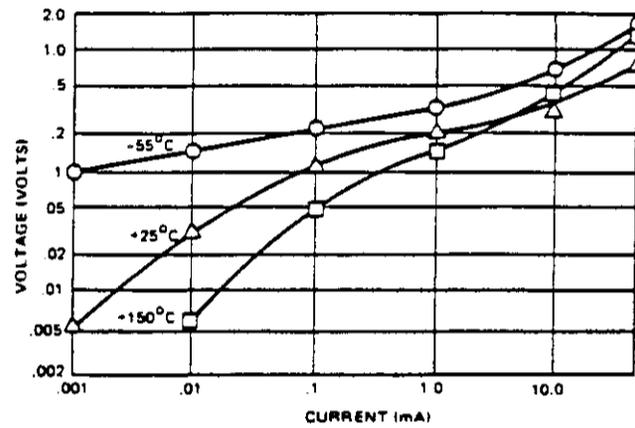


FIGURE 21. Typical forward dc characteristics vs temperature for point contact diodes.

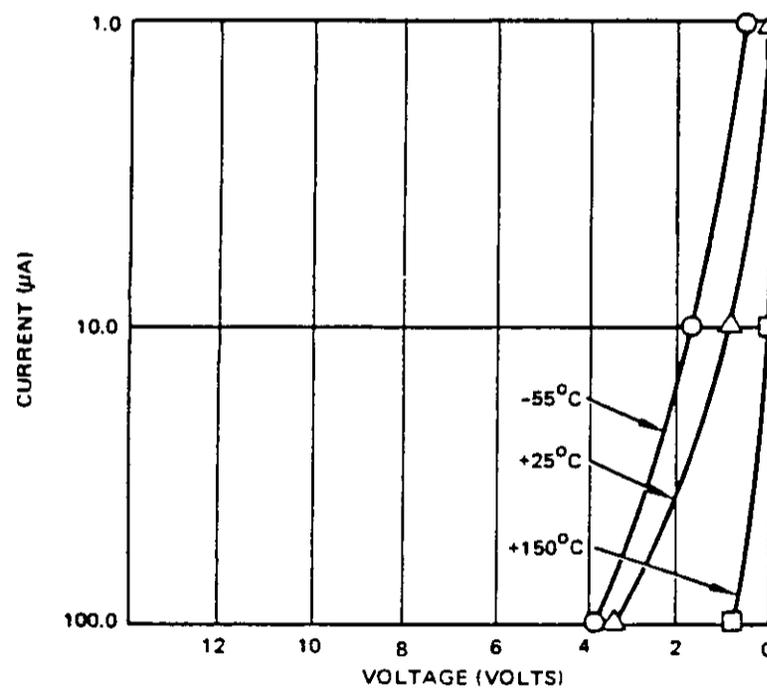


FIGURE 22. Typical reverse dc characteristics vs temperature for point contact diodes.

6.2 MICROWAVE DEVICES, DIODES

The hot-carrier diode or Schottky barrier diode has several advantages over the point-contact diode.

- a. More rugged. The point-contact diode depends on a mechanical contact between a sharp metal whisker and a semiconductor material. Shock and vibration often change or destroy this contact.
- b. Higher burnout. The uniformity and larger area of contact of the hot-carrier diode make it less susceptible to burnout from either cw power, peak power, or pulse energy. This is a common mode of failure for detectors and mixers.
- c. More reproducible. Each hot-carrier diode is very nearly like every other hot-carrier diode in forward characteristics. This actual forward characteristic is closer to the ideal diode law characteristics than any other type of diode. Point-contact diodes, however, vary widely. See Figure 23 for a comparison of current voltage characteristics of a batch of point-contact diodes and a batch of hot-carrier diodes.
- d. Lower series resistance. The lower series resistance of hot-carrier diodes leads to lower losses and better noise performance.
- e. Higher off-to-on ratio. The ratio of reverse current (at a given reverse voltage) to forward current (at a given forward voltage) is much higher for the hot-carrier diode. This leads to better performance as a switch, rectifier, or detector.
- f. Lower noise figure. If an average hot-carrier diode is compared with good selected-point contact the hot-carrier diode exhibits slightly better overall NF_0 . However, compared with the average point-contact, the average hot-carrier diode is superior.
- g. Low frequency (lf) noise. Figure 24 shows typical lf noise performance for a hot-carrier diode at several bias points and a good point-contact diode. The lf noise of the hot-carrier diode is significantly better.

6.2.6 Environmental considerations. Microwave diodes generally conform to the environmental specifications of conventional pn-junction diodes (see section 4.2, Diodes, switching). For more detailed requirements see applicable slash sheets of MIL-S-19500.

Typical environmental conditions and screens that microwave diodes are capable of withstanding generally conform to the requirements of MIL-S-19500 for JANTXV or JANS diodes.

6.2 MICROWAVE DEVICES, DIODES

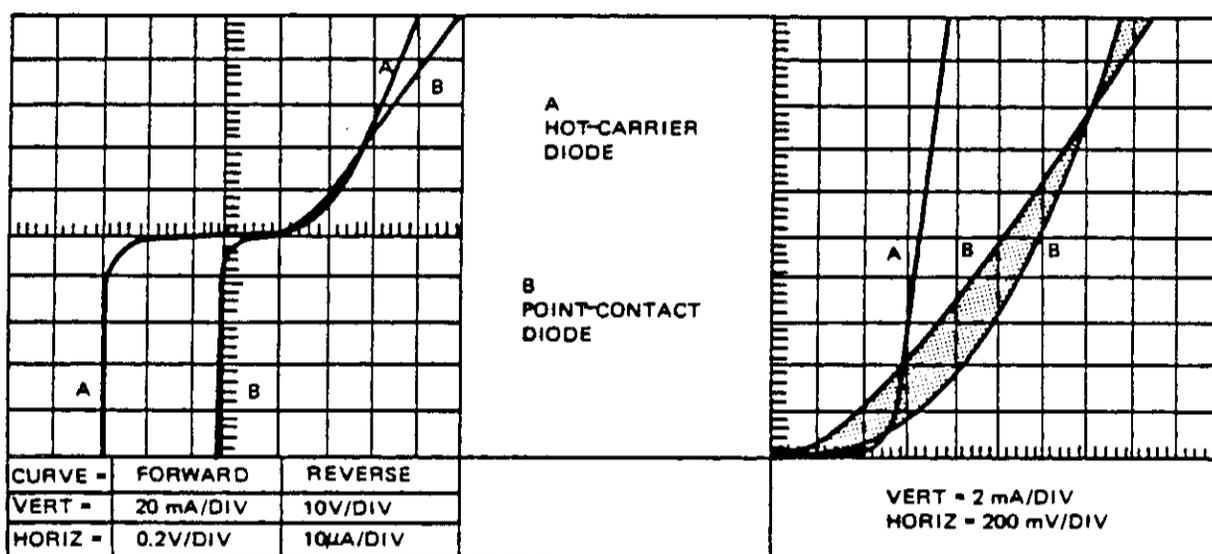


FIGURE 23. Current and voltage characteristics for hot-carrier and point-contact diodes.

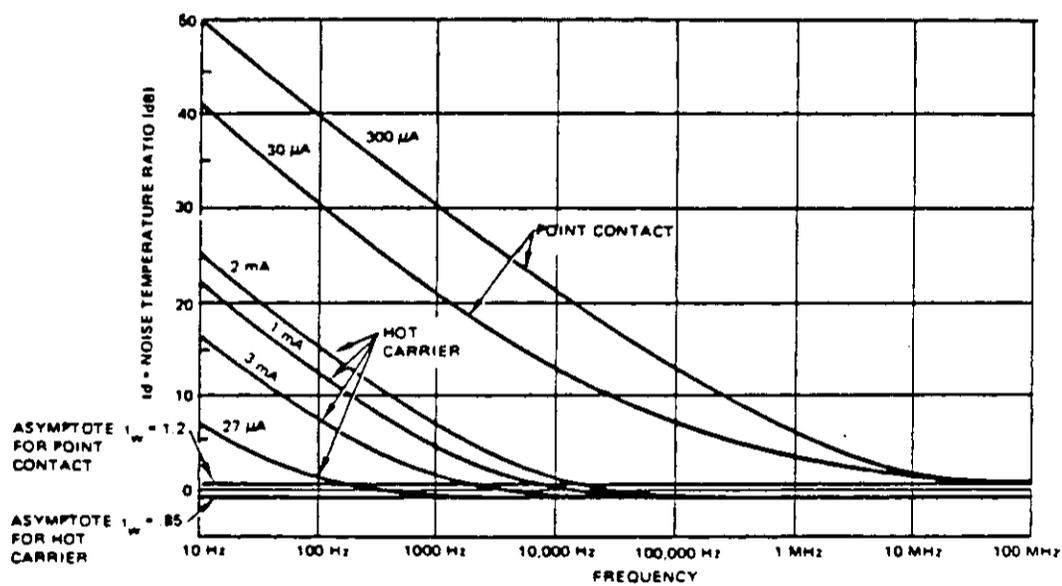


FIGURE 24. Hot-carrier and point-contact diodes--1/f noise.

6.2 MICROWAVE DEVICES, DIODES

6.2.7 Reliability considerations.

6.2.7.1 Failure modes. The primary mode of failure for microwave diodes is burnout. Burnout is defined as a catastrophic failure, such as an open or short circuit, or a 3 dB degradation in electrical performance. Burnout may be due to excessive energy, either continuous or transient, which is dissipated in the diode. In some circuits where a diode acts as a detector over a wide dynamic range and maximum sensitivity is sought, the diode capability will be exceeded at the high-power levels. Transient energy is more difficult to protect against, and transient spike leakage power is damaging because the heat generated in the diode during the duration of the spike is not dissipated throughout the bulk material of the diode semiconductor. Proper circuit design and selection of correct device characteristics can eliminate the preceding failure modes.

Another reliability consideration in the use of microwave diodes is that the point contact can damage the junction, depending on the amount of pressure exerted on the contact or external shock-induced oscillation of the whisker.

6.2.7.2 Derating. This is the intentional reduction of electrical stresses on a device to reduce the operating temperature of the diode for the purpose of extending its life. History has shown that the largest single source of failure of diodes is due to operating above allowable levels of stress. Accordingly, it is imperative that derating of parts be invoked to enhance the reliability of military systems. Derating conditions specified in MIL-STD-975 for diodes apply to microwave diodes. In general, derating for voltage, current, and surge current should be 50 percent of the value published on the vendor data sheet. The junction temperature should be limited to 125 °C maximum. The design and use of a part should always be within the thermal and electrical stresses defined. High-temperature operation is the most destructive stress a semiconductor could be subjected to. High temperature will cause early destruction life, electrical parameter drift, and general degradation of the electrical and mechanical characteristics of the device.

6.2.7.3 Screening. Development of the JANTXV process for several microwave diodes has done much to reduce the incidence of the failures discussed in this section. A description of the JANTXV process is given in section 4.1, Diodes, general.

6.2.7.4 Static sensitivity. Schottky diodes and point-contact diodes are electrostatic-sensitive devices. The following procedures should be adhered to when handling high frequency devices:

- a. The operator, as well as tweezers or any other pickup tool, must be grounded to the test or inspection station. This prevents the build-up of static charge, which can damage the diode if allowed to pass through it.

6.2 MICROWAVE DEVICES, DIODES

- b. All test fixtures should be equipped with a short across the terminals that is disconnected after the diode is inserted.

For leakage measurements where a short across the terminals is not practical, a series resistor may be used. The series resistor (approximately 10 K) should be physically close to the test terminal.

This will prevent discharge through the diode of any charge built up on the capacitance that is present in the fixture, leads, and test equipment. It is also advisable to minimize this capacitance.

- c. Spurious pulses generated by test equipment, contact bounce during switching, induced voltage in the leads, etc., must be eliminated.
- d. When passing a diode from one operator to another, the receiving operator should grasp the lead held by the passing operator. If the opposite lead is taken, there is a possibility of passing static charge difference between the operators through the diode.
- e. All soldering apparatus should be transformer isolated from the power line and should be free of leakage.

6.2.7.5 Radiation considerations. To insure that a circuit functions properly in space applications, the design engineer must consider the effects of radiation exposure. Requirements in military programs for systems that are survivable in a radiation environment adds to the challenge. The design engineer must know how radiation affects the circuit and components. When semiconductor devices are exposed to radiation environments, changes occur in their rated electrical parameters. The magnitude of the changes is a function of such things as the type of radiation, neutron or gamma ray, and time or duration. Generally, permanent damage is associated with displacement effects resulting from neutron irradiation, and transient effects are the result of ionization from gamma ray radiation.

Diodes exposed to radiation will display changes in the following electrical characteristics:

- a. Breakdown voltage increases
- b. Leakage current increases
- c. Forward voltage increases.

These changes are a result of changes to the bulk material and are permanent.

Four radiation hardness assurance levels (RHA) are provided for diodes screened and tested to JANS requirements in accordance MIL-S-19500. These are designated by the letters M, D, R, and H.

6.3 MICROWAVE DEVICES, TRANSISTORS

6.3 Microwave transistors.

6.3.1. Introduction. There are no microwave transistors in MIL-STD-975. They are included here for information only. However, a few rf transistors appear in MIL-STD-975.

The design of modern microwave amplifiers, oscillators, and occasionally multipliers and mixers incorporate a microwave transistor as the active, two-port element. The majority of these devices are either npn silicon bipolar or gallium arsenide metal electrode semiconductor FET (GaAs MESFET). A comparison of the relative gain, power, and noise figure is summarized in Table II. The GaAs MESFET has lower noise, higher gain, and higher power output than the silicon bipolar transistor. The MESFET can operate at higher frequencies than the bipolar transistor because the latter suffers from minority carrier storage effects. The MESFET is a majority carrier device and is not slowed down by this mechanism. The MESFET also has a higher input impedance and a negative temperature coefficient that prevents thermal runaway from occurring. The MESFET's main disadvantage is higher 1f noise, which can be significant for oscillator noise.

TABLE II. Comparison of microwave transistors

	Si Bipolar Transistors			GaAs MESFET			
	4 GHz	8 GHz	12 GHz	4 GHz	8 GHz	12 GHz	18 GHz
Gain typical (dB)	15	9	6	20	16	12	8
Power output (W)	6	2	0.25	25	8	4	1
NF min (dB)	2.5	4.5	8	1.0	1.8	2.2	2.5

The physical principles of microwave transistors are similar to those of low frequency transistors. Operation at rf and microwave frequencies puts more severe requirements on active area dimensions, process control, heat sinking, and packaging. The construction of most types is a planar, epitaxial process. Microwave transistors also include pnp bipolar, insulated gate (IG or MOS) FET and JFET; however, those types will not be covered in this section.

The first microwave transistors were fabricated using the interdigitated geometry. These were followed by the overlay metal matrix, sometimes called mesh or emitter grid, and the FET. Each of these technologies represents a distinct advancement of the state of the art. Also, each was developed with a certain end result in mind.

The following discussion explains the development, applications, advantages, and disadvantages of the four basic microwave transistor types: interdigitated, overlay, metal matrix, and GaAs FET. Also discussed are the refinements of each technology with respect to the materials, design optimization, and manufacturing methods.

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6.3.1.1 Device geometries.

Interdigitated. Devices with interdigitated geometry are still in the vast majority. The primary reasons for this are relative ease of fabrication and the broad spectrum of applications. The surface topography is shown in Figure 25.

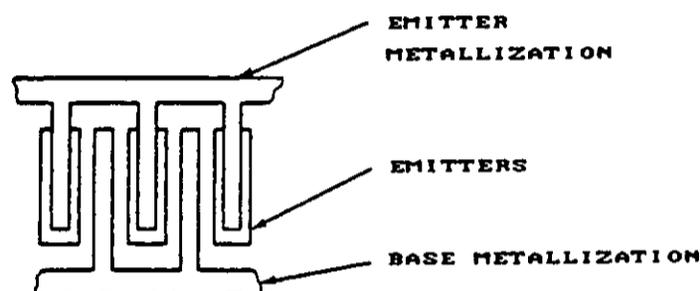


FIGURE 25. Surface geometry of an interdigitated transistor.

The greatest advantage of the interdigitated device is its inherently high ratio of emitter periphery to base area ($P_e:A_b$). This number is a figure of merit of transistors and is usually on the order of 3 to 5 cm^{-1} . The significance of this ratio is that it is an indicator of the device's gain-bandwidth product (f_t). Generally, the higher this ratio can be made, the higher the f_t of the device.

Through refinements of today's technology small signal interdigitated devices can achieve the following limits:

$$f_t = 8 \text{ GHz}$$

$$f_{\text{max}} = 25 \text{ GHz}$$

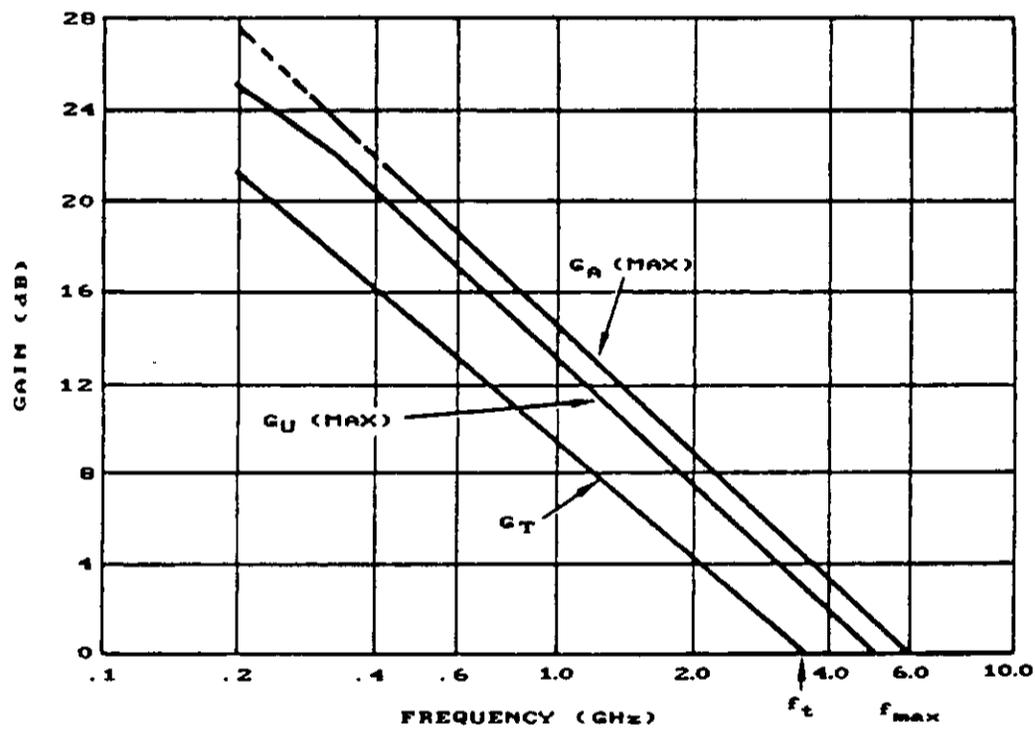
The difference between f_t and f_{max} is as follows: f_t is the frequency where the transducer power gain $|S_{21}|^2$ (known as G_T) goes to 0 dB, while f_{max} is the frequency above which oscillation cannot occur (Figure 26).

These values were significantly improved by the development of different diffusion technologies. Both arsenic-diffused emitters and ion implantation will be discussed later in this section. Basically, the key to improving f_t lies in the ability to control the depth of the base diffusion, the final base width, and the uniformity of dopant concentration. These newer technologies both have significant advantages over the conventional phosphorus-doped emitter.

The noise figure is one of the most critical parameters of a small signal microwave amplifier. Key factors for keeping the noise figure to a minimum are minimizing the collector-base capacitance and careful consideration of

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circuit design. Generally, for microwave devices, the noise figure is lower in the common emitter configuration than common base. Negative feedback through the collector-base capacitance tends to reduce the intrinsic noise figure. However, this effect will also reduce the forward gain G_T . Obviously, this trade-off must be considered in light of the application.



$G_U(\text{MAX})$ = Unilateral gain
 $G_A(\text{MAX})$ = Max available gain

FIGURE 26. Definition of f_t and f_{max} .

A typical noise figure versus frequency and emitter current for an npn bipolar transistor in the common emitter configuration is shown in Figures 27 and 28, respectively.

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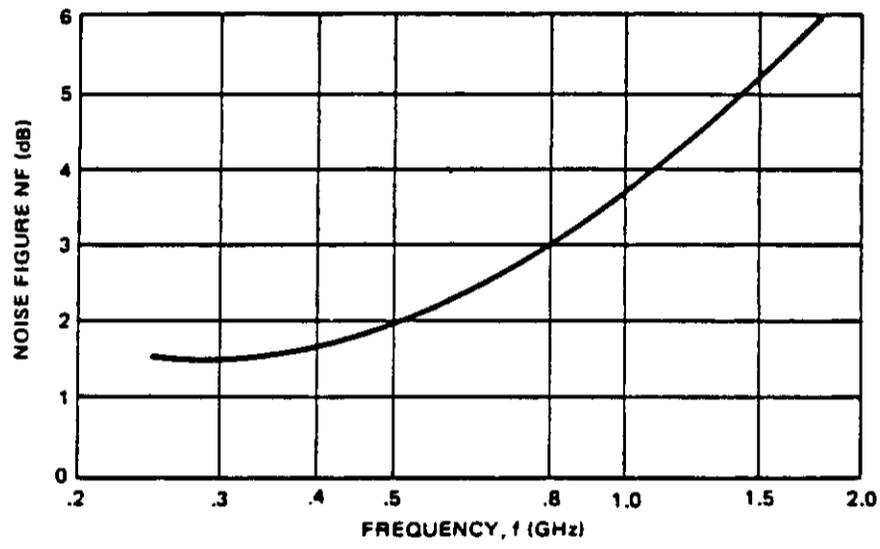


FIGURE 27. Noise figure vs frequency.

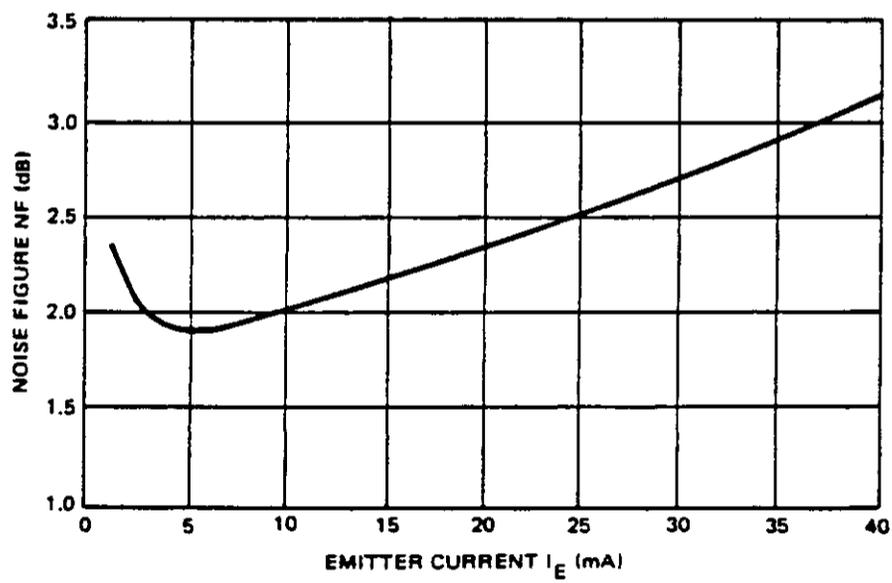


FIGURE 28. Noise figure vs emitter current.

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Overlay. The intrinsic ability of the overlay device to withstand greater current densities at a time when aluminum was the only metal in use for contact metallizing gave the industry a significantly more reliable product than any other yet produced.

The overlay geometry is shown in Figure 29 and the construction is shown in Figure 30. The details of fabrication will be covered fully in the section on fabrication and assembly. However, several features merit mention now.

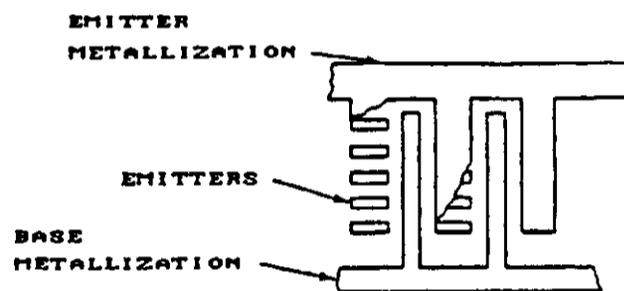


FIGURE 29. Overlay geometry.

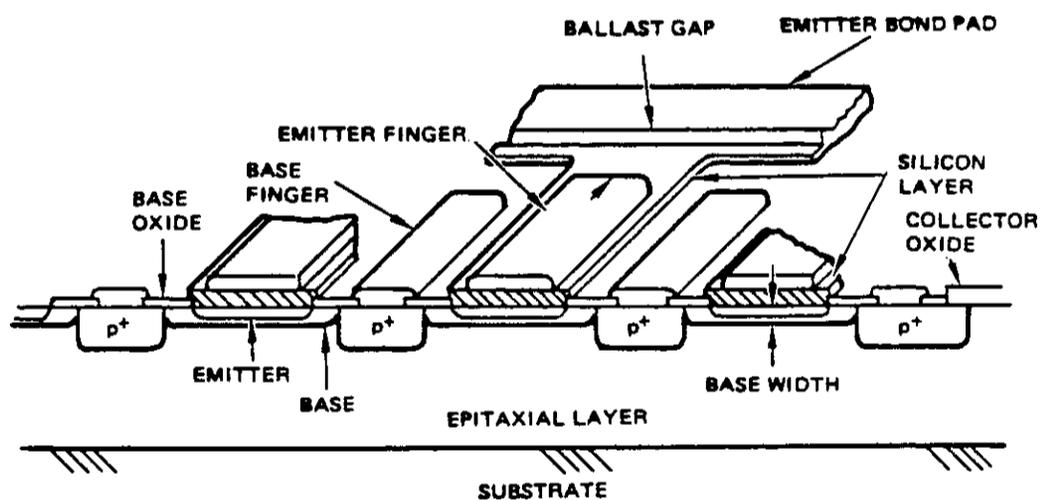


FIGURE 30. Typical construction of an overlay device.

The first is the layer of polycrystalline silicon under the emitter metallization. This technique was developed in an effort to hinder the spiking of aluminum through the device to the base and collector.

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At first, the silicon was sputtered at high temperatures ($>1000\text{ }^{\circ}\text{C}$). However, this caused further diffusion of the emitter. A method for depositing the silicon near room temperature was eventually discovered. This silicon barrier vastly improved the ruggedness of the overlay device.

The second feature was the development and use of ballast resistors on overlay devices. The purpose of the ballast resistor was to equalize the current in each emitter finger. (Note--in an overlay device, an emitter finger contacts many emitter sites as opposed to the fingers of an interdigitated device). Thus, more current would tend to flow in fingers closer to the bonding pad, causing potential overheating problems, whereas the ballast resistor design would evenly distribute the current before it could cause any thermal damage.

The material used for ballast resistors is most often nichrome, because it has higher resistivity than most contact metals. The formation of the ballast resistor is accomplished by depositing the nichrome in one long strip so that the fingers are at 90 degree angles to it. Doped silicon has also been used for this purpose.

The ballast technique is now used on practically every microwave power device regardless of geometry. Thermal effects in ballasted and unballasted devices will be discussed later in this section.

Metal matrix. The metal matrix geometry can best be described as a sibling to the overlay geometry. However, several distinct advantages are evident. This geometry affords the highest ratio of emitter periphery to base area of any other fabrication technology. Thus, the upper frequency range can be extended considerably. Power and current handling capability also remains high. The surface geometry of the metal matrix device is shown in Figure 31.

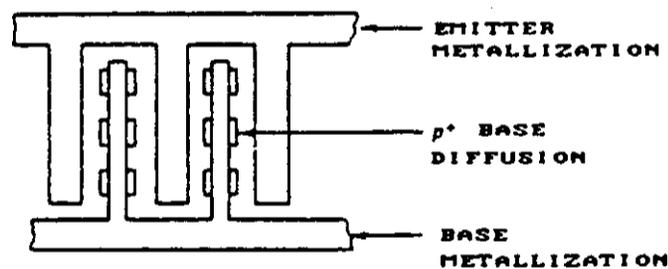


FIGURE 31. Metal matrix geometry.

The matrix structure achieves its high emitter periphery due to the fact that the emitter runs lengthwise across the chip as opposed to the interdigitated configuration.

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The wide metal fingers enhance the current handling capability of the device. Even where aluminum metallization is being used, the increase in metallization cross sectional area significantly reduces the rate of aluminum migration. Where refractory metals are used, migration is virtually eliminated as a problem.

These advantages are somewhat offset by the extreme difficulty in processing the matrix transistor. The accuracy with which the regions are aligned is probably the most crucial factor, along with precise diffusion control.

6.3.1.2 GaAs Schottky barrier field effect transistor (FET). To date, the most successful microwave field effect transistors are the n-type GaAs metal semiconductor field effect transistors. Devices can be fabricated that meet the following specifications:

Unilateral power gain (G_U) > 25 dB at 1 GHz (See paragraph 6.3.5 for a definition of G_U .)

f_{max} > 100 GHz

Noise factor < 3dB at 4 GHz

Probably the greatest initial problem in the development of these devices was the difficulty in controlling the thickness of the epitaxial layer. Factors affecting this thickness include dopant concentration, source temperature, and substrate temperature. Experimentation with these variables resulted in the capability to reduce the growth rate drastically. However, other problems such as poor vapor etching arose at lower source temperatures.

The orientation of the substrate determined whether the epitaxy was <100> or <111>. Although the <111> structure has a more controllable growth rate, <111> contributed to low yields in the scribe and break operation. Therefore, the <100> structure is being used almost exclusively.

Performance of the GaAs FET is extremely dependent on geometry. The gain-bandwidth product, f_t , is effectively maximized by reducing the dimensions of the gate and gate-source/gate-drain spacing. Initial configurations included an "enclosed drain" mesa geometry and an interdigitated mesa geometry.

Dimensions associated with the gate were kept in the 2- to 4-micron region. Figure 32 shows f_{max} and G_U as a function of gate length for various values of pinch-off voltage, V_p . These predicted values, however, could not be obtained with the interdigitated geometry due to two major problems: 1) formation of parasitic transistors and 2) resistors in the fingers of the interdigitated geometry devices.

Redesign of the enclosed drain geometry was pursued. A more refined geometry evolved that essentially eliminated the more serious problems. A comparison of performance of the interdigitated and enclosed drain designs is shown in Figure 33.

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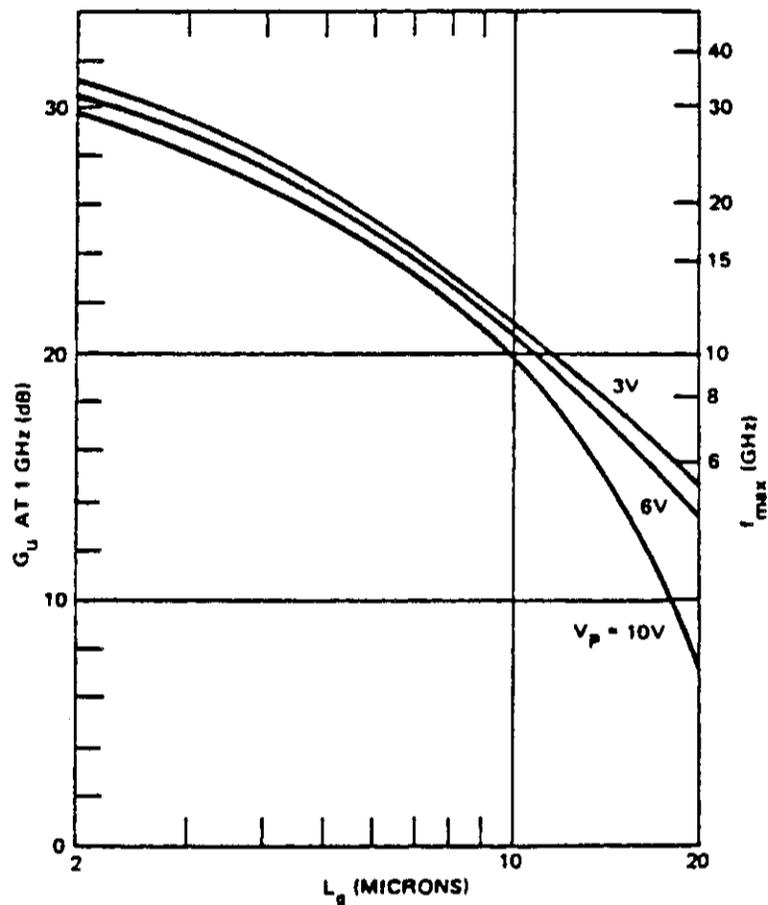
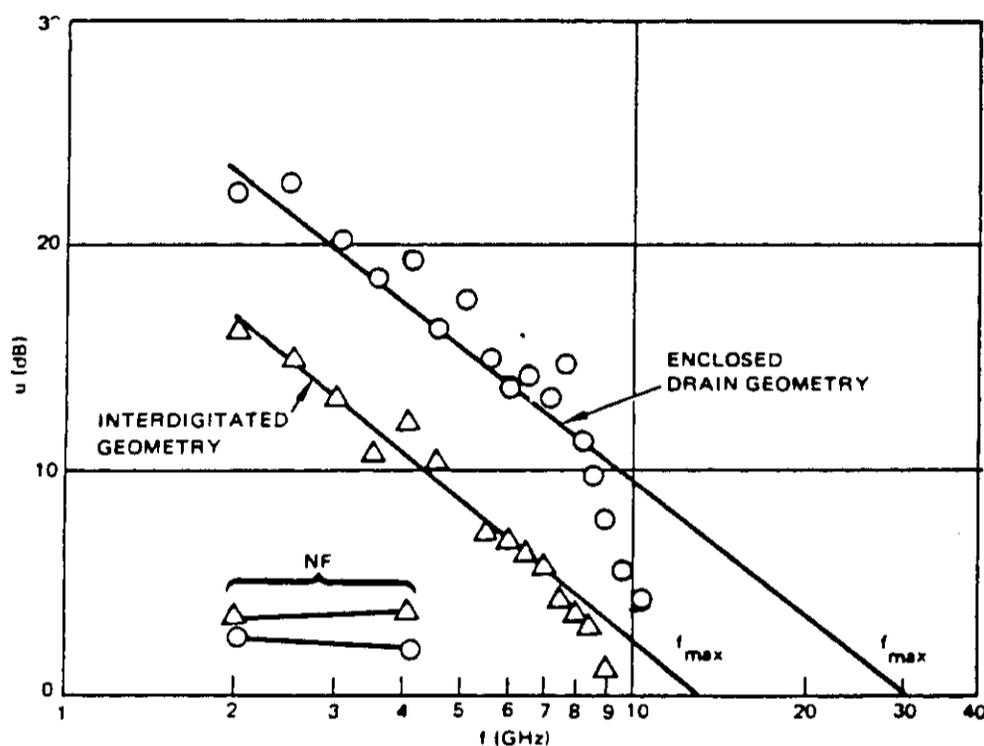


FIGURE 32. Performance prediction for the GaAs FET as a function of gate length.

6.3.1.3 Thermal considerations. Microwave power transistors, if they are to have a useful mean lifetime, must have an adequate thermal design. Thermal efficiency must be taken into account at dc levels and at the frequencies of operation.

From a dc standpoint, the most critical fabrication process is the attachment of the transistor chip to the header. The formation of the silicon-gold eutectic, so that 100 percent of the underside of the chip is wetted, ultimately determines the dc thermal capability of the device.

In many instances, tiny voids in the eutectic are created. These voids eliminate the heat transfer path that would be present if the eutectic had been properly formed. The rapidity of device degradation under dc load is proportional to the percentage of voided area under the chip.

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$V_{DS} = +6 \text{ V}$, $I_{DS} = 40 \text{ mA}$

FIGURE 33. Actual performance of the two GaAs FET models.

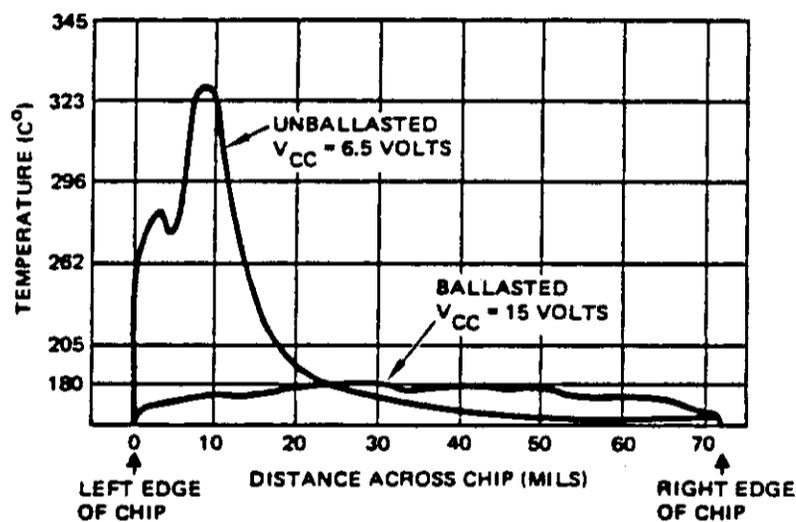
The reason for this is quite simple. The thermal impedance, θ_{jC} , of a void (hot spot) can be on the order of 20 to 50 °C/W. The measured θ_{jC} of the entire device is an average value, perhaps about 5.0 °C/W. The current density around the void raises the temperature in that vicinity much higher than the average junction temperature. This phenomenon is self-accelerating and is commonly known as thermal runaway.

The value of ballast resistors for a good thermal design has already been mentioned. This is applicable to any microwave transistor geometry. The equalizing of current in each metal finger is a great aid in slowing thermal runaway.

Thermal analysis of a transistor chip can be performed with great accuracy through the use of an infrared scanning microscope. A thermal profile, such as the one in Figure 34, can give a direct indication of the temperature anywhere on the surface of the chip. Figure 34 also shows the advantage of ballast

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resistors in keeping the junction temperature at a significantly lower level than that of an unballasted device. The resulting increase in safe operating area can be seen in Figure 35.

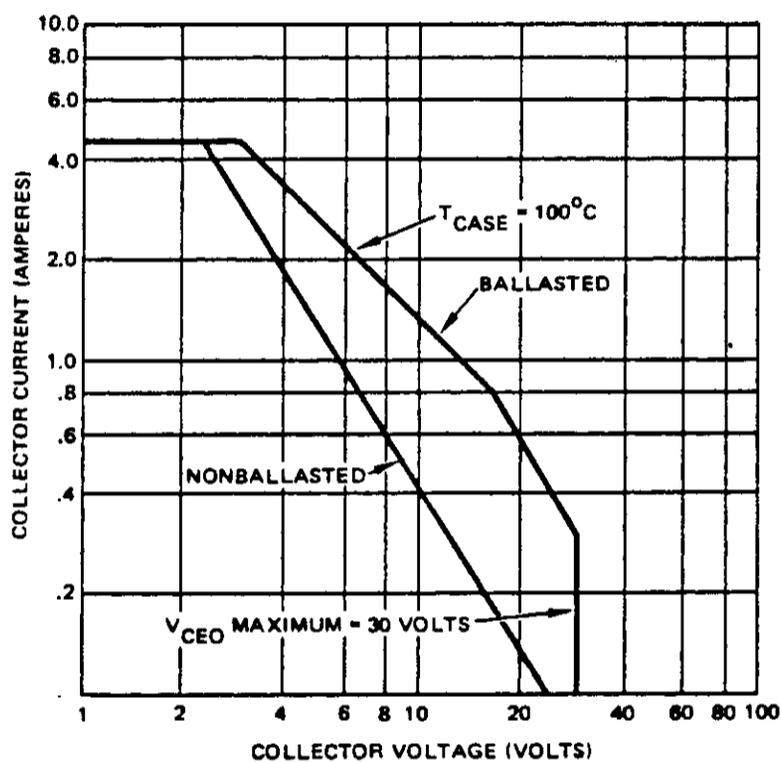


$P_{DISS} = 13 \text{ W}$
 $T_{CASE} = 100 \text{ }^{\circ}\text{C}$

FIGURE 34. Thermal profiles of a ballasted transistor vs an unballasted transistor.

It was previously mentioned that conventional θ_{JC} measurements yielded the average value for the device. The IR technique offers the unique ability to determine what can be called the hot spot, θ_{JC} . This means that the θ_{JC} is calculated from the temperature of the hottest spot on the chip. The advantage of this is a conservative specification of a device for a longer mean lifetime.

Adverse thermal effects at rf levels are generally attributed to the amount of mismatch (VSWR) the output of the transistor encounters. Figure 36 shows the variation in collector efficiency, hot spot temperature, and hot spot θ_{JC} over a range of frequencies for two values of output VSWR. Power reflected back into the transistor at high VSWRs causes excessive dissipation. This raises the average junction temperature as well as the hot spot temperature, causing an overall loss in collector efficiency.



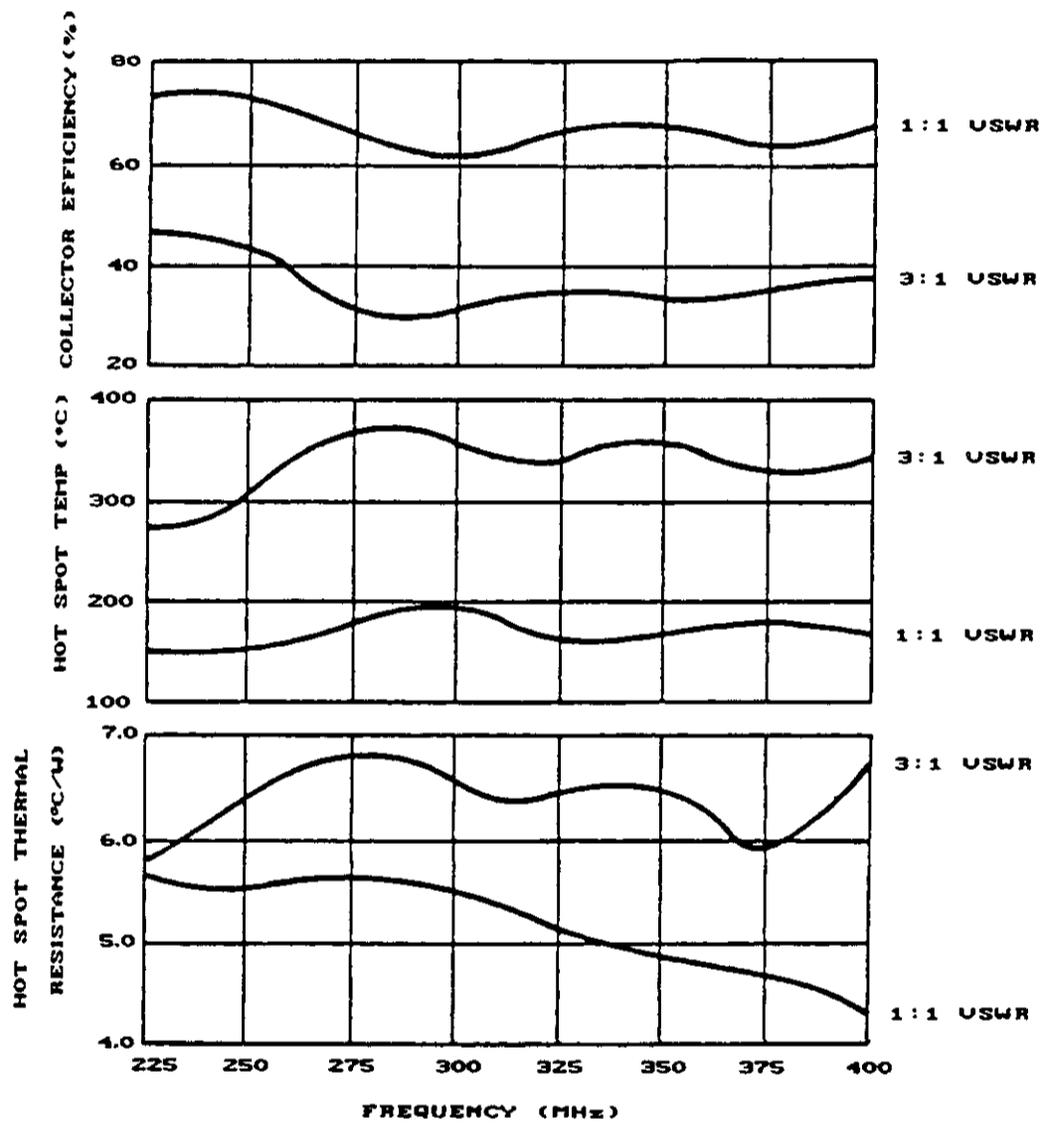
$T_{HOT \text{ SPOT}} = 200^{\circ}C$ maximum
 I_C maximum = 4.5 A

FIGURE 35. Safe operating area ballasted vs unballasted.

6.3.1.4 Metallization systems. Since the beginning of the semiconductor age, the contact metallization for transistors has classically been aluminum. In general applications, virtually no shortcomings could be cited. Basic advantages are as follows:

- Excellent conductivity
- Easily sputtered to contact areas
- Formed good eutectic with silicon
- Easily bonded to
- Adheres well to SiO_2 .

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VCC = 28 V
 T_{CASE} = 85 °C
 P_{OUT} = 25 W at VSWR of 1:1

FIGURE 36. Hot spot thermal resistance, hot spot temperature and collector efficiency as a function of frequency.

The discovery of aluminum's propensity to migrate in masses into silicon under certain conditions caused a great deal of alarm in the industry.

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Aluminum migration as a failure mechanism was not discovered until power devices in the 100 MHz region were in general usage. Because the migration is a function of current density and microwave devices use extremely fine geometries, the conditions for failure were perfect. Since then it has also been discovered that junction temperature and time also play an important role. J.R. Black, in 1969, related all of these factors by the equation

$$MTBF = \frac{1}{AJ^2 e^{-\phi/kT}}$$

where:

MTBF = Mean time between failure (hours)

A = a constant containing a factor equal to the cross sectional area of the metallization

J = current density (amperes/cm²)

ϕ = activation energy (ev)

k = Boltzman's constant

T = metallization temperature (K)

This left only two possible choices: redesign the devices to lower the current density or redesign using a metal which has less of a propensity to migrate. In actuality, both of these alternatives were used.

The conclusion reached by J. R. Black and widely accepted in the rf semiconductor industry is the following:

For an unpassivated device with

$$\begin{aligned} T_J &= 200 \text{ }^\circ\text{C} \\ A &= 10^{-7} \text{ cm}^2 \end{aligned}$$

the current density must be kept under 1×10^5 amperes/cm² for a 2-year mean lifetime. For higher current densities, the maximum rated junction temperature must be reduced accordingly.

Development of other metal systems has taken place and many are in use. Some examples are:

- a. Molybdenum-gold
- b. Nichrome-tungsten-gold
- c. Platinum silicide-tungsten-gold
- d. Chromium-silver-gold.

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All of these metals are heavy, or refractory. In essence, they are capable of resisting mass transport via momentum exchange due to their high atomic weight. The majority of products today employ these metallization systems.

In summing up the introduction to microwave transistors, it is clear that, of the many current and new technologies discussed, each has its relative merits and shortcomings. The metal matrix geometry is widely used in power applications due to its higher frequency limit. For small signal applications, the only feasible conventional geometry is the interdigitated geometry. Low-current requirements and high-frequency capability ($f_t < 6$ GHz) are the primary reasons. However, if the application requires amplifying small signals up to the frequency of x band and above with a low noise figure, the GaAs MESFET is the best possible three-terminal device.

Device failure due to metal problems can be considerably reduced through the use of refractory metals such as gold, tungsten, and chromium instead of aluminum. In addition, power devices should use emitter ballast resistors to equalize current sharing. The fact that metal failure mechanisms are self-accelerating make these requirements mandatory. These technologies are currently being used in high reliability microwave devices.

6.3.2 Usual applications. The primary use of microwave transistors is in strip transmission line amplifier and oscillator applications for low-noise receiver front ends, transmitters, and local oscillator operation. Figure 37 shows a microwave transistor-amplifier circuit with input and output matching networks. The unilateral power gain is the sum of the additional power gain (or loss) resulting from the input impedance matching network between the device and the source, the forward power gain of the device with the input and output terminated in matching loads, and the additional power gain (or loss) due to the impedance matching network between the output of the device and the load: $G_u = G_s + G_f + G_e$ in dB.

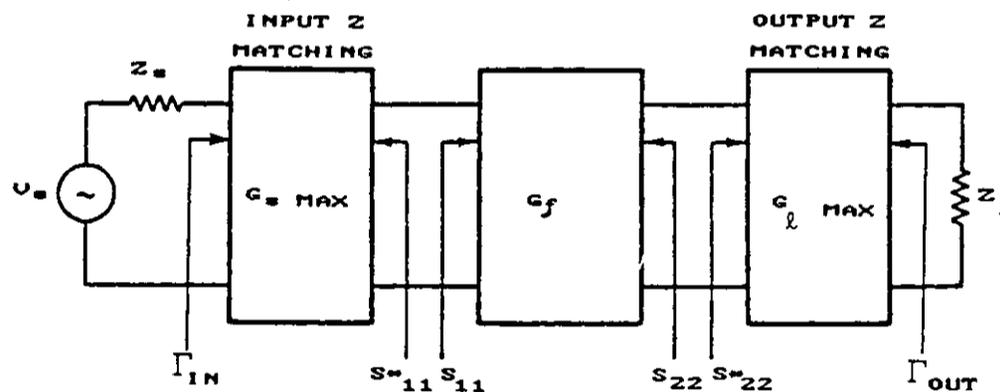


FIGURE 37. Amplifier circuit for maximum gain.

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The maximum unilateral gain will be obtained by choosing impedance matching networks so the reflection coefficients of the source and the load are equal to the complex conjugates of the input and output reflection coefficients of the device, respectively: $\Gamma_s = S_{11}^*$ and $\Gamma_L = S_{22}^*$, $Z_s = Z_0$, $Z_L = Z_0$.

If a transistor amplifier is to be unconditionally stable, the magnitudes of S_{11} , S_{22} , Γ_{in} , and Γ_{out} must be smaller than unity and the transistor's inherent stability factor K must be greater than unity and positive. K is computed from

$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}S_{21}|} > 1$$

where $\Delta = S_{11}S_{22} - S_{12}S_{21}$

the input and output reflection coefficients are given by

$$\Gamma_{in} = S_{11} + \frac{S_{21}S_{12}\Gamma_L}{1 - S_{22}\Gamma_L}$$

and

$$\Gamma_{out} = S_{22} + \frac{S_{21}S_{12}\Gamma_s}{1 - S_{11}\Gamma_s}$$

Also, the boundary conditions for stability are given by

$$|\Gamma_{in}| = 1 = \left| S_{11} + \frac{S_{21}S_{12}\Gamma_L}{1 - S_{22}\Gamma_L} \right|$$

and

$$|\Gamma_{out}| = 1 = \left| S_{22} + \frac{S_{21}S_{12}\Gamma_s}{1 - S_{11}\Gamma_s} \right|$$

Substitution of real and imaginary values for the S parameters in the previous two equations and solving for Γ_L and Γ_s yields

$$R_s \text{ (radius of } \Gamma_s \text{ circle)} = \frac{|S_{12}S_{21}|}{|S_{11}|^2 - |\Delta|^2}$$

$$C_s \text{ (center of } \Gamma_s \text{ circle)} = \frac{C_s^*}{|S_{11}|^2 - |\Delta|^2}$$

$$R_L \text{ (radius of } \Gamma_L \text{ circle)} = \frac{|S_{12}S_{21}|}{|S_{22}|^2 - |\Delta|^2}$$

$$C_L \text{ (center of } \Gamma_L \text{ circle)} = \frac{C_L^*}{|S_{22}|^2 - |\Delta|^2}$$

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where

$$\Delta = S_{11}S_{22} - S_{12}S_{21}$$

$$C_s = S_{11} - \Delta S_{22}^*$$

$$C_l = S_{22} - \Delta S_{11}^*$$

The reflection coefficient of the source impedance required to match the input of the transistor conjugately for maximum power gain is

$$\Gamma_{sm} = C_s^* \frac{B_s \pm \left(B_s^2 - 4|C_s|^2 \right)^{1/2}}{2|C_s|^2}$$

where $B_s = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2$

The reflection coefficient of the load impedance required to match the output of the transistor conjugately for maximum power gain is

$$\Gamma_{lm} = C_l^* \frac{B_l \pm \left(B_l^2 - 4|C_l|^2 \right)^{1/2}}{2|C_l|^2}$$

where $B_l = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$

If the computed values of B_s and B_l are negative, the plus sign should be used in front of the radical in the equations for Γ_{sm} and Γ_{lm} . Conversely, if B_s and B_l are positive, the negative sign should be used.

Stability circles can be plotted directly on a Smith chart. These circles separate the output or input planes into stable and potentially unstable regions. A stability circle plotted on the output plane indicates the values of all loads that provide negative real input impedance, thereby causing the circuit to oscillate. A similar circle can be plotted on the input plane which again indicates the values of all loads that provide negative real output impedance and causes oscillation. A negative real impedance produces a reflection coefficient that has a magnitude greater than unity. The regions of instability occur within the circles whose centers and radii are R_s , C_s , R_l , C_l .

By using an appropriate sign, only one answer is possible in either of the equations for Γ_{sm} or Γ_{sl} and a value of less than unity is obtained. The maximum available power gain possible is expressed as

$$G_{max} = \frac{|S_{21}|}{|S_{12}|} |K \pm (K^2 - 1)^{1/2}|$$

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Hence, maximum available power gain is obtained only if the microwave transistor amplifier is loaded with Γ_{sm} and $\Gamma_{\ell m}$ as reflection coefficients. The maximum frequency of oscillation is determined after the maximum available power gain is achieved.

It is possible to design more than one stage of amplification to achieve high gain and gain flatness over a specified bandwidth (Figure 38A). However, for multistage amplifiers where gains of 30 dB or greater are required, the use of such "single-ended" designs can lead to difficulty in producing the required gain flatness and distortion. This is because of the mismatches occurring between gain stages causing gain and phase variations. It is possible to minimize mismatches by designing interstage networks that will match the output impedance of the previous stage to the input impedance of the following stage and provide a positive sloped gain over some bandwidth to compensate the transistor's roll-off (Figure 38C) and give an overall flat gain (Figure 38E). Isolators can be added to the input, output, and between stages. Isolators can produce significant loss which will increase the noise figure.

A balanced amplifier shown in Figure 39 can operate with both high gain and wide bandwidth. Two FET amplifier units of the same performance are arranged between the output and input ports of two 3-dB, 90-degree hybrid dividers or two Wilkinson power dividers with 90-degree phase shifters.

The input and output VSWRs are predominately those of the hybrid dividers. If the individual amplifiers are not perfectly matched at certain frequencies, a signal in the zero degree arm of the divider will be reflected from the corresponding amplifier and a signal in the $\pi/2$ arm of the divider will be similarly reflected from its amplifier. The signals will be phased by 0 or $\pi/2$ again by traveling through the divider after reflection, and the signals will be fed to the load at the isolated port.

The total reflection and power transmission can be expressed as

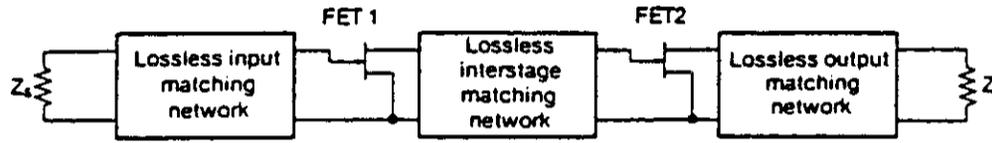
$$S_{11} = 1/2 (S_{11a} - S_{11b})$$

$$S_{22} = 1/2 (S_{22a} - S_{22b})$$

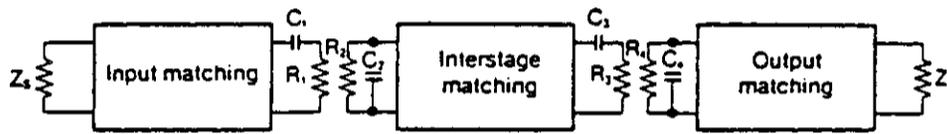
and
$$\text{Gain} = |S_{21}|^2 = 1/4 |S_{21a} + S_{21b}|^2$$

where a and b indicate the two GaAs MESFETs. The 1 and 2 refer to the input and output ports, respectively. The input and output reflections are reduced to one-half of the corresponding difference of reflection coefficients of the two MESFETs, and the total power gain is equal to the individual power gain if the two MESFETs are identical.

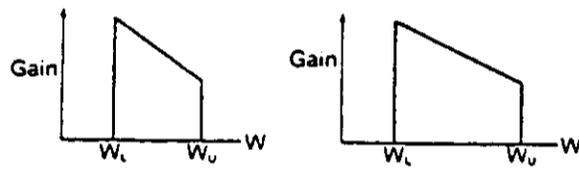
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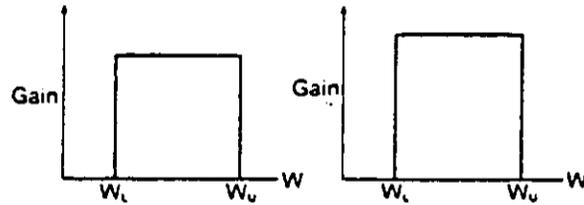
A. Amplifier schematic



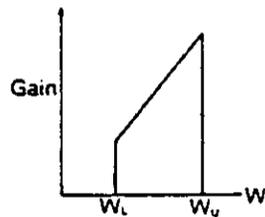
B. FET impedance models



C. Frequency response of transistors

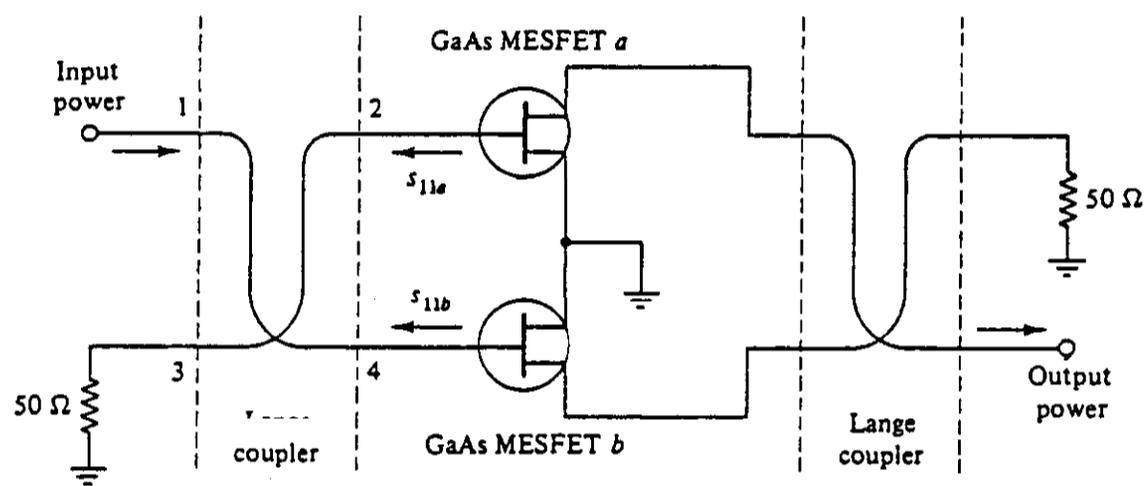


D. Frequency response of input and output matching networks

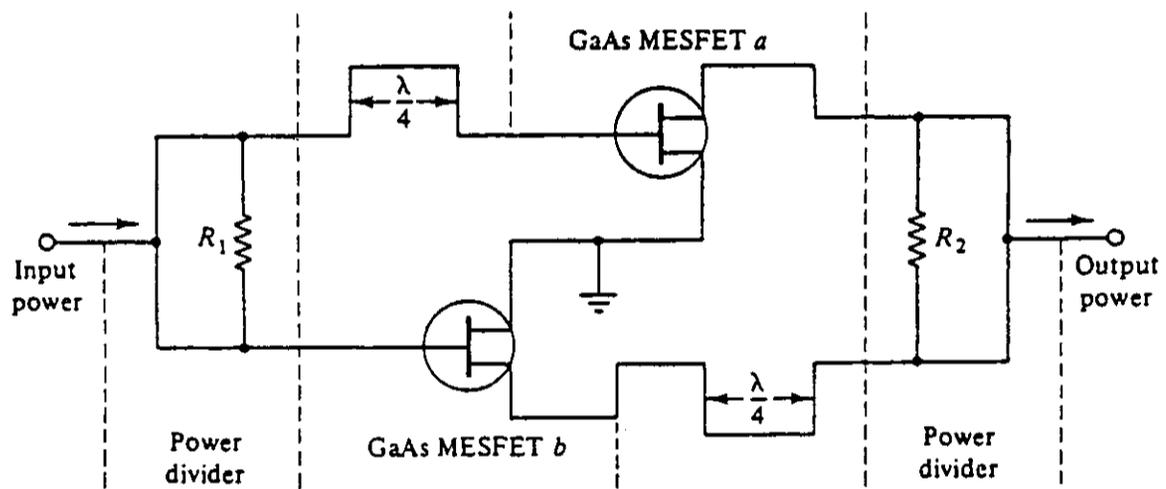


E. Frequency response of interstage

FIGURE 38. Multistage amplifier with input, output, and interstage matching network.



A. Balanced amplifier with Lange couplers



B. Balanced amplifier with Wilkinson power dividers and phase shifters

FIGURE 39. Two types of balanced amplifier circuits.

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If one MESFET fails, the loss for the balanced amplifier would be only 6 dB. The balanced amplifier configuration is by far the most common design in modern microwave integrated circuits. Its advantages are:

- a. Good input and output VSWRs
- b. Good stability
- c. High reliability
- d. Low tuning work
- e. Linear output power increases by 3 dB
- f. Up to 20 dB power gain with 10 percent bandwidth at X band.

Both silicon bipolar transistors and GaAs MESFETs can be designed as an amplifier or an oscillator, depending on whether the stability factor K is greater or less than unity. Common-emitter or common-source configuration is preferred for amplifier design and, alternatively, common-base or common-gate configuration for oscillator design. The stability factor K is greater than unity for an amplifier, but the input and output reflection coefficients (Γ_{in} and Γ_{out}) are less than unity. For an oscillator, however, the stability factor K is less than unity, but the input and output reflection coefficients are equal to or greater than unity. In amplifier design, two matching networks are required to match the input and output ports of the device to give a maximum transducer power gain, a minimum noise figure, and low input and output reflection coefficients. In a small-signal oscillator design the same two-port (or one that is properly modified with a feedback path) can be designed to deliver nearly the same output power to the same 50- Ω load.

Two-port oscillator design may be summarized as follows:

- a. Select transistor with sufficient gain and output power capability for the frequency of operation. This may be based on oscillator data sheets, amplifier performance, or S-parameter calculation.
- b. Select a topology that gives $k < 1$ at the operating frequency. Add feedback if $k < 1$ has not been achieved.
- c. Select an output load matching circuit that gives $|S'_{11}| > 1$ over the desired frequency range. In the simplest case, this could be a 50- Ω load.

$$|S'_{11}| = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

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- d. Resonate the input port with a lossless termination so that $r_G S'_{11} = 1$. The value of S'_{22} will be greater than unity with the input properly resonated.

$$S'_{22} = S_{22} + \frac{S_{12}S_{21}r_G}{1 - S_{11}r_G}$$

The transistor will oscillate with any of the six configurations given in Figure 40. In all cases, the transistor delivers power to a load and the input of the transistor. Practical considerations of circuit design and dc biasing will determine the best design. A bipolar transistor oscillator circuit is shown in Figure 41.

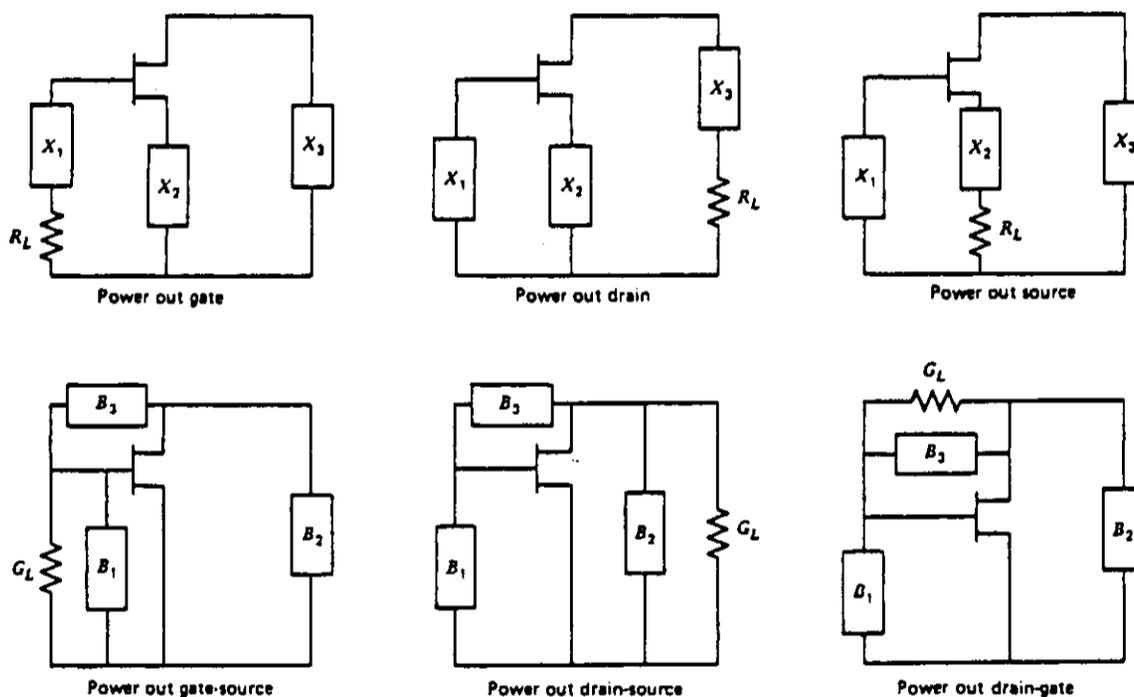


FIGURE 40. Six oscillator structures.

Microwave transistors can be used as mixers as shown in subsection 6.8. The non-linearities of bipolar transistors and FETs that produce mixing are used in frequency multiplier applications. The GaAs MESFET and, in particular, the dual-gate FET work exceptionally well. The major advantage over SRD multipliers, which is the greater than 100 percent efficiency, results in the replacement of complicated varactor diode chains and resultant pre- and post-amplification. Table III compares the performance of varactor diode, bipolar transistor, and FET multipliers.

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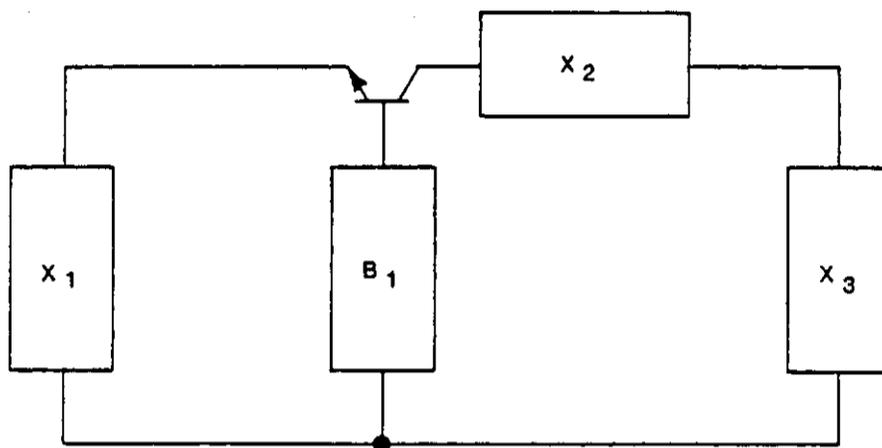


FIGURE 41. Bipolar transistor oscillator circuit.

TABLE III. Comparison of microwave frequency multipliers

Device Characteristic	Diodes	BJT	FET ^{1/}
Bandwidth	Narrow	Medium	Wide
RF driving power	300 mW	100 mW	1 mW (10 mW)
Output maximum frequency	SRD: 18 GHz Varactor: 120 GHz	11 GHz	30 GHz
Isolation	Poor	Medium	Good
Idlers	Critical	Critical	Less Critical
Power handling (X-band)	1 to 4W	0.5W	1 to 4 W
Stability	Good	Poor	Excellent
Higher harmonic distortion	High	Low	Low
RF efficiency	-1.5 to -3 dB	-2 dB	Up to +10 dB (up to 3 dB)

^{1/} Figures in parentheses are results for a single-gate FET.

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6.3.3 Physical construction.

6.3.3.1 General. Microwave transistor fabrication hinges most critically on the three items in the introduction of this subsection. It is the accuracy with which these processes are conducted which has the most profound effect on the electrical characteristics of the final device. These will be discussed one by one, paying particular attention to the problems involved in microwave transistor fabrication.

Masking. The function of a photo mask is to define an area on a silicon wafer which is to be subsequently etched. In concept the process is very simple.

- a. The wafer is coated with approximately 4000 Å of photo resist. A layer this thin is achieved using centrifugal action.
- b. The photo mask is placed over the wafer, and light passing through the clear areas of the mask hardens the photo resist by polymerization.
- c. The pattern is further developed and hardened so it will withstand a chemical etch. The etching procedure removes the unwanted, unexposed regions of the wafer.

In practice, several difficulties must be settled. First, there is the problem of producing masks for the extremely fine geometries necessary for microwave transistors. The fingers of a small signal interdigitated device in the 1- to 3-GHz region are on the order of 1.0 to 2.5 μ wide. In addition, the tolerances required for subsequent operations complicate matters even more. For instance, the alignment of emitters in an overlay device requires a registration accuracy of approximately 1 μ.

To achieve these levels of accuracy, many refinements in equipment and technology were required. Today, photolithography and self alignment can produce reliable widths to submicron dimensions in association with ion implantation. Direct writing techniques can increase accuracy and further reduce the critical dimensions.

Diffusion and dopants. The key factor regulating the f_{max} of a microwave transistor is the emitter-collector transit distance or base width. The narrower this region, the higher the f_{max} . Typical base widths are on the order of 1,000 to 2,000 Å. Ability to control diffusion depth from one run to the next is essential to the production of relatively homogeneous devices.

The dependence of the gain-bandwidth product, f_t , on the emitter-collector transit distance can best be shown by the equation:

$$\tau_{ec} = \frac{kT C_{te}}{qI_e} + \frac{W^2}{nD} + \frac{x}{2V} + C_{cr}c$$

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where

T_{ec} = emitter-collector transit time

$\frac{kT}{q}$ = constant, 26 mv at 300 K

I_e = emitter current

C_{te} = emitter transition capacitance

W = base width

n = doping profile constant

D = base carrier diffusion constant

X = collector space charge width

v = carrier velocity

C_c = collector-base capacitance

r_c = collector series resistance

This equation can also be written as

$$T_{ec} = T_c + T_b + T_{sc} + T_e$$

where

T_e = emitter charging time

T_b = base transit time

T_{sc} = collector space charge transit time

T_c = collector charging time.

It can be seen that each term in the equation has either distance or capacitance as the independent variable. Capacitance is extremely dependent on geometry and materials, whereas distance is controlled by the diffusion. Therefore, the key to producing devices with the highest possible f_t depends largely on the controls established for the diffusion process.

Diffusion, in practice, is an extremely delicate operation, especially when final base depths of only a few thousand angstroms are desired. Subsequent high temperature operations will cause the initially diffused base to be further diffused in both a downward and lateral direction.

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Emitter diffusion also has its problems. The phenomenon of "emitter push," that is, the deepening of the base diffusion directly under the emitter diffusion, destroys the f_{max} characteristic of the device.

Recently, many advances have been made in the technology and materials used in the diffusion process. The remainder of this section will be devoted to discussing them and their relative merits.

Ion implantation is a technique through which ions are driven into substances due to energy received from particle accelerators. Obvious advantages of this technology are the great degree of control over implantation depth, because the depth is determined by the energy level of the ions (Figure 42) and secondly, the ability to obtain extremely fine line widths. From Figure 42, another obvious advantage can be seen. That is the absence of lateral diffusion for an implanted junction.

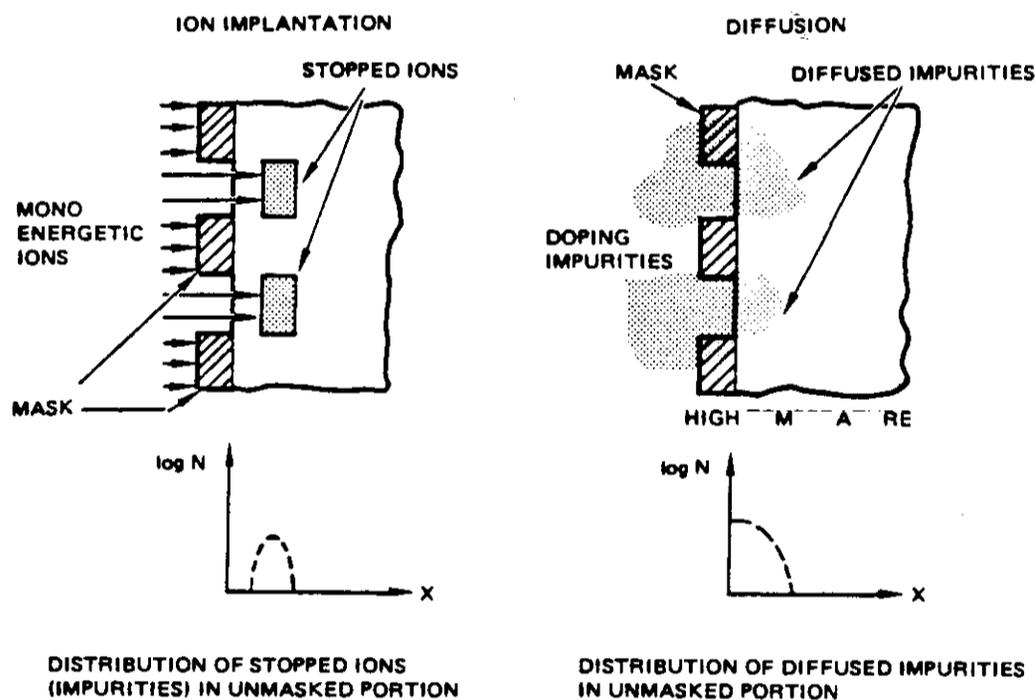


FIGURE 42. Impurity profiles--ion implantation vs diffusion.

The major drawback of ion implantation for use in microwave transistor fabrication is the possible damage of the lattice structure by radiation. Though ion implantation has definite possibilities, it seems that more favorable results are being obtained by using arsenic as an emitter dopant.

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The classic n-type dopant used in transistor fabrication is phosphorus. It is easily obtained and relatively simple to use. With the development of microwave transistors its use continued until it was discovered that many problems in device performance were linked directly to phosphorus. Both lateral diffusion and the "emitter push" phenomenon were associated with the properties of phosphorus itself or the processing conditions which had to be used.

The next n-type element on the periodic table is arsenic. Its use was initially avoided due to inability to obtain a satisfactory diffusion process. After a successful process was developed and experimental devices fabricated, the results of testing were conclusive. The heavier arsenic atom exhibited an unprecedented superiority over phosphorus. The depth of diffusion was controllable to such an extent that ion implantation efforts were temporarily tabled.

A brief summary of microwave transistor types and their relative merits and shortcomings is shown in Table IV.

TABLE IV. Summary of the four microwave technologies

Emitter Periphery	Inter-digitated	Overlay	Metal Matrix	GaAs FET
Base area (frequency capability)	Higher	High	Highest	Highest
Metal cross-sectional area	Low	High	High	Low
Processing	Easy	Easy	Difficult	Difficult
Chip size	Small	Variable	Large	Small
Current/power handling capability	Low	High	High	Low
Available MIL parts	Several	Several	None	None
Availability (no. of manufacturers)	All	Several	One	Many

When viewed from the more or less ideal situation depicted on a flow chart, the fabrication of a microwave transistor seems relatively simple. Mask tolerances, diffusion depths, oxide thicknesses, and other variables are tightly controlled so that the outcome is, theoretically, a high yield lot of homogeneous devices.

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However, real world forces can change the theoretical. Producing microwave devices on a repeatable basis is, in actuality, one of the most difficult tasks to accomplish. The product designer starts with a list of desired parameters, as follows:

- P_d Power dissipation
 - f_t Gain bandwidth product
 - S_{21} Forward transmission coefficient
 - S_{12} Reverse transmission coefficient
 - S_{11} Forward reflection coefficient
 - S_{22} Reverse reflection coefficient
- Noise figure.

From this the product designer must then decide the qualities of the required raw material. The primary considerations are the resistivity of the bulk silicon and the resistivity and thickness of the epitaxial layer that will be grown. Then, depending on the required limits on the electrical parameters, the designer calculates the optimum values for the following:

- Diffusion depths
- Dopant concentrations/profiles
- Oxide thickness
- Line widths/layout

After these are established, the designer generates a flow chart to insure process control.

In-process testing of the product at specified intervals is an integral part of this flow chart. In this manner, the product engineer can see how the established controls are affecting the parameters of the devices. (This testing is of prime importance because it affects percentage yield).

So the reader may develop a feeling for what is actually involved in processing a microwave transistor, the key steps will be described in detail. Because the interdigitated geometry is the simplest type and the most common, it will be used for illustrative purposes.

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6.3.3.2 Interdigitated device fabrication. The fabrication of an npn microwave device begins with the preparation of a slice of n^+ -type material. Initial slice thickness is approximately 15 mils and the resistivity is on the order of $0.01 \Omega\text{-cm}$. The crucial first step is the growth of a thin epitaxial layer of high resistivity n-type material, Figure 43A. Epitaxial growth is nominally $10 \mu\text{M}$ thick and has a resistivity of approximately $1\Omega\text{-cm}$. Exact values are determined by the desired parameters. For instance, devices with thinner epitaxial layers have higher values of f_t and lower collector-base breakdown voltages than devices with thicker epitaxial layers. True microwave devices, by our definition, are those devices that operate at 1 GHz or higher and must have this epitaxial growth. Without it obtaining the required values of f_t would be impossible.

On top of the epitaxial layer, silicon dioxide is grown to a thickness between 8000 and 12,000 Å, Figure 43B. This layer of oxide serves a dual function. First, the layer passivates the entire surface of the epitaxial layer. Secondly, in conjunction with photoresist, masking, exposure, and etching operations the layer allows for the definition of the base region, shown in Figure 43C. The opening of small windows through the oxide provides the means through which the p-type base dopant (usually boron) can be diffused into the epitaxial layer. For microwave devices, the initial depth of this diffusion is shallow, no more than a few thousand angstroms. The final depth of the base diffusion, which depends on the next process step, is instrumental in determining the f_t of the device.

Following the diffusion of the base region, another layer of silicon dioxide is grown, Figure 43D. This oxide passivates the entire base region. For the fabrication of microwave devices, however, this process has one definite disadvantage. As the silicon dioxide is being grown, further diffusion of the base is occurring simultaneously. This is caused by the elevated temperatures (approximately $1,000^\circ\text{C}$) required for oxide growth. At these temperatures thermal excitation moves the p-type impurity atoms in both a downward and lateral direction, as indicated by the two arrows in Figure 43D. The downward diffusion degrades the device's capability to attain the highest possible f_t because it effectively increases the distance through which charge carriers must travel from emitter to collector. Several techniques, such as plasma chemical vapor deposition, have been developed to grow silicon dioxide at reduced temperatures.

The emitter stripes and bonding pad are the next regions to be defined. Again, selective etching, following the usual photoresist, masking, and exposure cycle, creates windows through which the n-type emitter dopant (phosphorus or arsenic) can be diffused as shown in Figure 43E. Following the emitter diffusion, these emitter stripes also serve as the contact area for the emitter metallization. However, before the metallization process can take place, contact areas to the base region must be formed.

So, again, the procedure of photoresist, masking, exposure, and etching must be conducted to create the metallization contact areas for the base stripes and bonding pad. The base contact area is actually a p^+ region diffused into the p region.

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The final process, evaporation of metal (sputtering) deposits from 8,000 to 20,000 angstroms of metal to the contact areas, is shown in Figure 43F. For a more complete view of the device refer to Figure 44.

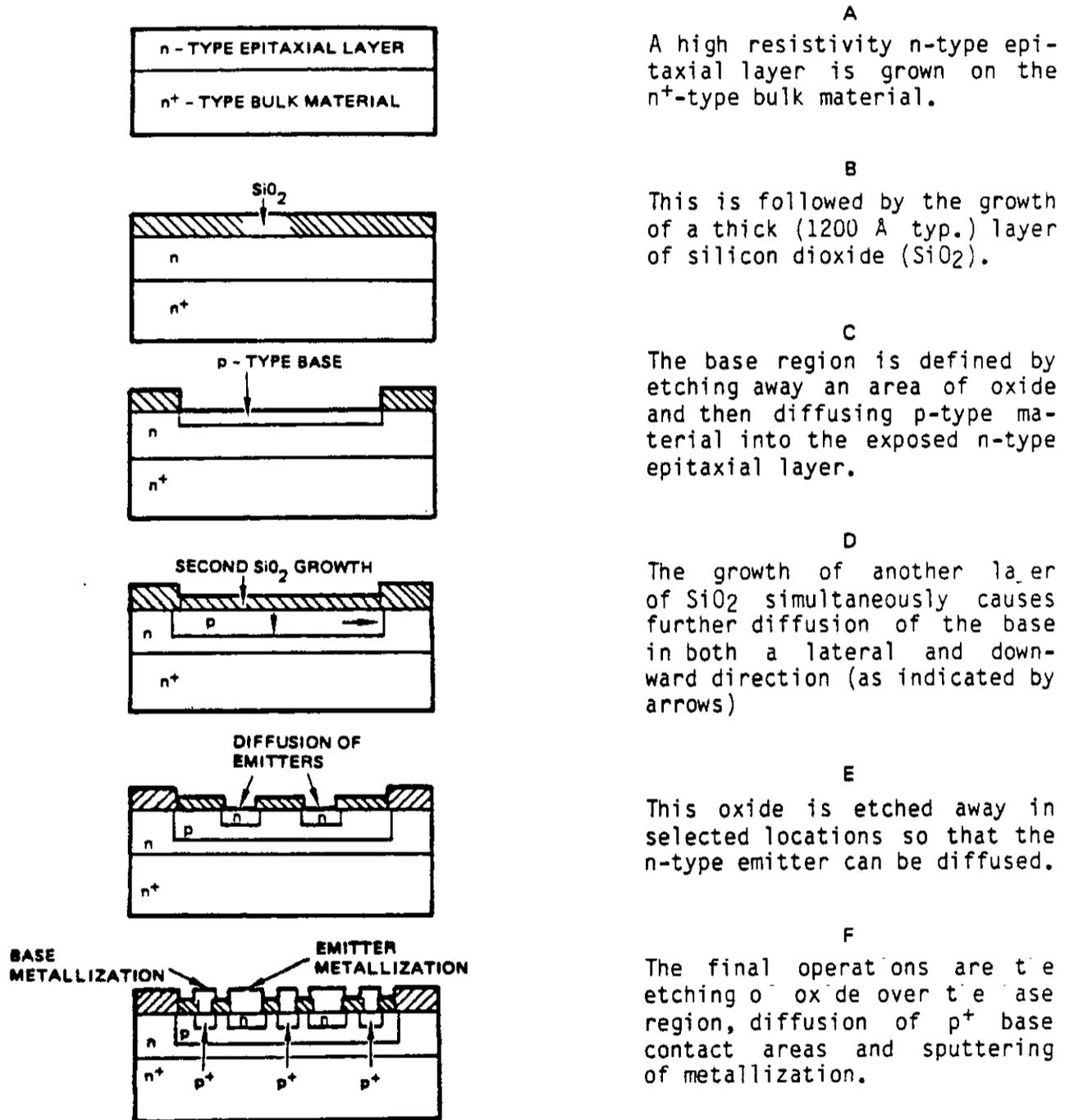


FIGURE 43. Interdigitated device fabrication.

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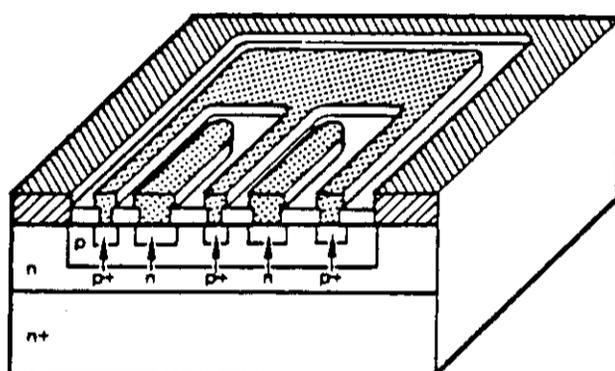


FIGURE 44. Cross-section of an interdigitated device.

The processes just described were done in an oversimplified manner. Though they are commonly used throughout the industry, the inherent difficulties of processing are always present. The processes should be examined in light of some of the necessary properties of a microwave transistor and the precision that must be maintained with respect to the following: epitaxial thickness, epitaxial resistivity, base diffusion depth, emitter diffusion depth, mask fabrication, masking alignment tolerances, emitter stripe width, metallization widths, and overall chip size.

The following variables that must be controlled throughout the entire fabrication operation should be considered: diffusion furnace temperatures, dopant concentration and flow rate, process time duration, etch rates, growth rates, and diffusion rates.

Each variable can shift the parameter distribution for a given production lot entirely out of specification. Hence, their control within precise limits is mandatory for product uniformity and repeatability.

6.3.3.3 Overlay device fabrication. The fabrication of the overlay transistor begins to differ from that of the interdigitated transistor during the diffusion of the p-type base dopant. The mask incorporated causes the dopant to be diffused in the shape of a grid.

Emitter sites are diffused precisely at the center of each square formed by the p-type base grid. Before the metallization process, the entire chip is passivated with polycrystalline silicon. Contact areas are then etched through the

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6.3.3.5 GaAs FET fabrication. The two techniques used for GaAs MESFET fabrication that have gained the most popularity are the self-aligned and the etched-channel technologies; both can exploit either optical or electron beam lithography to define the gate stripe, depending on the transistor's gate length. Electron beam lithography is used generally for gate lengths of less than 0.5μ .

Self-aligned gate technology. Figure 46 shows the basic processing steps of the self-aligned gate technology. The first step is defining the isolation mesa by etching away the active n layer until the semi-insulating substrate is reached. The gate metal, (i.e., aluminum) is then evaporated over the active area as shown in Figure 46A.

Source and drain areas are defined in photoresist and the exposed gate metal is removed by etching. Over-etching is used to undercut the resist, as shown in Figure 46B, to allow the necessary space between gate and drain and gate and source. Gold ohmic contact metallization, usually In-Ge-Au or Au-Ge-Ni, is then evaporated (Figure 46C). The resist, which is protecting the gate stripe, is now covered with this ohmic metallization, but this is conveniently removed by "floating off" the gold by dissolving away the resist. Thus, the remaining thin gate is left situated between source and drain contacts as shown in Figure 46D.

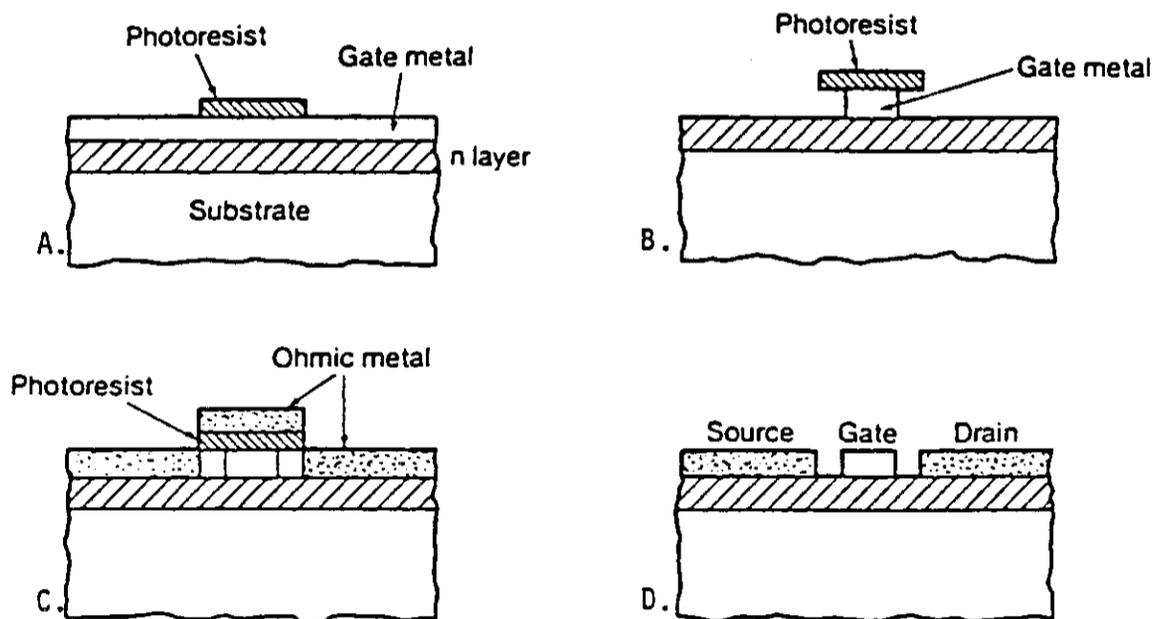


FIGURE 46. Processing steps of self-aligned gate technology.

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Rather than define the active channel thickness by the thickness of the n-type epitaxial layer, a thicker layer is grown and the channel region under the gate is defined by etching. This removes the high tolerance in thickness required for the epitaxial layer when the channel region is not etched. Most companies use a preferential etch that gives a flat bottomed recess. Source and drain contacts are deposited first, as in Figure 47A, and the gate is defined in photoresist.

A channel is etched in the GaAs until a specific current is measured between the source and drain contacts. Gate metal is then evaporated and the excess metal (Figure 47C) removed by using the "float-off technique." This basic method works equally well with both photolithographic and electron beam resist exposure techniques. Gate lengths as low as 0.25μ have been produced.

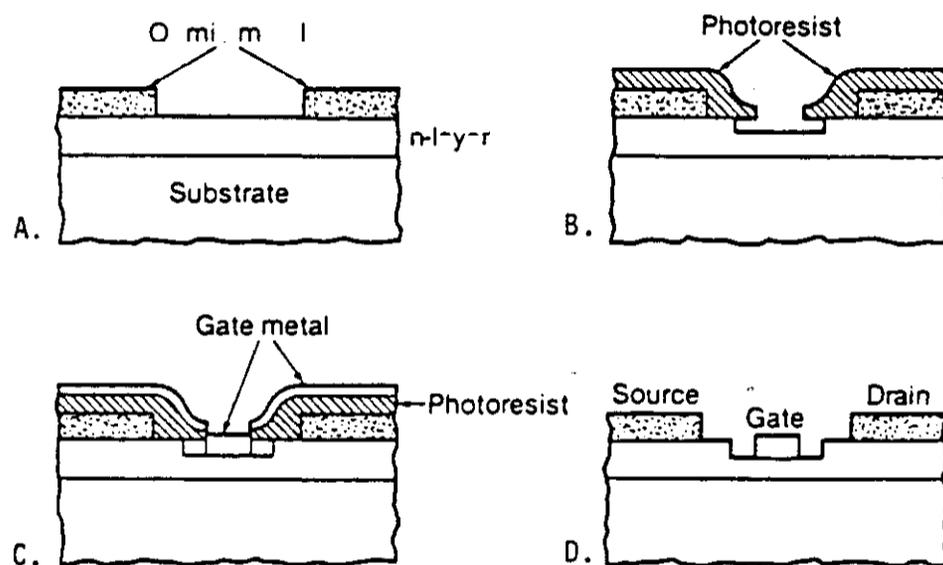


FIGURE 47. Processing steps of etched-channel technology.

6.3.3.6 Packaging of microwave transistors. Until microwave transistors were developed, packaging was simple. Ordinary switching transistors or low-frequency amplifiers could be sealed in epoxy for commercial use or hermetically sealed in metal cases for high reliability applications. In essence, the package did nothing to adversely affect the performance of the chip it contained.

Unfortunately, the same cannot be said about the microwave transistor. Conventional packages introduce stray capacitances and inductances which can cause such problems as parasitic oscillations, degradation of gain, poor broadband response, and impedance mismatch.

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Therefore, it was necessary for new packaging concepts to be developed as fast as higher frequency devices were being produced. This task became somewhat easier with some of the breakthroughs in materials and in microwave techniques.

The most significant advances in microwave technology were the development of the stripline and microstripline techniques. These two methods, which were developed to achieve low loss transmission of microwave signals, proved to be extremely instrumental in the development of microwave transistor packages. Package development proceeded rapidly with the advent of the microstripline transistor package.

Overall package design and compatibility with the microstripline technique virtually eliminated performance problems associated with conventional packages and circuits. The major advantage that the microstripline technique offered was an excellent impedance matching network.

Consequently, the VSWR was kept quite close to 1:1. Package/circuit compatibility can be readily seen from the high collector efficiency and low reflected power.

A microwave transistor chip will always exhibit some degree of performance degradation when placed inside a package. Although refinements in packaging technology have reached a point where the actual degradation is minute, further refinements are always being sought. At microwave frequencies, designers strive for every 0.1 dB of gain they can possibly obtain while trying to keep the noise figure at a reasonable level. These are the two parameters that exhibit the greatest degradation due to interaction between package and chip.

For example at 2 GHz, the particular package in which a chip is mounted can account for a difference of 0.5 dB in gain.

Materials used at microwave frequencies have a profound effect on device performance. This point has already been established in the discussions on metallization systems and dopants. With the packaging of a microwave chip comes another critical decision in material selection.

Substrate material must exhibit such properties as good heat transfer capability, negligible effect on transistor performance, and ease of die attachment. Several materials perform well above 1 GHz. Among them are beryllium oxide, alumina, spinel, and sapphire. The latter is rarely used on discrete devices but is used more often in microwave integrated circuits.

Overall package technology has advanced to a point of successful high reliability application. Hermetic seal and ability to withstand rigid military environmental requirements are considered standard among the larger microwave packages.

However, a definite problem exists with certain extremely small packages. Structurally, they are incapable of surviving environmental tests such as vibration or centrifuge.

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Recent developments in microwave integrated circuits and hybrids make use of another versatile package, the microstrip chip carrier. An example is shown in Figure 48. The chip is attached to the collector metallization, and wire bonds are made to the input and common, depending on the desired configuration, common emitter, or common base.

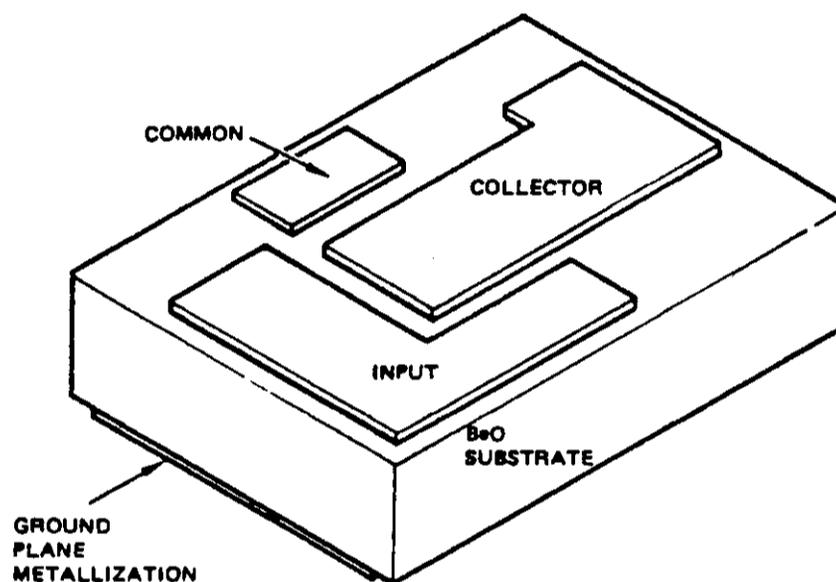
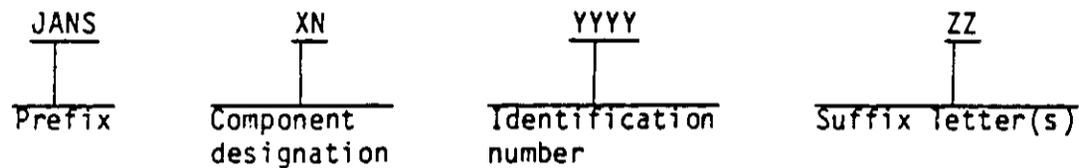


FIGURE 48. Microstrip chip carrier.

The advantage of the carrier over a chip for microcircuit purposes is simply described as follows. A chip-on-carrier is characterized after die attachment to the carrier. At microwave frequencies this is important because a device's electrical characteristics undergo a drastic change during the die attachment operation due to a piezoelectric effect. The extreme flexibility of the carrier makes it a desirable part in the microwave component repertoire.

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6.3.4 Military designation. The military designation for microwave transistors is formulated as follows:



The prefix includes the level of product assurance. For NASA applications, the level of product assurance is restricted to JANTXV or JANS in accordance with MIL-STD-975. A radiation hardness assurance (RHA) code, if applicable, is placed after the prefix. See MIL-S-19500 for details.

The component designation is "2N" for microwave transistors.

The identification number is assigned in order of registration and has no other significance.

The suffix letter symbolically describes matched devices (suffix "M"), reverse polarity (suffix "R"), or any other letter to indicate a modified version.

6.3.5 Electrical characteristics. This section shall concern itself primarily with the basic concepts and considerations required for accurate rf measurements on microwave transistors. At frequencies greater than 1 GHz, the methods used for conventional transistors or h or y parameters yield no meaningful information. The following basic difficulties are evident:

- a. H or y parameter measurements require open or short-circuited terminations and resonant lines. The resulting frequency dependency eliminates the advantage of swept frequency, broadband measurements.
- b. Measurements must take into account the parasitic effects of the package of the transistor as well as the effect of transmission lines.
- c. Bias networks that could cause the transistor to oscillate must not introduce reactive impedances into the circuit.
- d. Small signal measurements are not easily adaptable to power devices.

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6.3.5.1 Definition of S-parameters. Consider the two-port circuit in Figure 49. The response of this circuit, in s-parameter representation, establishes two independent variables:

$$a_1 = \frac{1}{2 (Z_0)^{1/2}} (v_1 + Z_0 i_1)$$

$$a_2 = \frac{1}{2 (Z_0)^{1/2}} (v_2 + Z_0 i_2)$$

and two dependent variables:

$$b_1 = \frac{1}{2 (Z_0)^{1/2}} (v_1 - Z_0 i_1)$$

$$b_2 = \frac{1}{2 (Z_0)^{1/2}} (v_2 - Z_0 i_2)$$

where $Z_0 = Z_{01} = Z_{02}$ is the characteristic impedance of the transmission line.

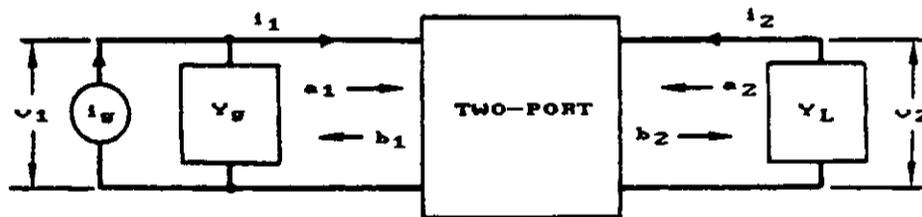


FIGURE 49. Two-port circuit.

From this the s-parameter matrix is defined as:

$$b_1 = S_{11}a_1 + S_{12}a_2$$

$$b_2 = S_{21}a_1 + S_{22}a_2$$

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or in matrix form:

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

For simplicity, we will refer to a_1 and a_2 as incident voltage waves and b_1 and b_2 as reflected voltage waves. It can be seen that:

$$\text{when } a_2 = 0, S_{11} = \frac{b_1}{a_1} \text{ and } S_{21} = \frac{b_2}{a_1}$$

and

$$\text{when } a_1 = 0, S_{12} = \frac{b_1}{a_2} \text{ and } S_{22} = \frac{b_2}{a_2}$$

Typical block diagrams for measuring the forward and reverse transmission coefficients, S_{21} and S_{12} , and the forward and reverse reflection coefficients, S_{11} and S_{22} , are shown in Figures 50 and 51, respectively.

In Figure 50, a swept frequency source feeds a power divider. The outputs of the power divider feed both a reference channel and a test channel. A line stretcher is placed in the reference channel and the device to be characterized is placed in the test channel. The function of the line stretcher is to compensate for the extra electrical length of the transistor. A test set and analyzer then compares the two signals, and yields S_{21} for the forward direction and S_{12} for the reverse direction.

The reflection coefficients, S_{11} and S_{22} , are the most easily measured by the reflectometer technique shown in Figure 51. The swept frequency source feeds a directional coupler. The transistor is placed at the measurement port of the coupler and is terminated in a 50- Ω load. The incident wave is measured at the reference port of the coupler. The reflected wave is measured at the test port of the coupler. The analyzer again compares the ratios of the signals. Representative plots of S_{21} , S_{12} , S_{11} , and S_{22} can be seen in Figures 52 through 55.

6.3.5.2 Gain. While the measurements of the s-parameters of a transistor characterize it as far as transmission and reflection coefficients are concerned, these measurements do not yield the complete device performance picture. The measurement of gain, however, has several unique advantages. The value of gain is not affected by package parasitics. Furthermore, the location of a reference plane is of no consequence.

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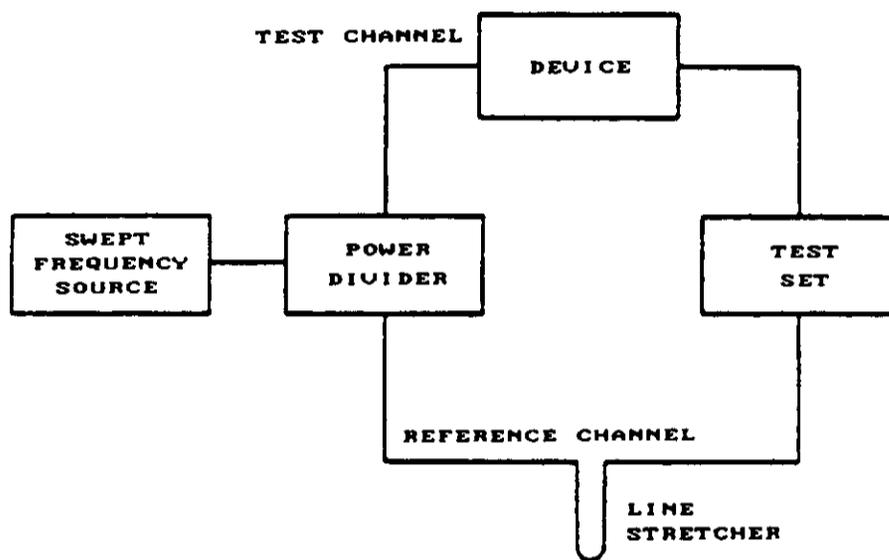


FIGURE 50. Block diagram for measurement of S_{12} and S_{21} .

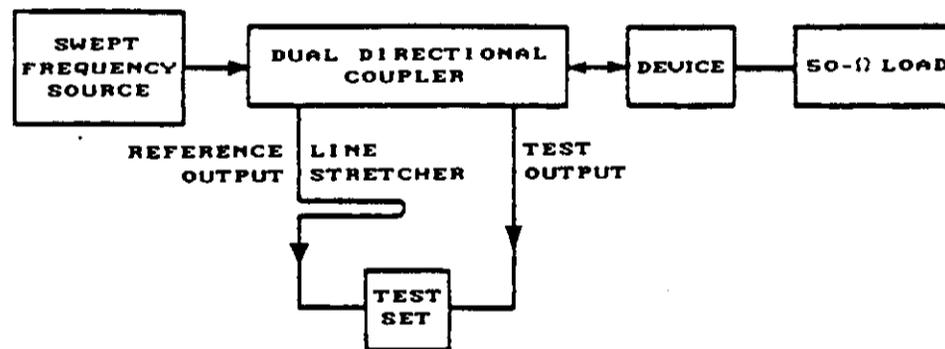


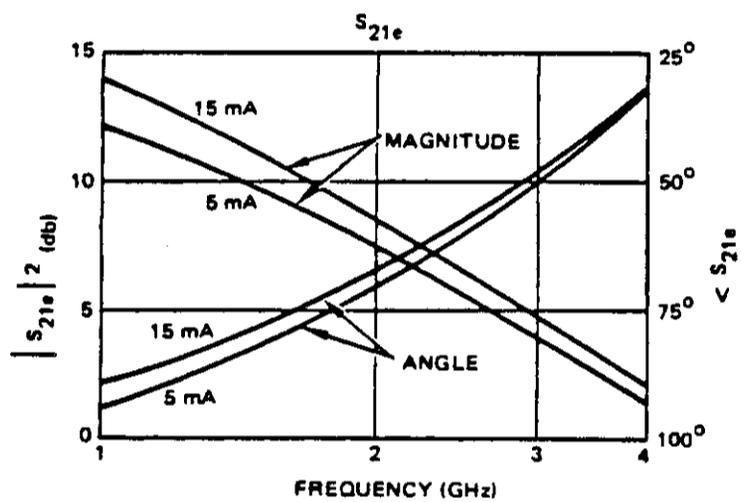
FIGURE 51. Block diagram for measurement of S_{11} and S_{22} .

The gain can be measured under several different conditions. The definitions of these are as follows:

- a. Maximum stable gain GMS is equal to

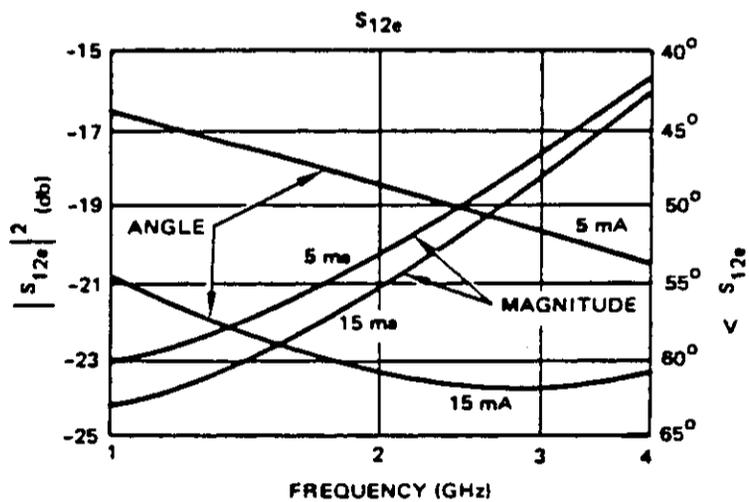
$$\left[\frac{|S_{21}|^2}{|S_{12}|^2} \right]^{1/2} \quad \text{or} \quad \left| \frac{S_{21}}{S_{12}} \right|$$

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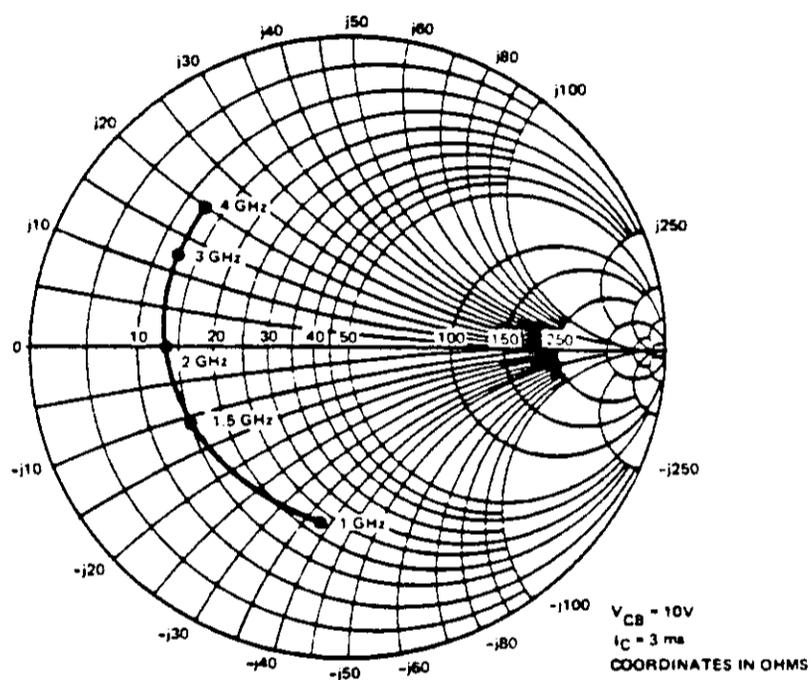
$C_B = 7$ V

FIGURE 52. S_{21} vs frequency.



$V_{CB} = 7$ V

FIGURE 53. S_{21} frequency.

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- b. Maximum available (tuned) gain (G_{MAX}) is measured with the input and output conjugately matched. The only other condition is stability. The stability factor K must be greater than 1 (unconditional stability).
- c. Unilateral gain (G_U) is the forward power gain of an amplifier whose reverse gain has been adjusted to zero.
- d. Stability factor (K) determines the tendency of a transistor to oscillate. For $K > 1$, stability is guaranteed. For $K < 1$, oscillation can be induced by introducing reactive components in to the load.

Each of these four parameters can be calculated from the measured s-parameters. The circuit is tuned so the magnitude and phase of G_{MAX} are recorded. Reversal of ports A and B and repeating the procedure yields.

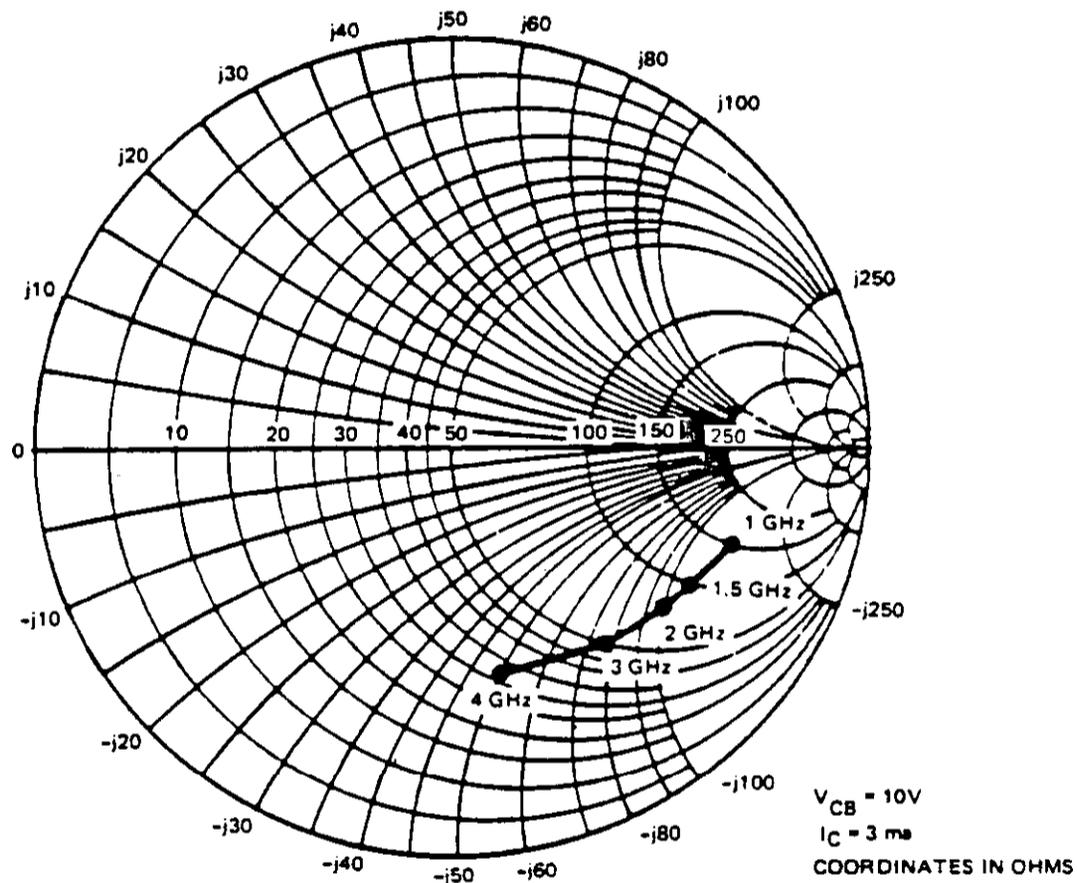
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$$\text{Then } G_{MS} = \left[|S_{21}|^2 |S_{12}|^2 \right]^{1/2} = |S_{21}/S_{12}|$$

$$K = \frac{1}{2} \left[\frac{G_{MS}}{G_{MAX}} + \frac{G_{MAX}}{G_{MS}} \right] \text{ FOR } K > 1$$

$$G_U = \frac{G_{MS} - (2 \cos\theta) + (G_{MS} - 1)}{2 (K - \cos\theta)}$$

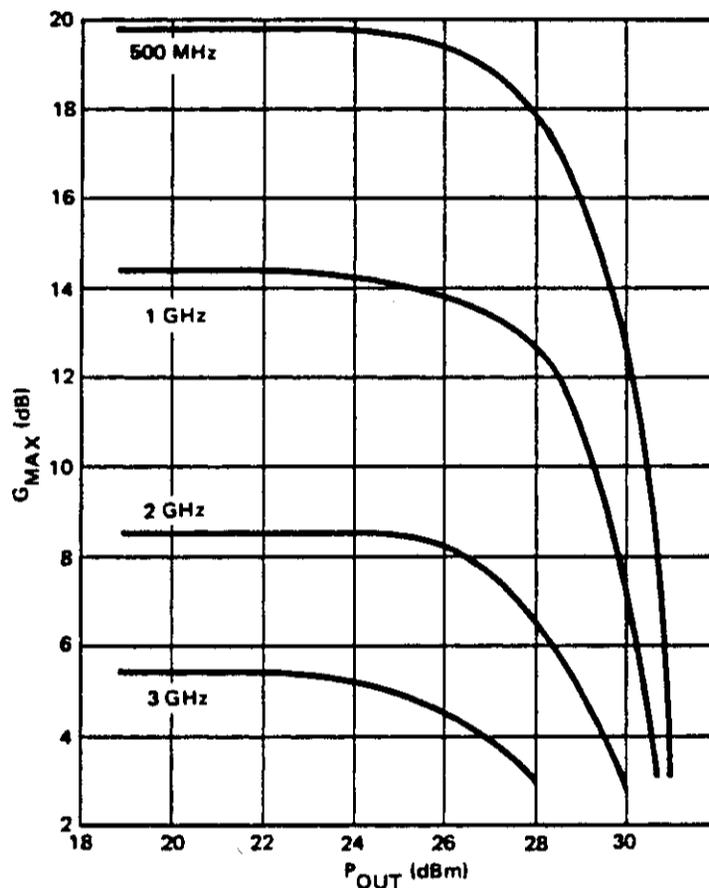
(θ is the phase difference recorded during the measurement of $|S_{21}|^2$ and $|S_{12}|^2$.)



$V_{CB} = 10 \text{ V}$
 $I_C = 3 \text{ mA}$
 Coordinates in ohms.

FIGURE 55. S_{22} vs frequency.

In addition to the swept frequency measurement of s-parameters and the various fixed frequency gain measurements, one other family of curves is important. This is the G_{MAX} as a function of actual rf power output. A typical family of curves is shown in Figure 56. The flat areas of each curve represent regions of uniform gain over a wide range of power output.

6.3 MICROWAVE DEVICES,
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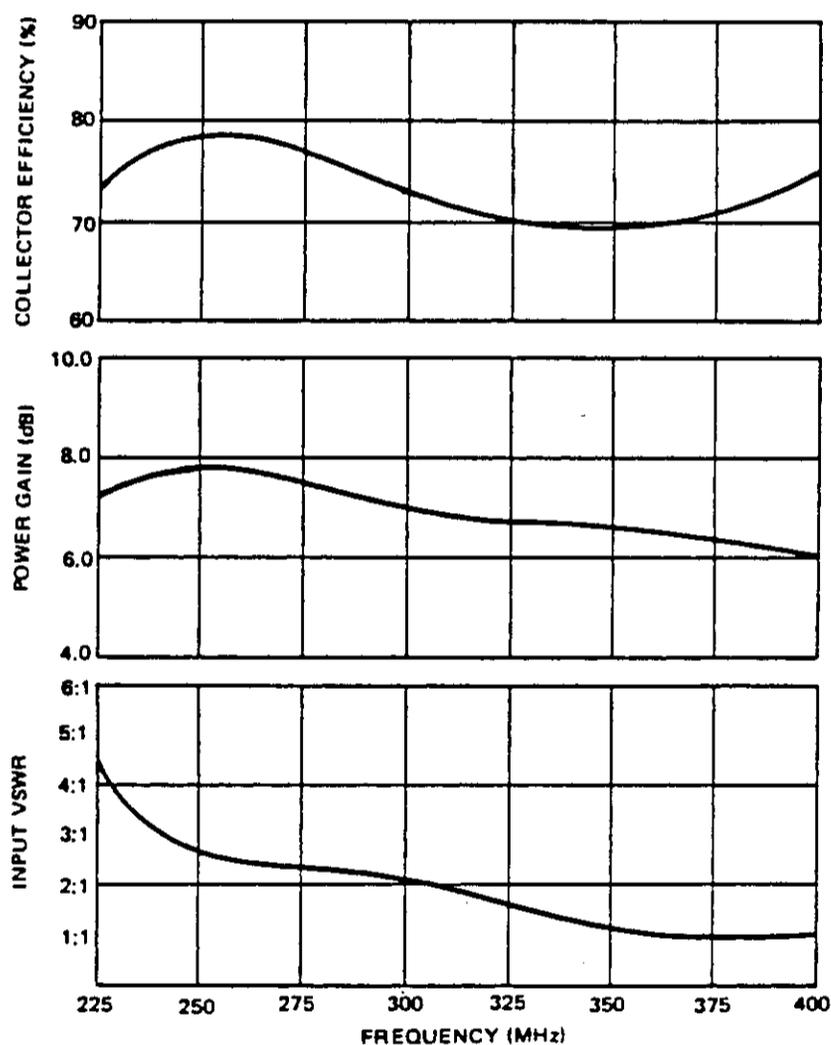
6.3.5.3 Noise figure. The measurement of noise figure is especially applicable to small signal devices. This parameter is a measure of the device's sensitivity to low level signals. It is a function of many variables, including frequency, source impedance, and current level.

Thus far, the discussion has been limited to the measurement of discrete devices. Of equal importance is the measurement of devices in a circuit. Because most microwave amplifiers are broadband, swept frequency measurements can be used to a great extent. Primary response characteristics include collector efficiency, input and output VSWR, power gain, and noise figure.

Collector efficiency is a term more commonly used for a power amplifier than a small signal amplifier. The major factor in determining efficiency is the class of operation. Class A amplifiers, because they draw current and dissipate heat 100 percent of the time, have the lowest efficiency, about 20 to 25 percent.

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In Class C operation the transistor is biased so conduction occurs only when a signal of sufficient magnitude is fed into the input. The quiescent current is zero when the driving signal is not present. Therefore, because current flows for less than 50 percent of each cycle, the collector efficiency can be as high as 70 to 80 percent. See Figure 57 for a typical plot of efficiency in a Class C amplifier over a frequency range.



$P_{OUT} = 30 \text{ W}$
 $V_{CC} = 28 \text{ V}$
 $T_{CASE} = 35 \text{ }^{\circ}\text{C}$

FIGURE 57. Typical broadband response of a UHF power amplifier.

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The amount of reflected power from the input of an amplifier is best demonstrated from the measurement of the VSWR. Several factors can affect the VSWR of a device in a given circuit. Most critical is the impedance matching both the input and output.

The input impedance of a power amplifier is generally low compared with the output impedance of the preceding driver. The real (resistive) component is inversely proportional to the area of the transistor on the order of 0.5 to 5 Ω . The imaginary (reactive) component of the input impedance is determined by the inductance of the package and the input capacitance of the device. At microwave frequencies, inductive reactance is dominant.

The output impedance is an extremely important factor in determining the amount of power gain of an amplifier stage. Although an optimum match is not always achieved due to other circuit considerations, the amount of mismatch adversely affects both gain and VSWR. Figure 57 depicts typical variations in gain and VSWR for an amplifier over a frequency range.

6.3.5.4 Amplifier parameters.

6.3.5.4.1 Intermodulation intercept. The intermodulation intercept is an expression of the low-level linearity of the amplifier. The intermodulation ratio (IMR) is the difference in decibels between the fundamental output signal level and the generated distortion product level. The relationship between intercept and intermodulation ratio is illustrated in Figure 58, which shows product output levels plotted versus the level of the fundamental output for two equal strength output signals at different frequencies. The upper line shows the fundamental output plotted against itself with a 1 dB to 1 dB slope. The second and third order products lie below the fundamentals and exhibit a 2:1 and 3:1 slope, respectively. The intercept point for either product is the intersection of the extensions of the product curve with the fundamental output.

The intercept point is determined by measuring the intermodulation ratio at a single output level and projecting along the appropriate product slope to the point of intersection with the fundamental. When the intercept point is known the intermodulation ratio can be determined by the reverse process. The second order IMR is equal to the difference in decibels between the second order intercept and the fundamental output level. The third order IMR is equal to twice the difference between the third order intercept and the fundamental output level. These are expressed as:

$$IP_2 = P_{out} + IMR_2$$

$$IP_3 = P_{out} + 1/2 IMR_3$$

where P_{out} is the power level in units of decibel above a milliwatt (dBm) of each of a pair of equal level fundamental output signals, IP_2 and IP_3 are the second and third order output intercepts in dBm, and IMR_2 and IMR_3 are the second and third order intermodulation ratios in decibels.

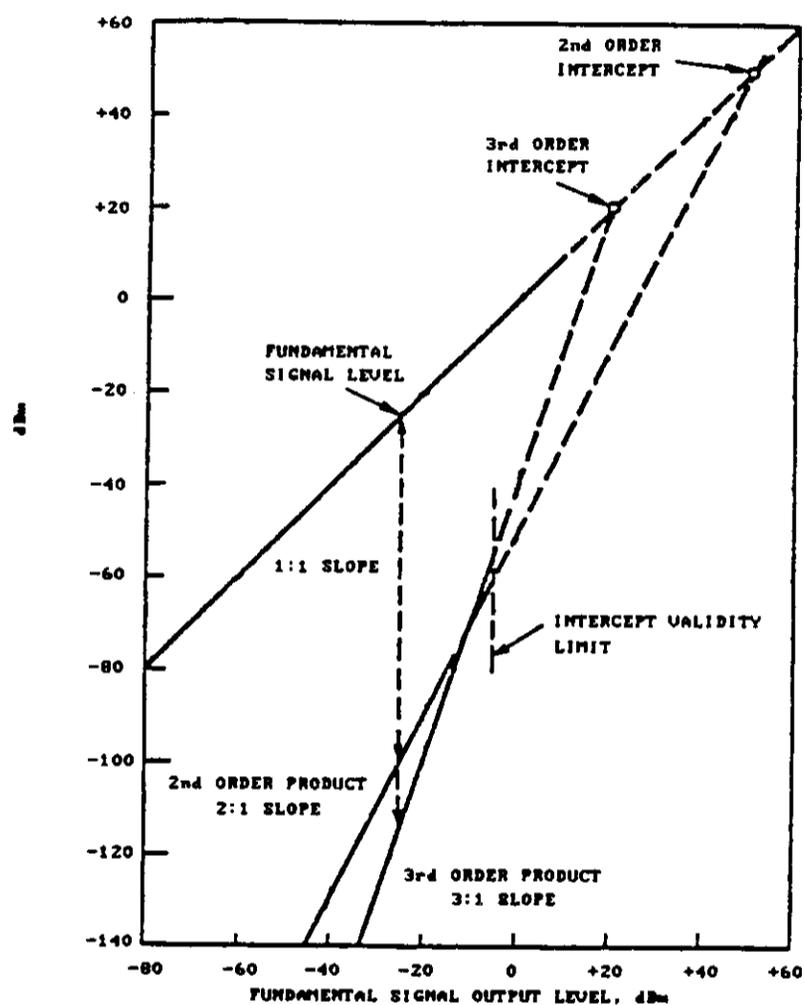
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FIGURE 58. Intercept diagram.

The intermodulation intercept is a valid indicator of intermodulation performance only in the small signal operating range of the amplifier. Above some output level which is below the 1 dB compression point, the active device moves into large signal operation attended by signal strength dependent bias level shifts. At this point, the intermodulation products no longer follow the straight line output slopes, and the intercept description is no longer valid.

The intermodulation ratios are determined by measurement using a conventional spectrum analyzer. The measurement dynamic range is enhanced using appropriate cancellation techniques when required to accommodate high dynamic range amplifiers.

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6.3.5.4.2 Compression. The 1 dB compression point is the output level at which the amplifier gain drops 1 dB below its small signal value. It is an indication of the signal level at which small signal conditions no longer apply. At this level the intermodulation intercepts no longer adequately predict the amplifier distortion behavior.

The measurement of the compression level is made using a frequency discriminating detector to detect the gain decrease at the fundamental frequency. This is preferable to using a total power detection measurement, since the fundamental frequency is probably of most interest to the user.

6.3.5.4.3 The most obvious conclusion that can be made regarding device selection is that circuit requirements should govern device selection, and not vice versa. Two seemingly identical devices will not necessarily perform comparably in a given circuit. Therefore, it is the entire network of transistor, package, parasitics, microstrip runs, and other components that must be considered as one fundamental unit.

The importance of S-parameter measurements cannot be overemphasized. S-parameter characterization gives the design engineer specific information at microwave frequencies on the performance of the device. Coupled with computer aided circuit design techniques, it enables a designer to optimize his circuit rapidly and accurately.

Complete S-parameter characterization is usually only meaningful for small signal devices. Power microwave devices are considered differently. The most important characteristic of a power device is its ability to operate in a mismatched load. Response curves for collector efficiency and junction temperature, as a function of frequency for various values of VSWR, can be considered to be of primary importance to a designer.

6.3.6 Environmental considerations. Typical environmental conditions and screens that microwave transistors are capable of withstanding can be found in MIL-S-19500. For the specific device selected, consult the applicable MIL-S-19500 reference sheet.

More information on environmental considerations can be found in section 3, Resistors, paragraph 3.8.7, Reliability considerations, and paragraph 3.1.7, General reliability considerations.

6.3.7 Reliability considerations.

6.3.7.1 Introduction. Until a few years ago there was a void in the transistor market area in the rf and microwave regions. This void is gradually being filled by the industry today using both old and new techniques: interdigitated, overlay, and metal matrix geometries. However, due to the lack of the volume in sales in this area, there has been only a minimum effort to fully understand the failure mechanisms associated with these devices and the necessary corrective action required to significantly reduce the failure mechanisms.

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The dominant failure mechanisms of a device vary according to the peculiarities of the design, fabrication process, and application of the device. Following is a list of the major failure mechanisms for rf and microwave transistors.

- a. Aluminum migration
- b. Die attach failure
- c. Metal-over-oxide-step coverage
- d. Lead bond failures
- e. Device inability to operate into a mismatched load.

6.3.7.2 Failure mechanisms for rf and microwave transistors.

6.3.7.2.1 Aluminum migration. This well documented phenomenon relates to the migration of aluminum in the presence of high temperature and high current densities. The transport of mass in metals when the metals are stressed at high current densities was recently recognized as a potential wear-out failure mechanism in semiconductor devices. On most semiconductor devices this failure mode is indicated by an electrical open due to voids or localized loss of conductor metal. In the case of a transistor, this normally begins on the emitter fingers. When one of the emitter fingers opens due to this migration, the adjacent fingers are required to carry this additional current. These fingers, now operating at an increased current, will accelerate their own failure rates, due to the increased current. Figure 59 illustrates this failure mechanism.

However, due to the shallow diffusion depth required on high frequency transistors, this failure mode can, on occasion, also be indicated by a shorted emitter-base junction. Figure 60 shows the cause of this shorted condition.

Studies have shown that aluminum migration is dependent on current density, surrounding material, and aluminum grain structure. Consequently, to reduce the possibility of aluminum migration, these phenomena must be closely scrutinized in the selection and application of rf and microwave transistors.

The current density problem is directly related to the cross-sectional area of the metallization; that is, the smaller the cross section, the smaller the MTBF. J.R. Black, in his paper on "Electro Migration Failure Modes in Aluminum Metallization for Semiconductor Devices," in the IEEE, Volume 57, No. 9, September 1969, used the following equation as the basis for calculating MTBF for large-grained films 6000 Å thick, with the pre-exponential constant normalized to the conductor cross-section areas of $9.65 \times 10^{-8} \text{cm}^2$.

$$\text{MTBF} = \frac{(w) (t) \exp (\theta/kT)}{(5 \times 10^{-13}) \text{ J}^2}$$

w = conductor width in centimeters

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t = conductor thickness in centimeters

ϕ = an activation energy in electron volts, for large grained films 0.84

k = Boltzman's constant

T = film temperature in degrees Kelvin

J = current in amperes per square centimeter

It can readily be observed from this equation that the MTBF of the film is directly related to the cross-section (w and t) of the conductor.

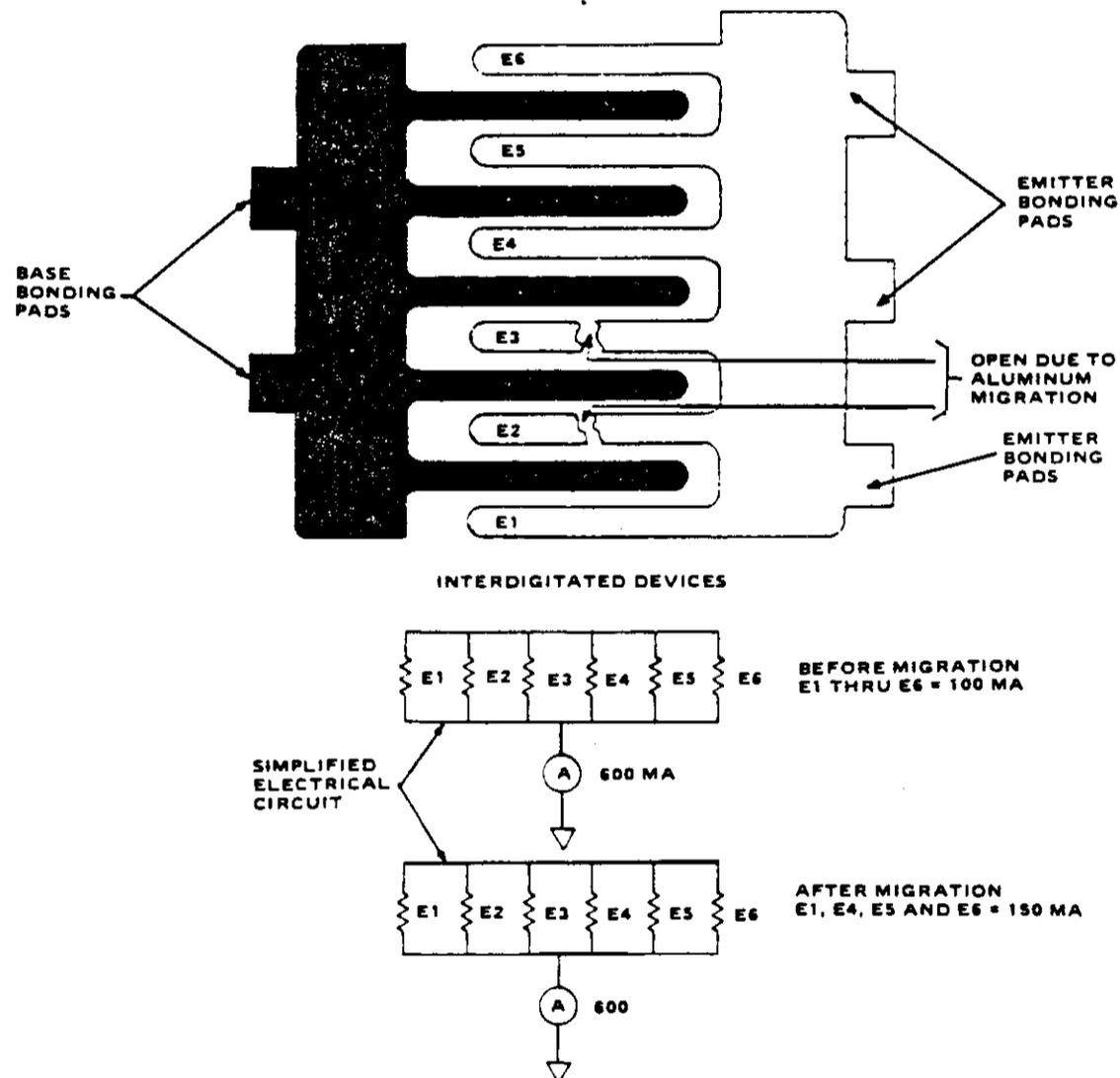


FIGURE 59. Current density, secondary problem.

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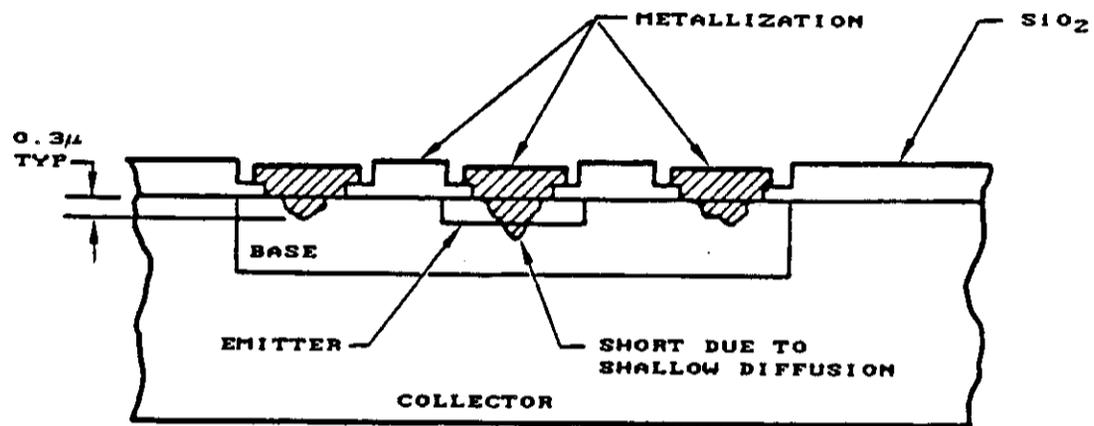


FIGURE 60. Short due to shallow diffusion.

6.3.7.2.2 Die attach failure. This defect is prevalent in any power transistor because of problems associated with attaching a die that has a large mass area; it is difficult to attach a die to the header without leaving any voids between the die and header. Figure 61 shows a typical cross section of a die bonded to the header. An X-ray photograph taken through the bottom of the header of a transistor package clearly shows the voids between the header and die. However, X-ray testing is potentially destructive and, therefore, cannot be used as a 100 percent screen.

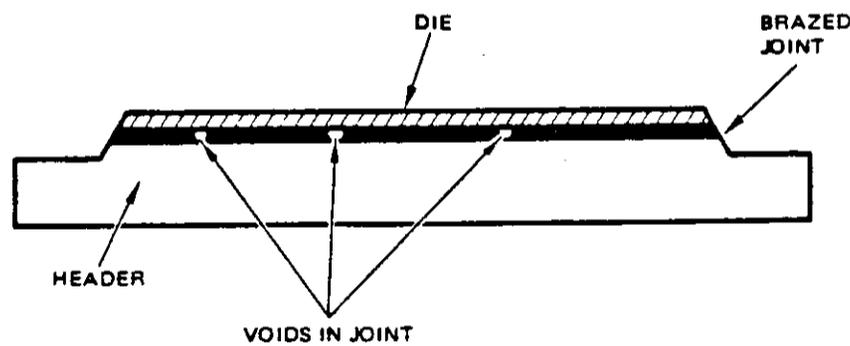
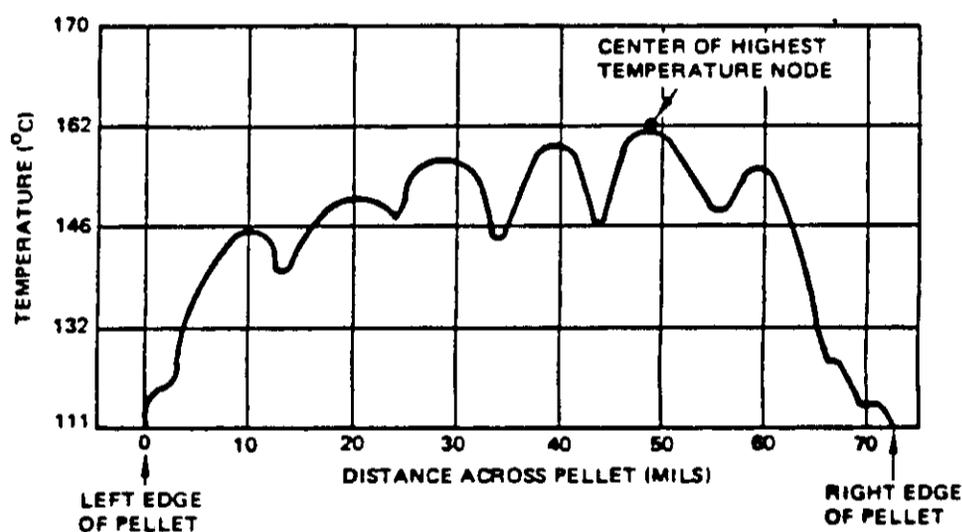


FIGURE 61. Cross-section of a die bonder to a header.

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These voids cause hot spots on the surface of the die, due to current injection, which, in turn, can cause thermal runaway. The measurement frequently used to characterize the transistor thermal properties is thermal resistance, θ_{JC} or θ_{JA} . These characteristics, θ_{JC} and θ_{JA} , are normally calculated by techniques that use average die temperature to arrive at their values. The result is an average value of thermal resistance when, in actuality, the thermal profile across the die varies greatly. Figure 61 shows a typical thermal profile across a die in operation. If a void is beneath a section of the die, as indicated in Figure 61, this will cause surface hot spots that cause greater variations than those shown in Figure 62.



$f = 400 \text{ MHz}$
 $P_o = 25 \text{ W}$
 $V_{CC} = 28 \text{ V}$
 $T_{CASE} = 85 \text{ }^\circ\text{C}$
 $V_{SWR} = 1.0$

FIGURE 62. Thermal profile of a die during operation.

Two main approaches have been taken to overcome the problems associated with these voids. The first and most significant is employing current ballast resistors on the multiple emitter fingers (refer to Figure 30). With this structure the effects of a hot spot can be cancelled by the voltage drop developed across the ballast resistor, which feeds current to the emitter sites.

The use of ballast resistors has been successful in significantly reducing the failure rate due to hot spots on transistors operating below 100 MHz on Class A and B power amplifiers.

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The second action being taken by the industry to reduce the effect of hot spots is improving the initial thermal resistance measurements. This improvement is being completed using infrared scanning techniques. This technique arrives at a worst case thermal resistance instead of an average value as the older method does. Therefore, if the vendors use these values for rating their devices, the users will be able to realistically design their circuits with these ratings.

Since infrared measurement is a sample test done only periodically for economic reasons, this measurement does not take into account the day-to-day variations of die attaching. Therefore, die attach failure remains an industry problem.

6.3.7.2.3 Metal-over-oxide-step coverage. Surface metal interconnects are subject to some of the more insidious failure modes. The more common symptom of metal problems is the total or intermittent open caused by fracturing or migration. It is now an established fact that aluminum migrates as the current density in the metal approaches 10^6 A per cm squared. Fracturing can be caused by thermal mismatch between the metallization and the silicon or silicon dioxide substrate. Thinning of the metallization at the metal-over-oxide step is caused by the method of metallizing the transistor.

In electron-beam evaporation, the source from which the metal is deposited is normal to the surface of the wafer. As a result the thickness of the metal over the oxide steps is far thinner than the metal deposited on the flat surface. It is much easier to measure the thickness of the metal on the flat surface. Therefore, all control of metallization revolves around the surface thickness. A typical cross section of a contact area is shown in Figure 63. It can be observed that at the metal-over-oxide step the metallization is only approximately 10 percent of the surface metal thickness. When the metal is caused to expand and contract by current heating or operating at thermal extremes, micro cracks can develop, causing an open or intermittent open.

Many approaches have been taken to improve the thickness of metal-over-oxide steps. Two of the approaches contour the oxide steps. They are: 1) tapering of the oxide steps, and 2) having a double step at each oxide step. Although tapering the oxide step by controlling of the etch rate is the better approach, it requires additional etching steps. If optimized properly, the results are worth the extra time and cost. Figure 64 shows a cross section of contact area with tapered oxide. As the chip complexity increases, especially with double metallization devices, this tapering of oxide becomes very important.

The other method, a double step, accomplishes the oxide height minimization by growing the oxide in two steps which require additional masking steps. This is also shown in Figure 63.

Also, the industry has been depositing the metal at more than one angle by electron-beam evaporation while universally rotating the wafers. This has helped to limit this problem. However, the metal can still migrate down into the contact area because current density causes the metal-over-oxide step to

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TRANSISTORS

become the weak area again. Therefore, most integrated circuit manufacturers have modified their design rules to allow a maximum current density of 10^2 A per cm squared. This rule is based on the assumption that the metal thickness over-oxide step is only 10 percent of the surface metal thickness. However, due to the miniature geometry required for high frequency transistors the current density runs in the range of 10^5 A per cm squared.

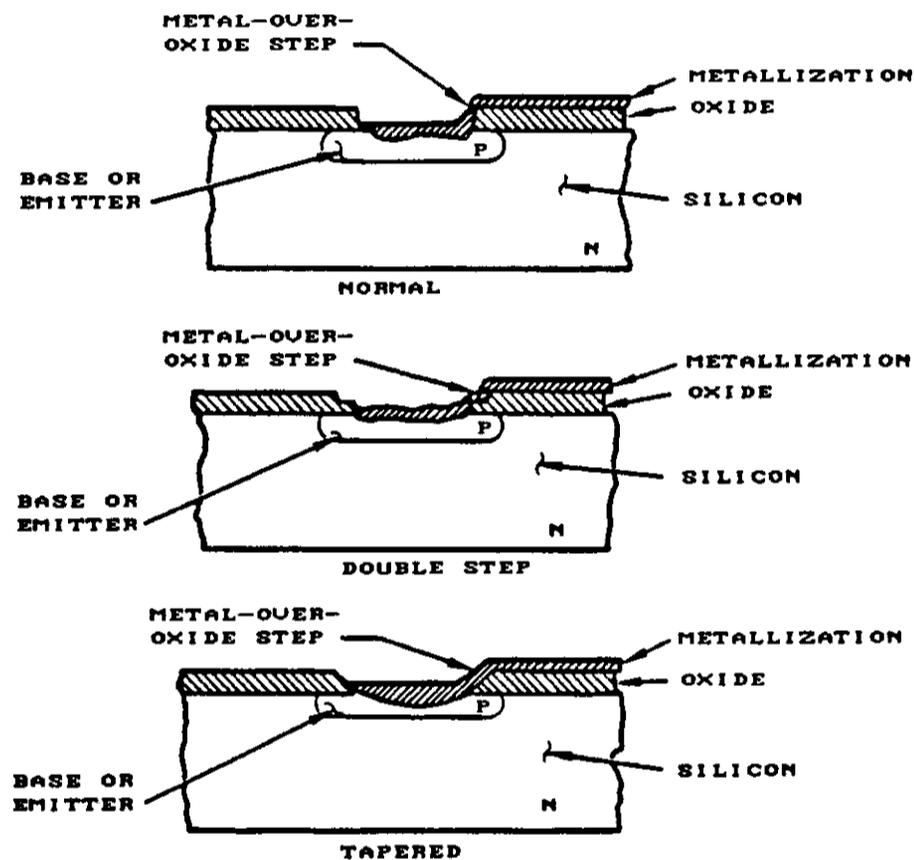


FIGURE 63. Cross-section of a contact area with tapered and double-stepped oxide.

The industry is presently developing metal systems other than aluminum for use on rf and microwave transistors. These metal systems are of refractory metal and, due to their high atomic weight, resist mass transport by way of momentum exchange.

Any other cause for the reduction of metal cross section can generate the same problems. Most common are photolithographic errors, scratches in the metal, and metal too thinly deposited. It is generally conceded that metal which opens due to thermal mismatch is caused by severely reduced cross section.

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Therefore, it can be assumed that improvements that increase metal thickness over steps will solve both failure modes simultaneously. The only other significant failure mode in the surface metallization itself is lack of adherence to the substrate either on the surface or down in the oxide cutouts. Lifting of metal can be caused either by residual contaminants present during metallization or by severe overalloying of the metal after deposition. These types of failure modes should be screened out by normal 100 percent tests such as burn-in and temperature cycling.

6.3.7.2.4 Internal lead bond failures. Defective internal lead bonds traditionally have been a large cause of device failure. The majority of these devices had seen a 100 percent screening, which included constant acceleration and temperature cycling. Lead bond failure will be more prevalent in rf and microwave transistors due to: (1) the increase in the number of bonds within a package, (2) the smaller bonding pads, and (3) the smaller wire diameter required.

Generally, the industry has settled on aluminum-aluminum and gold-aluminum metal systems for transistors. Both methods are satisfactory, although both can fail under certain circumstances.

Gold thermal compression bonding is inherently a stronger bonding method than the aluminum ultrasonic bonding technique. However, the gold ball bond is more massive and, therefore, more sensitive to high-g shock. An ultrasonic bond has a narrow cross section at the heel of each bond and it is difficult, if not impossible, to examine the depth of the wire deformation (Figure 64). A ball bond inherently exhibits no deformation at one end of the bonding wire, but the bond made at the package end suffers the same problem as an ultrasonic bond.

During storage, there is no concern over "purple plague" or other bimetallic formations using a gold bond method. These formations occur rapidly at temperatures above 200 °C. Below that temperature the bimetallic formation rate is not easily measured. The bimetallic formation or "purple plague" is of course not a failure mode set in aluminum bonding. Hermetic seal leakage that allows moisture to enter the package is very likely to destroy either class of bond, since the aluminum metallization corrodes regardless of bond type.

Presently, the industry uses three types of screens to reduce this cause of device failure: 1) constant acceleration, 2) temperature cycling, and 3) power cycling. However, these screening techniques are only partially effective; therefore, this failure mode will continue to plague the industry.

6.3.7.2.5 Mismatched loads. In rf and microwave transistor Class C power amplifiers, failures are readily caused when the output circuit is detuned at full power. This can happen in practical applications. The resulting mismatch causes a high VSWR condition at the collector of the transistor, subjecting the transistor to instantaneous voltage peaks many times the supply voltage. These voltage peaks will cause avalanching, which takes place within the collector depletion region. Therefore, there is no "current steering" effect by the emitter ballast resistors to reduce the localized current densities in the

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collector depletion region. Some devices are said to be capable of operating into an infinite VSWR; however close investigation of their specifications shows that this can only be done when the devices are derated greatly to accommodate the instantaneous voltage peaks.

The only form of corrective action to be approached by the industry is to use "collector ballasting" provided by a thick, undepleted collector layer. However, due to its effect on the electrical characteristics of the device, this approach is very limited.

6.3.7.2.6 Radiation effects. Radiation-induced damage in the form of permanent damage to the semiconductor crystal dominates other radiation-induced changes in microwave transistors. The type of damage produced by energetic electrons and gamma rays are called point defects because they are spread uniformly throughout the crystal. In this type of particle-atom collision an atom of semiconductor material will be displaced to an interstitial position, leaving behind a vacancy in the crystal lattice. The resulting vacancies and interstitials are mobile and will tend to unite with themselves or impurities in the crystal and form defect complexes. Different kinds of defect complexes are formed, depending on the temperature.

Neutron radiation can displace many more lattice atoms. As a result regions of localized damage occur. These regions are called defect clusters. Curves of the total number of defects per cubic centimeter of semiconductor per neutron per centimeter squared have been calculated. The effects of permanent radiation damage show on the dc, low frequency, and microwave properties of microwave transistors.

Low frequency radiation and dc effects have been thoroughly studied. The predominate effect is a decrease in current gain (β) due to recombination in the base region. For microwave transistors, the base width is sufficiently narrow that this effect can be neglected at neutron fluences up to 10^{17} n/cm². This value is moderated by degradation of current gain caused by excessive recombination in the base emitter space charge region and emitter bulk regions.

Permanent radiation effects on the microwave properties of small-signal bipolar transistors can be divided into two categories: 1) changes in the s-parameters, and 2) changes in noise performance.

The most significant changes are seen in the forward transfer gain (S_{21}). One study showed the f_t of a bipolar microwave transistor dropped from 4.5 GHz pre-radiation to 540 MHz postradiation. This was caused by a drastic increase in T_b due to a phenomenon similar to base pushout. The f_t increased from 540 MHz to 3 GHz when the collector bias was increased from 15 V to 30 V. Thus, the radiation tolerance of this type of transistor can be improved by operating it at a higher collector voltage and at lower collection currents.

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Experiments with small signal bipolar microwave transistors in an amplifier circuit have shown that the noise figure is fairly insensitive to neutron radiation. This is because the noise mechanisms introduced by neutron damage only contribute excess noise at low frequencies. The increase in the noise figure that does occur can be attributed to the decreased current gain (β_0).

The general military specification for semiconductor devices specifies four radiation hardness assurance levels (RHA) for JANTXV and JANS product assurance levels. Microwave transistors may specify RHA designation M, D, R, or H in accordance with the requirements of the military specification.

The major failure mechanisms on rf and microwave transistors, listed in order of importance are as follows:

- a. Aluminum migration
- b. Device inability to operate into a mismatch load
- c. Die attach void failures
- d. Internal lead bond failures
- e. Metal-over-oxide-step coverage.

**6.4 MICROWAVE DEVICES,
ATTENUATORS AND LOADS**

6.4 Microwave attenuators and loads.

6.4.1 Introduction. Attenuators and loads, which are known as dummy loads and terminations, are similar devices. A load terminates a line and dissipates the signal with low reflections. An attenuator lowers or attenuates the transmitted signal and has low reflections.

An attenuator is a device used to attenuate rf signals a given amount without introducing mismatch when operating within its specified frequency range. Frequency range is limited by the connection mechanism and the match of the dissipative element to the transmission media. Broadband devices are only achieved using resistive elements.

A load is a device used to match the transmission media and dissipate all signals presented at the load. A load used to terminate a line in place of additional circuitry is called a dummy load. A load that is designed into a circuit to terminate an unwanted signal path is called a termination.

6.4.2 Usual applications.

6.4.2.1 Attenuators. Attenuation, like insertion loss, is not usually desired in microwave design. In those cases where signal levels must be adjusted down to lower levels, attenuators or directional couplers can be used. Attenuators present low mismatch to the source, attenuate the signal a given amount, introduce little delay or phase shift, have relatively little change in attenuation, and present a low mismatch to the load. Attenuators may be coaxial, waveguide, or strip-transmission media.

Most present coaxial attenuators are designed to operate at 1 W or less from dc to 18 GHz, with some units available to 26 GHz. Lossy material attenuators can be made for slightly higher power but are bandwidth limited. High power attenuators use fins for air cooling or are liquid cooled. Variable attenuators are made using resistive or lossy elements, which are movable and coupled to the rf field. Such units have less than an octave bandwidth.

Waveguide attenuators are usually made of resistive material matched to the waveguide, with the resistive material mounting structure causing slight mismatch and small frequency variable phase shift. Variable waveguide attenuators can be made by changing the position of the resistive card in the waveguide. Precision variable attenuators are available for laboratory use only.

Attenuators are used to lower the power into amplifiers, detectors, and mixers to maintain operation in the linear region and reduce reflections at the input. Attenuators are used in the laboratory to substitute a specific amount of loss in a microwave circuit. Fixed attenuators (commonly called pads) ranging from 0 to 3 dB in 0.5 dB steps allow adjustment of signal levels in two paths to the same level. Attenuators can be used as matching devices where loss of power is not a concern. A 10 dB attenuator terminated in a short circuit returns a signal of -20 dB from the input signal that is equal to an input VSWR of 1.2:1.

6.4 MICROWAVE DEVICES, ATTENUATORS AND LOADS

6.4.2.2 Loads. Coaxial loads are available with various connectors and with frequency ranges as high as 18 GHz. Stripline loads and flange wall-mounted loads are available for use in stripline and microstrip applications. Power limitations are based on the dissipation element and its heat sinking. Ratings of 1 W are typical for most coaxial loads with some air cooled units available below 10 GHz at higher powers. Lossy elements are rarely used in coax except for a few applications at a few watts over limited frequency ranges. Typical applications are in loaded circulators and as terminations for hybrids and directional couplers.

Waveguide loads are predominately lossy material types and can have kilowatt ratings with liquid cooling. Stepped transition lossy material inserts are widely used in waveguide couplers. Waveguide loads are also used for circulators, hybrids, duplexers, and system self test circuits.

6.4.3 Physical construction

6.4.3.1 Attenuators. Coaxial attenuators construction has evolved from "Pi" and "T" types using rod and disk resistors to thin film resistor material on a substrate, Figure 64. Connector and transmission line matching into the resistive element allows VSWRs less than 1.3:1 at 18 GHz and attenuation flatness with frequency of less than 10 percent (typically ± 0.50 dB) of attenuation value.

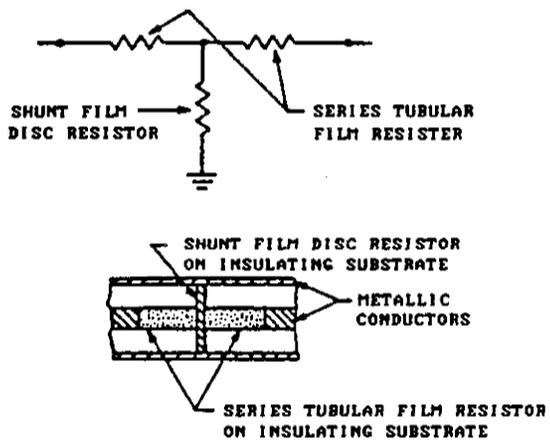
Variable attenuators are achieved by slot-line techniques with a lossy element movable in the slot. Frequency limitations are inherent in the length of the slot and the contour of the element.

Power limitations are based on the heat capacity of the element, its thermal resistance, and adequate heat sinking.

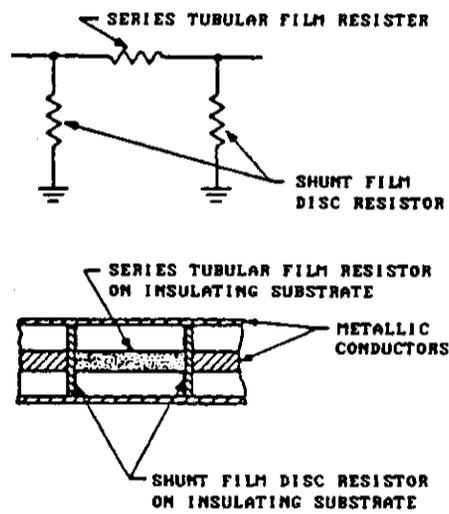
6.4.3.2 Loads. Loads are made in a similar manner as attenuators since a terminated attenuator is a load. Most coaxial loads use rod-type resistors with one end soldered to the center conductor and the other end soldered into the outer conductor. Lossy material coaxial loads are shaped material placed at the end of a coaxial launcher. Waveguide loads are usually lossy material formed to match the waveguide size. The load material can be mechanically held by epoxy or screws, or both. Tapered lossy material may be introduced directly to strip transmission media to terminate a line.

6.4.4 Military designations. The military selection standard for attenuators is MIL-STD-1352. The military selection standard for loads is MIL-STD-1637. Military attenuators are found in MIL-A-3933 and MIL-A-24215. Military loads are found in MIL-D-3954 and MIL-D-39030.

6.4 MICROWAVE DEVICES,
ATTENUATORS AND LOADS



A. Coaxial TEE section



B. Coaxial PI section

FIGURE 64. Construction of coaxial attenuators.

6.4 MICROWAVE DEVICES, ATTENUATORS AND LOADS

6.4.5 Electrical characteristics.

6.4.5.1 Attenuators. Attenuation is the primary parameter of an attenuator. The device is designed to have a nominal attenuation with a tolerance. A typical part may be specified as 30 dB \pm 0.50 dB at center frequency. Frequency sensitivity of such a device could be 1.0 dB from dc to 18 GHz for coaxial attenuators and 0.50 dB for waveguide. VSWR is also an important parameter. For coaxial devices the VSWR mismatch is set by the interface (connector) and the transition into and out of the dissipative element. VSWR usually increases with frequency and a 1.1:1 VSWR at 100 MHz may increase to 1.3:1 at 18 GHz. In this range, the values of VSWR will have periodic peaks as the two primary sources of reflections (connector and element transition) add in phase. Waveguide devices do not have the pronounced variations of the coaxial attenuators due to limited frequency coverage.

Either coaxial or waveguide attenuators will have both attenuation and VSWR changes with either temperature or power, or both. This is due to the resistance change of the element with temperature. VSWR changes are primarily affected and attenuation changes are secondarily affected. Changes due to power in an application is really a misapplication of the part. Proper derating with temperature should be observed. Applications requiring high peak power with low average power should be tested over frequency and altitude to assure no arcing.

6.4.5.2 Loads. The primary parameter of a load is VSWR. The power rating of the load is also important. These parameters are inherent to the matching of the dissipative element to the transmission media and the resistance in the thermal path to the heat sink.

6.4.6 Environmental considerations. Environmental considerations for loads and attenuators are mostly mechanical in nature. In coaxial devices, the stresses on the load or attenuator element result from movement of the center conductor at the interface. The body of the element needs to be supported to prevent lateral movement. Heat removal from the element is also important. Temperature cycling and vibration fatigue are the most stringent environments. Elements made of nichrome should be avoided unless the nichrome is adequately protected from moisture (disappearing resistor phenomena).

6.4.7 Reliability considerations.

6.4.7.1 Failure modes. Attenuator and load failures are usually mechanical in nature. Separation of the element at its interface is a predominate failure mode. Temperature induced changes in the dissipative element may occur if the element is not adequately heat-sunk. In high-peak power applications, arcing may occur due to spacing, burrs, or as a result of power and altitude conditions. Some high power waveguide loads may corrode due to brazing salts left in the device during manufacturing.

**6.4 MICROWAVE DEVICES,
ATTENUATORS AND LOADS**

6.4.7.2 Screening. Because the major failure modes are mechanical, temperature cycling is an effective screen to assure compatibility of materials. Electrical characterization of VSWR and attenuation (if appropriate) before and after such a screen can detect potential failures. Resistance measurements (center conductor to ground and to the other center conductor) are much more sensitive to changes than a VSWR test. Delta limits should be imposed. For resistor elements, delta limits should be representative of the capability of the resistive material.

Resistive element devices should undergo a dc burn-in at full rated power and temperature for 100 hours minimum (50 hours each end for attenuators). Delta limits of resistance change should again be imposed. When burn-in cannot be imposed, a high temperature storage for 100 hours at the maximum temperature of the device (above system use) should be done. High power testing by the manufacturer at rated conditions may not be possible and testing by the user must take its place. High-power testing at maximum rated temperature (both above system usage) should be done. A one-hour test under these conditions while observing for hot spotting of the element and body surface temperature monitoring can be effective.

6.5 MICROWAVE DEVICES, DIRECTIONAL COUPLERS

6.5 Directional couplers.

6.5.1 Introduction. Directional couplers (DCs) are devices that couple power via electric and magnetic fields from the primary signal path to a secondary signal path. Low frequency (signal processing) devices are coupled by transformer action. High frequency devices are coupled by proximity of transmission lines in either coaxial or waveguide. Coupling is the ratio of the power in the primary arm to the power in the secondary arm expressed in decibels. Directivity is the ability of the secondary arm to sense the direction of power flow in the primary arm.

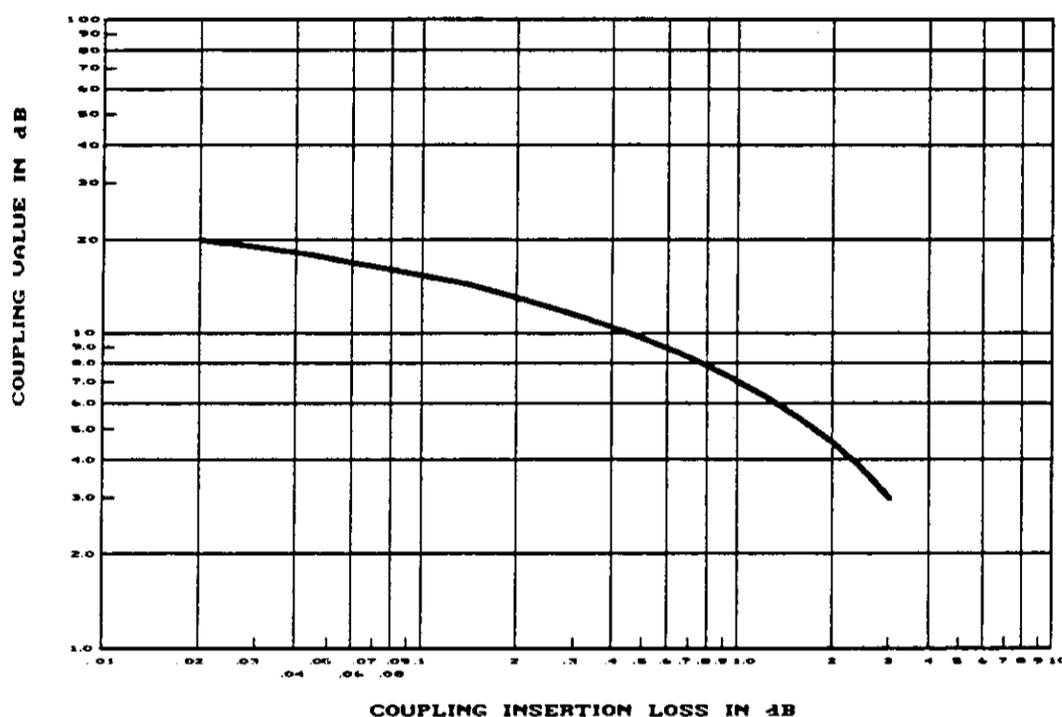
A very useful form of the DC is the microwave hybrid. The hybrid is a four-port, symmetrical device that can be made from transformers at high frequency and in stripline, microstrip, coaxial or waveguide at microwave frequencies. Power entering one port is split between two of the ports with the fourth port having a high isolation from the input signal.

6.5.2 Usual applications. The most common DC is the hybrid. It is used in amplifier chains to combine or split power, in balanced mixer designs for push-pull operation, and in double-balanced mixers to provide image rejection. It is also used in duplexers to separate signals, and in bridges for precision phase measurements, or anywhere a 3 dB, symmetrical, four-port, high-isolation device is needed.

Other uses of DCs are as signal or pulse samplers. For example, a DC that samples reflected power from a load indicates return loss or VSWR. In these applications, the sampled incident or reflected signal level is set by the coupling ratio to the power in the primary arm. Standard coupling values are 3, 6, 10, 20, and 30 dB. Any coupling value from 3 dB up is achievable. If the application involves sampling a high-power signal the worst possible load VSWR must be considered to avoid sending excess power into the secondary line reverse power termination. DCs span the frequency range from 1.0 MHz to greater than 100 GHz with limited operating frequency range per device.

Signal processing directional couplers are used in the frequency range of 1 to 1,000 MHz. Above this frequency, coaxial and waveguide couplers are available. Coaxial models are typically limited to a bandwidth of an octave. Waveguide couplers are limited by their respective waveguide transmission characteristics. Directivity in coaxial units is typically 15 to 20 dB, except in special designs. Directivity in waveguide can range from 20 to 40 dB, depending on the design. Insertion loss of couplers is affected by the coupling value for low coupling values as shown in Figure 65.

6.5.3 Physical construction. All DCs are essentially four-port devices with one port of the secondary either terminated or unterminated. Signal processing DCs are typically made using toroidal transformers with chip resistors and capacitors used for tuning, balancing, or terminating one-port. Package configurations include TO cans, DIPs, flat packs, and stripline with coaxial connectors. Hybrids of 180 and 90 degrees are common.

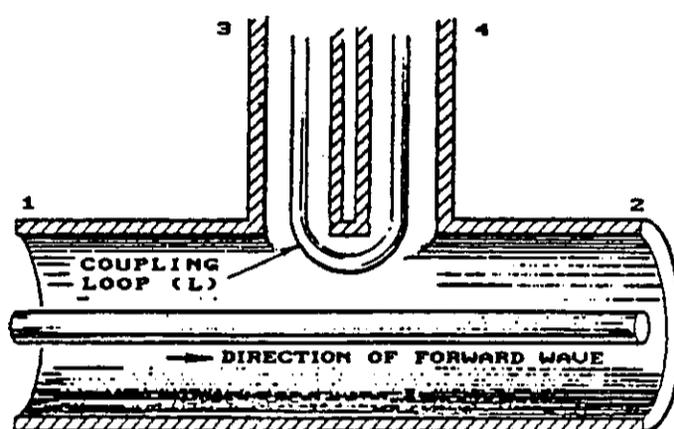
6.5 MICROWAVE DEVICES,
DIRECTIONAL COUPLERSFIGURE 65. Coupling value vs main line insertion loss.

Coaxial units are constructed using coupled line in coaxial, stripline, microstrip, and slab line. Secondary arm terminations are internal or screw-on coaxial loads, Figure 66A. Waveguide units are made by joining two pieces of waveguide with appropriate coupling apertures in the joining walls (i.e., there are broad wall, side wall and cross guide couplers), Figure 66B. A waveguide resistive loop coupler is made with a coaxial line as the secondary arm. A coupling slot in the outer conductor is embedded in the broad wall of the waveguide and is rotated and brought into proximity with a coupling hole in the waveguide until the desired coupling coefficient is obtained.

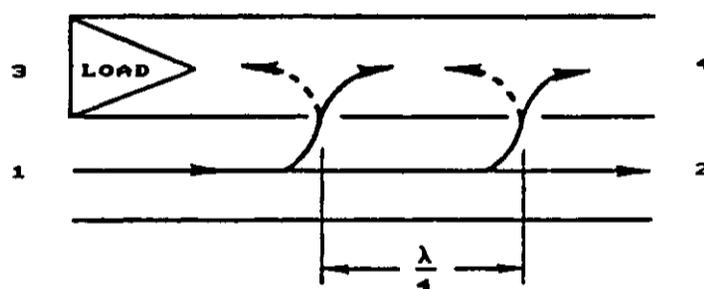
6.5.4 Military designations. The directional coupler selection standard is MIL-STD-1328. The military directional couplers are defined in MIL-C-15370.

6.5.5 Electrical characteristics. VSWR, insertion loss, coupling, coupling variation, and directivity are the primary parameters of a directional coupler. VSWR of the primary arm is measured with the device terminated in a matched load. VSWR of the secondary arm is measured with no load if the device is internally terminated. For DCs of less than 10 dB, the primary arm should be terminated in a matched load. Insertion loss of the primary arm is measured with a matched load on the secondary arm.

6.5 MICROWAVE DEVICES,
DIRECTIONAL COUPLERS



A. Coaxial coupler



B. Waveguide coupler

FIGURE 66. Directional coupler construction.

Coupling is usually a measure of the ratio of the power into the primary arm in the forward direction to the power out of the secondary arm at center frequency under matched conditions. The coupling coefficient is a directional parameter, because the reverse of coupling value is directivity. Coupling variation is the change in coupling value across the frequency band. A calibration chart plotting actual coupling versus frequency is often provided so that it is possible to dispense with extremely constant coupling.

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DIRECTIONAL COUPLERS**

Directivity is the ratio of the power at the secondary arm output with power into the primary arm in the forward direction, to power at the secondary arm with the same power as in the first condition going into the primary arm in the reverse direction at center frequency. The directivity equals the coupling coefficient under perfect short circuit conditions. Because the actual directivity is affected by load VSWR in the reverse direction, devices should be tested under conditions of system usage to assure adequate directivity.

6.5.6 Environmental considerations. Environmental considerations for directional couplers are both mechanical- and temperature-related. Because coupling is achieved by bringing two transmission media close to each other, any severe mechanical shock or vibration, or differences in coefficients of expansion and contraction can cause variations in the coupling coefficient. Braze joints, flanges, and bodies of waveguide DCs must withstand humidity and salt corrosion. High power waveguide DCs may have to be pressure tight and able to withstand high power at high altitudes.

6.5.7 Reliability considerations.

6.5.7.1 Failure modes. The dominant failure modes of directional couplers tend to be mechanical in nature. These failures are caused by leaks, structural separation, or displacement under stress. As previously mentioned, there is a possibility of electrical burn-out from secondary line termination overload.

6.5.7.2 Screening. Probably the best single screen is repeated temperature cycling (10 cycles). A 100-percent seal test such as MIL-STD-202, Method 112, condition A, provides an endpoint measurement of mechanical integrity.

If the secondary line termination is a discrete resistor, a power burn-in may be helpful. As with attenuators, there is little purpose in attempting power burn-in with distributed or bulk terminating resistors. However, in both discrete, distributed, and bulk terminating resistors, there is an advantage in a temperature-cycling screen of both the resistor and the method used to attach it to the structure. High-power waveguide DCs may require a pressure test and high-power, high-temperature or high-altitude, 1-hour burn-in to screen from arcing.

As with attenuators, while monitored shock and vibration may be helpful, such a procedure may be more indicative of weakness in the monitoring equipment than the coupler under test.

**6.6 MICROWAVE DEVICES, FLANGES
AND WAVEGUIDE ASSEMBLIES**

6.6 Flanges and waveguide assemblies.

6.6.1 Introduction. Microwave circuits are often configured in the waveguide transmission medium. Common usage ordinarily limits "waveguide" to some sort of metallic tubing and "waveguide assembly" to denote an assembly of waveguide sections. The hardware involved makes up the category "waveguide accessories" and includes flanges, adapters, and gaskets.

The waveguides discussed are air dielectric and may have cross sections that are coaxial, rectangular, rectangular with a re-entrant ridge in one or both of the wide sides, circular, or elliptical (see Figure 67). Waveguides may be further subdivided into rigid or nonrigid structures. Waveguide assemblies may be composed of one or more of these subdivisions.

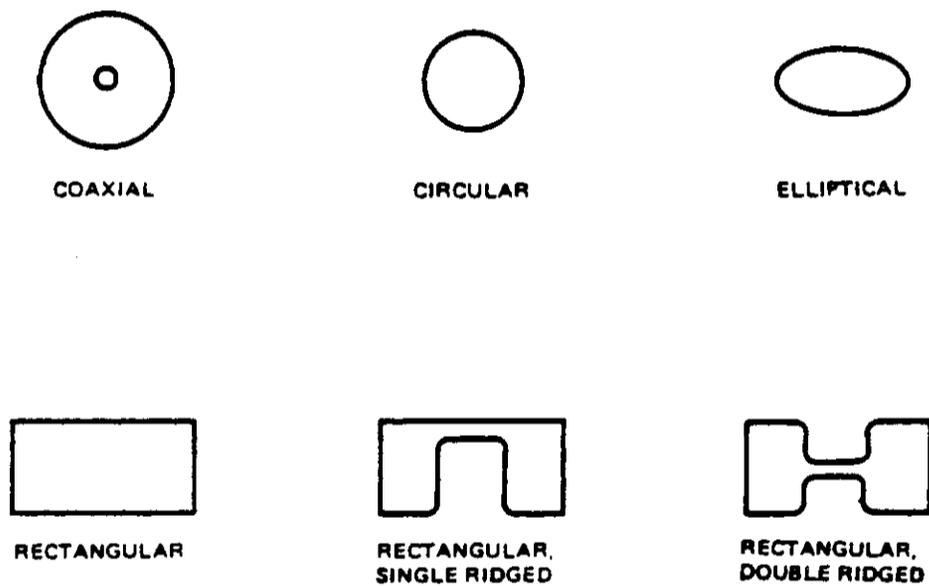


FIGURE 67. Waveguide cross-sections.

6.6.2 Usual applications. Waveguide flanges are the interface connections for waveguides. They are usually brazed onto waveguides and bolted or screwed together to make a waveguide interconnection. Two types of flanges generally used are cover and choke flanges. A choke to cover flange interface is used in pressurized systems. The choke flange is similar to a flat-face cover flange, but it has a recessed area for a gasket to maintain pressure and an rf choke area to prevent rf leakage.

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Waveguides have four standard shapes which offer the designer trade-offs in power handling capability, attenuation, and bandwidth. The rectangular waveguide, most commonly used, has the highest power capability. It has reasonably low attenuation, but limited bandwidth. Circular waveguide is usually used in rotary joints. It also has the advantage of having no upper frequency cutoff and is extremely wideband. Single ridge and double ridge waveguides offer wide bandwidth at low attenuation with limited power handling capability.

Waveguide assemblies are the building blocks of a waveguide system. Devices include bends (E and H), twists, and flexible waveguide--all with selected flanges. Also available are an assortment of devices to construct waveguide assemblies, including short slot hybrids, magic Ts, directional couplers, radiating devices, etc.

Adapters are waveguide transition devices which are used to make a transition to a different size or shape of waveguide or coaxial transmission line. Waveguide-to-waveguide transition must be several wavelengths long to match the complete bandwidth of the waveguide. Waveguide-to-coaxial adapters may have moderate VSWRs at the waveguide band edges.

Gaskets are designed to fit on waveguide flanges and prevent pressure leakage. Gaskets are typically silicone rubber and, when filled with silver-plated copper, prevent rf leakage.

6.6.3 Physical construction. Standard waveguides, flanges, assemblies (other than flexible) and adapters are constructed of either copper or aluminum alloys. Flexible waveguide assemblies are sometimes rubberized on the outside to reduce vibration fatigue and prevent pressure leakage. Fabricated waveguide parts and purchased waveguide parts should have finishes to prevent corrosion and reduce insertion loss. Military specifications describe in detail the materials and finishes used.

6.6.4 Military designations. Standard waveguide flanges may be selected from MIL-STD-1327. The general specifications for waveguide flanges are MIL-F-3922 and MIL-F-39000.

Rectangular, circular, and ridge waveguide selection is covered in MIL-STD-1358 and in MIL-W-85, MIL-W-23068, and MIL-W-23351.

Waveguide to coaxial adapters selection is covered in MIL-STD-1636 and MIL-A-22641.

Gaskets are usually supplied with flanges and are defined in MIL-F-3922 and MIL-F-39000.

6.6 MICROWAVE DEVICES, FLANGES AND WAVEGUIDE ASSEMBLIES

6.6.5 Electrical characteristics. The primary characteristics of waveguide hardware are VSWR and insertion loss. Secondary characteristics would include rf leakage or heating effects. Electrical characteristics of waveguide hardware depend primarily on mechanical attributes. VSWR is affected by interface mismatch due to tolerance build up in mounting, internal physical dimensions, surface roughness, and smoothness of transitions. Any of these mechanical faults causes discontinuities and results in power reflections. Surface finish, skin depths, and resistivity will affect insertion loss. Low insertion loss will keep heating effects to a minimum. The rf leakage occurs predominately at interfaces and depends on mechanical fit; it may be reduced with conductive gasketing. High-power applications may cause arcing or breakdown. Surface smoothness and finish may prevent arcing. Pressurization of waveguide increases the peak power rating of waveguide.

6.6.6 Environmental considerations. Metal corrosion of waveguide hardware may occur in the presence of moisture. Causes of the corrosion may be inadequate removal of brazing salts, improper or inadequate metal finishes, or contact of dissimilar metals. Metal porosity and inadequate seals can cause pressurization leakage. Metal fatigue due to usage or system vibration can occur in flexible waveguide. Thin wall waveguide may cause am and pm noise to appear on a conducted signal, due to high system vibration levels. High usage interconnections should use corrosion resistant, passivated steel inserts for durability.

6.6.7 Reliability considerations. In systems requiring pressurization, screening for air leakage by an air-bubble test or a dye penetrant test should be performed. Either thermal shock (10 cycles) or temperature cycling (25 cycles) is an excellent test for compatibility of materials and finishes. Physical appearance and delta VSWR or delta insertion loss measured in a standard fixture should be as acceptance criteria. Material and finish certification with periodic salt atmosphere testing should assure adequate finishes.

6.7 MICROWAVE DEVICES, RADIO FREQUENCY POWER DIVIDERS AND COMBINERS

6.7 Radio frequency power dividers and combiners.

6.7.1 Introduction. Communications equipment has often been designed to have the signal of interest travel a single path. As equipment has become more varied and elaborate, designers have found advantage in having signals travel two or more paths. This causes the twin problems of dividing the signal into the desired parts and, after processing, combining the results.

Because most power dividers can also be used as power combiners, the problem is simplified. There are two pairs of conditions commonly found: (1) Signals are at a sufficiently high level so that dissipation will not cause critical degradation of noise level and low enough so that dissipation will not cause heating problems, or the reverse. (2) The division and combination is at zero (or 180) degrees, or in quadrature (90 degrees) phase relation. In each of the first alternatives, the devices may be resistive; in the second alternatives there must be at least some reactive elements.

The designer may also want to provide some minimum degree of isolation between loads (in a divider) or sources (in a combiner). The degree of necessary isolation and impedance match will determine which of several choices should be used.

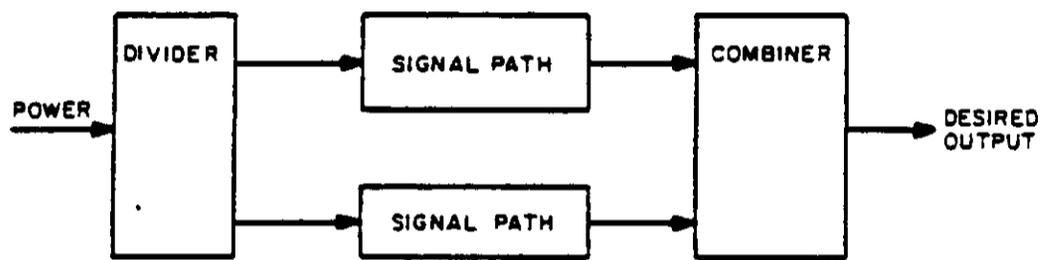
A two-way, in-phase (0 degrees) power divider is actually an internally terminated 180-degree hybrid. If the division and combination is at 180 degrees or in quadrature (90 degrees), the device is a hybrid.

6.7.2 Usual applications. The most common use of dividers/combiners is the division into two equal parts or the addition of two signals, (Figure 68A) One obvious solution, if more than two parts are desired or are to be combined (Figure 68B), is to "branch" divider/combiners, as in Figure 69. This method may have very persuasive advantages when the number of divisions/combinations is some power of 2 such as 4, 8, 16, or 32. The intermediate numbers, however, represent wasted power, for the unwanted paths should be terminated in dummy loads. Where there is no wasted power desired and the number of outputs and inputs is known, there are "n-way" dividers/combiners available. It is possible to combine 2-way and n-way devices, but for reasons related to the phase of the output or loss it may not be advisable.

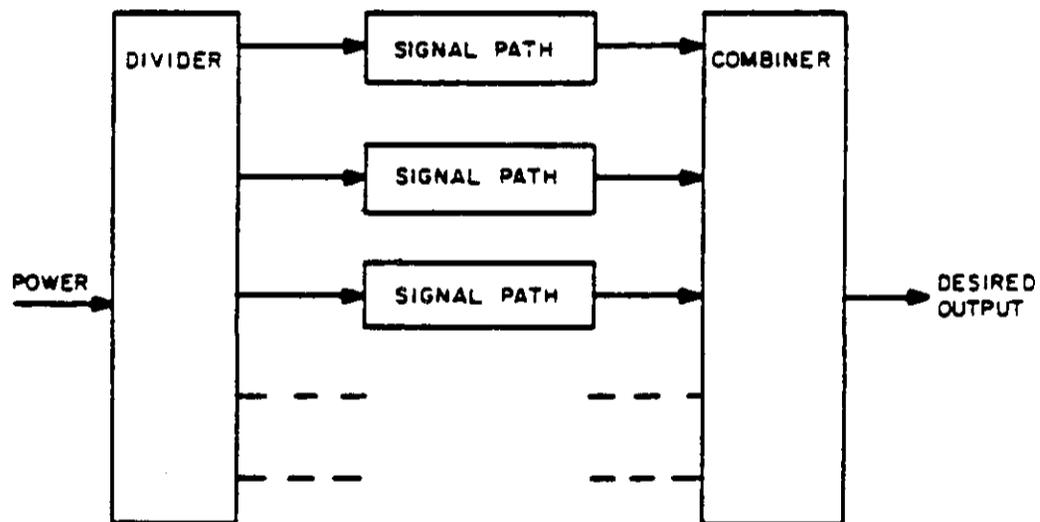
One obvious use of a divider is to proportion power to the various elements of an antenna array for transmission or, alternatively, to combine signals from the same elements for receiving.

Transistor amplification at high frequencies and microwaves is often accomplished with the use of power dividers and combiners. Signals can be amplified at lower power levels (if divided) and the power added from the several amplifiers (in the signal paths) in a combiner (see Figure 70). There is an additional advantage to this type of amplification if a form of divider/combiner is used that provides isolation between the various paths. One or more amplifiers can fail catastrophically without the array losing more than 1/n of the array gain/amplifier lost. The type of divider/combiner used in such application is commonly known as a "hybrid," and under some circumstances may provide 20 dB to 40 dB isolation.

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A. 2-way power divider/combiner



B. N-way power divider/combiner

FIGURE 68. Use of 2-way and n-way power divider/combine.

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POWER DIVIDERS AND COMBINERS

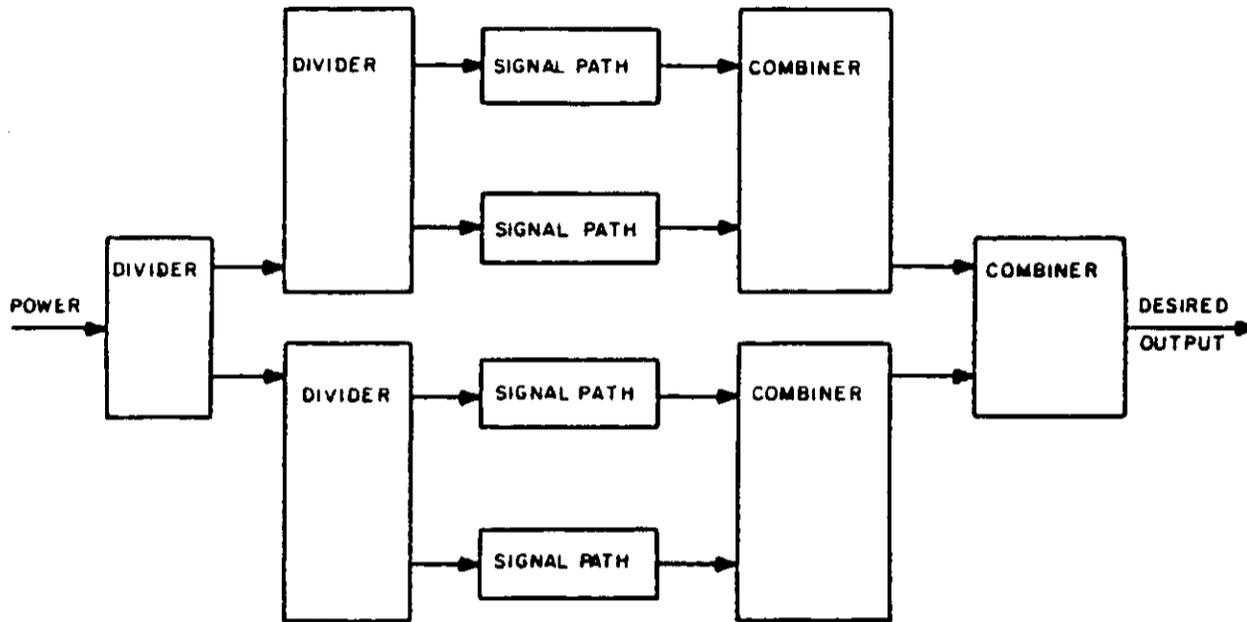


FIGURE 69. Use of 2-way dividers/combiners where $n = 2^x$.

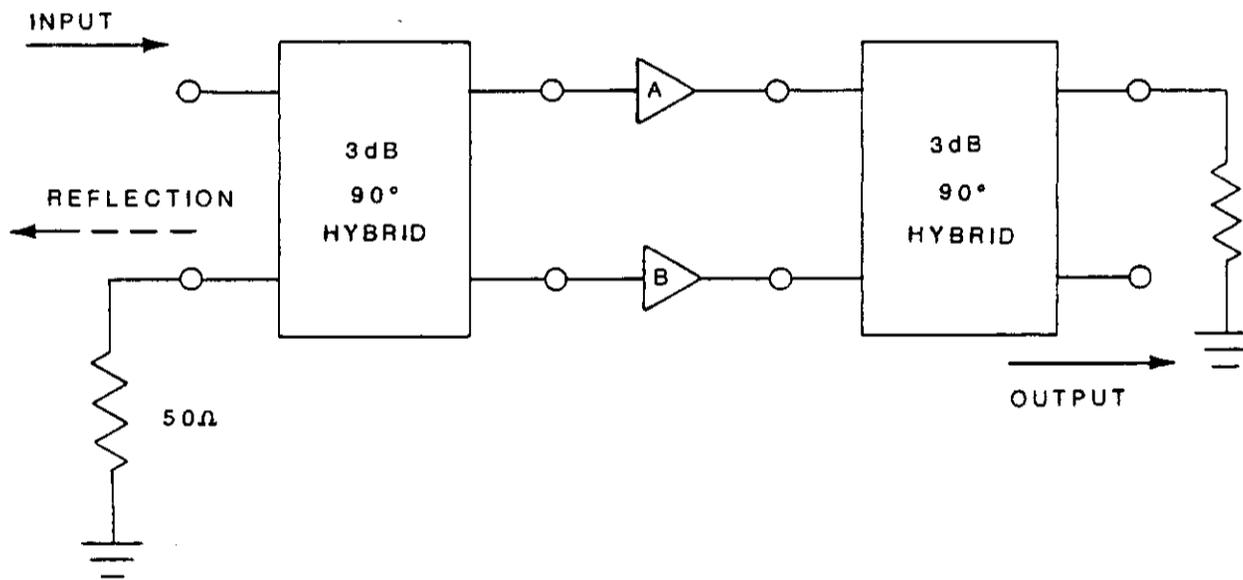


FIGURE 70. Balanced amplifier.

6.7 MICROWAVE DEVICES, RADIO FREQUENCY POWER DIVIDERS AND COMBINERS

A 180-degree hybrid is a reciprocal four-port device which provides two equal amplitude in-phase signals when fed from its sum port and two equal amplitude, 180-degree, out-of-phase signals when fed from its difference port. Opposite ports of the hybrid are isolated.

Utilizing the functional diagram of Figure 71, we can consider the application of signal at one or more of the ports of the hybrid. The cases that are important to consider are the following:

- a. Operation as a power divider--one source operating at ports A, B, C or D
- b. Operation as a power combiner--two sources operating at ports A and B, or C and D.

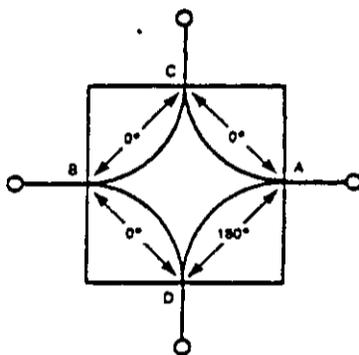


FIGURE 71. 180° Hybrid.

As a power divider, the hybrid will split equally the input signal and deliver one half the power to each load. Because all ports are considered to be at Z_0 impedance, the voltages at the outputs will be proportional to the square root of the output power and will be phase shifted by the amount indicated for that path of the hybrid, since $P_{out} = \frac{1}{\sqrt{2}} P_{in}$, $V_{out} = \frac{1}{\sqrt{2}} V_{in}$. For example, under

matched conditions, a source voltage of $2E \cos \omega t$ will supply a voltage of $E \cos \omega t$ to the input of the hybrid. If an input signal at Port A of $E \cos \omega t$ is injected, the resultant output is:

$$\text{At Port C } \frac{1}{\sqrt{2}} E \cos \omega t$$

$$\text{At Port D } \frac{1}{\sqrt{2}} E \cos (\omega t - 180^\circ)$$

No signal will appear at Port B. The various power divider relationships are summarized in Table V.

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TABLE V. Power divider relationships for 180-degree hybrids

Input Signal	Input Port	Output Signals			
		Port A	Port B	Port C	Port D
	A	-	0	$\frac{1}{\sqrt{2}} E \cos \omega t$	$\frac{1}{\sqrt{2}} E \cos(\omega t + 180^\circ)$
$E \cos \omega t$	B	0	-	$\frac{1}{\sqrt{2}} E \cos \omega t$	$\frac{1}{\sqrt{2}} E \cos \omega t$
	C	$\frac{1}{\sqrt{2}} E \cos \omega t$	$\frac{1}{\sqrt{2}} E \cos \omega t$	-	0
	D	$\frac{1}{\sqrt{2}} E \cos(\omega t + 180^\circ)$	$\frac{1}{\sqrt{2}} E \cos \omega t$	0	-

When the hybrid is used as a combiner, it performs the vector addition of two signals. For example, in the general case of two equal amplitude, equal frequency signals of arbitrary phase applied to Ports A and B; $E \cos(\omega t + \alpha) + E \cos \omega t$, the resultant outputs C and D will be

$$\begin{aligned} \text{Resultant C} &= \frac{E}{\sqrt{2}} [\cos \omega t + \cos(\omega t + \alpha)] \\ &= \frac{E}{\sqrt{2}} [2 \cos \frac{1}{2}(2\omega t + \alpha) \cos \frac{1}{2}(-\alpha)] \\ &= \sqrt{2} E \cos(-\alpha/2) \cos(\omega t + \alpha/2) \end{aligned}$$

$$\begin{aligned} \text{Resultant D} &= \frac{E}{\sqrt{2}} [\cos(\omega t + 180^\circ) + \cos(\omega t + \alpha)] \\ &= \frac{E}{\sqrt{2}} [2 \cos \frac{1}{2}(2\omega t + \alpha + 180^\circ) \cos \frac{1}{2}(180^\circ - \alpha)] \\ &= \sqrt{2} E \cos(90^\circ - \alpha/2) \cos(\omega t + \alpha/2) \end{aligned}$$

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Table VI lists the relationships for various combinations of signals applied to ports A and B or C and D.

A 90-degree hybrid functions in much the same manner as a 180-degree hybrid. Equal amplitude outputs result when a signal is fed to one of the inputs. Opposite ports of the 90-degree hybrid are also isolated as in the 180-degree hybrid. Figure 72 shows the functional diagram of the quadrature hybrid. As can be seen from this diagram, a signal applied to any input will result in two quadrature, or 90-degree outputs. Ports A and B or C and D are isolated.

Table VII shows the two equal amplitude outputs when a matched signal $E \cos \omega t$ is applied to each of the inputs. To use the hybrid as a combiner, apply two matched, equal amplitude signals to ports A and B, or C and D, with arbitrary phase. If, for example, the signal $E \cos \omega t$ is applied to port A and $E \cos(\omega t + \alpha)$ is applied to port B, the resulting outputs are

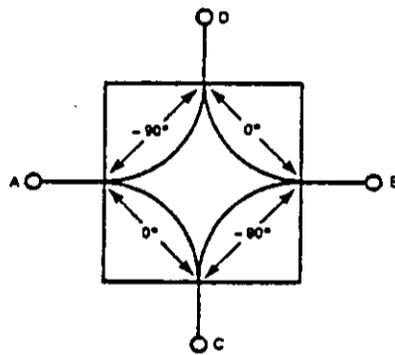


FIGURE 72. 90-degree hybrid.

$$\begin{aligned}
 \text{Resultant C} &= \frac{E}{\sqrt{2}} [\cos \omega t + \cos (\omega t + \alpha - 90^\circ)] \\
 &= \frac{E}{\sqrt{2}} [2 \cos 1/2 (2 \omega t + \alpha - 90^\circ) \cos 1/2 (-\alpha + 90^\circ)] \\
 &= \sqrt{2} E \cos (45^\circ - \frac{\alpha}{2}) \cos (\omega t + \frac{\alpha}{2} - 45^\circ)
 \end{aligned}$$

TABLE VI. Power combiner relationship for 180-degree hybrids

Input Signal	Input Port	Output Signals			
		Port A	Port B	Port C	Port D
*E cos ωt	A	-	-	$\sqrt{2} E \cos \omega t (-\alpha/2)$	$\sqrt{2} E \cos(90^\circ - \alpha/2)$
*E cos($\omega t + \alpha$)	B			$[\cos(\omega t + \alpha/2)]$	$[\cos(\omega t + \alpha/2 + 90^\circ)]$
E cos ωt	A	-	-	$\sqrt{2} E \cos \omega t$	0
E cos ωt	B				
E cos($\omega t + 180^\circ$)	A	-	-	0	$\sqrt{2} E \cos \omega t$
E cos ωt	B				
E cos ωt	C	0	$\sqrt{2} E \cos \omega t$	-	-
E cos ωt	D				
E cos ωt	C	$\sqrt{2} E \cos \omega t$	0	-	-
E cos($\omega t + 180^\circ$)	D				
E cos $\omega_1 t$	A	-	-	$\frac{1}{\sqrt{2}} E (\cos \omega_1 t + \cos \omega_2 t)$	$\frac{1}{\sqrt{2}} E (\cos \omega_1 t + \cos \omega_2 t)$
E cos $\omega_1 t$	B				

*See previous page for derivation of this example.

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$$\begin{aligned}
 \text{Resultant D} &= \frac{E}{\sqrt{2}} [\cos (\omega t - 90^\circ) + \cos (\omega t + \alpha)] \\
 &= \frac{E}{\sqrt{2}} [2 \cos 1/2 (2\omega t + \alpha - 90^\circ) \cos 1/2 (-\alpha - 90^\circ)] \\
 &= \sqrt{2} E \cos \left(-45^\circ - \frac{\alpha}{2}\right) \cos \left(\omega t + \frac{\alpha}{2} - 45^\circ\right)
 \end{aligned}$$

$$\text{Phase of Resultant C} = \text{Phase of Resultant D} = \left(\omega t + \frac{\alpha}{2} - 45^\circ\right)$$

TABLE VII. 90-degree hybrid as a power divider

Input Signal	Input Port	Output Signals			
		Port A	Port B	Port C	Port D
$E \cos \omega t$	A		$\frac{E}{\sqrt{2}} \cos (\omega t - 90^\circ)$	$\frac{E}{\sqrt{2}} \cos \omega t$	0
	B	$\frac{E}{\sqrt{2}} \cos (\omega t - 90^\circ)$		0	$\frac{E}{\sqrt{2}} \cos \omega t$
	C	$\frac{E}{\sqrt{2}} \cos \omega t$	0		$\frac{E}{\sqrt{2}} \cos (\omega t - 90^\circ)$
	D	0	$\frac{E}{\sqrt{2}} \cos \omega t$	$\frac{E}{\sqrt{2}} \cos (\omega t - 90^\circ)$	

The amplitudes of the resultant outputs at C and D vary, based on the phase of the inputs, whereas the phases of the outputs are always equal. This property is useful for receiving phase-modulated signals. The relationships for a 90-degree hybrid with signals applied to ports A and B or C and D are shown in Table VIII.

TABLE VIII. 90-degree hybrid as a power combiner

Input Signal	Input Port	Output Ports			Port B
		Port A	Port D	Port C	
$E \cos \omega t$	A		$\sqrt{2} E \cos(-45^\circ - \alpha/2)$	$\sqrt{2} E \cos(-45^\circ - \alpha/2)$	
$E \cos(\omega t + \alpha)$	B		$[\cos(\omega t + \alpha/2 - 45^\circ)]$	$[\cos(\omega t + \alpha/2 - 45^\circ)]$	
$E \cos \omega t$	A		0	$\sqrt{2} E \cos \omega t$	
$E \cos(\omega t + 90^\circ)$	B				
$E \cos(\omega t + 90^\circ)$	A		$\sqrt{2} E \cos \omega t$		
$E \cos \omega t$	B				
$E \cos \omega t$	D	0			$\sqrt{2} E \cos \omega t$
$E \cos(\omega t + 90^\circ)$	C				
$E \cos \omega t$	D	$\sqrt{2} E \cos \omega t$		0	0
$E \cos(\omega t + 90^\circ)$	C				
$E \cos \omega_1 t$	B		$\frac{E}{\sqrt{2}} \left[\cos \omega_1 t + \cos(\omega_2 t - 90^\circ) \right]$	$\frac{E}{\sqrt{2}} \left[(\cos(\omega_1 t - 90^\circ) + \cos \omega_2 t) \right]$	
$E \cos \omega_2 t$	A				

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The quadrature hybrid is often used to generate or assist in the detection of single sideband signals, in combination with mixers. One characteristic of mixers (actually algebraic multipliers) in heterodyning systems is the creation of sum and difference frequencies. A receiver is sensitive to an image frequency as well as a desired frequency. Processing a received signal through in-phase (I) and quadrature (Q) channels, however, allows addition of the desired signal components and reduces the image-frequency components -10 dB to -40 dB (see Figure 73). The reverse process can be used to generate a single-sideband signal.

The hybrid may be used as a directional coupler when one of the four ports of a 2-way hybrid is terminated in its design. The pair of ports isolated from each other may be used as the "input" and "reflected-power" ports of a directional coupler with the "load" connected to the combined port. Any reflection from the load will cause a proportional amount of power to appear at the reflected-power port. Power loss of 3 dB in the reactive hybrid or 6 dB in the resistive hybrid commonly eliminates consideration of use as a directional coupler, except for switching and instrumentation applications.

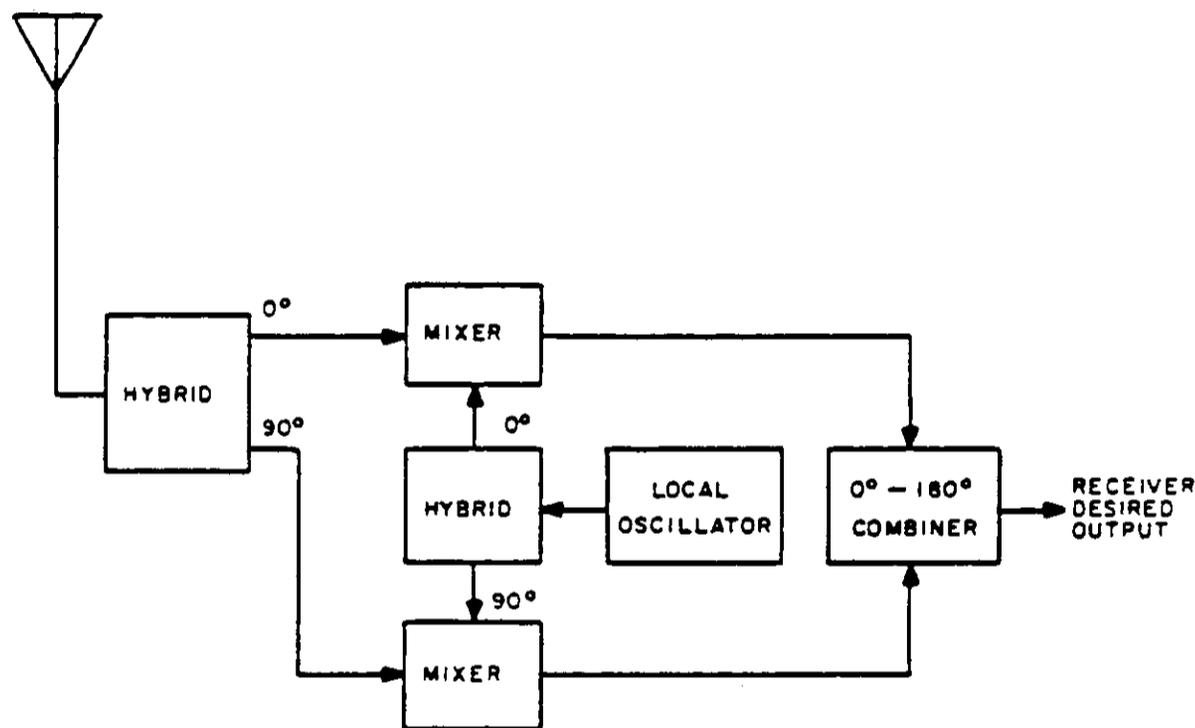


FIGURE 73. Phasing system of single-sideband reception.

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The two common switching applications for the reactive hybrid are to provide "lossless" TR and ATR switching between transmitter and receiver by total reflection from open or short-circuit terminations, or provide a digitally-determined phase shift as in an electronically-steerable antenna array (see Figure 74).

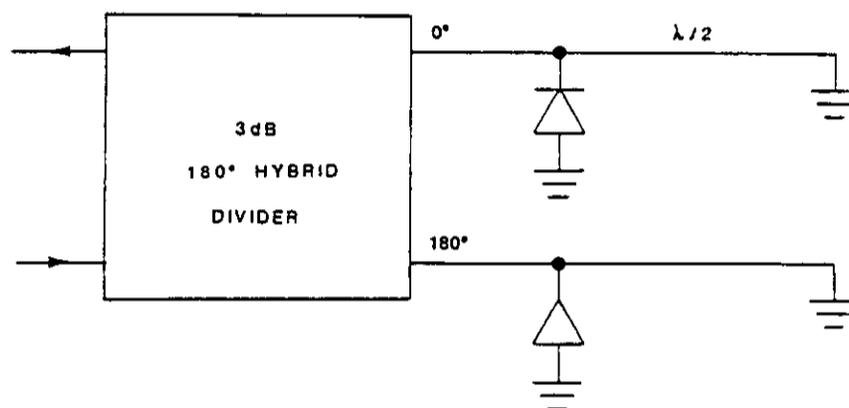


FIGURE 74. Single bit phase shifter.

The amplitude and phase of the reflection from a nonzero or noninfinity termination may be used to either discover what the termination is measurement, or to provide a phase shift (by reflection from a fixed or variable reactance or sliding short).

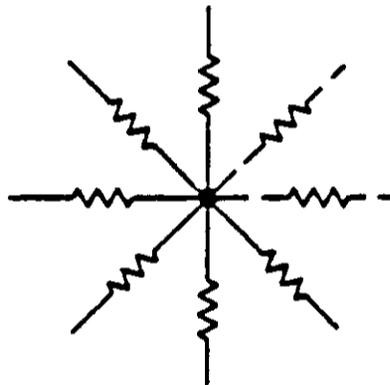
6.7.3 Physical construction. The most obvious physical features are controlled by the intended frequency range. At dc and relatively low frequencies, the parts making up the divider/combiner tend to be bulky with relatively open construction and wiring. When a high degree of isolation is required in the hybrids, balance of part values, wiring layout, and coupling between the windings of transformers become very important. A slight change of wire position may be the difference between 40 dB and 60 dB isolation. Shielding at low frequencies may be helpful, but even at low frequencies stray capacitance of wiring-to-shielding may cause problems.

As frequency rises, the parts tend to shrink and part types having less frequency sensitivity must be used. Film resistors replace composition and wire-wound types. Transformers become toroidal with bifilar (or 3-in-hand or even 4-in-hand) windings. Core material changes from iron to powdered iron to ferrite. Capacitors become low-inductance ceramic or mica. Connectors become coaxial or fitted with pins for insertion into the easily-duplicated wiring of a printed circuit board.

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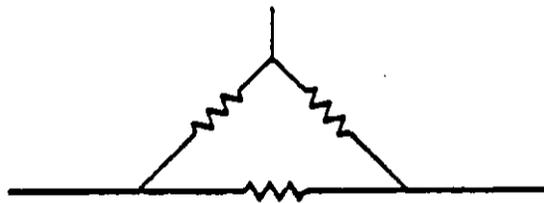
As frequency continues to rise above approximately 500 MHz, discrete parts give way to transmission lines. The transmission lines may be used singly to provide phase delay or reactance, or as coupled pairs to provide signal transfer. Center-tapped transformers are replaced with transmission-line baluns (balanced-to-unbalanced transforming networks). There is overlap in frequency of usage between coaxial transmission line and stripline, and then stripline to microstrip. Progressive increase of frequency or power mandates change to waveguide structures.

The simplest form of a divider/combiner is shown in Figure 75A. Any of the resistors may be used as the input port or ports with the remainder as output ports. The delta-circuit equivalent of the 3-port version of Figure 75A is shown in Figure 75B. The delta equivalent is rarely used with more than 3-ports, because for those cases, it requires more resistors than the star circuit and is physically more awkward. Some possible physical configurations are shown in Figure 76.



Each resistor = $Z_0 (N-2)/N$

A. Star network with n-ports



Each resistor = Z_0

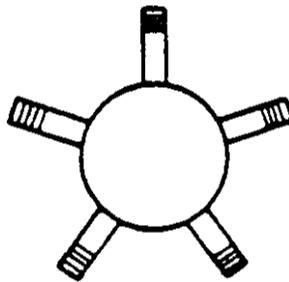
B. Delta network with ports

FIGURE 75. Resistive divider/combiners.

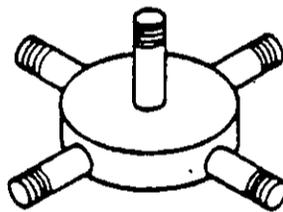
**6.7 MICROWAVE DEVICES, RADIO FREQUENCY
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A high frequency, 2-way, in-phase divider/combiner is shown in Figure 77. Transformer T2 is usually a bifilar-wound toroid and T1 provides a 2:1 impedance stepup to port 1. The resistor is an internal termination, and in dividing a port 1 input, it has zero required dissipation when the loads on ports 2 and 3 are equal to each other. When the loads on ports 2 and 3 are unequal, the resistor is called upon to dissipate power. The ghosted capacitor is often used to broaden the useful bandwidth of the device.

The most famous early waveguide hybrid is the Magic-T (Figure 78). Ports are made in straight waveguide so that one port receives the "electric" field and the other the "magnetic" field of any wave attempting passage. Because half the wave power is in each field, the power division is equal. Obversely, input to the E-plane port will result in 180-degree phase-difference output in the colinear ports, and input to the H-plane port will result in zero phase difference in the colinear port output. Because of mismatches at the junction, it is fairly common to provide an impedance match by such means as a post for H-plane compensation or an asymmetric diaphragm for E-plane compensation. This matching necessarily narrows bandwidth.



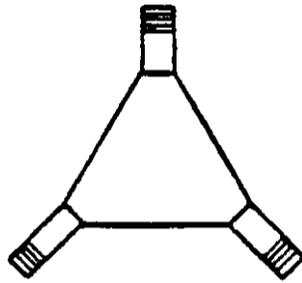
A. Five-port star (may have less or more branches)



B. Five-port star (may have less or more branches). It is a useful shape when one port is dedicated as input or output.

FIGURE 76. Some external configurations particularly adapted to resistive divider/combiner network.

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C. Frequent shape of a 3-port delta

FIGURE 76. Some external configurations particularly adapted to resistive divider/combiner network (continued).

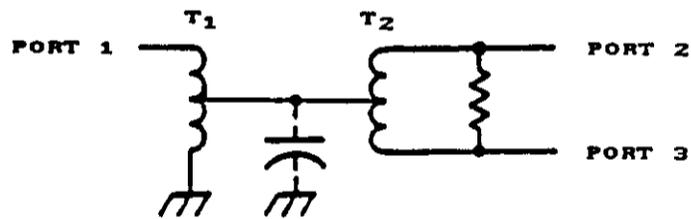


FIGURE 77. Simple zero-degree hybrid.

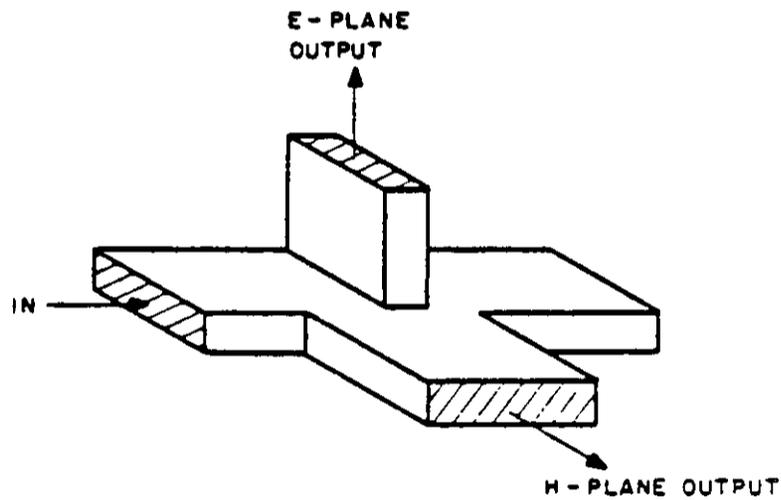


FIGURE 78. The "Magic-T."

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A hybrid that yields similar results may be obtained with the "6/4" rat-race (Figure 79). This configuration may be constructed in waveguide, coaxial cable, or stripline transmission lines. Again, the impedance of the transmission line in the rat-race should be different from the port impedances to provide impedance match.

A similar appearing quadrature, shown in Figure 80A, is known as a pi circuit. For stripline applications it has generally been superseded with the quarter-wave coupled quadrature hybrid as shown in Figure 80B. Where increased bandwidth is desired, a 3/4-wave coupled hybrid can be used (Figure 80C). The lines of the hybrid circuit are usually one above the other, but may be adjacent and are bent back and forth to save board space at relatively low frequencies. An example of this type of hybrid is shown in Figure 81.

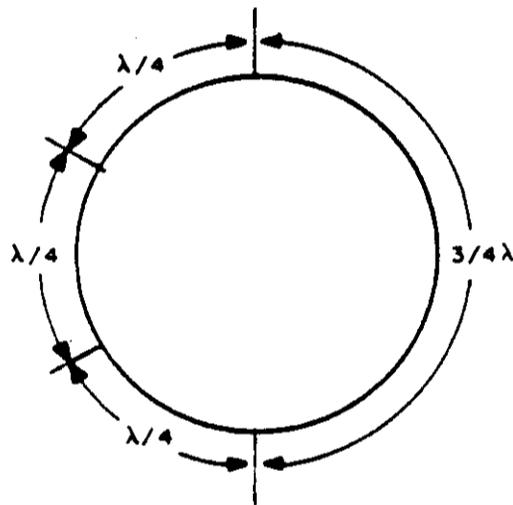


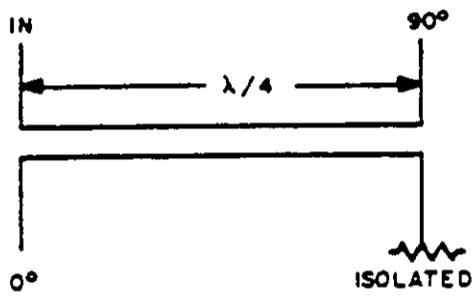
FIGURE 79. "6/4" rat race hybrid ring.



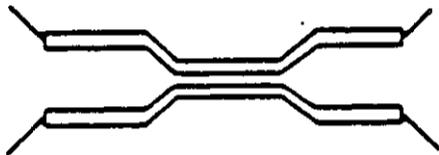
A. Branch line quadrature hybrid

FIGURE 80. Some forms of hybrid particularly adapted to stripline.

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B. Quarter-wave coupled quadrature hybrid



C. Three-quarter-wave coupled hybrid

FIGURE 80. Some forms of hybrid particularly adapted to stripline (continued).

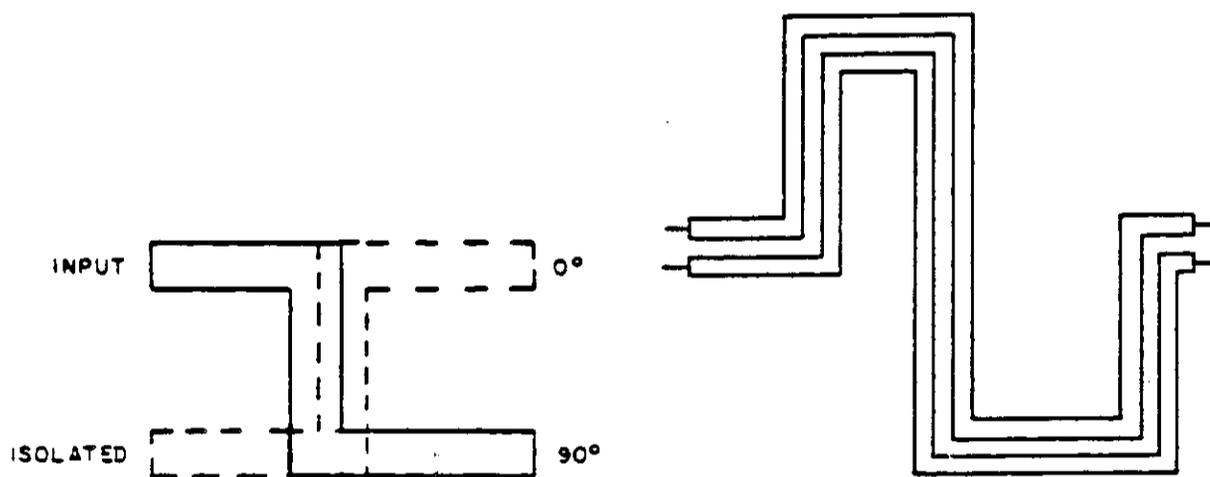


FIGURE 81. Stripline hybrids bent back to save space.

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At lower frequencies, quadrature hybrids are usually made of a mesh network of discrete inductors and capacitors intended to give equal-amplitude outputs having a near-constant phase difference of 90 degrees. Generally, the more complex the network, the nearer the ideal conditions can be approached.

Both low and high frequency "n-way" divider/combiners may be made; the simplest appearing approach being the Wilkinson configuration (Figure 81). In this configuration, sections of quarter-wavelength transmission line with an impedance equal to $Z_0 (n)^{1/2}$ are used to provide the impedance match. A somewhat similar configuration is used at lower frequencies, with the transmission lines replaced with transformer-appearing structures.

In each configuration, parts are commonly soldered into position, and when that is not possible, attachment to the case with epoxy or other adhesive is favored by the manufacturer. Use of adhesives requires careful consideration of the materials being bonded together, as well as the temperature, shock, and vibration environments to be encountered, and the electrical effect of the adhesive.

Most often, the divider/combiner is within a shielded enclosure. Sometimes this is only the pair of ground-planes of a stripline, but it is more often continuous. It should be remembered that a good electrical shield is not necessarily a hermetic seal. It should also be remembered that commonly used conductive epoxy may not be a very good rf shield.

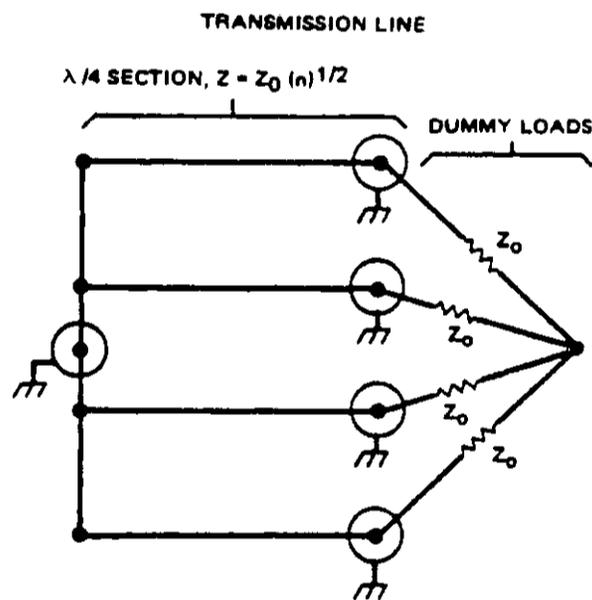


FIGURE 82. Wilkinson n-way hybrid all ports Z.

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Waveguide quadrature hybrids are constructed by paralleling two waveguides with a common wall, and having one or more openings in the common wall. If the narrow wall is common, the device is a "side-wall" hybrid; if the broad wall is common, the device is a "top-wall" hybrid. The size and positioning of the slot or slots between the waveguides affect the electrical characteristics.

6.7.4 Military designation. The military general specification for power combiners and divider/combiners is MIL-P-23971.

6.7.5 Electrical characteristics. The prime purpose of the hybrids and other power dividers/combiners under discussion is to provide equal known outputs with known phase relations. A secondary purpose is to provide isolation between several sources or loads is desired. Thirdly, though still important, is the desire to fit the device into a system having a certain characteristic impedance.

The impedance relations of Figure 74 define the target value of the resistances in the resistive power divider. The desire to match the characteristic impedance of the system results in the additional power loss shown in Table IX over that shown in Table X.

TABLE IX. Resistive dividers/combiners

Number of Ports	Output Level, dB
3	- 6.02
4	- 9.54
5	-11.14
6	-14.0
7	-15.56
8	-16.90

Note that the isolation from one port to another is no more or less than the tabulated value of output level. For identical construction in each of the legs of the star, all outputs would have the same amplitude and phase when feeding identical loads. Out of necessity, there will be small differences in the internal resistances and dimensions, and the loads will not be a perfect match. Therefore, it is necessary to specify how close to a perfect match of the nominal system characteristic impedance the test loads are. A maximum load VSWR of 1.01 is suggested.

The degree of amplitude and phase balance should be specified. The amount of loss in the combiner or divider in excess of the stated values of Table X should be specified. A fair approximation can be made by averaging the output levels and taking the difference between the average and the ideal.

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The same characteristics are important with the hybrids, except that they can be expected to be more frequency-dependent. Additionally, the hybrid has the isolation feature, which although high is not infinite. Combined with poor loads, this feature may cause reduced isolation proportional to the mismatch. It is important to note in the specification of hybrids that both the internal impedance of the generator and the termination of the isolated port affect isolation. Usually isolation is specified as "the insertion loss between a pair of ports with the other pair of ports terminated in the nominal impedance."

The simplest quadrature hybrid is very a narrow band for phase and amplitude. The common 1/4-wave coupled hybrid will have two crossover points and a fair match, otherwise, over a 2:1 frequency range. The 3/4-wave coupled hybrid will have three cross-over points, and a fair match over, perhaps, a 4.5:1 frequency range.

The zero/180 degree hybrids are somewhat more wideband in the low-frequency types, and component-limited at the higher frequencies. It is important to note in the specification that "zero-degree" ports may, in fact, be 360-degree ports, and the desired effect may not be obtainable on a nonrecurring wave.

Likewise, some forms of hybrid are "slow-wave" and others "fast-wave." Mixing the two forms in a single path may cause a nonlinear phase shift with frequency dispersion.

The linking in specification of the hybrid technology of directional couplers to that of the power dividers and combiners has caused some confusion in nomenclature. It is perhaps natural with a 1/4-wave quadrature power divider to think of the energy in the second line as "coupled" and the energy reduction in the first line as "coupling loss"; yet, in most cases, the truly significant factors are amplitude equality and phase deviation from ideal. The amplitude with ideal division (Table X) in "lossless" hybrids may be subtracted from the average of observed outputs of a number of hybrids to give close approximation of the net loss in the device.

TABLE X. Reactive (zero loss) dividers

Number of Inputs or Outputs	Output Level dB
2	-3.01
3	-4.77
4	-6.02
5	-6.99
6	-7.78
7	-8.45
8	-9.03

6.7 MICROWAVE DEVICES, RADIO FREQUENCY POWER DIVIDERS AND COMBINERS

The internal power rating of the combiner/divider is an important characteristic. There are three effects that limit the amount of power that these devices can withstand.

Insertion loss. The total power dissipated in the power divider under matched conditions with balanced outputs is limited by the losses in ferrite cores and wire windings in rf devices. In microwave devices, the heating of copper traces and board material causes increased resistance and more heat is generated. Exceeding the peak power rating can cause failures due to voltage breakdown of the dielectric.

Amplitude unbalance. Input power dissipation in a power divider under matched conditions takes place in the internal loads because of amplitude unbalance. Ideally, no power would be dissipated across this load, but because of imperfect amplitude balance a small voltage differential would exist.

Mismatched loads. Reflections from mismatched loads at the outputs of power dividers can cause a considerably larger voltage to appear across the internal load.

Equal amplitude signals are applied to the port when the device is used as a combiner, and little or no power is dissipated in the internal load. Possibly, a condition may occur where one or more of the signal sources fail, causing (in the case of a 2-way combiner, 50 percent of the power supplied at the remaining port to be dissipated in the internal resistor. Thus, the power injected at each port should not exceed twice the rating of the internal load in order to avoid this condition.

6.7.6 Environmental considerations. Environmental conditions may affect the equal division of power with high isolation. Those parameters are dependent on electric and magnetic fields within the power divider/combiner. Any environment that will affect those fields is potentially degrading.

Environmental conditions may cause electrical or mechanical effects that are temporary and reversible when the environment approaches again the initial conditions; or the effect may be permanent. Examples of the temporary effect might be the temperature coefficient of a capacitor, inductor, or resistor, whereas more permanent effects result from the dismounting or displacement of internal parts resulting from thermal shock, vibration, and physical shock.

The effect of moisture and humidity is somewhat open to question. The common power divider/combiner is designed for a fairly low impedance level, and a modest amount of condensation might not be operationally visible. Certainly there should be little possibility for moisture entrapment; for even if a modest collection is not electrically visible, it may promote corrosion, mold, or fungus which, in turn, will disable the device.

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POWER DIVIDERS AND COMBINERS**

Most dividers/combiners seem to be used at power levels of one watt or less. Under such circumstances altitude has little effect, except to cause stress on seals and sealed cases, and to cause unsealed enclosures to breathe. At higher power levels there may be a tendency for a voltage discharge of a Paschen or multipactor type, unless the device is sealed and/or pressurized. There is also a possibility that dummy resistor ratings, always less in lower-pressure air, may be exceeded.

One additional comment about shock is desirable: The electrical characteristics and environmental integrity of a device should always be suspect after a shock from dropping. A drop of a few inches onto a workbench may cause shock of a few thousand to several thousand G acceleration. A power divider/combiner should be handled with care.

In view of reliability, the divider/combiner should be considered to be an electronic assembly. The waveguide types are an exception, and should be treated as a moderately critical mechanical assembly.

Each electrical joint in the circuit and the shield should be inspected before it is hidden by subsequent assembly steps. This does not mean an inspection after each application of the soldering iron, but, rather, while the joint is still visible.

Prime failure modes noted to date have been the breaking loose or displacement of parts and wiring. This has most commonly been the result of improper soldering, improper wire treatment (anchoring or nicking), or improper choice or use of adhesives for anchoring. Design and assembly care and inspection are the only cures, but a regimen of thermal shock, vibration and mechanical shock tests also make a good screen. It is good to make "delta" electrical measurements (before and after, then calculate differences), but the limit to the delta is the repeatability possible in the measurement. For any requirement for small delta, there should be exact repetition of test equipment and personnel.

There is some advantage in the screening of parts that are to be used in a device--particularly discrete resistors and capacitors. At frequencies where inductors and transformers are more than a very few turns, these parts may also benefit. Please refer to the appropriate section of this manual for more details.

6.7.7 Reliability considerations. Specification and examination of seal integrity is not easy, and the result of poor seals is open to some discussion. A poor seal is usually a sign of poor design, poor workmanship, or abuse. Seal test, to the degree specified by the procurement documentation, is recommended on a 100-percent basis until it is determined that risk is sufficiently low to justify revision of the requirement.

6.8 MICROWAVE DEVICES, ELECTRICAL MIXERS

6.8 Electrical mixers.

6.8.1 Introduction. "Mixing" is a term commonly used in the subject of radio frequency to describe the practice of multiplying two electrical voltage or current waves or electromagnetic fields, to produce an output that is proportional to at least a part of the resulting product. This discussion is limited to devices that perform such a function.

Multiplication involving radio or higher frequencies almost invariably makes use of analogs which are functions of electrical circuits that provide continuous product output for inputs of two or more electrical functions. Probably the most common analog so used is the approximation to the logarithm. Two or more electrical functions are changed to their logarithms and added to derive the antilogarithm. This, if accurately done, gives the product of the original electrical functions. Some elements, such as particular classes of semiconductor diodes, have closely logarithmic current/voltage relationships over a ratio of perhaps 1,000,000:1, Figure 83. Often, the approximation shown will give sufficiently good results if the circuit includes wave filtering to remove products that are unwanted but produced by the actual relationship.

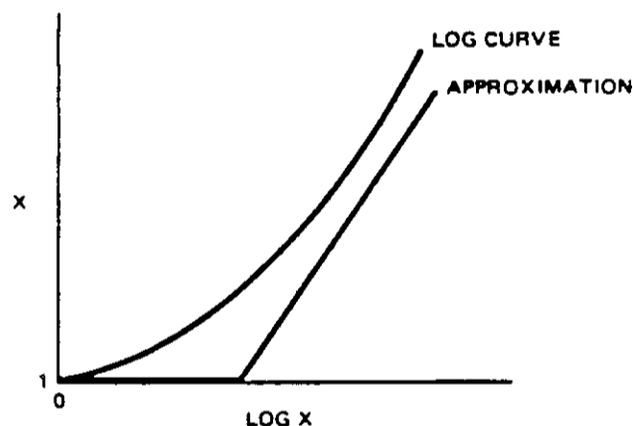


FIGURE 83. Current voltage curve for a mixer diode.

Often, unwanted products (including harmonics) can be minimized by connecting two such analog devices in a push-pull arrangement, commonly reducing even harmonics and other undesired products by 20 dB or more (balanced mixers).

The analog device is by no means limited to two-terminal components. Magnetic amplifiers, dual-gate FETs (transistors), and various connections and biasings of tubes and transistors will produce suitable characteristics. For instance, the Class C amplifier makes a good mixer. Mixers are available in monolithic and hybrid microcircuit forms; however, this discussion is limited to diode and transistor mixers.

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Though there are many applications for mixers, some of the characteristics often sought are that the generators of the functions not affect each other by connection to the mixer, that the mixer present a known and constant load to those generators, that the mixer produce a useful amount of output, and that the output be the product of the inputs with a minimum of other outputs produced. These are called isolation, VSWR, conversion loss, and harmonics.

Approaches other than the logarithmic or logarithmic approximations (including the switching approximation) are possible. One that has received little recent emphasis is the negative-resistance mixer. The first examples that come to mind are the regenerative detector and the tunnel diode mixer; Figure 84 shows a way in which the multiplication function of a mixer can be theoretically obtained. As one of the goals of using any device in equipment is to obtain stable results, and since negative resistance in a linear circuit is not easy to maintain, it is doubtful that much use will be made of this multiplying procedure.

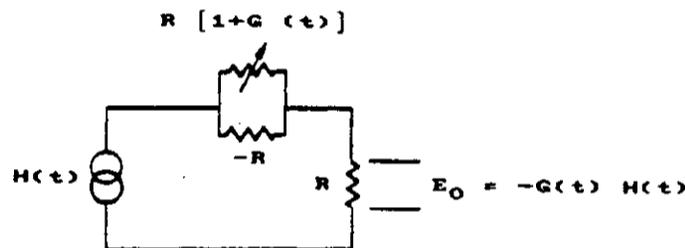


FIGURE 84. Tunnel diode mixer.

Time-varying electrical functions occupy bandwidth in the frequency domain. Though there is no bandwidth limitation in the concept of multiplication of electrical functions, there are limitations in the bandwidth of practical parts and circuits. Frequencies higher than the highest frequency present in the input to the mixer are present in its output, and frequencies approaching zero, or direct current are not uncommon.

Noise may be a problem at the desired frequencies and also at the harmonics. Electrical noise is usually described in terms of "noise figure," which is the ratio of actual extraneous noise power compared with the thermal agitation noise generated in an ideal resistor having the same impedance as the mixer input. Semiconductors and some other parts display "flicker" noise, characterized by an increase in per-Hertz bandwidth power inversely proportional to frequency. Another term for flicker noise is "1/f" noise. Above frequencies where 1/f noise is significant, the noise figure will rise slowly with the frequency, if the design is appropriately scaled. Finally, a frequency will be reached where parasitic inductances, capacitances, and resistances affect the circuit design, and the noise figure will begin a steeper climb. The noise figure is the limiting measure to the weakest electrical function input that will give useful output.

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Commonly, there is an advantage to setting the power level of one of the inputs much higher than that of the other inputs. This practice establishes the operating level of the mixer, hopefully, in some region that gives the desired outputs with a minimum of unwanted outputs. The ratio is usually 10 dB or more.

There is a tendency for some of the input energy to feed through to the output port; in many circuits the input currents will flow through the output circuit. In these cases, the user may have to terminate the circuit so that the input currents will develop no (or very little) power in the output load. These and other considerations are discussed later under Electrical characteristics, paragraph 6.3.5.

Establishing of the operating level of the mixer by the higher-power input will do much to reduce output distortions (intermodulation, compression, and "intercept point" - to be discussed later) and establish the other input and output impedances of the mixer.

6.8.2 Usual applications. The mixer function of multiplication has application from virtually zero frequency through millimeter waves. Some nonelectrical mixers are even used into the light-frequency regions in laser work.

Considering the simplest multiplication function where $C = k f(A) \times f(B)$, consider the case where $f(A)$ can have any value from zero to one. When $f(A) = 1$, $C = k f(B)$, so $f(B)$ may be said to have been passed without change, except for the amplitude factor k . As $f(A)$ approaches zero, the value of C will also approach zero, the output C will be zero, and the mixer will have acted as an electrically variable attenuator capable of giving infinite attenuation. (Mixers giving this characteristic are called "balanced," in that at least one input may be used to balance itself out in the event of a zero value for the other input.)

If, in the above example, $f(A)$ were allowed to progress from zero to -1, the output would reverse the polarity of $f(B)$ and increase to the same absolute magnitude as for $f(A) = 1$. Schematically, this could appear as in Figure 85. This is the foundation for three widely used applications.

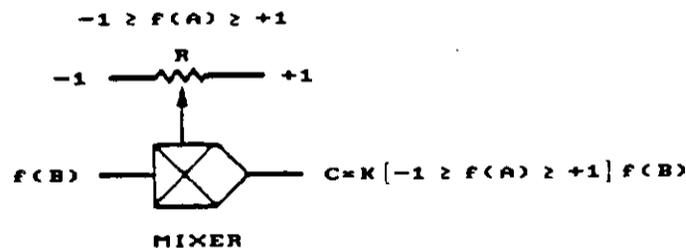


FIGURE 85. Simple mixer.

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On-off switching of $f(B)$ can be accomplished by switching $f(A)$ between +1 and zero.

Phase-reversal of $f(B)$ can be done by switching $f(A)$ between +1 and -1.

Amplitude modulation can be accomplished by considering the use of electrical sine waves for $f(A)$ and $f(B)$. Let $f(A) = \cos A$, $f(B) = \cos B$, then, consider the original multiplication function,

$C = \frac{k}{2} \cos (A+B) + \frac{k}{2} \cos (A-B)$, which is the expression for double-sideband amplitude modulation without carrier wave (Figure 86).

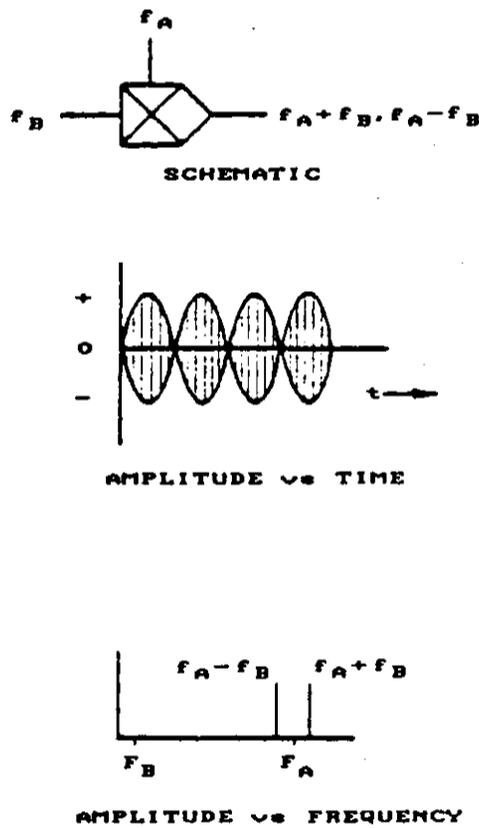


FIGURE 86. Double-sideband amplitude modulation.

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Amplitude modulation with carrier (conventional broadcast AM) can be obtained by letting

$$f(A) = 1 + \cos A, \text{ resulting in } C = k \cos B + \frac{k}{2} \cos (A+B) + \frac{k}{2} \cos (A-B)$$

The (A+B) and (A-B) terms are the "sum-and-difference" or "upper and lower" frequency sidebands, where the angles A and B are

$$A = 2\pi f_A t \text{ and } B = 2\pi f_B t, \text{ the frequency } f \text{ is in Hertz and time } t \text{ in seconds}$$

Note that in "amplitude modulation with carrier" the total bandwidth involved is from zero frequency (or dc, the "1" term) to $f_A + f_B$. A properly chosen and applied mixer must consider each of these frequencies (Figure 87).

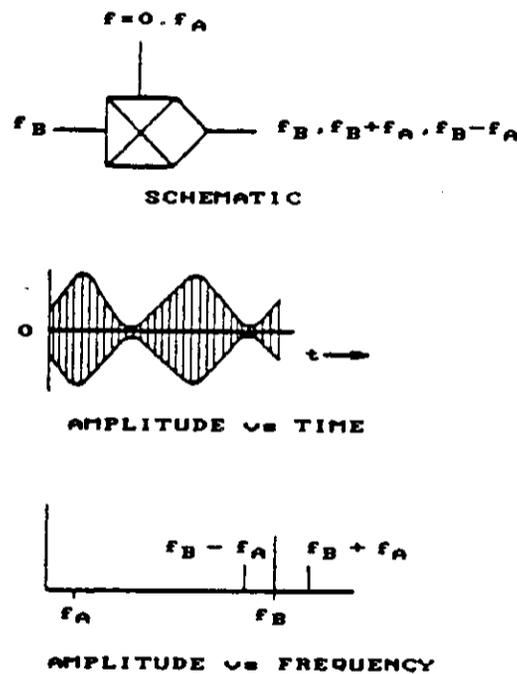


FIGURE 87. Modulator with carrier.

Phase detector use of mixers is possible by having one of the frequencies in $f(A)$ the same as one frequency in $f(B)$. A direct current component of the output results which is proportional to the cosine of the phase angle between the two inputs and the product of the magnitudes of the single common frequency (Figure 88).

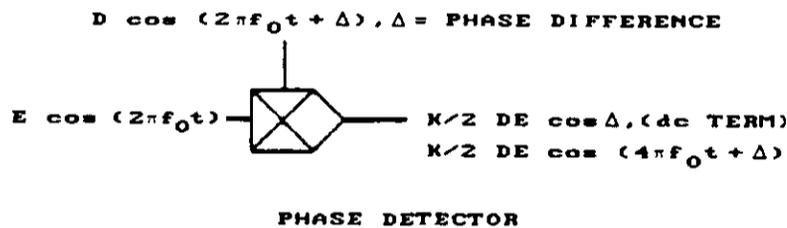
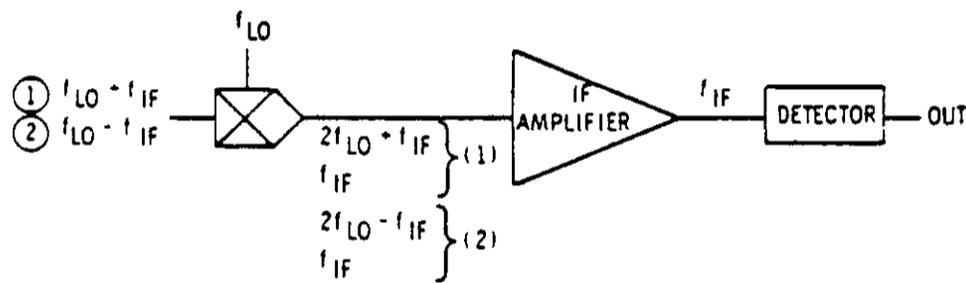
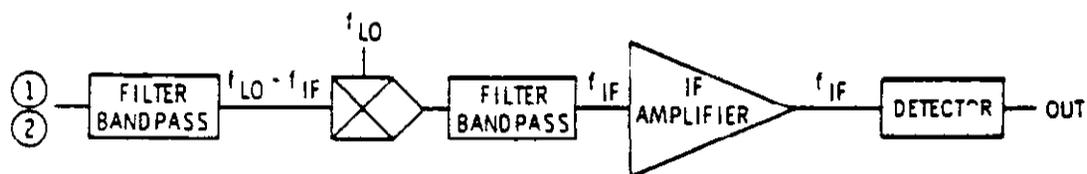


FIGURE 88. Phase detector.

Phase detector output from the mixer that is not a function of phase alone (except the zero output, or odd multiple of 90 phase difference) and is not proportional to phase angle is awkward. Filters and amplitude limiters on each input help select the frequency of interest. Clippers to make approximate square waves of the input waves will, with a wide-band mixer, give output proportional to phase angle with fairly good accuracy. The mixers must be wide-band, as making square waves of the basic frequency will require distortionless processing up to 10 or more times the basic frequency in the mixer.



A. Simple superheterodyne



B. Superheterodyne with filter selectivity

FIGURE 89. Superheterodyne with filter selectivity.

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Frequency mixing for use in superheterodyne receivers is probably the most common use of mixers (Figure 89). Here the received signal is mixed with a locally generated source of radio frequency power of much greater level, and the difference frequency (known as the intermediate frequency (IF)) is selected as the output. If the local oscillator is made variable, the receiver will be sensitive to various frequencies (for a constant intermediate frequency) by tuning the local oscillator. The IF may be amplified and narrowed by filtering to any desired degree, giving a receiving system of constantly high gain. Unfortunately, for each local oscillator (LO) frequency, there are two received frequencies giving the same IF. The unwanted frequency is called the "image frequency." Image responses may be reduced either by filtering the antenna line or by single sideband cancellation techniques.

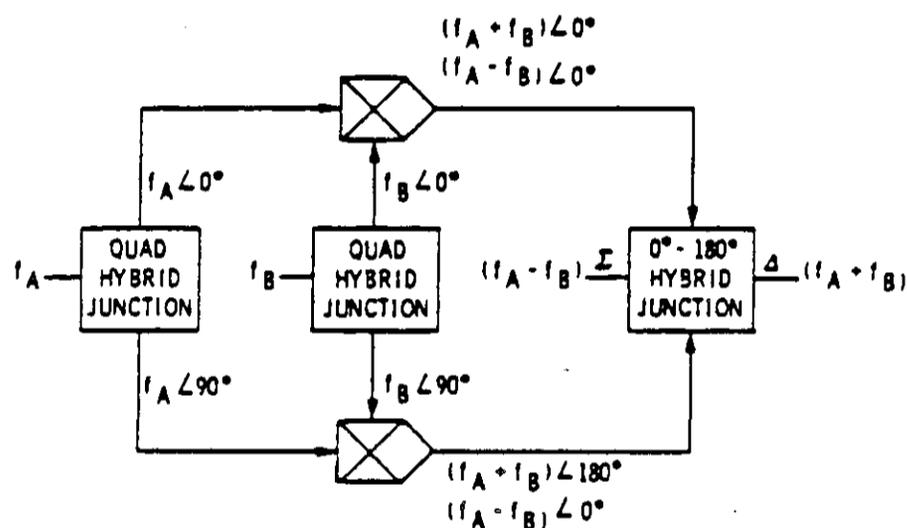


FIGURE 90. Single sideband generator.

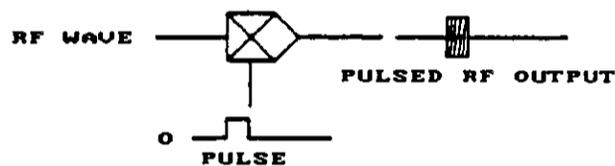
Single sideband signals may be generated by the phasing system as shown in Figure 90. Here, if the signal amplitudes are kept equal in the corresponding parts of the two paths, the lower sideband will emerge from the summation port, and the upper sideband will emerge from the difference port. Depending on the balance and phase accuracy of the hybrids, the suppression of the unwanted sideband can be 20 dB or more. Such a system may exceed 40 dB suppression of the unwanted sideband if operated on fixed frequencies, but this degree of suppression becomes more difficult as bandwidth and operating frequency increase.

The image frequency of superheterodyne reception may likewise be suppressed.

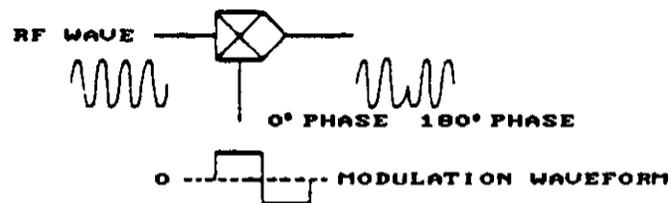
Coherent detection of signals having symmetrical sidebands or a carrier is likewise attainable by feedback control of the frequency and phase of f_B (a voltage-controlled oscillator) by locking f_B to either the optimum phase for

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detection of the sidebands or to the carrier. Noncoherent interference above or below f_b may be suppressed without loss of desired sideband content. (In voice communications circuits, the two outputs have been presented binaurally to allow additional mental discrimination by the apparent directional difference between the desired signal and interference.)



A. Pulse modulation or gating



B. NRZ (non return-to-zero) modulation

FIGURE 91. Modulators.

Pulse modulation or gating (Figure 91) may be easily accomplished by feeding the signal to be modulated or gated into one port and feeding the gating signal into the other. As the gating signal ratio of "on" to "off" time may range from extremely large to very small, the port to which the gating pulse is applied should be extremely wideband. Earlier when the switching possibilities of the mixers were discussed, this factor was not mentioned.

Nonreturn-to-zero (NRZ) modulation is often more useful pulse modulation than the on-off form of amplitude modulation. In Figure 91b, this principle is shown in a phase-reversal scheme frequently used in data communications. The modulation waveform bandwidth again must be considered. Some of the value of the approach may be seen in that the NRZ system virtually eliminates problems that might be caused by a moderate leakage of the rf wave through the mixer. The full-value output cancels or masks all but massive leakage.

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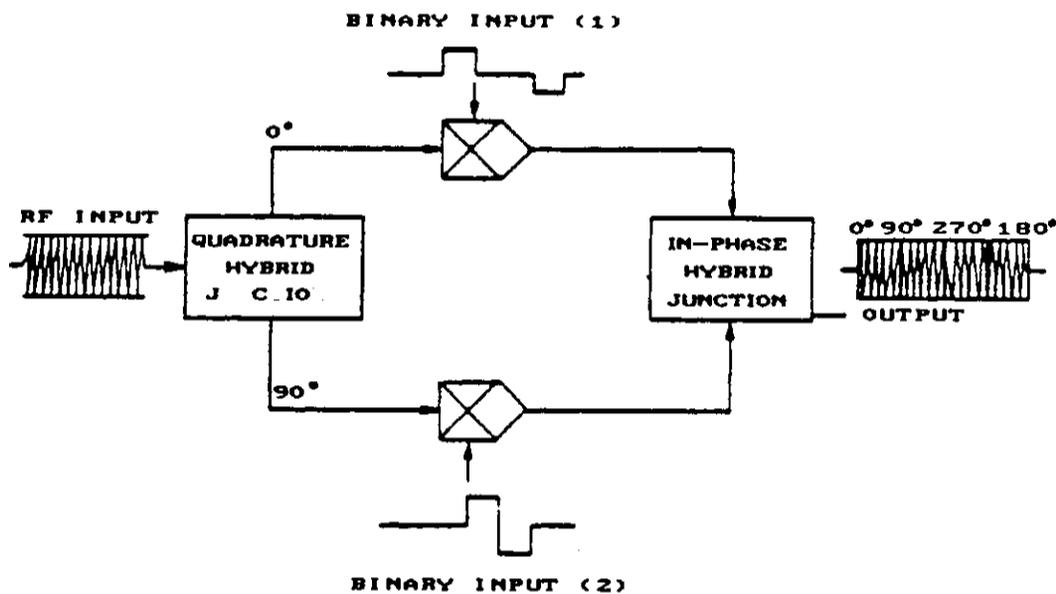


FIGURE 92. Quadrature four-phase modulation.

One attractive form of modulation is quadrature or four-phase modulation (Figure 92), where any quadrant can be electronically selected. The use of hybrid junctions not only supplies the 90 degree phase increments and easy addition, but also helps isolate the mixers from each other electrically.

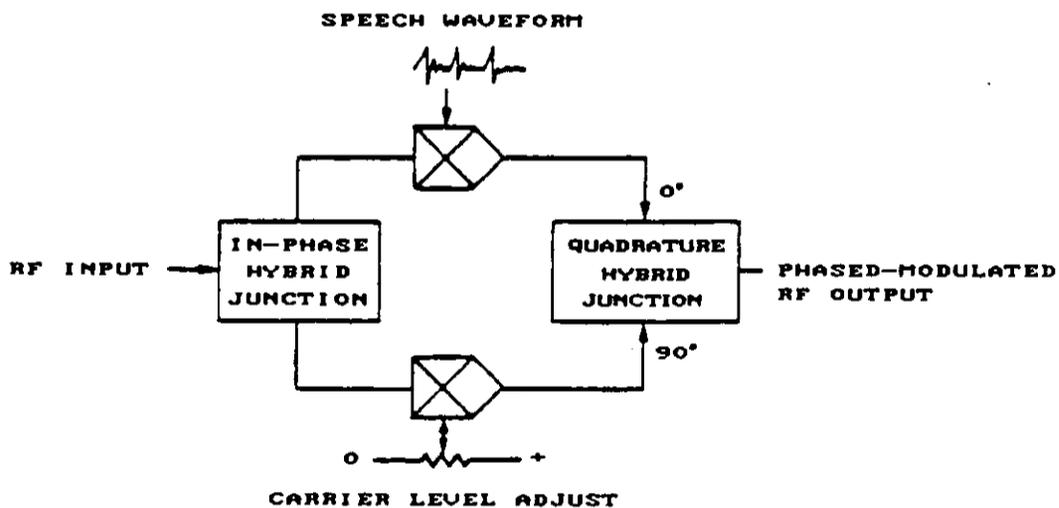


FIGURE 93. Phase modulation (voice communications).

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Voice communications phase modulation illustrates (Figure 93) how the combination of mixers and hybrids can provide simultaneous but different processing in additive paths. The upper path provides voice-frequency sidebands, whereas the lower path gives electrically variable attenuation and shifts the carrier wave 90 degrees to form a conventional phase-modulated communications signal.

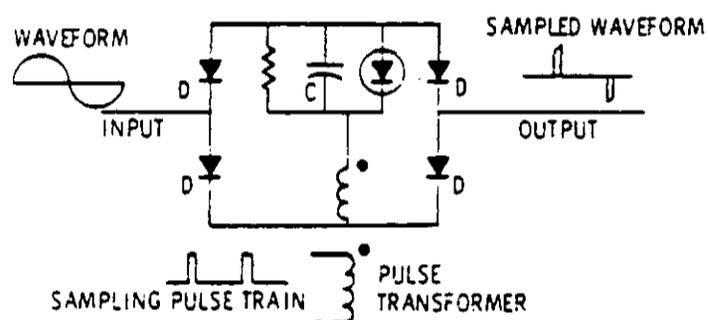


FIGURE 94. Sampling bridge.

A sampling-bridge (Figure 94) is often used to monitor very rapid periodic waveforms. This circuit (sometimes called a "boxcar" circuit) acts as a very high-speed switch, connecting input and output during a positive-going sampling pulse, and disconnecting during the remainder of the train until the next sampling pulse arrives. Followed by an integrator, the circuit is useful in automatic frequency control application. Similar circuits are often used in "sampling oscilloscopes." A more conventional mixer may also be used for sampling, but the configuration shown has advantages when the diodes "D" are well chosen for type and balance and the pulse transformer provides good isolation.

6.8.3 Physical construction. There is no single mixer configuration; this discussion is limited to diode and transistor mixers. Although the simplest mixer is no more than a single diode, this device normally contains additional elements to achieve the operations previously discussed.

At lower frequencies, the nonlinear element in which the product is generated is usually assisted with resistors, capacitors, inductors, and transformers to provide efficient coupling and filtering of the various inputs and outputs, to control the amount of power and biasing that the nonlinear element needs, and to confine the action of the mixer to the desired volume of space.

Such additional elements may not be so obvious at microwave frequencies, for pegs or ridges (by being significant parts of a wavelength) may act as capacitors, inductors, and transformers. The closed nature of the coaxial transmission line and waveguide makes the (equivalent circuit) complexity of microwave mixers even less obvious.

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The outward physical appearance of mixers will vary with frequency and power level. An RF mixer up to 2 GHz may be packaged in various microstrip or stripline flatpacks, TO cans or "bathtub" DIPs, with round pins. These parts may be further packaged in a housing with rf connectors, or connectors may be bolted right onto the flatpack. Microwave mixers are usually a stripline or microstrip assembly in a package with rf connectors, and of course they may be implemented in waveguide. To achieve high isolation, careful circuit design or packages with rf connectors should be used.

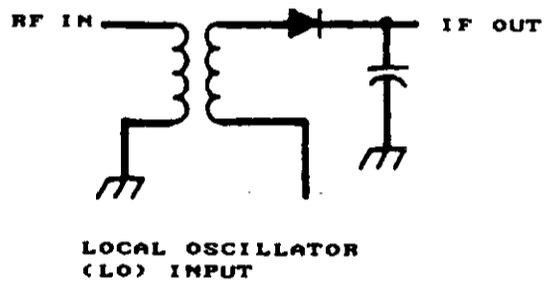
The package prevents the mixer from radiating unwanted signals and protects the mixer circuit from electromagnetic interference. Open components such as chip capacitors, thin film resistors, toroids and semiconductors are protected by the seal of the package. Flatpacks and TO cans are usually seam welded or resistance welded. DIPs are either welded or solder sealed. Microwave mixers tend to be more sensitive to overload, so mixer diodes are often mounted so that when burned out by excess power, they may be replaced without disassembling the mixer. Some diodes must have bias power from a dc connector or pin to place the operation in the best part of the diode characteristics.

6.8.4 Military designations. The military general specification for mixers is MIL-M-28837.

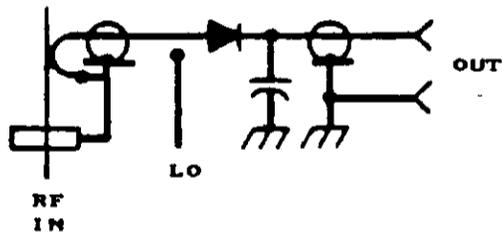
6.8.5 Electrical characteristics. The simplest form of two-input mixers uses a single diode. This is typical of use in superheterodyne receivers where varying input frequencies are translated into a fixed band of intermediate frequencies. Thus, in Figure 95A the two inputs are connected in series and the local oscillator (as previously mentioned) usually has a power level at least 10 times greater than the RF IN. A definite disadvantage exists; local oscillator current will flow back through the RF INPUT circuit. If filtering or other isolation against LO passage back to the RF source is not provided, LO energy may be radiated from the equipment antenna or other circuitry. This is presently illegal (in the general sense) in commercial and consumer equipment, and dangerously revealing in military equipment.

A practical form of shunt inputs is shown in Figure 95B. Here, the RF IN is coupled by a magnetic probe from a waveguide to the diode with a coaxial cable, whereas the LO is capacitively coupled to the coaxial line. The LO level signal determines the operating point of the diode mixer, and the difference in frequency is presented to the output. (The output circuit must provide a dc return path to ground.)

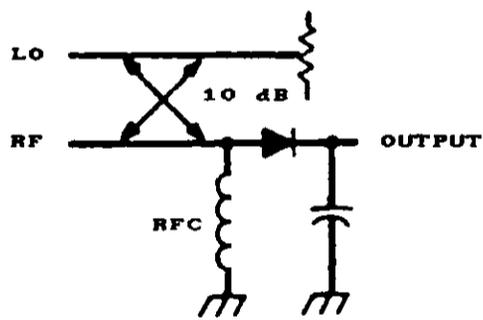
The series and shunt methods of coupling, as previously discussed, suffer from the possibility that the LO will feed back into the RF IN. Use of a directional coupler as in Figure 95C can reduce that effect. If the mixer assembly is perfectly matched to the directional coupler at the frequency of LO, all of the LO power will be absorbed in the mixer and none will reflect to the RF port of the directional coupler. (Most published examples show a 3 dB coupler, but if excess LO power is available, the 3 dB resulting loss in noise figure may be greatly reduced by using a 10 dB coupler.)



A. Series input series diode



B. Shunt inputs series diodes microwave



C. Shunt inputs series diodes directional coupler isolation

FIGURE 95. Single diode mixers.

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Several simple transistor mixers are shown in Figure 96. Bipolar transistors are shown in A and C, whereas field effect transistors are used in B and D. Shunt feed of RF INPUT and LO INPUT is used in A and B, with consequent possible interaction of the tuned RF INPUT circuit pulling the LO frequency. A degree of isolation is provided by the LO injection methods of C and D, but the results may not be satisfactory in critical applications.

Even less isolation has been used in some inexpensive transistor radios where the mixer transistor was also acting as its own local oscillator. In addition to the pulling effect, there is always the possibility of those mixers, and the circuits shown in Figure 96, permitting an undesirable amount of radiation at the LO frequency.

Transistor mixers are adjusted to the most efficient mixing point by a combination of dc bias (sometimes self-generated) and LO drive level. (As with other types of mixers, the LO INPUT level should be one or more orders of magnitude greater than the largest expected amount of RF INPUT.)

The field-effect transistor (FET) has become very attractive because of low-noise capabilities; however, in mixer service, the FET is not particularly distinguished by a low noise figure.

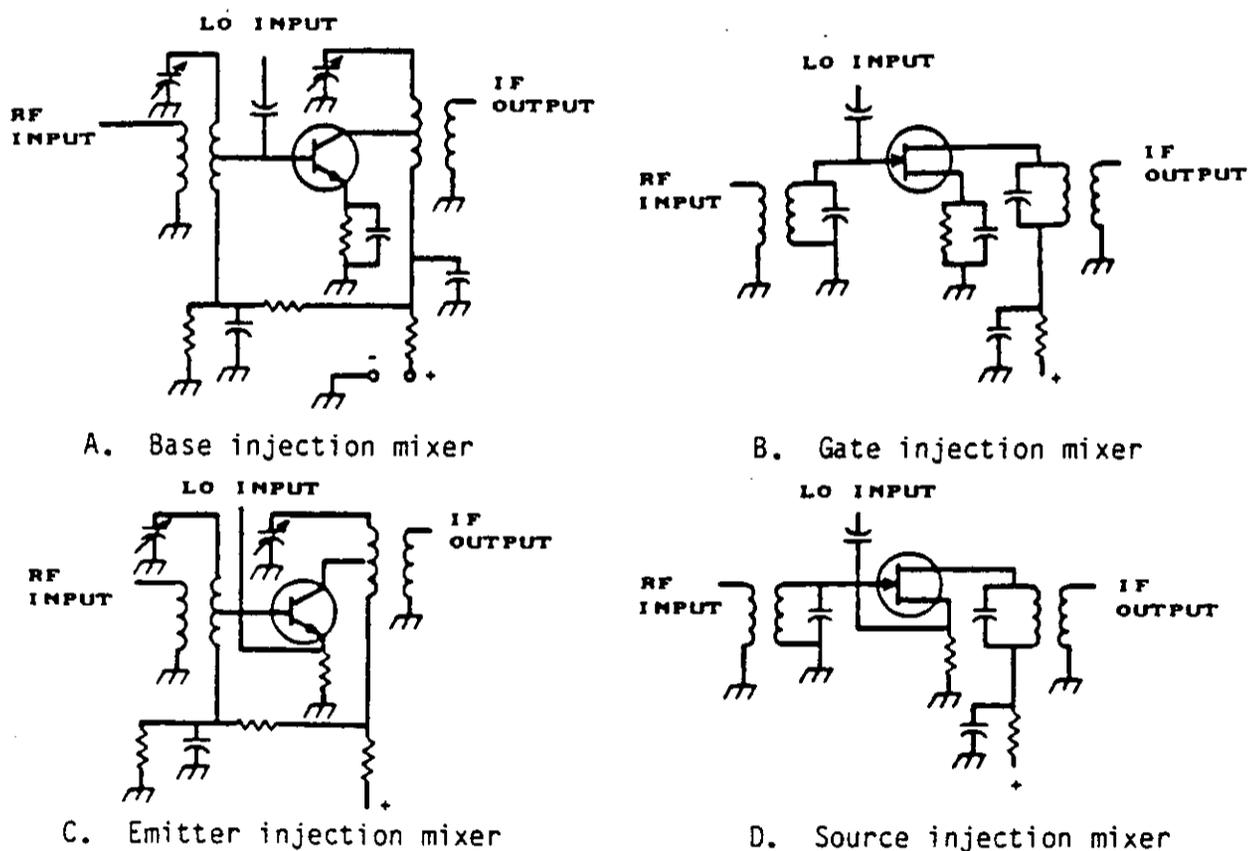


FIGURE 96. Transistor mixers.

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The noise figure is the ratio of noise contributed by a device or system to the noise produced by thermal agitation in the resistive parts of the characteristic impedance of that device or system. The noise figure is usually referenced to the input circuit to make easier the understanding of the unwanted competition that the incoming signal will face.

A more serious cause of noise in mixers intended for use at microwave frequencies is noise generated by the local oscillator. Balanced modulators or mixers can be used to reduce the effect of local oscillator noise by 15 dB or more by use of two mixers in which near-equal but opposing noise components are created by interaction of the local oscillator and incoming signal. The mixer outputs, when added, perform the cancellation.

Balanced mixers have an additional advantage. Either of the inputs may be electrically isolated from the output or the other input. The various advantages are not available in all combinations in all types of balanced mixers, but the need for one or more is the usual reason for choosing the balanced configuration.

A single-balanced mixer is shown in Figure 97A. The RF IN does not appear in the output, although the LO IN does. The LO IN must provide a dc path so the diodes may conduct, and it is well if the LO IN presents a very low impedance at the desired output frequencies. Similarly, for the maximum LO power to be presented to the mixer diodes, the output circuit should present a very low impedance to the LO frequencies.

A double-balanced mixer (ring modulator) is shown in Figure 97B. Here each of the three ports is isolated from the others. A number of the unwanted outputs are cancelled in this arrangement if each port is properly terminated. For test purposes it is best to terminate each port in its characteristic impedance at all frequencies, as otherwise there may be unpredictable reflections from the termination back to the diodes. With proper terminations and high values of LO IN, low values of two-tone intermodulation (-60 to -100 dB or more) and standing-wave-ratio of the RF IN and IF OUT ports are possible. The SWR of the LO IN port will probably be much higher (perhaps 3:1) if the mixer is used over a wide frequency range.

Note that there is no inherent frequency dependence in either the single- or double-balanced diode mixers so far discussed, except that transformer-coupled ports will not pass zero frequency.

Balanced mixers do not have to be inherently wideband. The single-balanced mixer often benefits from restricted-bandwidth ports. The 10-dB conversion loss listed by one author for the wideband mixer of Figure 97A drops to the order of 2-3 dB with frequency-selective port terminations.

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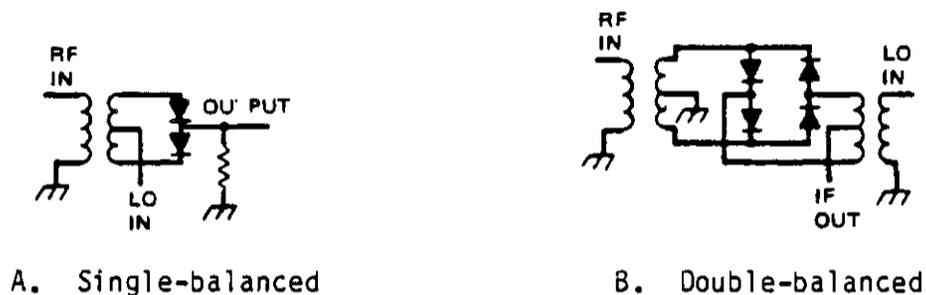


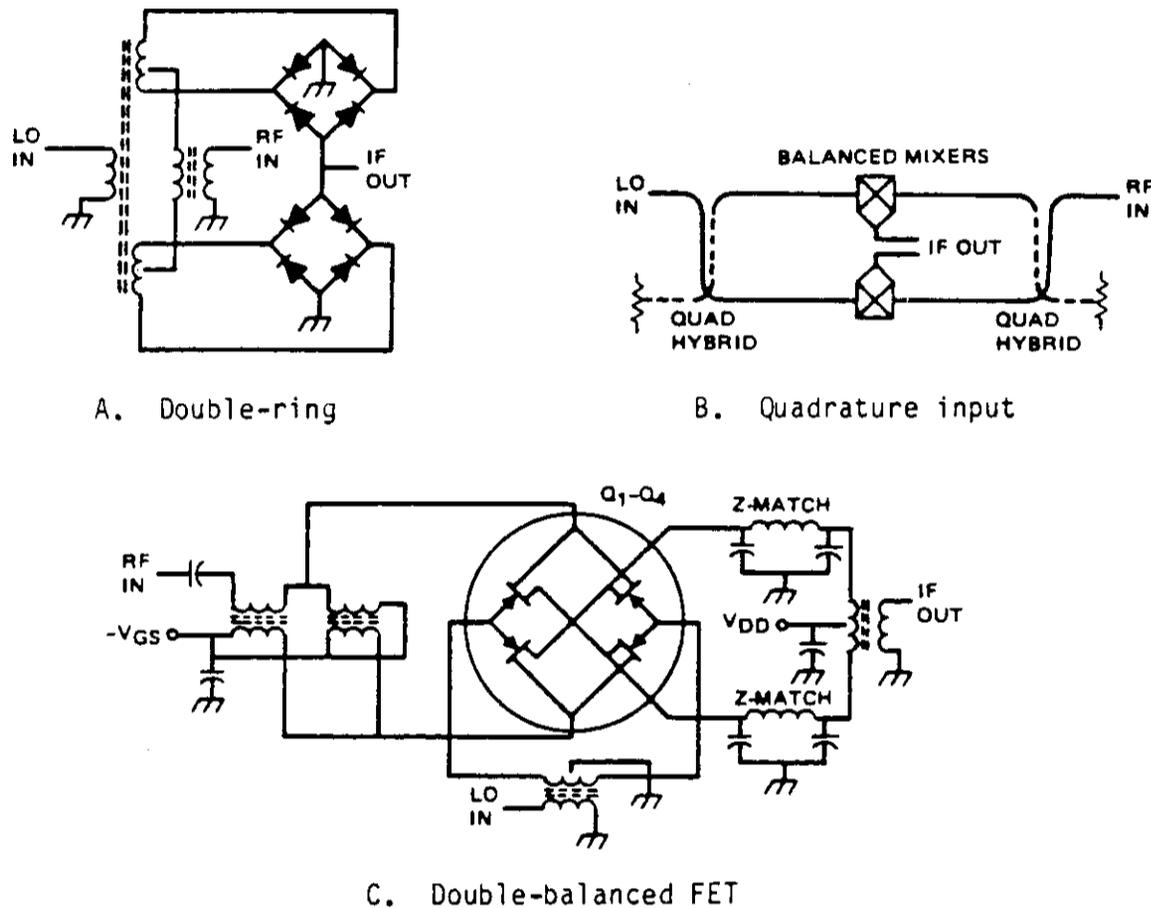
FIGURE 97. Balanced mixers.

Somewhat more complex double-balanced mixers are shown in Figure 98. A double-ring mixer is shown in Figure 98A, capable of operation from 10 MHz through 3 GHz. Though the IF OUT is dc-coupled, the response of this port falls off below 10 MHz because of dc shunting effects. This bridge is characterized by fairly high compression and desensitization level and third-order intercept point and reasonably low third-order intermodulation. These various attributes are desirable in the direction indicated, and in most types of mixers are--within limits--dependent on a high level of local oscillator power. The mixer shown has good impedance match over the frequency range on the RF IN and IF OUT ports, but will increase to perhaps 3:1 VSWR on the LO IN port.

The quadrature input, double-balanced mixer shown in Figure 98B is specifically designed to present a low VSWR over a full octave (2:1) bandwidth. The LO IN and RF IN ports will have a 1.3:1, or better VSWR because of the action of the quadrature hybrids, which diverts reflected power from the mixers to dummy loads R1 and R2, respectively. The quadrature hybrids also improve the LO IN to RF IN isolation by perhaps 20 dB above that provided by the individual balanced mixers. The IF OUT ports may be connected to provide an unbalanced output, or the diodes in one mixer reversed to provide two-conductor balanced transmission line output.

The double-balanced FET mixer of Figure 98C is relatively narrow band because of the impedance-matching (Z-match) networks transforming the low-impedance output power-combiner to a higher impedance for the drains of the FETs. The combination of four FETs may be procured either as a single sealed or a single unsealed assembly.

The double-balanced FET mixers are not inherently narrow band. One vendor advertises such a mixer with bandwidth limits of 200 KHz and 100 MHz. This unit, with 2 W of local oscillator power, is claimed to have a dynamic range of 155 dB, with -30 dBm third- and fifth-order intermodulation levels 140 dB below the desired products.

6.8 MICROWAVE DEVICES,
ELECTRICAL MIXERSFIGURE 98. Double-balanced mixers.

The simple mixer may be adequate for many applications. Relatively little power is required if frequency-selective networks are used at the input and output ports. Even if such networks are used, there is a strong tendency for the inputs to affect other inputs, either by energy feeding out the other ports or by "pulling" (shifting oscillator frequency) the circuits connected to other ports. Most distortion-free operation is usually obtained with one input (usually "local oscillator") having 10 or more times the power of the other input. Intermodulation and crosstalk tend to be high.

The single-balanced mixer requires perhaps twice the "local oscillator" power of the simple mixer, but offers the advantage that most local oscillator noise may be suppressed by utilizing a 90- or 180-degree hybrid divider at the input.

6.8 MICROWAVE DEVICES, ELECTRICAL MIXERS

The double-balanced mixer may require up to four times the local oscillator power of the simple mixer but offers the advantages of good isolation between all ports, much cancellation of noise from the local oscillator, and great reduction of intermodulation distortion. The port impedances of double-balanced mixers may be established within fairly close limits. The quadrature input double balanced mixer offers the above advantages regardless of the LO power and termination impedances.

None of the four mixer types is optimum for all parameters. The relative advantages and disadvantages of each type for a leading microwave mixer manufacturer is summarized in Table XI.

In each type of mixer, not only intermodulation but also compression and desensitization are to be expected. Each of these effects is dependent on the level of the local oscillator power available, the level or levels of one or more other inputs, and the terminations of each port. Compression deals with the effects of just one other input. Desensitization deals with one other desired input and one other undesired input. Intermodulation deals with two other desired inputs. ("Desired" here means that the inputs are in the passband of desired signals.)

Compression

Compression is usually defined as the minimum input level at which the consequent output will become 1 dB less than straight-line prediction of a proportional output-power/input-power ratio.

Desensitization level

Desensitization level is the minimum level at which the output level, corresponding to a weak desired input, is reduced by 1 dB, compared with when the undesired signal is absent.

Intermodulation

The two-tone method is the usual form of intermodulation test, wherein two equal-power desired frequencies are the other input, and particularly the third and fifth order intermodulation products are examined. Other products may on occasion be important, but the third and fifth order are most often annoying. The third order products have the frequencies $2f_1 \pm f_2$ and $2f_2 \pm f_1$. The fifth order products may include harmonics, the sum of whose orders add up to five. In some mixers it is possible to detect intermodulation products above the 250th.

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TABLE XI. Performance comparison for four basic mixer types*

Mixer Type <u>1/</u>	No. of Diodes <u>2/</u>	VSWR <u>3/</u>	Conversion Loss <u>4/</u>	LO/RF Isolation <u>5/</u>	Bias	Spurious Rejection $ m-n = 1$ <u>6/</u>	Harmonic Suppression <u>7/</u>	Intercept Point (dBm) <u>8/</u>	Dynamic Range	IF Bandwidth	Image Focusing Port <u>11/</u>
90° hybrid single balanced	2	Good	Lowest	Poor	No	Poor	Poor	+15	High	Wide	LO
	2	Good	Lowest	Poor	Yes	Fair	Fair	+30 <u>9/</u>	Highest	Wide	LO
180° hybrid single balanced	2	Poor	Low	Good	Yes	Fair	Good	+15	High	Wide	RF
	2	Poor	Low	Good	Yes	Fair	Good	+30 <u>9/</u>	Highest	Wide	RF
Quad input double balanced	4	Good	Low	Very good	No	Good	Fair	+18	High	Wide	Internal load
Double balanced	4	Poor	Low	Very good	No	Good	Very Good	+18	High	Extremely wide <u>10/</u>	RF
	4	Poor	Low	Excellent	No	Good	Very Good	+18	High	Extremely wide	RF

See following page for notes.

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ELECTRICAL MIXERS

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**6.8 MICROWAVE DEVICES,
ELECTRICAL MIXERS**

TABLE XI. (Continued)

NOTES:

- 1/ Data is from typical L-band or S-band units.
- 2/ 2-diode types typically require +7 dBm LO power for best performance;
4-diode types require +10 dBm LO power.
- 3/ VSWR: Poor: 2.5:1 typical
Good: 1.3:1 typical
- 4/ Conversion loss: Lowest: 5-7 dB typical
Low 7-9 dB typical
- 5/ LO/RF isolation: Poor: 10 dB typical
Good: 20 dB typical
Very Good: 25-30 dB typical
Best: 35-40 dB typical
- 6/ Spurious rejection ($m \times n$): where $|m - n| = 1$; i.e., 1 x 2, 2 x 1, 2 x 3,
3 x 2, etc.
Poor: partial rejection of most $|m - n| = 1$ spurs
Fair: partial rejection of most $|m - n| = 1$ spurs (bias adjust
will suppress some spurs even further)
Good: potentially rejects all $|m - n| = 1$ spurs.
- 7/ Harmonic suppression: Poor: partial rejection of LO/RF even harmonics
Fair: partial rejection of LO/RF even harmonics
(bias adjust will suppress some harmonics
even further)
Good: can reject all LO even harmonics
Very Good: can reject all LO and RF even harmonics
- 8/ Intercept point: Typical third-order intercept point is 6 to 9 dB above
the LO power.
- 9/ This intercept point can only be achieved by using the optimum load line
biasing technique and increasing the LO power to approximately +23 dBm.
- 10/ The IF OUTPUT bandwidth for this series of balanced mixers overlaps the
RF INPUT range of the units. Bandwidths from dc to 4 GHz are available.
- 11/ Image focusing: defines where image signal energy is directed to the LO
port, RF port, or the internal termination. Where this image signal is
focused can affect mixer performance in phase and amplitude tracking
systems.

**6.8 MICROWAVE DEVICES,
ELECTRICAL MIXERS**VSWR

The VSWR specifies the degree of impedance match between the mixer LO and RF ports and the 50- Ω system in which the mixer is used.

One of the most desirable characteristics of the 90-degree hybrid mixer is its excellent VSWR over the full performance range of the hybrid. When a signal is fed into either input, any reflections from the similar mixer diodes will combine at the other input, providing a low VSWR at either port. The input reflection coefficient for a 90-degree hybrid mixer having diode reflection coefficient, ρ_1 , ρ_2 is given by:

$$\rho(\text{mixer}, 90^\circ) = \frac{\rho_1 - \rho_2}{2}$$

If $\rho_1 = \rho_2$ (if the diodes are balanced), whatever their value, the input remains matched.

If the mixer uses a 180-degree hybrid, any reflections from the similar mixer diodes will be focused back to the input port. Unless the diodes look like a very good match, the input VSWR will be poor. The input reflection coefficient for a 180-degree hybrid mixer having diode reflection coefficients ρ_1 , ρ_2 is given by:

$$\rho(\text{mixer}, 180^\circ) = \frac{\rho_1 - \rho_2}{2}$$

Diode impedance is a function of LO power; therefore, VSWR will be affected by LO drive level for this mixer. If the diode impedances are equal, the input VSWR equals the return loss for one diode.

A quadrature input double-balanced mixer has the same good VSWR characteristics as the 90-degree hybrid mixer, because it uses 90-degree hybrids as the coupling mechanism to the diodes.

A double-balanced mixer exhibits poor VSWR characteristics. It uses 180-degree hybrids as the coupling networks to its diodes and, therefore, depends on a good diode match to provide good input VSWR. In addition, the LO VSWR is frequently degraded even further by the necessity for extracting the IF signal from the LO coupling network. VSWRs of 3:1 or greater are not uncommon for double-balanced mixers.

LO/RF leakage

The LO/RF isolation is a measure of the LO leakage at the RF port of the mixer.

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**6.8 MICROWAVE DEVICES,
ELECTRICAL MIXERS**

The 90-degree hybrid mixer has poor isolation characteristics because LO/RF isolation is dependent upon diode match. Any LO energy not absorbed by the diodes is directed to the RF port. The LO/RF isolation for a 90-degree hybrid mixer having diode reflection coefficients ρ_1 , ρ_2 is given by:

$$\text{Isolation (dB)} = 20 \log_{10} \left(\frac{2}{\rho_1 + \rho_2} \right)$$

If the diode impedances are equal, the isolation equals the return loss for one diode. Because diode impedance is a function of LO power, the isolation will be sensitive to LO drive.

A 180-degree hybrid mixer has good isolation characteristics; dependent only on the diode balance, regardless of their impedance values. The LO/RF isolation for a 180-degree hybrid mixer having diode reflection coefficients ρ_1 and ρ_2 is given by:

$$\text{Isolation (dB)} = 20 \log_{10} \left(\frac{2}{\rho_1 - \rho_2} \right)$$

LO power will not affect isolation if the diodes continue to track each other.

The quadrature input double-balanced mixer will have good LO/RF isolation, regardless of LO power, because the phasing of the LO and RF networks directs any unused LO energy to an internal termination; it will never show up at the RF port as LO/RF leakage. The good LO/RF isolation is independent of LO power level. The quadrature input double-balanced mixer is the only one of the four basic mixers that will provide, simultaneously, good VSWR and good isolation regardless of LO drive level.

The double-balanced mixer has the same good LO/RF isolation characteristics as the 180-degree hybrid-balanced mixer because it uses 180-degree hybrids as the coupling networks to its diodes. As long as the diodes track each other, the LO/RF isolation will be good and will remain independent of LO power.

For many receiving applications, the amount of LO power leaking out the RF port is important because it will ultimately be reradiated out of the antennas. A biased mixer can be very useful in such applications because the LO power and resultant reradiation can be significantly lowered while still maintaining good conversion loss. In fact, using biased 90-degree balanced mixers may actually result in lower radiated energy from the antenna and better overall performance than when using many double-balanced mixers. Double-balanced mixers, at first, would appear to offer better results because of their higher LO/RF isolation specifications, but the following comparison, shown in Table XII, will serve to illustrate why this may not be the case.

6.8 MICROWAVE DEVICES,
ELECTRICAL MIXERSTABLE XII. Application showing a comparison of biased versus unbiased mixer operation

Biased 90-degree Balanced Mixer		Typical Double-Balanced Mixer	
Conversion loss:	8.0 dB max (0 dBm LO) 8.5 dB max (-4 dBm LO)	Conversion loss:	9.0 dB max
LO power:	0 dBm at 1.8 mA bias -4 dBm at 1 mA bias	LO power:	+10 dBm (no bias option)
LO/RF isolation:	10 dB typical (6 dB min)	LO/RF isolation:	17 dB typical
LO leakage at RF port:		LO leakage at RF port:	-7 dBm (+10 dBm LO)
LO, RF VSWR:	1.65 typ	LO, RF VSWR	3.0 typ

The biased 90-degree balanced mixer radiates 7 dB less LO power at the antenna terminal than the double-balanced mixer. This improved leakage performance is even more impressive when the rest of the specifications are examined. When compared with the typical double-balanced mixer, the biased, 90-degree hybrid mixer has:

- a. Lower LO/RF leakage levels
- b. Equal or better conversion loss
- c. 10 to 14 dB less LO drive required
- d. Much lower VSWR.

Conversion loss and noise figure.

Conversion loss and noise figure are two of the most important and closely-related mixer parameters. The primary requirement for a mixer is to provide the maximum IF output power with the minimum RF INPUT power while generating the least amount of noise. Minimum losses in the RF, LO coupling networks, high quality Schottky-barrier (hot carrier) coupling networks to the diodes, broadband diode matching and optimum local oscillator drive, all contribute to the good conversion loss and low noise figure.

Conversion loss of the mixer is a measure of the power at the IF frequency relative to the power at the rf input frequency. It is a function of the mixer alone.

6.8 MICROWAVE DEVICES, ELECTRICAL MIXERS

The mixer noise figure is commonly stated at a particular IF frequency when measured in a system containing an IF amplifier following the mixer. This IF amplifier contributes to the measured noise figure. The mixer diodes also can contribute a small amount of noise energy to the noise figure. The measured noise figure (NF_0) in decibels is related to the mixer conversion loss (L_c), the IF amplifier noise figure (F_{if}), and the mixer diodes noise-temperature ratio (T) by:

$$NF_0 = L_c + 10 \log_{10} (F_{if} + T - 1) \quad (1)$$

Where L_c is mixer conversion loss in dB
 F_{if} is the noise factor (a numeric ratio) of the IF amplifier
 T is the effective diode noise-temperature ratio.

T can be as low as 0.85 for modern high-quality hot-carrier diodes, but a more typical value is 1.0. The relationship between measured noise figure and conversion loss then simplifies to:

$$NF_0 = L_c + NF_{if} \quad (2)$$

Where NF_{if} is IF preamplifier noise figure in dB.

When $T = 1$, the mixers' single-sideband noise figure is the same as its conversion loss. Equation (2) then shows that the overall noise figure is obtained by adding the noise figure of the IF preamplifier directly to the mixer conversion loss.

It is instructive to observe what happens to the overall noise figure when $T = 0.85$. By equation (1):

$$NF_0 = L_c + 10 \log_{10} (F_{if} + 0.85 - 1)$$

When NF_{if} is equal to 1.5 dB:

$$\begin{aligned} NF_0 &= L_c + (1.413 - 0.15) \\ &= L_c + 1.02 \text{ dB} \end{aligned}$$

The combined preamplifier/diode noise contribution is only 1.02 dB, which is 0.48 dB less than that produced by the preamplifier alone.

Biased mixers

In many instances the optimum LO power is not available for best conversion loss. The conversion loss of an unbiased mixer degrades rapidly as the LO power decreases. This is due chiefly to diode mismatch at the rf and IF frequencies. The diode mismatch can be minimized and the conversion loss dramatically improved by biasing the diode with a dc current.

**6.8 MICROWAVE DEVICES,
ELECTRICAL MIXERS**

6.8.6 Environmental considerations. Every adverse environment will unfavorably affect some type of mixer. Generally balanced mixers will keep more desirable characteristics in the face of adverse environments than unbalanced mixers, but even their characteristics may be badly degraded.

Heat or cold causes at least temporary dimensional change, causing corresponding changes in inductance and capacitance. Values of permeability, dielectric constant, resistance, and the characteristics of semiconductors, likewise, change with temperature change. It is particularly important that balanced mixers be built so that the balance is maintained over the operating temperature range.

Repetitive temperature changes tend to stress mechanical structures both to reduce the strength of joints and break seals. In some cases, stresses will be relieved less disastrously (as in coil interturn stresses), and will result in a more stable, though somewhat changed, product.

With the exception of high-power modulated radio amplifiers, most mixers do not involve high enough voltages to be affected by high-altitude breakdowns. Barometric pressure changes, thus, have their major effect in aneroid stressing of sealed cases. Temperature cycling will also cycle pressure, except that it is the pressure within the case that is changed. Where cases have poor seals, the refrigeration effect of gas expansion will tend to condense water vapor within the case.

If not properly finished and sealed, the mixers can be corroded by salt spray and damaged by sand and dust.

Mixers (generally being aggregations of parts) will, even in the best cases, be adversely affected by the forces of shock, vibration, and acceleration in direct proportion to the degree of the force effect on the component parts. Where balanced circuits are involved, the effect may seem greatly magnified if one member of a balanced pair is affected while the other is not.

6.8.7 Reliability considerations.

6.8.7.1 Failure modes. Reliability of mixers as a topic literally covers the field of electronic reliability. It is an oversimplification to say that the reliability of a mixer is the product of the reliability of the component parts, unless the parts count is taken in the same way as the parts count of an equipment. In equipments, the reliability of each connection--as well as the reliability of the discrete parts--must be considered.

A major consideration is the method of construction. For instance: is the assembly on a printed circuit board? Is there conformal coating? Are the parts embedded in plastic encapsulation? Is there stress relief to prevent embedment crushing of the parts? These and similar questions must be answered if an accurate estimate of reliability and reliability factors is to be considered.

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6.8 MICROWAVE DEVICES, ELECTRICAL MIXERS

A brief review of the basics of mixers reveals that the action of mixing takes place in one or more devices that do not have proportional current versus voltage relationships. A number of passive parts may be included in the mixer assembly to provide the proper magnitudes of voltages and currents, and proper phases to give the desired mixer action. As a general rule, the reliability of the mixing device is more in question than the reliability of passive (linear) parts.

The discussion of reliability may be divided into three parts. Most commonly reliability is discussed in terms of catastrophic failure, where a permanent and disabling fault occurs. There are also drift failures, where a desired function is permanently impaired because of a permanent shift of the characteristics outside of some critical parameter limit. Finally, there are reversible changes in a critical characteristic, usually in response to an environment such as temperature, often with full recovery of the desired function when the forcing stress is removed. All of these factors should be considered in a discussion of mixer reliability.

Because the field of mixers and the parts that go into making the assembly is broad, only some of the reliability factors are discussed in this section. The reader is urged to scan other sections of the manual covering parts used (diodes, inductors, and others) for discussion of some of the more subtle reliability considerations affecting those parts.

The input range within which diodes and transistors will operate properly is very limited. Certain minimum input levels (such as local oscillator input) are needed to bring the mixer device into the optimum operating range, but as little as two or three times the proper driving level may be sufficient to burn out the mixer device. Thus, an excessive drive level may cause catastrophic failure, and if the level is too low, it will cause a temporary degradation in operating characteristics until corrected.

Environments affect mixers, usually as a result of temperature changes, causing characteristics to shift, or by the adverse effect of moisture on structures or electrical resistances. Changes in temperature tend to cause stress and fatigue in mechanical joints, often resulting in the failure of hermetic seals. Moisture that reaches semiconductor junctions can cause direct or indirect contamination. These examples show the desirability of conducting seal leak tests after thermal cycling semiconductors.

Such screening is only a small part of the screening desirable with semiconductors. Screening provisions of MIL-S-19500 are very desirable for transistors and diodes. Although screening is intended to weed out individual devices of poor workmanship, it is also helpful in identifying poor designs or material. However, there is no substitute for good electrical and mechanical design as a foundation for reliability.

6.8 MICROWAVE DEVICES, ELECTRICAL MIXERS

Often, particularly in the microwave frequency region, the discrete parts of the mixer assembly are packaged in containers that are bulky enough to cause excessive stray inductance and capacitance.

It is tempting, and in many cases essential, to use unpackaged discrete parts, but the part protection previously supplied by the package must be furnished in some way. This goal is commonly met by arranging mixer ports to accommodate transmission lines carrying signals into and out of a sealed enclosure containing the mixer assembly. A properly terminated transmission line does not either narrow the bandwidth of the port or add reactance. The usual low impedance eases the metal-ceramic seal problem of added shunt capacitance to conductors passing through an insulator mounted in the wall of a metal container. The sealing of a large container, though, can be much more difficult than the sealing of a diode or transistor package.

After the reliability factors of the parts making up the mixer assembly have been considered, the design and workmanship of the assembly itself deserve investigation. A sealed package must have a good mechanical design. The sealing operation itself must not damage the assembly and the finish of the package must meet paint or plating requirements.

Some aspects of poor design are inherent to the assembly; others are triggered by external factors. One example of the inherent poor design might be the use of a point contact diode when a Schottky barrier (perhaps with bias) will provide equivalent electrical performance with increased reliability. Specification MIL-S-45743 gives a great deal of guidance on aspects of conventional solder assembly which affect reliability.

Poor design triggered by external factors is common, and high temperature is, perhaps, the leading cause. An extremely common example, is the use of lead-tin solder in assemblies later processed at temperatures above 180 °C. Temperature shock and vibration often reveal that a bad choice of structure or materials was made. Some apparently good materials in an enclosure may outgas and contaminate the enclosure. Age affects materials, too, usually adversely.

When considering high-temperature environments, the combination of age and heat has several adverse effects. In semiconductors this combination is the accelerator of "purple plague," where a joint of aluminum and gold conductors swells into a brittle compound of gold and aluminum.

Wire insulation similarly has a temperature rating, with the temperature being the combination of the ambient plus the temperature rise of the device used. (In low-level mixers, the temperature rise may usually be ignored.) The temperature codes often used are as listed in the table herein:

6.8 MICROWAVE DEVICES, ELECTRICAL MIXERS

TABLE XIII. Wire insulation temperature rating

Temperature	Military	Commercial	Notes
85 °C	Q	--	Minimum limit for some types of organic insulations
90 °C	--	O	--
105 °C	R	A	--
130 °C	S	B	Usual limit for organic insulations
155 °C	V	--	Usual limit to avoid outgassing and weight loss
170 °C	T	--	Inorganic insulation (mica, asbestos, etc.) only
180 °C	--	H	

The military code U and commercial code C are occasionally used, but usually must be individually and additionally described.

6.8.7.2 Screening. Ideally, screened parts should be used in a mixer assembly to achieve the highest possible quality. However, that is not always possible. Therefore, the intent of the military specification is to perform screening of the entire mixer to achieve a high degree of quality and allow the manufacturer to use unscreened parts. Screening steps should include internal visual examination, high temperature storage for at least 24 hours at the maximum rated storage temperature, thermal shock for 10 cycles at minimum and maximum rated storage temperature, hermetic seal (if applicable) fine and gross leak, and high temperature power conditioning with power injected directly into the diodes at 60 Hz, if possible, or at the frequency of operation. Initial and final electrical tests with calculation of delta limits (percent change) should be specified. Additional tests might include radiographic inspection, and electrical tests at high and low operating temperatures.

**6.9 MICROWAVE DEVICES, CIRCULATORS
AND ISOLATORS**

6.9 Circulators and Isolators.

6.9.1 Introduction. A circulator is a three-port device used to direct and separate rf signals. An rf signal entering a port of a circulator interacts with a magnetized ferrite and is directed out the next port in the direction of rotation. A very small signal leaks through to the port in the opposite direction of rotation. This leakage signal, when compared with the output signal in the direction of rotation, is called the isolation. The loss of input signal at the output port due to internal mismatch and dissipation is called the insertion loss. If a short circuit is put on the first port in the direction of rotation, the signal travels to the next port in the same direction. The power will be reduced by the insertion loss of the two paths of travel.

When the third port of a circulator is terminated in a matched load, the device is an isolator. An isolator is a unidirectional device, somewhat like a diode, that presents a matched impedance to a source when it is connected to a mismatched load. The worst case of mismatch presented by the isolator may be measured by placing a variable phase short circuit on the output of the isolator and measuring the input VSWR as a function of the phase of the output short circuit.

6.9.2 Usual applications. Isolators are used predominately as matching devices and are almost mandatory in power and frequency generating circuits. Certain rf power amplifiers, varactor multiplier chains, and Gunn device circuits would not be possible without these devices.

Isolators and circulators are relatively narrow band devices, as can be seen from a plot of isolation versus frequency. Typically, temperature shifts this curve in frequency. Device characteristics must be controlled somewhat beyond the frequency band used, and the isolation should be tested at the operating temperature. Interaction of the device's magnetic field with nearby parts should not occur, nor should nearby parts interact magnetically with the isolator or circulator.

Circulators and isolators are generally low-power to medium-power devices. Air- and liquid-cooled devices are possible when necessary. Special consideration must be given to achieve adequate cooling of the ferrite and to assure that the power level will not cause ferrite limiting or arcing. A duplexer is a special case of a circulator using hybrids and ferrite in a four-port device to combine both transmit and receive signals to the antenna.

6.9.3 Physical construction. Coaxial circulators are usually a stripline configuration with coaxial connectors. The ground planes are the circulator body. A ferrite is placed in the center of the center conductor with magnets mounted into the ground planes. Connectors are usually flange-mounted with solder connection to the center conductor. Small pieces of dielectric material are sometimes epoxied on or near the center conductor for matching. Isolators use the same general construction with a built-in or connector-mounted resistive load.

6.9 MICROWAVE DEVICES, CIRCULATORS AND ISOLATORS

Waveguide circulators have the ferrite, usually on a stepped transition, mounted in the waveguide. A three-port waveguide Y-junction circulator is shown in Figure 99A. Magnets are mounted on the waveguide walls and sometimes use a "C" type magnet for magnetic field concentration. Isolators use insert or flange-mounted loads. A four-part waveguide, Faraday rotation circulator is shown in Figure 99B. Energy entering port 1 will leave at port 2, but energy entering port 2 will leave from port 3. Similarly, energy entering port 3 will leave at port 4, and energy entering port 4 will leave at port 1. Such a device is useful for coupling an antenna to a receiver-transmitter system.

Drop-in isolators are used in microstrip circuits to provide input and output matching of the circuit and internal components. They are purchased complete with magnets and a load in place. The flat input and output leads are matched and may be soldered directly to the microstrip trace. A stripline drop-in Y-junction circulator is shown in Figure 99C.

Air-gap in the ferrite-to-magnet interface is minimized for all types of construction; and ferrite placement, spacing, and conductor interfaces are critical for high performance units. Magnetic field strength, choice and placement of ferrite material, and matching of materials are the key to a good design.

6.9.4 Military designation. Military circulators are found in MIL-C-28790. Military isolators are found in MIL-C-28791.

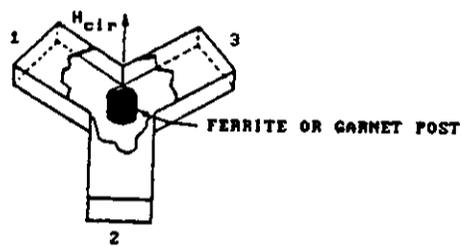
6.9.5 Electrical characteristics. Primary parameters of a circulator are VSWR, isolation, and insertion loss. Typical values for coax units are VSWR = 1.2:1 maximum, isolation = 20 dB (at band edge) minimum, and insertion loss = 0.4 dB maximum across the specified frequency range. Isolation should peak, and VSWR and insertion loss should be at a minimum at band center because the device is a tuned unit. Waveguide units are similar, except that VSWR will be less. Higher isolation can be achieved by cascading units, but at the expense of higher insertion loss. Broader bandwidth can be achieved by overlapping isolation curves of two units. Four-port (or more) units are available.

6.9.6 Environmental considerations. The environmental considerations most critical for circulators and isolators are mechanical- and temperature-related. Efficient rotation of the electromagnetic field will be degraded by anything which may cause the physical location of the ferrite or magnet to move. This includes mechanical shock, repeated thermal shocks or high accelerations.

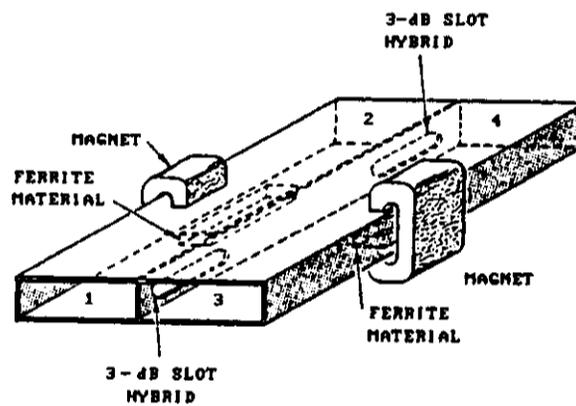
Solvents and extremely high or low temperatures may weaken adhesives used to mount ferrites and magnets. Braze joints, flanges, and bodies of waveguide circulators and isolators must withstand humidity and salt corrosion. High power circulators and isolators may have to be pressure tight and withstand high power at high altitudes. Temperature extremes affect VSWR, isolation, and insertion loss and derate the power handling capabilities of isolator loads.

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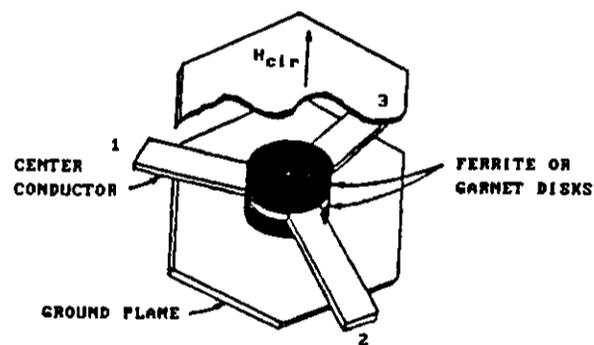
6.9 MICROWAVE DEVICES, CIRCULATORS
AND ISOLATORS



A. Waveguide Y-junction circulator



B. Four-port Faraday circulator



C. Stripline Y-junction circulator

FIGURE 99. Construction of circulator.

**6.9 MICROWAVE DEVICES, CIRCULATORS
AND ISOLATORS**

6.9.7 Reliability considerations.

6.9.7.1 Failure modes. Circulator and isolator failures usually exhibit themselves in reduced performance of isolation, VSWR, and insertion loss over the bandwidth of interest. The cause is usually mechanical in nature. Movement of the ferrite or magnet or tuning elements will cause rapid degradation of electrical parameters. Mismatch of an isolator load element will cause signal reflections to rotate to the output or input port. In high-peak power applications, arcing may occur due to spacing, burrs, or as a result of power and altitude conditions. Isolator load destruction may result from misapplication of power to the isolator, large reflections of power from the output, or inadequate heat dissipation due to lack of derating or poor heat sinking.

6.9.7.2 Screening. Initial electrical measurements of isolation, VSWR, and insertion loss followed by thermal shock (10 cycles) or temperature cycling (25 cycles) followed by final electrical measurements with calculation of delta parameters is an excellent screen for circulators and isolators. A high-power, high-temperature burn-in for isolators is effective for screening potential failures of load elements. If this is not possible, a high-temperature storage for 100 hours at the maximum rated temperature of the device should be done. Electrical testing of isolation, VSWR, and insertion loss at low and high temperatures may be required for extremely narrow band devices.

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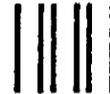
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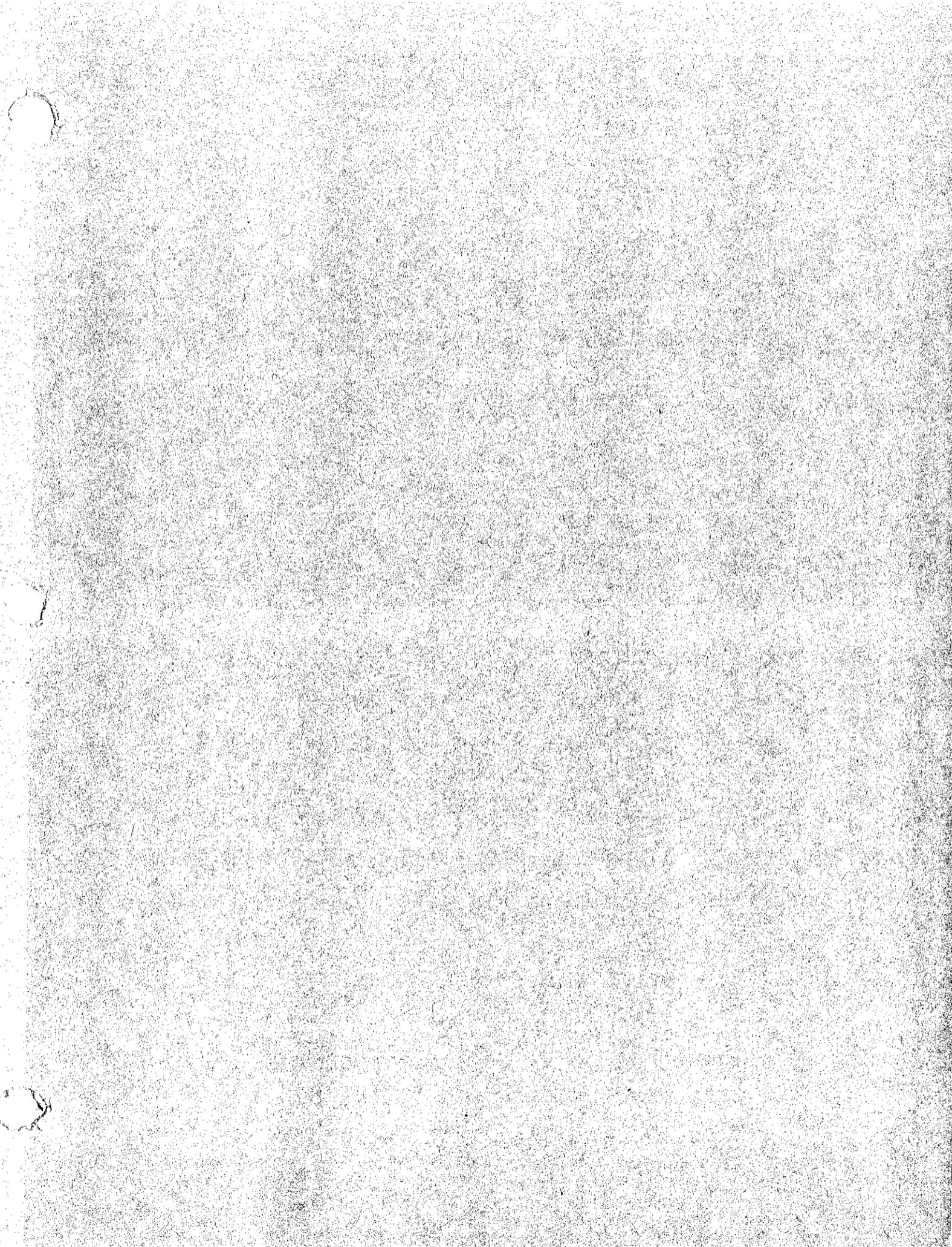
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1 MARCH 1988

SUPERSEDING
MIL-HDBK-978-A (NASA)
15 MARCH 1984

MILITARY HANDBOOK

NASA PARTS APPLICATION HANDBOOK

(VOLUME 3 OF 5)
MICROCIRCUITS



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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

WASHINGTON, D.C. 20546

NASA Parts Application Handbook

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2. Beneficial comments (recommendations, additions, deletions) and any pertinent data which may be of use in improving this document should be addressed to: Manager, NASA Parts Project Office, Goddard Space Flight Center, Greenbelt, Maryland 20771.

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FOREWORD

This handbook provides a technological baseline for parts used throughout NASA programs. The information included will improve the utilization of the NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List (MIL-STD-975) and provide technical information to improve the selection of parts and their application, and failure analysis on all NASA projects. This handbook consists of five volumes and includes information on all parts presently included in MIL-STD-975.

This handbook (Revision B) succeeds the initial release. Revision A was not released. The content in Revision B has been extensively changed from that in the initial release.

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7.1 MICROCIRCUITS, GENERAL

7. MICROCIRCUITS

7.1 General.

7.1.1 Introduction. The field of microelectronic devices is a rapidly advancing technology in which new devices become outdated in a relatively short span of time. In this application handbook, the following grouping is used:

Low-power Schottky transistor-transistor logic microcircuits -
subsection 7.2

Advanced low-power Schottky transistor-transistor logic microcircuits -
subsection 7.3

Advanced Schottky transistor-transistor logic microcircuits -
subsection 7.4

4000 series CMOS microcircuits - subsection 7.5

High-speed CMOS microcircuits - subsection 7.6

Interface microcircuits - subsection 7.7

Memories - subsection 7.8

Microprocessors - subsection 7.9

Linear microcircuits - subsection 7.10

Hybrid microcircuits - subsection 7.11

When used within their limitations, microcircuit devices offer several advantages over circuits made from discrete devices. Their high packaging densities permit reduced end product sizes and minimum propagation delay times. When produced in volume they are generally far less expensive. Complex circuits can be created with a minimum of connections within a single hermetically sealed package.

There are some disadvantages which should be examined when microcircuits are evaluated for an application. Because of their small size, microcircuits are power limited. The close proximity of active elements, unless carefully shielded, can sometimes result in crosstalk problems. The nonrecurring cost for development of monolithic circuits is very high. Therefore, when not made in volume production, the overall cost per device is also high. However, when considering all major factors such as reduced size, power consumption, powerful circuit implementation, and reliability, microcircuits are clearly the choice of circuit designers and application engineers.

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7.1 MICROCIRCUITS, GENERAL

Microcircuits may be regarded as capable of providing high reliability as a result of such factors as reduced component count, elimination of interconnections and high mechanical integrity.

7.1.1.1 Applicable military specifications. The following documents of the latest issue in effect are applicable:

<u>Mil Spec</u>	<u>Title</u>
MIL-M-38510 MIL-PRF-38535	Microcircuits, General Specifications for
MIL-M-55565	Microcircuits, Packaging of
MIL-STD-105	Sampling Procedures and Tables for Inspection by Attributes
MIL-STD-883	Test Methods and Procedures for Microelectronics
MIL-STD-975	NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List
MIL-STD-1331	Parameters to be Controlled for the Specification of Microcircuits
MIL-STD-1772	Certification Requirements for Hybrid Microcircuit Facilities and Lines
MIL-HDBK-108	Sampling Procedures and Tables for Life and Reliability Testing
MIL-HDBK-217	Reliability Stress and Failure Rate Data for Electronic Equipment
MIL-HDBK-251	Reliability/Design Thermal Applications
DoD-HDBK-263	Electrostatic Discharge Control Handbook

7.1.2 Definitions, abbreviations, conversion factors.

7.1.2.1 Term definitions.

Absolute accuracy. Allowable error of the full scale value in relation to the absolute voltage standard (also, see relative accuracy).

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Absorbed dose. See Dose.

Access time.

- a. The time required for a computer to move data between its memory (storage) section and its computation section.
- b. Time between application of address (input signal) and availability of data at the output (also called read time).

Accumulator. One or more registers associated with the arithmetic logic unit (ALU) which temporarily store sums and other arithmetical and logical results of the ALU.

Active elements. Those components in a circuit which have gain or which direct flow of current (diodes, transistors, SCRS, etc).

A/D. Analog-to-digital.

Adder. Switching circuits which combine binary bits to generate the sum and carry of these bits. Takes the bits from the two binary numbers to be added (addend and augend) plus the Carry from the preceding less significant bit and generates the sum and carry.

Address.

- (noun) A code label that identifies a specific location in a microprocessor's memory where certain information is stored.
- (verb) Selection of stored information for retrieval from a microprocessor's memory.

Amplifier, Darlington. A connection of two transistors that results in a current gain (hfe) which is the product of the individual transistor gains.

Amplifier, differential. A circuit which amplifies the difference of potential between two input signals.

Amplifier, operational (op amp). A linear amplifier circuit having high gain and input impedance and low output impedance.

Analog-to-digital converter (ADC). A device or a circuit which changes a continuous function (e.g., voltage, current, etc.) into digital outputs. The inputs may be dc or ac and the outputs may be parallel or serial and coded in various codes.

AND. A Boolean logic expression used to identify the logic operation wherein given two or more variables, all must be logic 1 for the result to be logic 1. The AND function is graphically represented by the dot (·) symbol.

7.1 MICROCIRCUITS, GENERAL

Arithmetic logic unit (ALU). The ALU is one of the three essential components of a microprocessor, the other two being the registers and the control block. The ALU performs various forms of addition and subtraction; the logic mode performs such logic operations as ANDing the contents of two registers, or masking the contents of a register.

Astable device. A device which has two temporary states and oscillates between these states.

Asynchronous. Nonclocked operation of a switching network in which the completion of one operation triggers the next.

Asynchronous, microprocessor. Operation of a switching network by a free-running signal which signals successive instructions, the completion of one instruction triggering the next. There is no fixed time per cycle.

Baud rate. A measure of data flow. The number of signal elements per second based on the duration of the shortest element. When each element carries one bit, the baud rate is numerically equal to bits per second (bps). The baud rates on UART data sheets are interchangeable with bps.

Beta, inverse. Resulting gain of a transistor when the emitter and collector bias connections are physically reversed in a circuit operation.

Beta particle. A negatively or positively charged electron.

Bidirectional. A term applied to a port or bus line that can be used to transfer data in either direction.

Binary code. A code which is based on a binary (two valued: "0" and "1") system of numbers; i.e., two distinct kinds of values, as for example, the presence and absence of a pulse. In the binary code 0000....0 corresponds to zero analog value, whereas 1111....1 corresponds to the most positive analog value (+ full scale).

Binary coded decimal (BCD). A binary numbering system for coding decimal numbers in groups of 4 bits. The binary value of these 4-bit groups ranges from 0000 to 1001 and codes the decimal digits 0 through 9. To count to 9 takes 4 bits, to count to 99 takes two groups of 4 bits, to count to 999 takes three groups of 4 bits, etc.

Bipolar converter. An A/D converter whose input range varies from a negative value to a positive value (i.e., -5 to +5 V) is referred to as a bipolar type. Similarly, a D/A converter whose output range varies from a negative value to a positive value is referred to as a bipolar type.

Bistable element. Another name for flip-flop. A circuit in which the output has two stable states (output levels 0 and 1) and can be caused to go to either of these states by input signals, but remains in that state permanently after

7.1 MICROCIRCUITS, GENERAL

the input signals are removed. This differentiates the bistable element from a gate also having two output states but which requires the retention of the input signals to stay in a given state.

The characteristic of two stable states also differentiates it from a monostable element which keeps returning to a specific state, and an astable element which keeps changing from one state to another.

Boolean algebra. The mathematics of logic which uses alphabetic symbols to represent logic variables and 1 and 0 to represent states. There are three basic logic operations in this algebra: AND, OR, and NOT.

Bremsstrahlung. German for "radiation resulting from a stopping process" or, literally, "from braking." This term designates electromagnetic radiation generated when high-energy charged particles are decelerated by electric and/or magnetic fields.

Bucket brigade. A circuit with capacitors for charge storage and means for transferring charge serially from capacitor to capacitor. Usually implemented on a silicon substrate using MOS processing.

Buffer. A circuit inserted between other circuit elements to prevent interactions, to match impedances, to supply additional drive capability, to delay rate of information flow, or to convert input and output circuits for signal level compatibility. Buffers may be inverting or noninverting.

Bulk damage. Radiation-induced defects in the crystal lattice of material which, in a semiconductor, act as additional recombination centers for minority carriers and thus decrease the lifetime of the minority carriers.

Byte write cycle. A timing diagram for writing a byte.

Capacity (memory). Total number of bits that can be stored within a memory. Usually a power of 2 (e.g., $2^{10} = 1024$, called 1K).

Cell.

- a. A set of interconnected elements (e.g., transistors, resistors, and capacitors) with a specific circuit function (e.g., 2-input NAND gate, inverter, flip-flop, etc.).
- b. Basic storage element to memorize one bit of information.

Charge coupled device (CCD). A transfer device that stores a charge in discrete regions in a semiconductor, utilizes charge transfer for readout, and charge injection into the semiconductor for charge removal.

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Charge pump. Device based on a MOS transistor, where a small charge (current) flows from source to substrate when a pump frequency is applied to the gate. A charge pump is used as a constant current source to replace load devices in some memory cells.

Charge transfer device. A device the operation of which depends on the movement of discrete packets of charge along or beneath the semiconductor surface.

Charged particle. Any nuclear particle (electron, proton, etc.) having an electrical charge associated with it.

Clear. An asynchronous input; also called reset. It is used to restore a memory element or flip-flop to a standard state, forcing the Q terminal to logic 0.

Clock. A pulse generator or signal waveform which controls and synchronizes the timing of information flow and processes in a digital network or system.

Common mode rejection ratio (CMRR). The ratio of the input common-mode voltage range to the peak-to-peak change in input offset voltage over this range.

Comparator, differential. Same as differential amplifier but having only two possible stable output states. The output is dependent upon the relative level of the two input signals.

Comparator, voltage. A circuit which generates a logic output that is dependent upon two signals at its input.

Complementary binary code. Complementary binary code is similar to the binary code, however in the complementary binary code 0000....0 corresponds to the positive analog value, whereas 1111....1 corresponds to the most negative analog value.

Complementary metal-oxide semiconductor (CMOS). A circuit design having both P and N channel field effect transistors and processed on the same wafer.

Compton effect. The interaction of a photon with an electron where some of the energy of the photon goes to the recoil electron, and the rest remains with the photon (degraded in energy) which may make still more collisions.

Conductive current (radiation controlled). An abnormally high leakage current--flowing in insulators or semiconductors because of a radiation-induced increase in their conductivity.

Conductor-insulator semiconductor (CIS). A general description used in MOS and other structures.

Conversion rate. The number of conversions an A/D converter is capable of making per second, usually expressed in MHz or KHz.

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Conversion time. The time a converter uses for one complete conversion.

Converter, A/D. Analog-to-digital converter (encoder) for changing an analog quantity to an equivalent digital word.

Converter, D/A (decoder). Device used for changing a digital word to an equivalent analog quantity.

Counter. A device capable of changing states in a specified sequence upon receiving appropriate input signals. The output of the counter indicates the number of pulses which have been applied. (See also Divider). A counter is made from flip-flops and some gates. The outputs of all flip-flops are accessible to indicate the exact count at all times.

Counter, binary. An interconnection of flip-flops having a single input so arranged to enable binary counting. Each time a pulse appears at the input, the counter changes state and tabulates the number of input pulses for readout in binary form. It has 2^n possible counts where n is the number of flip-flops.

Counter, ring. A special form of counter sometimes called a Johnson or shift counter which has very simple wiring and is fast. It forms a loop of circuits of interconnected flip-flops so arranged that only one is 0, and that, as input signals are received, the position of the 0 state moves in sequence from one flip-flop to another around the loop until they are all 0, then the first one goes to 1 and thus moves in sequence from one flip-flop to another until all are 1. It has 2^n possible counts where n is the number of flip-flops.

Counter, ripple. A binary counting system in which flip-flops are connected in series. When the first flip-flop changes it affects the second which affects the third and so on. If there are ten in a row, the signal must go sequentially from the first flip-flop to the tenth.

Cosmic ray. High-energy particles or electromagnetic radiation originating in interstellar space.

Cumulative dose (radiation). The total dose resulting from repeated exposures to radiation of the same region or the whole body.

Current mode logic (CML). Type of logic in which bipolar transistors operate in the unsaturated mode. It has very high switching speeds and low logic swings. Emitter coupled logic--ECL is the most common representative of CML.

Cycle time. Also called read-write cycle time. It is a measure of how long it takes to obtain information from a memory and then to write back information into the memory.

D/A. Digital-to-analog.

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7.1 MICROCIRCUITS, GENERAL

Damage threshold. The fluence or dose at which detectable degradation of a component parameter or parameters occurs.

Decoder. A device that translates a combination of signals into a single signal.

Delay. The slowing up of the propagation of a pulse either intentionally, such as to prevent inputs from changing while clock pulses are present, or unintentionally as caused by transistor rise and fall time pulse response effects.

Depletion mode. Describes a type of field effect transistor in which a conducting channel exists between source and drain in the absence of a gate bias voltage. To change the channel to the nonconducting state, the field established by the gate bias must deplete the channel of carriers.

Destructive read-out (DRU). A memory in which reading the contents of a storage cell destroys the contents of that location.

Detector, phase. A circuit that produces an output signal that is proportional to the phase difference between two input signals.

Diagram, logic. A picture representation for the logic functions such as AND OR, NAND, NOR, and NOT.

Digital-to-analog converter (DAC). A device or a circuit which changes digital inputs into a continuous analog signal (e.g., voltage, current, etc.).

Diode, Schottky. A metal semiconductor contact having diode characteristics.

Displacement damage. Degradation induced in a material by the displacement of atoms from their initial locations by collisions with bombarding nuclear radiation.

Displacement effects. The effects of displacement in the lattice structure of a material that results from particulate irradiation. See bulk damage.

Divider, frequency. A counter which has a grating structure added which provides an output pulse after receiving a specified number of input pulses. The outputs of all flip-flops are not accessible.

Dose. The energy deposited per mass of absorbing material. Used to describe ionization effects where the number of electron/hole pairs generated is proportional to the energy lost by the incident radiation. The unit, a Rad (material), corresponds to 100 ergs of energy absorbed per gram of material.

Dot AND. Externally connecting separate circuits or functions so that the combination of their outputs results in an AND function. The point at which the separate circuits are wired together will be a 1 if all circuits feeding into this point are 1 (also called WIRED OR).

7.1 MICROCIRCUITS, GENERAL

Dot OR. Externally connecting separate circuits or functions so that the combination of their outputs results in an OR function. The point at which the separate circuits are wired together will be a 1 if any of the circuits feeding into this point are 1.

Drain. The terminal in a field effect transistor which receives carriers from the conducting channel.

Driver. An element which is coupled to the output stage of a circuit to increase its power or current handling capability or fanout; for example, a clock driver is used to supply the current necessary for a clock line.

Driver, bus. An integrated circuit which is added to the data bus system to facilitate proper drive to the CPU when several memories are tied to the data bus line. These are necessary because of capacitive loading which slows down the data rate and prevents proper time sequencing of microprocessor operations.

Dual-in-line package (DIP). A microcircuit package having two rows of pins (usually spaced on 0.1-inch centers.)

Electromagnetic radiation. Radiation associated with a periodically varying electric and magnetic field that is traveling at the speed of light, including radio waves, X-rays, and gamma ray radiation.

Electron. An elementary particle with a mass of 9.1×10^{-31} kg and a negative charge of magnitude 1.6×10^{-19} coulombs. Electrons are stable, point like, and possess an intrinsic angular momentum (spin) of magnitude $(h/2\pi)$.

Electron charge. The magnitude of the charge of an electron. The fundamental unit of charge.

Emitter coupled logic (ECL). A variety of current mode logic.

Emitter follower. A transistor circuit in which the input signal is applied to the base and the output is taken at the emitter.

Enable. To permit an action or the acceptance or recognition of data by applying appropriate signals (generally a logic 1 in a positive logic) to the appropriate input.

Encoder. A unit which changes inputs into coded combinations of outputs.

Energy spectrum. The distribution of radiation, such as X-rays, neutrons, electrons, and protons, as a function of energy.

Enhancement type. Describes a type of field effect transistor in which no conducting channel exists between source and drain in the absence of a gate bias voltage. To change the channel to the conducting state, the field established by the gate bias must enhance or create carriers in the channel.

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Environmental component. Any specific type of radiation contributing to a radiation environment consisting of mixed radiation types.

Etching. A process by which a material is selectively removed, usually by chemical means.

Exclusive OR. A Boolean logic operation which produces a logic level 1 output if at least one, but not all, of the inputs is at a logic level 1.

Fall time. A measure of the time required for the output voltage of a circuit to change from a high voltage level to a low voltage level after a level change has started. Current could also be used as the reference, that is, from a high current to a low current level.

Family, logic. A group of integrated circuits based on the same basic circuit design and with compatible input-output characteristics.

Fan-in. The number of inputs available to a specific logic stage or function.

Fan-out. The number of input stages that can be driven by a circuit output.

Field effect transistor (FET). A circuit and process technology in which conductivity of a defined channel is changed by means of an electric field.

Film, thick. A method of manufacturing integrated circuits by depositing thin layers of materials on an insulated substrate (often ceramic) to perform electrical functions; usually only passive elements are made this way.

Filter, active. A device using electronic gain elements for selective treatment of frequencies in a signal.

Flip chip. A semiconductor die suitable for mounting on a substrate with the circuit-side down.

Flip-flops (storage elements). A circuit having two stable states and the capability of changing from one state to another with the application of a control signal and remaining in that state after removal of signals.

Flip-flop, D. D stands for delay. A flip-flop the output of which is a function of the input that appeared one pulse earlier; for example, if a 1 appeared at the input, the output after the next clock pulse will be a 1.

Flip-flop, J-K. A flip-flop having two inputs designated J and K. At the application of a clock pulse, a 1 on the J input and 0 on the K input will set the flip-flop to the 1 state, a 1 on the K input and a 0 on the J input will reset it to the 0 state; and 1s simultaneously on both inputs will cause it to change state regardless of the previous state. J=0 and K=0 will prevent change.

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Flip-flop, R-S. A flip-flop consisting of two cross-coupled NAND gates having two inputs designated R and S. A 1 on the S input and 0 on the R input will reset (clear) the flip-flop to the 0 state, and 1 on the R input and 0 on the S input will set it to the 1. It is assumed that 0s will never appear simultaneously at both inputs. If both inputs have 1s it will stay as it was. The 1 is considered nonactivating. A similar circuit can be formed with NOR gates.

Flip-flop, R-S-T. A flip-flop having three inputs, R, S, and T. This unit works as the R-S flip-flop except that the T input is used to cause the flip-flop to change states.

Flip-flop, T. A flip-flop having only one input. A pulse appearing on the input will cause the flip-flop to change states. It is used in ripple counters.

Fluence. The number of particles or photons or the amount of energy that enters an imaginary sphere of unit cross-sectional area. The time integrated flux. The units are particles/cm².

Flux. At a given point, the number of photons or particles or energy incident per unit time on a small sphere centered at that point, divided by the cross-sectional area of that sphere. The units are particles/cm²--sec.

Gain (scale factor) error. The difference between a measured output and the ideal output in a D/A converter.

Gamma ray. A quantum of short wavelength electromagnetic radiation emitted by a nucleus in its transition to a lower energy state. The range of wavelengths is from about 10⁻⁸ to 10⁻¹¹ cm. Gamma rays have zero rest mass and zero charge but have energies in the range of approximately 1 MeV.

Gate. A control element for field effect transistors and for silicon controlled rectifiers.

Gate, AND. All inputs must have level 1 signals at the input to produce a level 1 output.

Gate, exclusive OR. A logic level 1 at one single input will produce a logic level 1 output. If more than one input is logic level 1 or if all inputs are logic level 0, the output logic level is 0.

Gate, NAND. All inputs must have level 1 signals at the input to produce a level 0 output.

Gate, NOR. Any one input or more than one input having a level 1 signal will produce a level 0 output.

Gate, OR. Any one input or more than one input having a level 1 signal will produce a level 1 output.

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Gate, oxide. Used in metal-oxide semiconductors. The oxide gas is the very thin silicon oxide which isolates the gate from the semiconductor channel between the source and the drain.

Gates (decision elements). A circuit having two or more inputs and one output. The output depends upon the combination of logic signals at the input.

Glassivation. The protective coating, usually silicon dioxide or silicon nitride, deposited on the entire die surface exclusive of bonding pads.

Glitch. When turn-off and turn-on times of bit switches are not precisely equal, a spike (or glitch) is induced in the output. The magnitude of this spike is dependent upon the amount of mismatch in the switching times.

Gray. A unit of total dose. One gray is equal to 1 joule per kilogram of absorbed energy. One gray is thus equal to 100 rads.

Hardened. A system or part that has been specially designed and built to survive a specified radiation environment.

Hardness. Radiation resistance.

Hexadecimal. Whole numbers in positional notation using 16 as a base. (See octal and compare.) Because there are 16 hexadecimal digits (0 through 15) and there are only 10 numerical digits (0 through 9), an additional six digits representing 10 through 15 must be introduced. The extra digits will be provided from the alphabet. Hence, the least significant hexadecimal digits read: 0, 1, 2, 3, 4, 5, 6, 7, 8, 9, A, B, C, D, E, F. The decimal number 16 becomes the hexadecimal number 10. The decimal number 26 becomes the hexadecimal number 1A.

High threshold logic (HTL). Logic circuits designed to operate in electrical noisy environments.

Hybrid integrated circuit. A mating of two or more elements; e.g., a class of integrated circuits where two or more silicon chips are interconnected within the package, or a combination of the monolithic and thick- or thin-film methods of manufacture.

IGFET. Insulated gate field effect transistor.

Inhibit. To prevent an action or acceptance of data by applying an appropriate signal to the appropriate input (generally a logic 0 in positive logic).

Input, clock. That terminal on a flip-flop whose condition or change of condition controls the admission of data into a flip-flop through the synchronous inputs and, thereby, controls the output state of the flip-flop. The clock signal performs two functions: it permits data signals to enter the flip-flop and after entry, it directs the flip-flop to change state accordingly.

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Input/output (I/O). Package pins which are tied directly to the internal bus network to enable I/O to interface the microprocessor with the outside world.

Inputs, asynchronous. Those terminals in a flip-flop which can affect the output state of the flip-flop independent of the clock. Called set, preset, reset or dc set and reset, or clear.

Inputs, synchronous. Those terminals on a flip-flop through which data can be entered but only upon command of the clock. These inputs do not have direct control of the output, such as those of a gate, but only when the clock permits and commands. Called JK inputs or ac set and reset inputs.

Instruction set. Constitutes the total list of instructions which can be executed by a given microprocessor and is supplied to the user to provide the basic information necessary to assemble a program.

Integrated circuit (EIA definition). The physical realization of a number of electrical elements inseparably associated on or within a continuous body of semiconductor material to perform the functions of a circuit.

Integrated flux. Cumulative number of particles per square centimeter over an interval of time.

Interstitial atoms. Atoms which are displaced from their equilibrium positions into a nearby vacancy.

Ionization. The separation of a normally electrically neutral atom or molecule into electrically charged components.

Ionization damage. Damage caused by interaction of incident radiation with orbital electrons.

Ionization effect. An effect resulting from material being ionized by incident radiation, ionization damage.

Ionizing radiation. Electromagnetic radiation (gamma rays or X-rays) or particle radiation (neutrons, electrons, etc.) capable of producing ions, i.e., electrically charged atoms or molecules, in its passage through matter.

Interrupt. The suspension of the normal programming routine of a microprocessor in order to handle a sudden request for service.

Interrupt mask bit. The interrupt mask bit prevents the CPU from responding to further interrupt requests until cleared by execution of programmed instructions. It may also be manipulated by specific mask bit instructions.

Interrupt, vectored. A microprocessor system in which each interrupt, both internal and external, has its own uniquely recognizable address. This enables

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the microprocessor to perform a set of specified operations which are pre-programmed by the user to handle each interrupt in a distinctively different manner.

Inverter. A circuit whose output is always in the opposite state from the input. This is also called a NOT circuit.

Isolation, dielectric. A process in which the individual components are separated through dielectric isolating layers (usually silicon dioxide).

JFET. Junction field effect transistor.

Junction leakage current. See reverse leakage current.

Large scale integration (LSI). A large number of interconnected integrated circuits manufactured simultaneously on a single slice of semiconductor material (usually over 100 gates or basic circuits with 1000 circuit elements or more).

Latchup. A situation where a device goes into a low impedance state, drawing high current until the device is powered off. The latch current may for some devices be high enough to damage the part. High dose-rate environments can cause latchup in nondielectrically isolated devices that have internal pnpn structures.

Lateral pnp (transistor). A pnp transistor used in linear monolithic integrated circuits in which current flow occurs parallel to die surface, rather than vertical.

Layer, buried. A layer of heavily doped material (high conductivity) under the collector region, generally applied to reduce the collector saturation resistance of a transistor and to reduce parasitic pnp transistor action to substrate. The buried layer is diffused on the wafer before epitaxial growth.

Lead, beam. A type of connection lead which is cantilevered beyond the edge of a completed silicon die. These leads may be used to bond the die directly into the circuit without need for fine wire bonding.

Leakage currents. See reverse leakage current.

Least significant bit (LSB). In converters, the bit which corresponds to the smallest analog increment is called the least significant bit. In an 8-bit converter, for example, it represents $(1/2)^8$, or $1/256$ of the total analog range. The lowest weighted digit of a binary number.

LET (Linear energy transfer). The stopping power of a particle, dE/dx , often expressed in units of $\text{Mev/gm} \cdot \text{cm}^{-2}$.

Linearity. Linearity of a converter is defined as the maximum deviation from the ideal straight line drawn between the end points as derived from the trans-

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fer function. Alternately, the linearity can be described as the deviation from the "best straight line" value for any given bit.

Logic. A mathematical treatment of formal logic in which a system of symbols is used to represent quantities and relationships. The symbols or logical functions are called AND, OR, NOT, etc. Each function can be translated into a switching circuit, more commonly referred to as a gate. Because a switch (or gate) has only two states--open or closed--it makes possible the application of binary numbers for the solution of problems. The basic logic functions obtained from gate circuits are the foundation of complex computing machines.

Logic, binary. Digital logic elements which operate with two distinct states. The two states are called true and false, high and low, on and off, or 1 and 0. In computers they are usually represented by two different voltage levels. The level which is more positive (or less negative) than the other is called the high level, the other the low level.

Logic, combinational. A circuit arrangement in which the output state is determined by the present state of the input. Also called combinatorial logic.

Logic, negative. Logic in which the more negative (less positive) voltage represents the 1 state and the less negative (more positive) voltage represents the 0 state.

Logic, positive. Logic in which the more positive voltage represents the 1 state; for example, 1 = +3V, logic 0 = +0.45V.

Logic, saturated. Logic in which the transistors operate in the saturated region of their characteristic. TTL is an example of this logic.

Logic, sequential. A circuit arrangement in which the output state is determined by the previous state of the input.

Logic upset. Change of state of a digital device caused by either a high dose rate environment or the passage of a single highly ionizing particle through the device (see also SEU).

Look ahead. (1) A feature of the CPU which allows the machine to mask an interrupt request until the following instruction has been completed. (2) A feature of adder circuits and ALUs which allow these devices to look ahead to see that all carries generated are available for addition.

Loop, phase locked. A circuit used for synchronizing a variable local oscillator with the phase of a transmitted signal.

Medium scale integration (MSI). A circuit smaller than LSI but having at least 10 gates or basic circuits with at least 100 circuit elements.

Memory, dynamic. Memory where the parasitic capacitance of MOS-FET gates within a storage cell is used for temporary storage of information. Due to junction

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leakage currents this is possible for only a finite time. Prior to the loss of data the information must be refreshed by some electrical method.

Memory, serial. Memory in which information is stored in series and is written or read in time sequence, as with a shift register. Compared with a RAM, the advantages of a serial memory are slow to medium speed with lower cost.

Memory, static. A memory which needs no refresh cycle to keep the data stored. With simple operation and less control circuitry, but usually is slower than dynamic memories.

Metal-oxide-semiconductor FET (MOS, MOSFET). A circuit in which the active region is a metal-oxide-semiconductor multilayer. The oxide acts as the dielectric insulator between the metal and the semiconductor. The current flow in the semiconductor is controlled by an application of a potential to the metal (called gate) thus forming a FET.

Metallization. A metal film deposited on a substrate and selectively etched to serve as conductive interconnects between elements of the integrated circuit and as bonding pads for external connections.

Microprocessor. The microprocessor is a central processing unit fabricated on one or more chips. When joined to a memory storage system, it can be programmed with stored instructions to process a wide variety of functions consisting of arithmetic and logic units, control blocks, and register arrays. Each microprocessor is supplied with an instruction set which may be just as important to the user as the hardware.

Microprocessor, clock. A generator of pulses which controls the timing of switching circuits in a microprocessor. Clocks are a requisite for most microprocessors and multiple phased clocks are common in MOS processors.

Monolithic. Refers to the single silicon substrate in which an integrated circuit is constructed.

Monotonicity. The output of a converter is monotonic when it moves always in an increasing direction in response to an increasing input stimulus (or an always decreasing direction in response to a decreasing input stimulus).

Most significant bit (MSB). In converters, it is the bit which carries the most weight and corresponds to one-half of the analog range. The highest weighted digit of a binary number.

Multiplexing. Multiplexing describes a process of transmitting more than one signal at a time over a single link, route, or channel.

Multiplier. A linear circuit which accepts two input signals and produces their algebraic product.

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N-channel. The conducting channel region induced in p-type material of a field effect transistor by a positive gate voltage.

N-channel metal-oxide semiconductor (N-MOS). The use of negative charge carriers (electrons) for operation.

N-type material. Semiconductor material that has been doped with impurity atoms (arsenic, phosphorus) so as to have an excess of free electrons.

NAND. A Boolean logic operation which yields a logic 0 output when input signals are logic 1.

Neutron. An elementary particle with no net charge and a mass of 1.675×10^{-27} kg. Though not having a net charge, neutrons do possess a magnetic moment. Neutrons are highly penetrating particles. They are unstable in free space, decaying into a proton, an electron, and an antineutrino. Neutrons have a spin of magnitude $(h/2\pi)$.

Neutron fluence. Time integrated neutron flux (Unit: n/cm^2).

Neutron flux. The product of the neutron density (number per cubic centimeter) and the neutron velocity; the flux is expressed as neutrons per square centimeter per second. It is numerically equal to the total number of neutrons passing, in all directions, through a sphere of 1 cm^2 cross-sectional area per second.

Noise immunity. A measure of the insensitivity of a logic circuit to triggering or reaction to spurious or undesirable electrical signals or noise, largely determined by the signal swing of the logic. Noise can be either of two directions, positive or negative.

Nondestructive read-out memory (NDROM). A memory where the read operation does not cause the storage cell to lose the stored information. Almost all semiconductor memories are of this type.

NOR. A Boolean logic operation which yields a logic 0 output with one or more true 1 input signals.

NOT. A Boolean logic operation indicating negation, NOT 1. Actually an inverter. If input is 1 output is NOT 1 but 0. If the input is 0 output is NOT 0 but 1. Graphically represented by a bar over a Boolean symbol such as \bar{A} .

Nuclear radiation. Neutrons, alpha, beta, and gamma rays from primary or secondary power plants and natural space radiation. Only neutrons and gamma rays penetrate shielding.

Octal. Whole numbers in positional notation using 8 as a base. The decimal (base 10) number 125 becomes 175 in octal or base 8.

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Offset binary code. The offset binary code is similar to binary code; however, in the offset binary code, 000....0 corresponds to the most negative analog value (-full scale), 111....1 corresponds to the most positive analog value (+ full scale), and 100....0 corresponds to zero analog value.

Offset error. Offset error is the deviation of the "all bits off" condition from the ideal. It must normally be compensated (by an actual adjustment) in application.

One shot. A monostable device which, when triggered by an external pulse (or level), will switch into a temporary state (period of time is determined by time constant circuitry) then revert to its stable state.

OR. A Boolean logic operation wherein two or more true 1 inputs only add to one true 1 output. Only one input needs to be true to produce a true output. The graphical symbol for OR is a plus sign (+).

Organization. Indicates how the storage cells are organized within the memory. Expressed in "words by (bits per word)" (e.g., 4096 x 1 = 4096 words of one bit per word; 64 x 4 = 64 words of four bits per word, etc.).

P-channel. The conducting channel region induced in the n-type material of a FET by a negative gate voltage.

P-channel metal oxide semiconductor (P-MOS). The use of positive charge carriers (holes) for operations.

P-type material. Semiconductor material that has been doped with impurity atoms so as to have an excess of free holes.

Parallel. This refers to the technique for handling a binary data word which has more than one bit. All bits are acted upon simultaneously.

Parameter. A definable and measurable electrical characteristic of a circuit or a device.

Parasitic elements (or parasitics). Unavoidable stray electrical effects (e.g., capacitive, resistive) which limit the performance of theoretically ideal circuit elements.

Passivation. The surface coating of the die (usually thermally grown silicon dioxide) through which contact and diffusion windows are opened.

Passive elements. Elements without gain; i.e., resistors, inductors, or capacitors.

Particle radiation. Radiation consisting of energetic particles such as electrons, protons, neutrons, and alpha particles.

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Permanent damage. Occurs when displacement and/or rearrangement of atoms or groups of atoms takes place in a material. Degree of permanent damage depends on total or integrated dose received, type of radiation, and temperature.

Photon. The quantum of electromagnetic radiation. It has an energy of $E = h\nu$ where ν is the frequency of the radiation. For some purposes, photons can be viewed as massless particles traveling at the speed of light with momentum $h\nu/c = h/\lambda$.

Power supply rejection ratio. The ratio of the change in input offset voltage to the change in power supply voltages producing it.

Preset. An input like the set input and which works in parallel with the set.

Probing. Electrical testing of integrated circuit elements using probes (pressure contacts). Most commonly, it is done before devices are broken from the wafer.

Programmed logic arrays (PLA). The PLA is an orderly arrangement of logical AND logical OR functions. Its application is similar to a ROM. It is primarily a combinational logic device.

Propagation delay. A measure of the time required for a change in logic level to be transmitted through an element or a chain of elements.

Propagation time. The time necessary for a unit of binary information (high voltage or low) to be transmitted or passed from one physical point in a system or subsystem to another. For example, from input of a device to output.

Proton. An elementary particle of mass 1.67×10^{-27} kg and positive charge equal to an electron charge. The proton is stable and has a spin of $1/2(h/2\pi)$. Low energy protons are highly ionizing but are easily stopped. Ultra high energy cosmic ray protons are far more penetrating.

Q output. The reference output of a flip-flop. When this output is 1 the flip-flop is said to be in the 1 state; when it is 0 the output is said to be in the 0 state. (See also state and set.)

\bar{Q} output. The second output of a flip-flop. It is always opposite logic level to the Q output.

Quantization uncertainty. For a given digital code there is a range of analog values associated with it. The midpoint of this range is usually specified as being the value associated with the digital code. However, because the range is 1 LSB wide, the uncertainty is $\pm 1/2$ LSB.

RAD (Roentgen absorbed dose). A unit of absorbed dose. One rad is equal to 100 ergs of absorbed energy per gram of absorbing material (e.g., H₂O, C, Si). This unit cannot be used to describe a radiation field.

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Radiation hardening. The application of certain semiconductor design and fabrication techniques which produce devices capable of operating during exposure to radiation.

Radioactivity. Spontaneous nuclear disintegration occurring in elements such as radium, uranium, and thorium and in some isotopes of other elements (e.g., Co⁶⁰). The process is usually accompanied by the emission of alpha and beta particles and/or gamma rays.

Random access. The memory words can be selected in any order; access time is independent of storage cell location.

Random access memory (RAM). A memory that has the stored information immediately available when addressed, regardless of the previous memory address location. As the memory words can be selected in any order, there is equal access time to all. (See also read/write memory.)

Random logic design. A system design using discrete logic circuits. Numerous gates are required to implement the logic equations until the problem is solved. Even then, the design is not completed until all redundant gates are weeded out.

Read-only memory (ROM). A memory in which the information is stored at the time of manufacture. The information is available at any time, but it cannot be modified during normal system operation.

Read time. Time between application of read control (and address) and availability of data at the output, commonly called access time.

Read/write memory. A memory in which the stored data is available at any time and can be changed in normal system operation (versus a read-only memory, in which stored information cannot be changed during normal system operation).

Ready/busy (R/B). A pin indicating the completion of a write operation.

Refresh cycle. The time required for the process used to keep data stored in dynamic memory. Restores a charge in a dynamic storage unit to desired voltage level.

Register. An interconnection of computer circuitry, made up of a number of storage devices (usually flip-flops), to store a certain number of digits. For example, a 4-bit register requires 4 flip-flops.

Register, shift. An arrangement of circuits (e.g., flip-flops) that are used to shift serially or in parallel. Binary words are generally loaded in parallel and then held temporarily or serially shifted.

Regulator, voltage (VR). A circuit which maintains an output voltage at a predetermined level independent of input voltage and load impedance.

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Relative accuracy. This is the deviation of actual analog values from nominal analog values for a given digital input.

Reset. Also called clear. Similar to set except it is the input through which the Q output can be made to go to 0.

Resolution. The resolution of a converter is the ratio of the value of the LSB to the full analog range, or $(1/2)^n$ where n is the number of bits.

Reverse leakage current. The current that flows when the diode is biased in the direction of greatest resistance. The reverse current, as normally measured, is a combination of reverse saturation current, carrier generation current, and surface leakage current.

Ripple. The transmission of data serially. It is a serial reaction analogous to a bucket brigade or a row of falling dominoes.

Rise time. A measure of the time required for the output voltage of a state to change from a low voltage level (0) to a high voltage level (1) after a level change has been started.

Roentgen (R). Quantity of X or gamma rays which will produce, as a result of ionization, one electrostatic unit of electricity (either sign) in 1 cc of dry air at 0 °C and standard atmospheric pressure. One roentgen = absorption of 83.8 ergs of energy per gram of air. It is the quantity of radiation that produces 2.083×10^9 ion pairs per cubic centimeter of air at standard pressure, 760 millimeters, and standard temperature, 25 °C or 77 °F at sea level. The rate of energy release is expressed in roentgens per second.

Schmitt trigger. A circuit having two output states, where the output state is determined by the level of the input, with a hysteresis loop intentionally included.

Serial. This refers to the technique for handling a binary data word which has more than one bit. The bits are acted upon one at a time.

Set. An input on a flip-flop not controlled by the clock (see asynchronous inputs) and used to affect the Q output. It is this input through which signals can be entered to get the Q output to go to 1. Note it cannot get Q to go to 0.

Settling time. The amount of time required by the output of a D/A converter to settle to within a certain percentage (usually 1 or 0.1 percent) of final value.

Single event upset (SEU). A change in logic state of a digital device due to the passage of a single ionizing particle through the device. This is referred to as a soft error as it does not in itself damage the part.

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Shift. The process of moving data from one place to another. Generally many bits are moved at once. Shifting is done synchronously and by command of the clock. An 8-bit word can be shifted sequentially (serially)--that is, the 1st bit goes out, the 2nd bit takes the 1st bit's place, the 3rd bit takes the 2nd bit's place, and so on. Generally referred to as shifting left or right. It takes 8 clock pulses to shift an 8-bit word or all bits of a word can be shifted simultaneously. This is called parallel load or parallel shift.

Shift register, half. Another name for certain types of flip-flops when used in a shift register. It takes two of these to make one stage in a shift register.

Silicon-on-sapphire (SOS). An integrated circuit structure which utilizes epitaxially grown single-crystal silicon on an insulating (sapphire) substrate.

Skewing. Refers to time delay or offset between any two signals in relation to each other.

Slew rate. Maximum rate at which the output can be driven from limit to limit over the dynamic range. A measure of how fast the output can respond to a required change in voltage.

Small scale integration (SSI). Used to describe the relative complexity of an integrated circuit with less than 10 gates and less than 100 circuit elements.

Solar flares. Chromospheric eruptions occurring in the vicinity of sun-spot groups. These eruptions are observable in certain lines in the visible and far ultraviolet ranges. They consist of intense streams of X-rays, ultraviolet rays, protons, and electrons ejected from the sun at irregular intervals by electromagnetic storms associated with sun-spots. Most of these streams are absorbed by the earth's atmosphere.

Solar winds. Streams of protons that have been ejected by the sun and are traveling through space.

Source. The terminal in a field effect transistor which provides (sources) current to the conducting channel.

State. This refers to the condition of an input or output of a circuit as to whether it is a logic 1 or a logic 0. The state of a circuit (e.g., gate or flip-flop) refers to its output. The flip-flop is said to be in the 1 state when its Q output is 1. A gate is in the 1 state when its output is 1.

Stopping power. Rate of energy lost by incident particle per unit distance. It has units of energy per length, dE/dx .

Successive approximation. This is a method frequently used in high speed analog to digital converters. The analog input is presented to an array of comparators, in succession. Each comparator is set at a precise reference level. The digital output code is generated from the comparator outputs.

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Swing, logic. The voltage difference between the two logic levels 1 and 0.

Synchronous. Operation of a switching network by a clock pulse generator. All circuits in the network switch simultaneously. All actions take place synchronously with the clock.

Temperature coefficient. Certain converter characteristics vary over the operating temperature range. Temperature coefficients are usually expressed as ppm/°C (parts per million per °C).

Temporary degradation. Radiation induced damage or degradation which recovers at room temperatures upon termination of the radiation, usually in times of a few hours or less.

Threshold dose. The minimum dose fluence that will produce a detectable degree of any given effect.

Transient effects. A phenomena which occurs when radiation causes electronic excitation without atomic displacement in a material. It usually results from ionizing radiation and is a function of the dose rate.

Throughput. The speed with which problems or segments of problems are performed.

Toggle. To switch between two states as in a flip-flop.

Transfer characteristic. The relationship of the output and input of a device or a circuit.

Transistor-transistor-logic (TTL, T²L). A logic system which evolved from DTL, wherein the multiple diode cluster is replaced by a multiple-emitter transistor but is commonly applied to a circuit which has a multiple emitter input and an active pullup network.

Trigger. A timing pulse used to initiate the transmission of logic signals through the appropriate circuit signal paths.

Truth table. A chart which tabulates and summarizes all the combinations of possible states of the inputs and outputs of a circuit. It tabulates what will happen at the output for a given input combination.

Two's complement numbers. The ALU performs standard binary addition using the 2s complement numbering system to represent both positive and negative numbers. The positive numbers in 2s complement representation are identical to the positive numbers in standard binary.

Unipolar converter. A converter is unipolar if its analog range is entirely one side of zero.

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Universal asynchronous receiver transmitter (UART). This device will interface a word parallel controller or data terminal to a bit serial communication network.

Van Allen radiation belts. Belts of energetic charged particles, mainly electrons and protons, orbiting in an equatorial plane around the earth and confined there by its magnetic field. The spectrum of particles varies with distance from the earth. Due to solar activity, the flux and energy spectrum at fixed distance from earth, varies with time.

Volatility of data. Stored information is lost in case of power shutdown.

Voltage controlled oscillator (VCO). An oscillator, the frequency of which is a function of the input voltage.

Voltage, offset. The change in input voltage required to produce a zero output voltage (or other specified output condition) in a linear amplifier circuit. In digital circuits it is the dc voltage in which a signal is impressed.

Voltage, threshold. In logic circuitry, the input voltage level required to cause the output of a device to switch from one logic level to another.

Wired OR. Externally connected separate circuits or functions arranged so that the combination of their outputs results in an AND function. However, common usage is, that the point at which the separate circuits are wired together will be a 0 if any one of the separate outputs is a 0; it is the same as a dot AND.

Word. Set of bits, usually 8, 12, 16, 24, 32, etc., treated as a unit. A word may contain one or more bytes.

Word length. The number of bits in a sequence that is treated as a unit and that can normally be stored in one memory location. Longer words imply higher precision and more intricate instructions.

Write/erase endurance cycle. Number of times device data can be written and erased.

Write time. Time needed to store information safely into the memory, after presence of data input, write control, and address.

X-ray. A form of penetrating electromagnetic radiation (zero charge, zero mass) having wave lengths shorter than those of visible light (approximately 10^{-8} cm). Usually produced by bombarding a metallic target with a particle in a high vacuum. In nuclear reactions, it is customary to refer to photons originating in the nucleus as gamma rays and those originating in the extra-nuclear part of the atom as X-rays. Often called roentgen rays.

Yield. The ratio of usable components at the end of a test or process to the number of components which were submitted to the test or process.

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7.1.2.2 Symbol definitions.

The following symbols are intended to supplement those found in subsection 5.1 Transistor general.

ICCL, IDDL, etc.	Low-level supply current drain
ICCH, IDDH, etc.	High-level supply current drain
IIH	High-level input current
IIL	Low-level input current
IINH	High-level node current
IINL	Low-level node current
IOH	High-level output current
IOL	Low-level output current
IOS	Output short circuit current
tPHL	Propagation delay high to low level output
tPLH	Propagation delay low to high level output
tTHL	Transition time high to low level output
tTLH	Transition time low to high level output
VN	Noise margin
VOH	High-level output voltage
VOL	Low-level output voltage
VT	Threshold voltage
VDD	Power supply voltage

Linear Electrical

AGC	Automatic gain control range
AVC	Common-mode voltage amplification
AVD, Avd	Differential voltage amplification
AVS, Avs	Single-ended voltage amplification
BOM	Maximum output swing bandwidth
CMRR	Common-mode rejection ratio
Gp, Gp	Power gain or insertion power gain
IIB	Input bias current
$\Delta I_{IB}/\Delta T$	Input bias current temperature sensitivity
IIO	Input offset current
$\Delta I_{IO}/\Delta T$	Input offset current temperature sensitivity
PD	DC power consumption
PSRR	Power supply rejection rate
SR	Slew Rate
THD	Total harmonic distortion
tor	Overload recovery time
TR	Transient response
VI	Quiescent input voltage
VICR	Common-mode input voltage range
VIO	Input offset voltage

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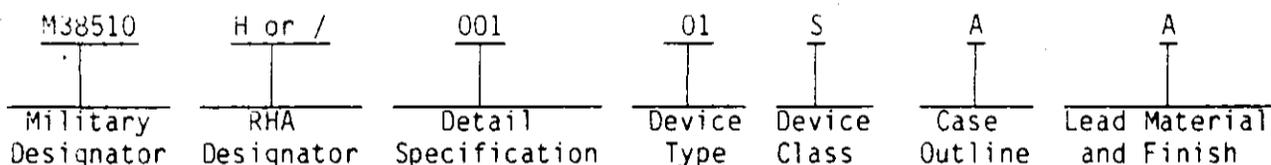
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$\Delta V_{IO}/\Delta T$	Input offset voltage temperature sensitivity
V_{ISR}	Single-ended input voltage range
V_O	Quiescent output voltage
V_{OC}	Common-mode output voltage
V_{OO}	Output offset voltage
V_{OPP}	Maximum output voltage swing
V_{OU}	AC unbalance voltage
Z_{id}	Differential input impedance
Z_{is}	Single-ended input impedance
Z_{od}	Differential output impedance
Z_{os}	Single-ended output impedance
ϕ_m	Phase margin

7.1.3 NASA standard parts.

7.1.3.1 NASA Standard Parts Program. The purpose of the NASA Part Standardization Program is to provide the designer with a list of acceptable parts and specifications for space flight missions and to reduce the quantity of part numbers used in order to derive standardization benefits. See the General Section 1.1 for additional information regarding the NASA Standard Parts Program. In addition to this manual the principal element of this program is MIL-STD-975, which is a standard parts list for NASA equipment, with Section 7 containing a summary of standard microcircuits.

7.1.3.2 Military designation. The military microcircuits designation uses the military General Microcircuit Specification number (MIL-M-38510) in conjunction with individual device specification "slash sheet" numbers. The complete part number must also include the reliability screening class, the package configuration and the lead material and finish. The designation for lead material and finish is selected from QPL-38510. An example of the complete military part number is as follows:



Whenever any one of the three lead finishes can be used in the application the lead material and finish designator should be "X."

It should be noted that fully qualified devices require JAN or J marking in front of the military designator (e.g., JM38510...); however, the military logistics system is limited to 15 characters and therefore this requirement is usually not reflected in most of the documentation. A more elaborate description of the complete military part number is given in Section 7 of MIL-STD-975.

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All microcircuits included in MIL-STD-975(NASA) are specified by existing MIL-M-38510 device detail specifications "slash sheets." The two levels of military product assurance for microcircuits are class S and class B. The product assurance class S has been established for devices that are intended for Grade 1 NASA applications and class B for Grade 2.

Generally, the use of integrated circuits specified by MIL-M-38510 reduces the overall cost of systems in which they are being used. They eliminate the cost of parts documentation and impose standardization in the form of part types available to the MIL specification. They further reduce cost over the period of system use because of the higher availability and lower maintenance costs associated with higher-reliability components.

7.1.4 General device characteristics. This section will give a broad outline of microcircuit device characteristics which the NASA applications engineer must be aware of in microcircuit applications. More detailed references will be given for actual applications use.

7.1.4.1 Available technologies. The microcircuit technologies available to the NASA applications engineer are Schottky bipolar, linear bipolar, NMOS aluminum and silicon gate, and CMOS aluminum and silicon gate. Specific circuit examples of each of these technologies are given in MIL-STD-975.

7.1.4.2 Levels of integration. The integration levels available in the microcircuits listed in MIL-STD-975 are SSI, MSI, and a relatively few LSI devices. With technological trends and further technology maturity, more LSI and eventually VLSI will be available. To answer the question of what constitutes SSI, MSI, etc., Table I presents a classification of integration levels with respect to device count per chip, and alternatively, chip functions associated with each level.

TABLE I. Classification of integrated circuits by complexity and functionality

Name	Device Count	Chip Functions
SSI	1 - 100	Gates
MSI	100 - 1000	Register parts
LSI	1000 - 100,000 - 100,000	Bit slices processors
VLSI	100,000	Computers systems

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7.1.4.3 Available packages. The majority of packages (with the exception of special packages which are designed for a specific application such as voltage regulators) can be classified into four basic styles with variations in dimensions and number of leads:

- a. Transistor style metal packages
- b. Flat packages
- c. Dual in-line packages
- d. Chip carriers.

Devices of each of these styles have certain common characteristics which allow them to be treated as a group. Refer to Appendix C of MIL-M-38510 for case outlines of specific devices.

7.1.4.3.1 Transistor style metal packages. These packages generally utilize a kovar header with kovar external leads to which the circuit is attached and over which a metal cap is welded. These packages are generally 6-, 8-, or 10-leaded variations of the popular T0-5 case (such as T0-99 and T0-100) or 12- and 16-leaded packages on the same diameter base as the T0-8. These packages offer the advantages of relatively good heat conduction, radiation hardness, ease of obtaining a good hermetic seal, ruggedness, and long manufacturing experience. The principal disadvantages of these packages are their relatively inefficient use of space and the limited number of external leads. For these reasons the transistor style metal packages are generally used only for devices which require a limited number of external leads (such as linear circuits) and are not limited by space or weight limitations.

7.1.4.3.2 Flat packages. Flat packages come in a variety of body sizes, with the most common types providing 10, 14, 16 or 24 external leads. The body can be metal (except for seals), ceramic and metal, or all ceramic. Precautions should be taken as some ceramic packages may contain beryllium oxide (BeO). In a finely powdered form (from grinding or cutting) or at high temperatures (above 900 °C) where fumes are given off, beryllium oxide is extremely toxic. Inhalation of small amounts can be fatal.

Flat packages offer the advantage of very high packaging density, but have relatively poor heat dissipation (except where the case is in contact with a good heat sink) and require more care in handling because of their thin ribbon leads. Metal flat packages can develop short circuits between leads and case because of the small clearances. Also die bond separation can occur as a result of the flexibility of the package. Flat packages are used almost exclusively in military applications.

7.1.4.3.3 Dual-in-line packages. Dual-in-line packages are made in both plastic and hermetically sealed versions (normally with leads spaced on 100 mil centers). The most common type provides 14 leads, but devices with 8, 10,

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16 and 24 leads are also common, and some dual-in-line packages are being made with much larger numbers of leads. Body materials are generally the same as for flat packages, but the dimensions are larger and the external leads heavier. Advantages include ease of handling and a greater number of external leads than modified TO-5 packages. The larger seal perimeter of these packages makes them more vulnerable to hermetic seal problems and even lid separation.

7.1.4.3.4 Chip carriers. The basic design of the chip carriers is similar to the design of flatpacks. The main differences are that the external connections in chip carriers are made on four sides (instead of only two sides made on flatpacks) and there are no leads. There are a number of variations of chip carriers of different package construction, sizes, and types and arrangements of external connections and their spacing. An example of a chip carrier is shown in Figure 1. A majority of devices have 0.050-inch center-to-center connection spacing, with a small number of devices using 0.040-inch spacing.

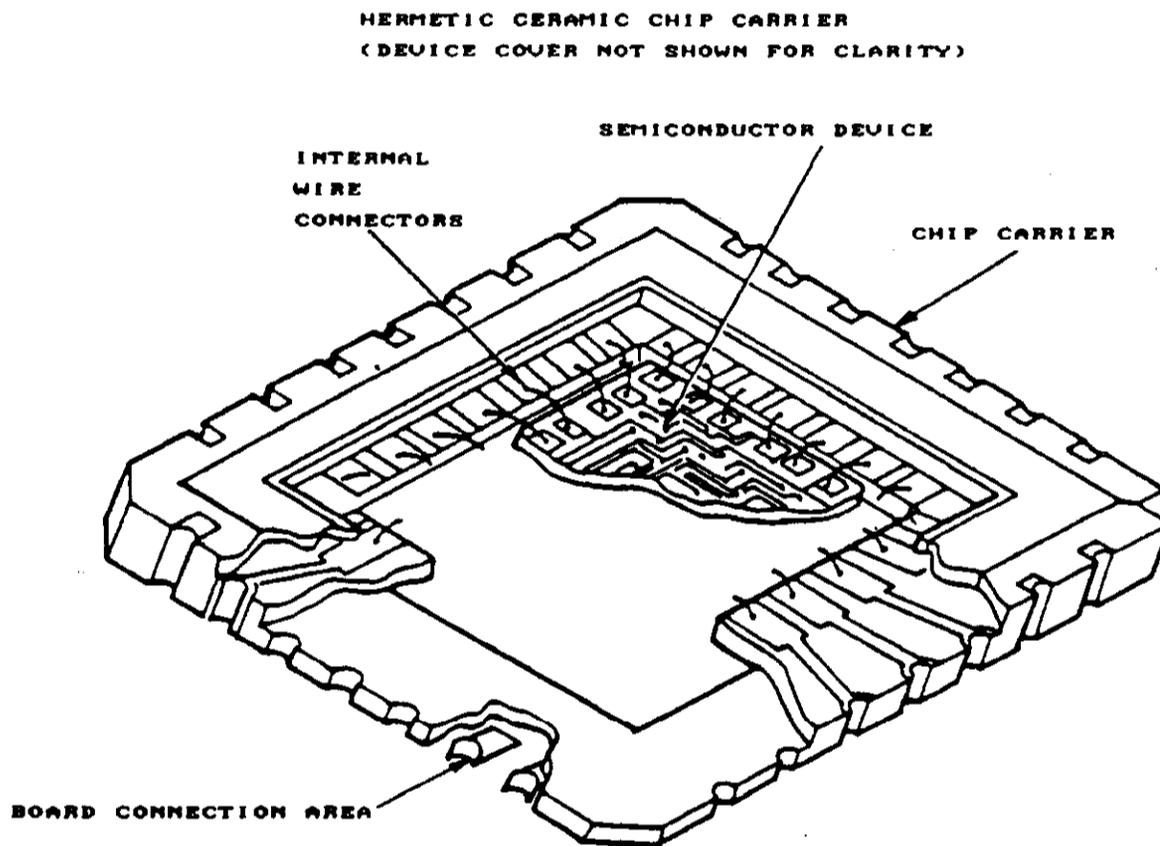


FIGURE 1. Example of a chip carrier package.

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7.1.4.4 Thermal considerations. Thermal considerations are an important part of microcircuit application engineering. Thermal resistances of the microcircuit chip and package should be considered in estimating the maximum device junction temperature under worst case conditions. The maximum junction temperature should not exceed the specified values. Good thermal design must be followed as outlined in such documents as MIL-HDBK-251 Reliability/Design Thermal Applications.

7.1.4.5 Reliability, major factors, derating. Derating is the reduction of electrical, thermal, and mechanical stresses applied to a device in order to decrease the degradation rate and prolong the expected life of the device. Derating increases the margin of safety between the operating stress level and the actual failure level for the part, providing added protection from system anomalies unforeseen by the application engineer. When derating, the application engineer must first take into account the specification environmental and operating condition rating factors, consider the actual environmental and operating conditions of the application, then apply some recommended derating factors. For microcircuits, the major derating factors are supply voltage, power dissipation, signal input voltages, and output voltages and output currents. Refer to MIL-STD-975 for recommended numerical derating factors for microcircuits.

7.1.4.6 Radiation hardness assurance. Microcircuits certified as radiation hardness assurance (RHA) microcircuits must pass group E test of test Method 5005 of MIL-STD-883. This group of tests consists of testing the microcircuit devices' susceptibility to neutron radiation and to gamma or electron radiation.

7.1.4.7 Electrostatic discharge (ESD) sensitivity and handling. All microcircuits must be considered to some extent sensitive to electrostatic discharge damage. Test Method 3015 of MIL-STD-883 establishes a method for determining the electrostatic discharge sensitivity classification of microcircuit devices. DOD-HDBK-263, Electrostatic Discharge Control Handbook, is a very complete guide for all aspects of electrostatic discharge control for microcircuits and should be used with the results of Method 3015.

7.1.4.8 Testability. In general, testability of microcircuits is best when it is designed in at the start of the chip design process. If done after the chip design, the test engineer often has inaccessible areas of functions of the chip and the test process can become very involved and convoluted. The basic principles of design for testability are:

- a. Controllability and observability. A fundamental rule is to provide a general reset to all internal states for initialization.
- b. Partitioning of complex circuits. The scan-path technique provides a structure oriented partitioning on the circuit level.

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- c. Easily testable circuits. Certain types of faults can be avoided by a design which eliminates them. Use regular structures such as PLA's and PROM's instead of random logic.
- d. Implementation of self-test and self-checking capability. Redundant hardware and on-line tests are implemented in critical functions.

7.1.4.9 Comparative attributes. Comparative attributes of the semiconductor technologies involved in NASA standard parts are discussed in the following subsections of this General Section and in the subsections on digital and linear microcircuits.

7.1.5 General parameter information. Because of the great diversity of circuit functions of microelectronic devices no meaningful treatment of parameter relationships can be presented at the general level. The digital microcircuit subsection makes a comparison of several logic types on speed, power, fan-out and other characteristics. Even these comparisons must be regarded as a broad generalization since the range of values which can be obtained from a logic type usually allows considerable overlap with other types. Lack of commonality of linear and hybrid devices makes even this type of treatment impractical.

7.1.5.1 Device specifications. The details of device specifications required for Grade 1 or Grade 2 parts are listed in Appendix F of MIL-M-38510. The requirements include content, format, interchangeability, and any special requirements. Individual item requirements are enumerated.

7.1.5.2 Decoding a manufacturer's data sheet. The most important information of a data sheet is the specification which tells the guaranteed characteristics that will result when the given test conditions are applied to the device. Adjacent to the guaranteed specifications is usually a column labeled "typical" which usually relies on the manufacturer's interpretation. Absolute maximum ratings are given which indicate limits beyond which damage to the device may occur. Neither dc nor ac electrical specifications apply when operating the device beyond its rated operating conditions. Absolute maximum ratings generally include operating conditions such as maximum operating and storage temperatures and maximum supply voltages and over-voltages or currents on various input pins. Characteristic curves and applications sections are usually given but must be used by the application engineer only as a general guide since these are not guaranteed or well documented by the manufacturer.

7.1.5.3 Design precautions. The absolute maximum ratings of the device must be strictly adhered to. The application engineer shall make a thorough worst-case analysis of the device application in terms of thermal management, latch-up possibility with CMOS devices, and ESD exposure. Avoid running logic signal lines near clock lines because of possible crosstalk. Systems that operate at speeds above 30 MHz should have ground planes on the boards. The application section of the device data sheet often gives useful design precautions and should be consulted.

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7.1.5.4 Technology interfaces. Interfacing between different technologies is covered by the manufacturer's general information sections in their data books. The application engineer should carefully follow the manufacturer's specifications but electrical tests should also be done to verify the interfacing design.

7.1.5.5 Noise. The problem of noise can be minimized if ground planes are used on circuit boards containing the subject microcircuits. Ground planes are solid metal sides on the circuit board which have the effect of surrounding the active elements on the board with a noise shield. Extra by-pass capacitors should be used on power lines and in noise sensitive circuit elements. Specific noise types encountered in microcircuits, their sources, and methods of control are as follows:

- a. External noise. These are noises radiated into the system. Sources include circuit breakers, motor brushes, arcing relay contacts, and magnetic-field generators. Methods of control are shielding, grounding, or decoupling.
- b. Power-line noise. These are coupled through the ac or dc power distribution system. The sources and control methods are the same as for external noise.
- c. Cross talk. This is noise induced into signal lines from adjacent signal lines. Control methods are shielding, grounding, decoupling, and if possible, increasing the distance between signal lines.
- d. Signal-current noise. Noise generated in stray impedances throughout the circuit. Control methods are shielding, grounding, decoupling, and reduction of stray capacitance in the circuit.
- e. Transmission-line reflections. Noise from unterminated transmission lines that cause ringing and overshoot. Method of control is to terminate transmission lines.
- f. Supply-current spikes. Noise caused by switching several digital loads simultaneously. Control method is to design systems so that digital loads are not switched simultaneously.

7.1.5.6 Grounding. The applications engineer will not be called upon to design individual microcircuits but rather systems using one or more microcircuits. Generally, a common-ground-plane structure should be used to minimize noise problems. The more closely the chassis and ground can approach to being an integral unit, the better the noise suppression characteristics of the system. For grounds and decoupling on circuit boards, the most desirable arrangement is a double-clad or multilayer board with a solid ground plane or mesh.

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7.1.5.7 Clock signals. Do not run logic signal lines near clock lines because of the possibility of cross talk in either line. This mutual coupling can be reduced by using coaxial cable or shielded twisted pairs. Coaxial cable eliminates all cross talk but because of high drive requirements, it should be used only in the noisiest environments. Twisted pairs are adequate for most applications and are easier to use.

7.1.6 General guides and charts. Generally, less complex devices with a proven performance record in a variety of circuits should be used in preference to more complex parts lacking a reliability history, except where the part without the proven performance record offers substantial improvement in other areas. These rules must be used with caution, however, since parts about to become obsolete usually fit them quite well.

7.1.6.1 Speed versus power by technology. Figure 2 shows the approximate speed versus power dissipation of the logic families listed in MIL-STD-975.

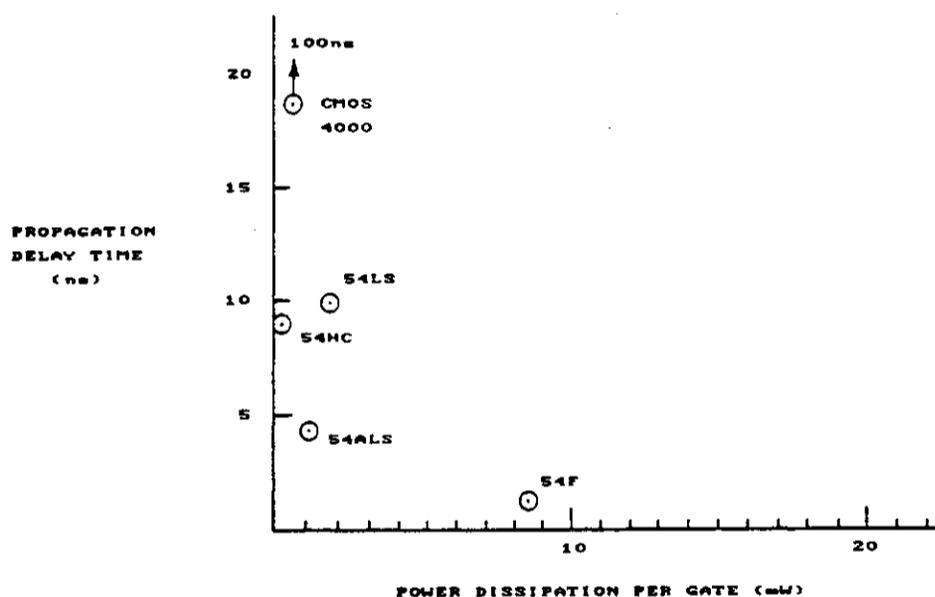


FIGURE 2. Speed versus power dissipation of MIL-STD-975 logic families.
Conditions: $C_L = 15$ pF, 25°C , $V_{CC} = 5$ V, and frequency = 100 KHz.

7.1.6.2 Package versus reliability by technology. Because all microcircuits identified in MIL-STD-975 as Grade 1 have a class S quality military rating, they are of the highest reliability that is currently possible. The class S qualification process ensures that only mature technologies are approved with the highest quality level achievable by the industry. Only the strongest and highest quality manufacturing lines are approved for class S (or Grade 1) production.

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7.1.6.3 Programmable versus non-programmable parts. At this time, the only programmable parts qualified as NASA standard are some programmable memories (PROMs). It is anticipated that in the near future some form of programmable array logic devices will be qualified to Grade 1 requirements. Memory microcircuits are discussed in detail in subsection 7.5.

7.1.6.4 Technological maturity versus availability. The technological maturity and availability of NASA standard microcircuits must meet the requirements of MIL-M-38510 for microcircuits with no exception.

7.1.6.5 Level of integration versus reliability versus cost. Generally higher levels of integration produce an increase in reliability and a concomitant decrease in system cost. A major factor of microcircuit device failures are interconnections and solder joints. Higher integration such as LSI and VLSI decrease the number of interconnection failure possibilities. Because packaging and interconnection hardware account for the major cost factor in microcircuits, higher integration levels also decrease the overall system cost.

7.1.6.6 LSI or VLSI versus hybrid. Table I demonstrates that LSI and VLSI microcircuits have individual device counts in excess of 1000 and 100,00 respectively. In some applications, LSI and VLSI devices may replace hybrid circuits in terms of functionality and reliability. Only other special considerations such as high power requirements, low quantities, or other unusual circuit requirements would make hybrids the choice over LSI and VLSI devices. A major question here could be the availability of manufacturers willing to go through the class S qualification procedure for their LSI and VLSI production lines.

7.1.6.7 Comparison of fan-out for MIL-STD-975 technologies. Schottky transistor-transistor logic devices are functionally compatible with each other, that is, a device from one series may drive or be driven by a device from another series. The input and output voltage levels illustrating this are shown in Table II. Selection of one series over another would be based primarily on the propagation delay, power consumption, and the system requirements. Advanced Schottky and advanced low-power Schottky offer the designer gains in reduction of power consumption and lower propagation delays while maintaining functional capability of the TTL family.

CMOS devices offer moderate propagation delays per gate and very low static power consumption. However, the power consumption increases rapidly with increase in frequency. The devices will tolerate a very wide VDD range, and their interfamily dc driver capabilities are very large.

The interfacing between TTL and MOS devices is voltage compatible. The VDD range of CMOS devices includes the operating VCC range for TTL devices, and interfacing between them can in many cases be accomplished with no buffering.

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CMOS devices have adequate output capabilities to drive approximately 10 LS TTL loads. For devices not capable of driving TTL directly, CMOS buffers are available to perform the CMOS to TTL interface.

The Schottky TTL output high level (V_{OH}) of 2.7 V is not sufficiently high to guarantee recognition as a logic 1 at a CMOS input. A pullup resistor from the TTL output to V_{DD} provides an adequate interface for many applications by pulling the driven CMOS input nearly to V_{DD} and restoring an adequate noise margin.

TABLE II. Comparison of MIL-STD-975 technologies

Characteristic	LS	ALS	AS (or F)	CMOS 4000	CMOS HC
Power supply requirements, Vdc	5	5	5	3 to 18	2 to 6
Input voltage, V high	2.0	2.0	2.0	3.5	3.5
low	0.8	0.8	0.8	1.5	1.0
Output voltage, V high	2.7	2.7	2.7	2.5 to 15	4.9
low	0.4	0.4	0.4	0.1 to 0.5	0.1
Fan-out (LS loads) Standard output	20	20	50	4	10
Minimum output drive, mA ($V_O = 0.4$ V) Standard output	8	8	20	1.6	4
Propagation delay, ns $C_L = 15$ pF	10	4	1.5	105	8
Power dissipation mW, at $f = 100$ kHz	2	1	8.5	0.1	0.17
Worst-case noise margin	0.4	0.4	0.4	1 to 2.5	0.9

7.1.6.8 Circuit design cycle. Because of the uniqueness of the LSI and VLSI design and development cycles vs off-the-shelf SSI/MSI microcircuits there are several alternate approaches. For low-volume orders (500 to 5000 per year) it is more advantageous to buy standard off-the-shelf items (ROMs, RAMs) and to utilize programmed arrays. Computer aided design arrays can be "programmed" either by mask adaptation techniques (ROM) or by electrically opening certain gates (PROM).

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While many working relationships between the manufacturer and the user are possible, there are four primary types:

a. Performance specification:

The user provides a system design and generates specifications. The LSI manufacturer then performs the logic design for minimum cost implementation. This may involve the use of one or more arrays, MOS standard products, interface circuits, multichip packages, etc.

b. Logic diagram only:

The user provides a logic diagram to the manufacturer, who analyzes the design, generates test criteria, and performs computer simulation. The user then analyzes computer simulation data and verifies the design.

c. Logic diagram and functional test criteria:

The user provides logic diagrams and functional test data. The logic is simulated and verified with test criteria. This approach requires less manufacturer involvement than does (b) above.

d. CAD interface:

If available, the manufacturer's computer simulation program provides a compact language for specification of arrays and associated test criteria. The user may elect to supply design data to the manufacturer via their means of computer communication. The logic simulation generated with this program eliminates the need for breadboarding. Breadboarding with discrete MOS is often impractical because the parasitic capacitance between logic functions is almost eliminated when functions are subsequently integrated.

The foundation of the CAD is a computerized library of standardized logic circuits, usually called standard cells. This library comprises from 30-200 cells and includes all of the standard logic functions that are normally required to produce complex digital hardware, such as binaries, shift registers, gates, inverters, buffers, etc. New cells are added to the library as the need for them arises. The dimensions of the cells vary according to their capacity.

The standard cells are designed in accord with the manufacturer's own rules and may not be compatible with the production techniques of other LSI semiconductor manufacturers. With respect to such factors as the spacing of p-material to p-material, the diffusion depth of the dopant, the metal-to-metal spacing, etc., glass master plates are suitable for use by only those manufacturers with compatible design rules.

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The benefits of fast turnaround time and systematic design procedures would be negated if the standardization required to realize them resulted in such inefficient area usage that subsystems of true LSI complexity could not be economically integrated. Several factors ensure that area efficiency will be maintained in the development of custom arrays. First, there is a wide variety of standard functional blocks so that the various internal functions of a custom array are concisely implemented. Moreover, each of these cells has been topologically optimized. The extent of the library ensures sufficient flexibility to let the array family be useful in widely differing types of applications. Second, the cell patterns are not prediffused, and the location of each functional block is selected during the design phase to minimize the total length and physical complexity of the interconnection system. Finally, both the cell patterns and the interconnection system take full advantage of current technology to conserve surface area.

7.1.7 General reliability considerations. High potential reliability is one of the chief inducements to the use of microcircuits. Precise statements of reliability in terms of failure rates are more difficult to obtain for these devices than most other components. The complexity of individual devices, the wide variety of device types and the rapid change of the state of the art tend to make testing complicated and expensive. Differences from manufacturer to manufacturer and from lot to lot for a given manufacturer for devices of a given type may be more significant than the variables controlled. On long life tests with numerous measurements, handling problems, such as physical damage or inferior contact to one or more external leads, can greatly complicate interpretation of test results. For reliability estimates in numerical form, publications of Rome Air Development Center (such as Microelectronic Failure Rates) are probably the best source.

7.1.7.1 Microcircuit reliability. Microcircuits with NASA Grade 1 quality level must meet class S quality and reliability requirements specified in paragraph 3.4 and 4.0 of MIL-M-38510 and as additionally referenced in MIL-STD-883. Microcircuits with NASA Grade 2 quality level must meet class B requirements of the same documents.

7.1.7.1.1 Screening. Screening procedures are a group of tests imposed on a microcircuit production lot to assist in achieving the desired levels of quality and reliability for the device commensurate with the intended application. Method 5004 of MIL-STD-883 and the general requirements of MIL-M-38510 establish the screening procedures for class S and for class B microcircuits. Microcircuit devices satisfying Grade 1 NASA requirements must meet class S screening requirements of these documents. Grade 2 NASA requirements are satisfied by class B screening requirements of the same documents.

Screening begins with the wafer lot acceptance tests of Method 5007 of MIL-STD-883. Although it is not possible to test several key parameters (mostly ac), development and selection of good correlation testing is essential to maintain high yields later on when the device is assembled. On larger arrays (especially LSI circuits) special wafer or die test transistors or patterns are prerequisites.

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As with wafer electrical testing, wafer visual inspection is a cost efficient screen. This is primarily because little time has been invested so far on a per die basis, and add-on assembly parts (e.g., package, wire, etc.) are not yet involved. Thus tightening visual criteria at this point is an effective way to reduce failure rates experienced later on. Included in the wafer level test is a scanning electron microscope (SEM) inspection of quality and acceptability of the wafer metallization.

Precap visual inspection is usually the next screening performed on monolithic microcircuits. This consists of a visual examination of the component after it has been attached to the package and wired up. Depending on the magnification power used, this inspection examines the assembled parts, checks for gross die defects caused during assembly (e.g., metal scratched open), or takes a final look at the die for fabrication defects that might have been overlooked previously. The precap visual inspection is normally the last screening step before the sealing of the device.

Hybrid microcircuits also have a precap inspection but usually go through additional other intermediate visual and electrical screens where portions of the hybrid microcircuit are tested. This is primarily because of the number and variety of components assembled in the package. For example, a hybrid microcircuit of average complexity might have ten transistors, ten diodes, and ten integrated circuits along with thick- or thin-film circuitry. Thus waiting until precap visual inspection to make repairs or sort out rejects becomes extremely expensive.

Temperature storage is designed to accelerate any surface chemical or inter-metallic reaction in the device which might ultimately cause failure. Generally this reaction is a result of contaminants remaining on the die surface which have escaped previous visual inspections. Although metallization patterns are usually the target of such reactions, contamination of insulator materials used in some processes can result in device failure by parametric drift or by catastrophic failure.

Temperature cycling is intended to alternately stress the part under test to its rated temperature extremes. This is a particularly effective screen for LSI and discrete resistors and capacitors in hybrid circuits. Due to the relatively large sizes of the packages and die or chip, the temperature excursions point up any mismatch in thermal coefficients of the materials used to produce the device. Marginal defects such as hairline cracks in the metallized leads, metal-to-silicon contacts, peeling metal, pin holes, etc., will usually be detected by this screen also.

Hermetic seal tests for semiconductor devices are well known and commonly used in the semiconductor industry. They are primarily designed to assure package integrity. The only special consideration for LSI and hybrid microcircuits relates to leak criteria for fine-leak tests as a function of the generally larger cavities of these packages; the hermetic-leak-rate criteria is usually set higher for the larger packages.

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Centrifuge (constant acceleration) is designed to test the mechanical aspects of devices and packages, including the bond strengths of the attachment and the wire bonds. These have special significance in hybrid packages where there are large numbers of die and wire bonds.

Usually, intermediate electrical tests are conducted at this point. The main intentions are to detect parameter degradations caused by the previous environmental stresses and to act as reference to burn-in. Devices usually have external leads shorted during assembly, to provide rigidity to the leads. Thus, the devices cannot be electrically tested until the shorting connections are removed.

High-temperature burn-in (electrical functioning or biased devices at temperature--usually 125 °C--for an extended period of time--usually 160 to 240 hours) has long been recognized as an effective screen for microcircuits. Its primary purpose is to detect infant failures, and in this role it will significantly increase the reliability of any product for which it is specified. High-temperature reverse bias burn-in (HTRB) is a version of burn-in that is gaining more widespread acceptance especially in MOS circuits. This is due to the fact that HTRB is especially effective in screening out ion (surface) related defects.

The purpose of X-ray is to detect foreign objects or materials that have inadvertently been sealed within the package. X-rays also show the integrity of the glass seal.

External visual inspection is the last screen of the sequence. This is a final inspection for external package or lead defects.

A failure analysis program, with provision to feed back information to the responsible sources for corrective action, is an important part of a good reliability program. Failures from production (in the case of a manufacturer), testing and field usage are examined intensively to determine the failure modes and, if possible, their causes.

Failure modes of microcircuits are generally the same as those of other semiconductors except that their greater complexity causes some modes to be more of a problem than for discrete devices. Bulk defects and masking faults are more likely to cause failures because of the small sizes and greater numbers of semiconductor junctions. Aluminum migration is more likely because of the necessarily narrow interconnection runs. The smaller bonding areas generally available make lead bonding more critical. The greater number of external leads and the use of package types which may sacrifice mechanical strength for compactness make these devices more prone to hermetic seal failures. Lead bond failures and masking defects are the leading failure modes for microcircuits.

7.1.7.1.2 Handling of electrostatic discharge sensitive (ESDS) devices. All microcircuits can be damaged or destroyed by high-voltage pulses. MOS devices are especially susceptible to overstress because the thin gate oxide can be ruptured easily. To minimize the electrical overstress of MOS, as well as other ESDS devices, special precautions should be employed when handling these micro-

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circuits (e.g., at inspection, test, or assembly). DOD-HDBK-263, Electrostatic Discharge Control Handbook, provides useful information for developing, implementing, and monitoring an ESD control program for all aspects of ESD phenomena and should be consulted and adhered to.

The following is a list of recommendations for handling these components on an individual basis.

- a. Require all devices to be shipped from the sources in containers that clearly identify them as ESDS (ESD Sensitive) devices.
- b. Do not separate the devices from the containers before their use.
- c. Incoming inspection and electrical testing should be conducted only in a special ESD controlled facility.
- d. Follow precisely the recommendations of the device manufacturer.
- e. Minimize the number of people handling these devices in all phases of manufacturing.
- f. Train people by explaining the phenomena of static electricity and how it can damage or destroy these devices.
- g. Minimize the use of nonconductive plastics for shipping containers.
- h. If the use of nonconductive plastic containers cannot be eliminated, then line the inside of the containers with a conductive foil such as aluminum foil.
- i. When handling the devices, use cotton gloves (not nylon or rubber).
- j. Use a manufacturer-supplied shorting mechanism wherever possible.
- k. Do not pile or stack the devices.
- l. If the devices are to be removed from carriers, wrap them in aluminum foil.
- m. Label containers and transportation boxes "FOR ESDS DEVICES ONLY."
- n. If it is necessary to handle the devices, avoid touching the leads. That is, handle by the case only.

The following is a list of recommendations for handling ESDS device assemblies:

- a. Before touching the devices with hands, tools or probes, momentarily touch a grounding plate.

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- b. Devices must never be plugged into test equipment or next higher assemblies with power applied.

The following is a list of recommendations for a facility to assemble and test ESDS devices and their assemblies:

- a. Nonconductive plastics should not be used in the area with the exception of aluminum-foil-covered tote trays.
- b. All cabinets, benches and equipment must be properly grounded.
- c. All soldering irons must be permanently grounded prior to soldering any part of the assembly.
- d. The soldering iron tips must be checked on the first day of each work week for resistance. When the resistance exceeds 5 Ω , then, it should be replaced.
- e. Planar joining equipment should be checked for its potential to ground. If the potential exceeds 10 V then isolate the equipment from the line and install a ground strap to the head.
- f. The area should be designated as "RESTRICTED TO AUTHORIZED PERSONNEL ONLY" and enforced accordingly.
- g. The number of people authorized to handle ESDS devices and assemblies will be an absolute minimum.
- h. Avoid the use of smocks made of material that is susceptible to static electricity. Smock material such as nylon and dacron should not be used.
- i. Maintain the relative humidity at around 50% percent.

7.1.7.1.3 Microcircuit failure rate model.

Failure rates. Monolithic microcircuits delineated within MIL-HDBK-217 are described by the failure rate model:

$$\lambda_p = \pi_Q [C_1 \pi_T \pi_V \pi_{PT} + (C_2 + C_3) \pi_E] \pi_L \text{ failures}/10^6 \text{ hours}$$

where λ_p is the overall microcircuit failure rate. Of the factors in the model, π_Q , π_E , and π_L normally are fixed, reflecting the reliability and application requirements of the individual program. Programming technique factor (π_{PT}) is applicable only for PROMs and ROMs. The remaining factors offer measurable degrees of freedom for reducing the estimated failure rates for a given application. Some in the model will be addressed as to how and to what extent each contributes to a device's estimated failure rate. For a complete discussion, reference should be made to MIL-HDBK-217.

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Learning factor (π_L) is 1 for NASA programs where device selections are made per MIL-STD-975 or are otherwise controlled. Quality factor (π_Q) deals with the level of screening and procurement. Prime equipment contracts for NASA require use of microcircuits procured in full accordance with MIL-M-38510, class S or class B, and have quality factors respectively of 0.5 and 1. A class B device would have a failure rate twice that of a class S device. But if relaxed screening levels were allowed the quality factors become progressively larger. For commercial plastic encapsulated devices it is 35 and the resultant failure rate would be 70 times larger than for class S devices.

Temperature acceleration factor (π_T) is dependent on the device technology, actual operating junction temperature, and package type and has a value of .1 for junction temperatures of 25 °C for all monolithic microcircuits for all package types. The following π_T factors are for junction temperatures of 125 °C listed by technology represented by MIL-STD-975:

<u>Technology</u>	<u>π_T for Package Type</u>	
	<u>Hermetic</u>	<u>Nonhermetic</u>
TTL, DTL, ECL	5.0	8.1
LTTL & STTL	8.1	13.0
LSTTL, PMOS	13.0	22.0
MNOS, IIL	35.0	248.0
NMOS, CCD	22.0	248.0
CMOS & LINEAR	57.0	659.0

Note: The nonhermetic listing is included for reference only.

Temperature acceleration factor is based on junction temperature which is determined by the following:

$$T_J = T_C + \theta_{JC} P_{MAX}$$

where

T_C is the case temperature

θ_{JC} is the package thermal impedance

P_{MAX} is the device's maximum power dissipation

Devices of technologies dissipating low power will have lower junction temperatures and π_T factors. Additionally, by controlling the junction temperature through the use of heat sinks, air flow, board layout, etc., and by prudent selection of devices, a moderate factor can be achieved.

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Voltage derating stress factor (π_V) is 1.0 for all technologies other than CMOS and for CMOS with maximum recommended operating V_{DD} of 12 V. CMOS devices with worst-case junction temperature of 175°C and operating at 90 percent of the maximum rated supply voltage (V_{DD}) of 15, 18, and 20 V would have a voltage derating stress factor respectively of 3.3, 1.8, and 2.6. However, if the operating V_{DD} was reduced to the 5- to 6-V range, the π_V stress factor, each case, would be less than 1. (Reference Table 5.1.2.7-14, MIL-HDBK-217)

7.1.7.2 Hybrid reliability. MIL-STD-975 does not specifically give detailed requirements for hybrid microcircuits. However, hybrid microcircuits should conform to NASA quality requirements of Grade 1 and Grade 2 microcircuits by satisfying class S and class B quality and reliability requirements of paragraphs 3.4 and 4.0 of MIL-M-38510. General quality and reliability requirements for custom hybrids are detailed in Appendix G of MIL-M-38510.

7.1.7.2.1 Screening. Method 5008 of MIL-STD-883 and the general requirements of MIL-M-38510 establish the screening procedures for class S and class B hybrids and should be satisfied for hybrids used in NASA programs.

7.1.7.2.2 ESDS devices-hybrids. All hybrids should be considered as electrostatic discharge sensitive and the same handling procedures apply as for monolithic microcircuits discussed in paragraph 7.1.7.1.2.

7.1.7.2.3 Hybrid failure rate model. According to MIL-HDBK-217, the hybrid failure rate model is:

$$\lambda_p = \{ \sum N_C \lambda_C \pi_C + [N_R \lambda_R + \sum N_I \lambda_I + \lambda_S] \pi_F \pi_E \} \pi_Q \pi_D \text{ failures}/10^6 \text{ hours}$$

where λ_p is the hybrid microcircuit failure rate and the symbols are defined in the reference handbook.

Design and layout of the hybrid microcircuit, including part selection, derating criteria, and package selection are important areas in controlling the hybrid failure rate.

If the hybrid package encloses more than one substrate, each substrate should be treated as a separate hybrid. Each substrate shall include its own density and function factor and only the largest area substrate or the substrate mounted on or serving as the package header shall include a package factor. The net hybrid failure rate will be the sum of the failure rates for the individual substrates.

7.1.7.3 System design choices. After the basic architecture of a system design is set, the next basic steps are logic design, circuit design, memory design, software, and peripheral equipment. In all of these areas, the application engineer should use only mature proven designs, processes, and devices. For example, circuit implementation should only use Grade 1 or Grade 2 type microcircuit devices.

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7.1.7.4 Device choices affecting part and system reliability. The NASA application engineer should only choose microcircuits listed in MIL-STD-975 as these will be of mature designs and proven reliability.

7.1.7.5 Redundancy and simplicity. Two proven design techniques to ensure reliability are redundancy and simplicity of design. Redundancy should be incorporated both at the system level and at appropriate circuit design levels. Design simplicity, if possible, is a much practiced reliability enhancing technique.

7.1.7.6 Worst-case analysis. A worst-case analysis of a microcircuit application should be done to ensure reliability under all operating conditions. Worst-case constraints that should be considered are power supply extremes, worst-case temperature depending on circuit use, noise, and any other applicable circuit conditions. A careful study and judgement should be performed to determine which parameter extreme is actually the worst-case mode. It is possible that a minimum value is worst-case for certain conditions and a maximum value is worst-case for other conditions. In such cases, both parameter extremes should be analyzed.

7.1.7.7 Radiation considerations. When microcircuits are exposed to sufficient space radiation, changes can be observed in some electrical parameters and/or functional operation. These are due to two types of radiation damage effects induced by charged particles in passing through the device, namely total dose effect and single event phenomena. The total dose effect is due to the cumulative effect of all the ionizing particles present in the space environment and is uniform over the device. The basic damage mechanism is the result of the generation of e-h pairs by ionizing particles. Both electrons and holes, so created, drift under the influence of local electric fields. However, the electrons are more mobile than holes, so that a net build up of holes occurs in the silicon dioxide at the silicon-silicon dioxide interface. This causes a change in the threshold voltage of MOS transistors making it difficult for N channels to "turn off" and p-channels to "turn on". In the bipolar devices, the charge deposited near the emitter-base junctions results in an enhanced recombination of minority carriers in the base with a resultant decrease in current gain. In addition, in some technologies, interface states are introduced in the silicon at the silicon-silicon dioxide interface. These too can modify the overall charge dependent properties of MOS devices causing an increase in propagation delay times, and changes in threshold voltages. The interface state generation acts in opposition to the effects of hole trapping, tending to compensate for threshold shift, and can even overcompensate in some (newer) device types.

Single event upset is a localized effect which occurs when a single heavy ion of high energy causes a false signal in semiconductor memory devices, microprocessors, etc. This effect occurs when a single charged particle can cause sufficient ionization to supply the "critical charge" needed for the flipping of the memory state from '1' to '0' or vice-versa. This type of error is called a "soft error" as it causes no permanent damage in the cell and the cell can be reprogrammed for correct functioning. However, in addition to the soft errors, the single heavy ions can cause "latch-up" in some technologies, such as bulk CMOS. Latch-up results in a massive number of bit errors on the device and in

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the device drawing excessive supply current. If not limited, this excessive current can result in device destruction. Functionality of the device can be restored by power cycling.

In any given microcircuit, the magnitude and characteristics of radiation damage is highly dependent on the spacecraft's orbital parameters which dictate the type of the radiation environment, its intensity and energy spectrum as well as its time history. Other factors that must be considered are the electronic system's materials, circuit application, device biasing conditions, and spacecraft shielding.

Radiation units. Radiation is usually measured and expressed in terms of absorbed energy per second (dose rate), absorbed energy (dose), particles per square centimeter per second (flux), or particles per square centimeter (fluence). Fluence may be considered as the time integrated flux. For ionization effects the most commonly used unit is absorbed dose expressed as rad, (Roentgen absorbed dose) which is defined as 100 ergs absorbed from the radiation field per gram of irradiated material. In defining an absorbed dose, the material being irradiated must be defined. For example, 234 rads deposited in silicon would be expressed as 234 rads (silicon). For displacement effects the radiation by particles such as neutrons, protons, and electrons is generally measured in fluence.

7.1.7.7.1 Radiation environment. A satellite or deep space probe must contend with a range of radiation environments that contain potentially high levels of radiation and the attendant damage from natural and possibly manmade sources. In this latter category is any radiation stemming from radioactive sources on the spacecraft such as radioisotope heater units and radioisotope thermoelectric generators. In addition, there may be radiation environments resulting from a nuclear weapon, but these environments and their effects are not a constraint on operation of NASA spacecraft and will not be considered here. The natural environment must be considered for all spacecraft.

7.1.7.7.2 Natural environment. There are three sources of natural radiation an earth orbiting satellite will encounter: the Van Allen trapped radiation belts, solar radiation, and galactic cosmic radiation. Deep space probes must also face these or similar sources associated with other planets.

- a. Van Allen radiation belts. The earth's magnetic field traps incoming charged particles giving rise to radiation belts surrounding the planet. An incoming charged particle with insufficient energy to escape this field will go into a helical type path around the earth's magnetic field lines. The exact motion will depend on the initial conditions of entry, the energy, and the angle with respect to the magnetic field of the incident particle. Subsequent collisions will also affect this motion. The makeup of the Van Allen belts exhibit spatial and temporal variations. Their composition is in a large part

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due to solar activity and consists of electrons, protons, alpha particles and some cosmic rays. Thus, the radiation levels a spacecraft will be subjected to from the Van Allen belts must be calculated for each specific spacecraft mission based on the exact orbit, mission life, and assumptions regarding expected solar activity.

Figure 3 provides a graphic representation of the Van Allen belts for protons and electrons. These curves are given only for illustrative purposes. Because the composition of the belts varies, the flux density numbers on the graphs should only be used for comparison of the zone or proton to electron fluxes. Where more quantitative work is required the latest model of this environment must be used.

- b. Solar radiation. Solar radiation is usually broken down into two categories, solar wind and solar flare activity. The solar wind is a plasma of protons, electrons, and alpha particles that is continuously emitted from the sun. The particles travel radially outward. Near earth measurements indicate particle velocities ranging between 350 to 700 km/s. At the high end this corresponds to an energy of less than 3 keV for protons. Although solar wind particle fluxes can be quite large, of the order $10^9/\text{cm}^2\text{-s}$, their low energies mean little penetration, hence the only damage is to directly exposed surface material.

During solar flare activity large numbers of protons, electrons, and alpha particles are emitted with some more massive particles. The protons emitted during solar flare activity can have energies of many million electron volts. Thus, solar flare activity can contribute significantly to the radiation levels a spacecraft can encounter. Solar flare activity is not a consistent phenomenon. Some assumptions as to the expected level of solar flare occurrences must be made when predicting the radiation levels to which a spacecraft will be subjected.

- c. Galactic cosmic rays. Galactic cosmic rays consist mainly of protons with an energy range of many orders of magnitude up to billions of electron volts. Cosmic ray fluxes of carbon, oxygen, or iron nuclei groups are several orders of magnitude less than proton fluxes. Cosmic ray interception by microcircuits is a source of single event upsets (SEU) in the affected microcircuit.

The total dose contribution a satellite will encounter from cosmic ray radiation is typically small due to the small fluxes. Cosmic ray fluxes are isotropic in galactic space.

- d. Packaging and circuit material. The circuit function of microcircuits can be affected by alpha particles emitted by the decay of trace radioactive elements such as uranium and thorium that may be present in the ceramic packaging materials. The effect is mainly a localized charge

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packet generation which can charge or discharge a circuit node causing a circuit malfunction. These elements may also be present in materials actually used in circuit construction such as the metallization. Coatings on the surface of the microcircuit die can provide effective shielding against alpha particles because of the limited range of the particles.

7.1.7.7.3 Manmade environment. The manmade or nuclear radiation environment can differ substantially from the natural environment. The radiation from onboard radioactive sources depends on the nature of the source. Radioisotopes used for calibration could be alpha, beta, and gamma emitters. Typically, they would be very low level sources and of little consequence. In the case of radioisotope heater units or radioisotope thermoelectric generators, the radiation consists mainly of gamma rays and neutrons. They would be found only on deep space probes where distances from the sun would make solar cells ineffective thus requiring auxiliary power sources. Although the radiation levels are typically low, especially for radioisotope heater units, over a long duration mission into deep space they could become significant.

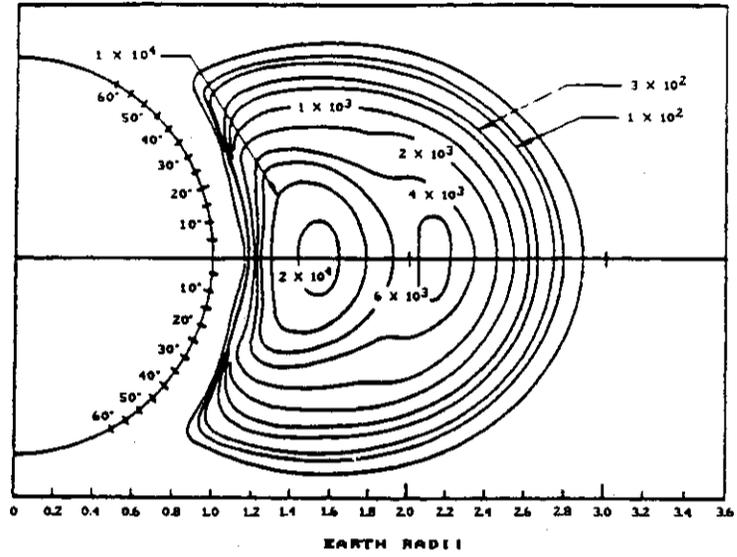
7.1.7.7.4 General usage considerations. Radiation test experience shows that radiation damage effects can vary widely and unpredictably among semiconductor part types, manufacturers of the same part type, and different lots of the same part type produced by the same manufacturer. Nevertheless, general trends have been observed for microcircuit parts which should be considered during the design of electronic units that must take radiation damage into account. Also, it should be noted that the effect of on-going annealing in devices may lead to some recovery of the radiation damage. The extent of recovery can vary from little or no recovery to substantial recovery of induced radiation damage. Some of the newer technologies have also exhibited "rebound" effect, where the recovery on long term annealing exceeds the induced radiation damage.

The following items should be considered by the application engineer when using microcircuits in radiation environments.

MOS microcircuits.

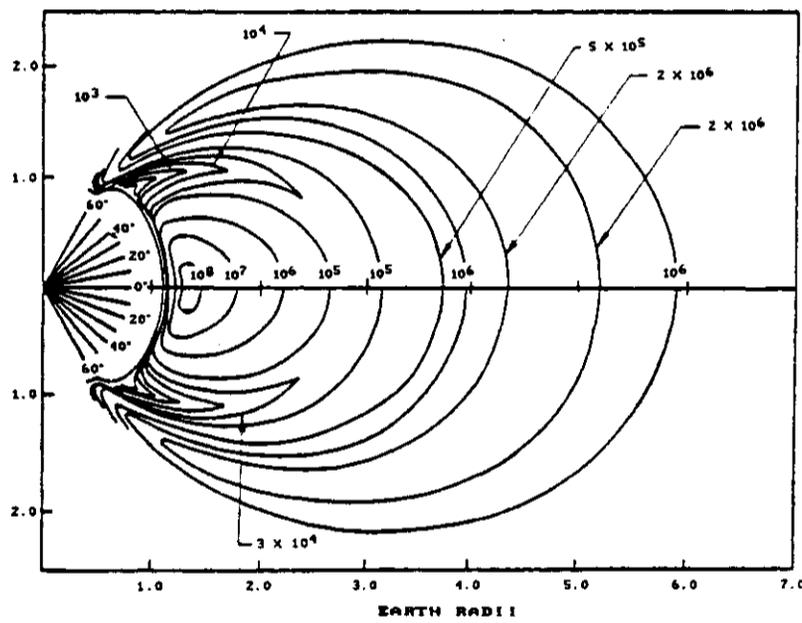
- a. Threshold voltages of n-channel devices may increase or decrease depending on dose level and dose rate.
- b. Threshold voltages of p-channel devices increase with absorbed radiation dose.
- c. Quiescent device current increases with absorbed radiation dose.
- d. Propagation delay and transition times increase with absorbed radiation dose.
- e. Annealing effects may be significant.

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A. PROTON ISOFLUX CONTOURS ($E > 34$ MeV)
CONTOURS ARE LABELED IN UNITS OF PROTONS/ cm^2 -SEC

A. Proton isoflux contours ($E > 34$ MeV). Contours are labeled in units of protons/ cm^2 -s



B. TRAPPED ELECTRON ISOFLUX CONTOURS ($E > 0.5$ MeV)
CONTOURS ARE LABELED IN UNITS OF ELECTRONS/ cm^2 -SEC

B. Trapped electron isoflux contours ($E > 0.5$ MeV). Contours are labeled in units of electrons/ cm^2 -s

FIGURE 3. Van Allen radiation belts for protons and electrons.

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Operational amplifier microcircuits.

- a. Input offset voltage increases with absorbed radiation dose.
- b. Input offset current increases with absorbed radiation dose.
- c. Input bias current increases with absorbed radiation dose.
- d. The dc open loop gain decreases with absorbed radiation dose.

Bipolar TTL digital microcircuits.

- a. Propagation delay times increase with absorbed radiation dose.
- b. Leakage currents increase with absorbed radiation dose.

In circuit designs the above radiation effects on microcircuits should be considered and design guidelines established to assure device functionality and operation within parametric limit values after exposure to radiation environments. These design guidelines should include, but not necessarily be limited to, the following design implementation:

Resistive current limiting to preclude device destruction under latchup;
power cycling to allow restoration of functionality

Derating of electrical parameters

Shielding

In all applications where radiation is expected, a radiation specialist should be consulted for current radiation hardness data, latest data on expected radiation, and methods of minimizing radiation effects on microcircuits. Also, the indirect effects of radiation damage on other system components, such as capability of power supplies to provide increased current drive to compensate for radiation induced degradation in electronic parts, should be considered.

7.1.7.7.5 Radiation summary. Although there can be considerable variation in the radiation tolerance of different parts of the same type, some device types and technologies are more resistant to radiation damage than others. Digital bipolar microcircuits are generally more radiation tolerant to total dose than linear bipolar microcircuits which, in turn, are generally harder than NMOS microcircuits.

Table III gives an approximate comparison of the susceptibilities of different technologies to total dose effect and single event phenomena.

The application engineer concerned with radiation should consult with a radiation specialist concerning current radiation hardness data for the device or technology family being considered and the planned radiation environment.

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TABLE III. Comparison of Radiation Susceptibility for Microcircuits of Different Technologies

Technology	Total Dose Hardness Level Rads (Si) (Note 1)	Relative Susceptibility To: (Note 2)	
		Soft Error	Latch-Up
<u>DIGITAL</u>			
NMOS	$5 \times 10^2 - 10^4$	High	Immune
CMOS/Bulk (unhardened)	$10^3 - 10^5$	Moderate to high	Moderate
CMOS/Bulk (hardened)	$2 \times 10^3 - 10^6$	Low	Low
CMOS/SUS	$10^3 - 10^5$	Very Low	Immune
TTL, Low Power TTL	$10^5 - 10^7$	Low to High	Low
Schottky TTL, Low Power Schottky TTL	$10^5 - 10^7$	Low to High	None to Low
Advanced Low Power Schottky TTL	$2 \times 10^4 - 10^6$	Moderate	Low
I ² L	$2 \times 10^4 - 10^6$	Moderate	None to Low
ECL	$\geq 5 \times 10^6$	Low	None to Low
<u>LINEAR</u>			
CMOS (unhardened)	$10^3 - 10^5$	- No Data Available -	
CMOS (hardened)	$3 \times 10^3 - 10^6$	- No Data Available -	
Bipolar, BI-FET	$6 \times 10^3 - 10^7$	- No Data Available -	

Notes:

1. These figures define process averages. However, some devices may not meet these levels while others may exceed them. For example, some Schottky TTL RAM's fail much below the lower limit listed in the Table while most other devices with this technology fall within the range shown.
2. The single event susceptibility "ratings" listed here are relative to each other. However, a "moderate" error rate in a specific application may be unacceptably high if the application is critical. Also, circuit organization and/or use of error detection and correction can considerably "harden" soft parts in some applications.

**7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY
TRANSISTOR-TRANSISTOR LOGIC**

7.2 Low-power Schottky transistor-transistor logic.

7.2.1 Introduction. For applications where the high speed of standard transistor-transistor logic (TTL) or Schottky transistor-transistor logic (STTL) is not needed, the low-power Schottky transistor-transistor logic (LSTTL) family is more applicable. With regard to nomenclature, the abbreviations LS and LSTTL both represent low-power Schottky transistor-transistor logic.

The LS series has a power dissipation of approximately 2 to 20 mW per gate and an approximate gate delay of 10 to 20 ns. Compared with Schottky TTL, LSTTL processing produces more shallow diffusions and smaller transistors with greater bandwidth. The LSTTL circuits operate at about one-half the speed of STTL while using only about 20 percent as much power. The lower power dissipation of the LSTTL is achieved by using larger resistor values than the STTL series, which also causes longer delay times due to larger RC time constants.

The lower power needs of the LS series require smaller size and weight of power supplies, which in turn, results in lower heat dissipation. Lower heat dissipation means greater device density on one printed-circuit board and reduction in size, or elimination of cooling fans. Lower power dissipation also allows a greater density of components within an integrated circuit of a given size without exceeding the power dissipation capabilities of the integrated circuit package. It also results in lower junction temperatures for the transistors and diodes, which increase the reliability of the device. The LS series switches in approximately 25 percent less current than the standard TTL series, resulting in lower amplitude current spikes; i.e., less internal noise generation. Consequently, fewer and lower-valued decoupling capacitors are required.

The low input current requirements of the LS series make it an ideal interface between TTL-compatible MOS devices and TTL. The high output drive current allows the LS series to drive capacitive loads, other TTL series, and CMOS logic devices.

7.2.2 Usual applications. LSTTL microcircuits are used where moderate speeds and low-power dissipation are required in logic functions. Some general guidelines will be presented here; however, the application engineer should refer to device data sheets for specific guidance.

7.2.2.1 Available functional types. MIL-STD-975 (NASA), NASA Standard Electrical, Electronic, and Electromechanical Parts List, presents the available LSTTL part types that must be used for NASA microcircuit applications. This standard should be consulted for part selection. See subsection 7.1 General for discussion of Grades 1 and 2.

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**7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY
TRANSISTOR-TRANSISTOR LOGIC**

At the present time, STTL is available to the NASA microcircuit applications engineer in three families; low-power Schottky (LS), advanced low-power Schottky (ALS), and advanced Schottky (AS).

Table IV presents the available functions for the LS family that shall be used for NASA programs. This table summarizes information found in MIL-STD-975.

TABLE IV. LS functions available in MIL-STD-975

Gates	AND, NAND, OR, NOR, INV
Buffers	NAND, NOR, INV
Drivers	Bus
Counters	Binary, decade, dividers
Flip-flops	D, J-K, Monostable multivibrators
Latches	Bistable, addressable, S-R
Shift registers	Parallel, bidirectional, asynchronous
Adders/comparators	Binary, magnitude
Decoders	BCD-to-decimal; BCD-to-seven segment, line
Encoders	8-to-3 priority
Multiplexers	2-, 4-, 8-input selector/multiplexers
Transceivers	Bus

7.2.2.2 Comparative attributes specific to LSTTL microcircuits. Four general attributes will be described that the microcircuit applications engineer must be aware of when designing with LS or any type of digital microcircuits for NASA applications. The designer should refer to the specific microcircuit data sheets for exact information.

Attributes specific to all digital microcircuits including LSTTL are:

- a. Interfacing of one family type to another
- b. Noise immunity
- c. Power dissipation
- d. Switching and propagation times.

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**7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY
TRANSISTOR-TRANSISTOR LOGIC**

A general discussion of these attributes will be presented in this section. A more detailed discussion will be presented in paragraph 7.2.5 Electrical characteristics.

For proper interfacing, the following four parameters must be considered when the output terminal of one device is connected to the input terminal of another:

- V_{IH} = minimum HIGH input voltage
- V_{IL} = maximum LOW input voltage
- V_{OH} = minimum HIGH output voltage
- V_{OL} = maximum LOW output voltage

The devices will interface properly if the inequalities

$$V_{IH} < V_{OH} \text{ and } V_{OL} < V_{IL}$$

are satisfied. The inequalities must remain satisfied for all expected temperatures and worst-case combinations of input and output currents. Table V presents worst-case values of these interface parameters for the LSTTL family. Specific data should be obtained from device data sheets. Additionally, drive capability or fan-out must be satisfied. This is discussed in paragraph 7.2.5 Electrical characteristics.

TABLE V. Worst-case values of primary interfacing parameters for LS logic devices

Parameter	LS value
V _{IH} min	2 V
V _{IL} max	0.8
V _{OH} min	2.7
V _{OL} max	0.4
High-level noise margin V _{OH} - V _{IH}	0.7
Low level noise margin V _{IL} - V _{OL}	0.4

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Noise immunity or noise margin is a parameter describing the allowable noise voltage on the input terminals of the digital device. Noise immunity is specified using two parameters. The low dc noise immunity V_{NIL} is the difference in magnitude between the maximum low output voltage of the driving device and the maximum input low voltage recognized by the receiving device:

$$V_{NIL} = |V_{IL,max} - V_{OL,max}|$$

The high dc noise immunity or noise margin V_{NIH} is the difference in magnitude between the minimum high output voltage of the driving device and the minimum high input voltage recognized by the receiving device:

$$V_{NIH} = |V_{OH,min} - V_{IH,min}|$$

The above parameters V_{NIL} and V_{NIH} are dc noise margins. In addition, ac noise and transients are always present on signal lines between microcircuits. Noise margins for ac conditions are not provided on data sheets because they depend on the speed of the device and capacitance to ground for the input considered. Careful board layout and good bypassing will minimize this noise. Table V also presents a comparison of high level and low level noise margins for the LS microcircuit family.

Power dissipation is usually given in LS data sheets as power dissipated per gate as a function of frequency. In general, all logic devices show an increase in power consumption as operating frequency increases. Individual device data sheets should be consulted for specifications because the power dissipation varies widely with frequency and load.

The most common switching parameter usually given in LS and other logic data sheets is the propagation delay which is defined as the time required for an output to change state after an input has changed state. The measurement is usually taken at 50 percent amplitude (see Figure 4). Usually the delay of changing states in both directions is given. The two propagation delay parameters are:

- t_{PLH} = delay for output changing from low to high,
- t_{PHL} = delay for output changing from high to low.

The propagation delay times vary with device temperature as well as resistance and capacitance values of the driven load, but generally, LS devices have an approximate range of delay times from 10 to 20 ns. Individual data sheets should be consulted for exact specifications.

7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

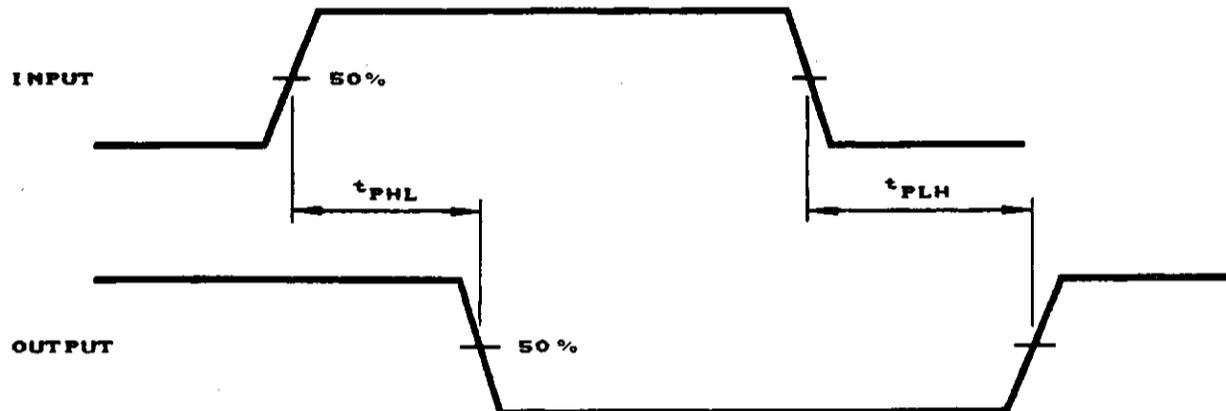


FIGURE 4. Definition of propagation delay as applied to an inverting buffer waveform.

A second common switching parameter for digital microcircuits is the maximum toggle frequency or maximum clock frequency (f_{max}). This parameter applies only to devices with storage elements such as flip-flops, registers, counters, and latches. The value of the maximum toggle frequency for LS devices varies from device type to device type, but representative worst-case values are 30 MHz for a flip-flop and 25 MHz for a counter. Another switching parameter related to f_{max} is the minimum clock-pulse width (t_w) for device operation. Rise time (t_r) and fall time (t_f) for device output waveforms are sometimes specified. These parameters indicate the time required for the output to switch between 10 and 90 percent of full amplitude. Specific device data sheets should be consulted for exact t_w , t_r , and t_f values.

7.2.2.3 Critical parameters. The attributes discussed in paragraph 7.2.2.2 of proper interfacing, noise margins, power dissipation, and timing are critical in microcircuit applications. If these are not adhered to in the application, the microcircuit will malfunction, or at the minimum, operate unreliably. In addition, absolute maximum ratings for a number of parameters are always given in device data sheets and should be followed. Some of the parameters are supply current, input voltage and current, lead temperature, ambient temperature under bias, and storage temperature. Any exposure of the device to more stressful conditions than the absolute maximum rating may permanently damage the device. Exposure to absolute maximum rated conditions for extended periods may affect device reliability.

7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

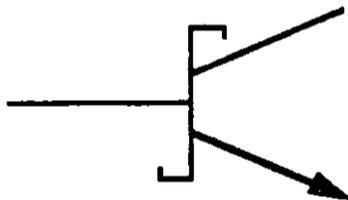
7.2.2.4 Specific design precautions. All unused input pins of LS devices should be tied to the power supply or to an output of an unused gate that is forced HIGH rather than be left floating. Resistors are recommended for certain connections. The application engineer should refer to the manufacturer's recommendations for each device type. Ground planes are sometimes preferred instead of ground wires.

Interfacing LS to other logic families should be accomplished in accordance with the data sheet parameters of both systems. The output and input requirements of both systems should be understood. Sometimes interface circuitry is required such as pull-up resistors or drivers. Drive capability of the LS family is discussed in paragraph 7.2.5.

All digital microcircuit families, including LSTTL, are electrostatic-discharge sensitive. Proper handling and manufacturing design precautions should be followed. Electrostatic-discharge sensitivity of microelectronic devices is discussed in subsection 7.1 General.

7.2.3 Physical construction. Digital microcircuits of the LSTTL family are constructed with the planar diffusion process. Basic descriptions of fabrication, assembly, thermal characteristics, reliability, and other details are given in the general sections for transistors and microcircuits.

7.2.3.1 Schottky diodes and Schottky diode clamped transistors. The low power Schottky TTL digital microcircuit uses the Schottky diode clamped transistor. This is an npn transistor produced by conventional photolithography diffusion, an epitaxial growth techniques with a metal-silicon barrier Schottky diode in parallel with the collector-base junction. The base contact of the transistor serves as the metal contact of the diode and the collector region of the transistor serves as the n-region of the diode. It is convenient to regard the npn transistor Schottky diode combination as a single device and to represent it by a single symbol:



To form the Schottky diode clamped transistor, the base-contact opening is extended beyond the base diffusion and over the collector region. When the aluminum metallization is deposited, it simultaneously functions as the contact to the base region of the transistor and the anode of the Schottky diode. To prevent high field concentration, the aluminum can either be extended over the passivating oxide or terminated over a p⁺ guard ring, which can be diffused into the die at the same time as the base, as shown in Figure 5.

7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

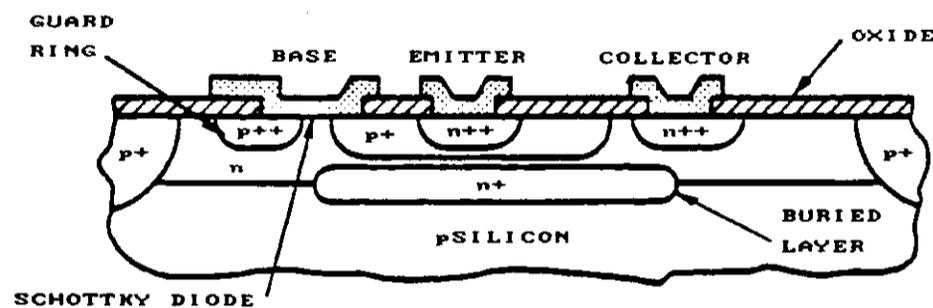


FIGURE 5. Cross section of Schottky transistor.

In the forward-biased pn junction, a current is carried by holes flowing from the p-type material into the n-type material, and electrons flowing from the n to p material. Either case results in an excess of minority carriers near the junction. If the voltage is reversed these carriers will flow back across the junction, creating a high current until the supply of carriers is exhausted. In other words, the pn junction cannot be turned off immediately.

In a Schottky diode made of aluminum on n-type silicon, essentially all of the forward current is carried by electrons flowing from the semiconductor into the metal. They quickly come into equilibrium with the other electrons in the metal, so there is effectively no stored charge to prevent rapid switching. Another major point of difference between the Schottky and the pn junction diode is that the former has lower forward voltage for a given current. In a practical circuit the Schottky diode is placed in parallel with the base-collector junction of an npn transistor as shown in Figure 5; the metal electrode is connected to the base and to the n-region of the collector, where it forms a rectifying contact. Because the Schottky diode has a lower forward voltage compared with the collector-base junction, the diode clamps the transistor and diverts most of the excess base current through the Schottky diode, preventing the transistor from deep saturation. There is no charge storage in either the transistor or the diode.

The advanced Schottky (AS or F) and the advanced low-power Schottky (ALS), both discussed later, use ion implantation of impurities instead of diffusion, oxide isolation of transistors, and full Schottky clamping to achieve improved switching times at reduced speed-power products.

7.2.3.2 Packaging. Digital microcircuits differ from other microcircuits mainly in method of operation, which is controlled by design and production by steps prior to breaking the wafer into dice. Physical construction of the die package in which it is encased is usually fundamentally similar for all microcircuits.

7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

Although microcircuits are encased in a wide variety of hermetically sealed and plastic packages, they are generally limited to three basic styles with variations in dimensions and number of leads:

- a. Flat packages (flatpacks)
- b. Dual in-line packages (DIPs)
- c. Leadless chip carriers (LCC).

The description of these packages is given in subsection 7.1 General.

7.2.4 Military designation. Specifications and requirements for standard microcircuits used on military programs are called out in the General Specification for Microcircuits (MIL-M-38510) and associated detail specifications (MIL-M-38510 slash sheets). Descriptions of the designation for military microcircuits are given in subsection 7.1 General.

Military microcircuit parts which have the potential for being used in NASA programs shall have the MIL-M-38510 Class S or Class B qualification. NASA Grade 1 and Grade 2 parts are listed in MIL-STD-975 (NASA).

7.2.5 Electrical characteristics. Logic circuits, including LSTTL, are generally characterized by their high and low output voltage states, speed, and power consumption. In the latter two factors, an improvement in one may be often gained at the expense of the other for a given type of logic. The difference between the high and low output voltage is a factor in determining the fanout and the noise immunity of a device. Speed is generally measured in terms of the propagation delay time. For flip-flops, the maximum toggle frequency is an important characteristic. Other parameters which are needed to fully characterize a circuit vary from one logic type to another. However, for high reliability applications, all critical parameters should be specified to include all worst-case conditions using threshold voltages at the inputs, maximum and minimum temperatures, maximum and minimum supply voltages, maximum and minimum load currents, etc. For best results, the circuit speed should be appropriately matched to the needs of the application. The fastest circuit is not always the best circuit, since it may waste power or be susceptible to oscillations.

Table VI gives typical values of power dissipation, propagation delay, maximum toggle frequency, and speed/power product for the LS family.

Table VII gives the fan-out capability for the LSTTL family. The fan-out numbers given in Table VII are the number of loads of the various Schottky families which the LSTTL driver family can drive. For example, the LS output high state can drive 20 AS devices in the high state, and the output low state can drive 8 AS devices in the low state. The lowest figure is taken so that the interconnection is reliable for both logic states. Therefore, one LS gate can reliably drive 8 AS gates. Output and input current per gate at high and low logic levels are also given in Table VIII.

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**7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY
TRANSISTOR-TRANSISTOR LOGIC**

TABLE VI. Typical values of performance characteristics of low-power Schottky TTL at 25 °C

Parameter	LS
Power dissipation per gate (mW)	
Static	2
At 100 kHz	2
Propagation delay time (ns), $C_L = 15$ pF	10
Maximum toggle frequency (MHz) $C_L = 15$ pF	40
Speed/power product (pJ) at 100 kHz	20

TABLE VII. LSTTL fan-out capability to selected Schottky families

LSTTL Driver	I/O	Output Current (mA)	Driven Family		
			LS	ALS	AS (F)
			H 0.02 L 0.4 <u>1/</u>	0.02 0.1 <u>1/</u>	0.02 0.5 <u>1/</u>
LS	H L	0.4 4	20 10	20 40	20 8

1/ Maximum input currents for high and low logic states.

The characteristics discussed here are meant to show only typical values of LSTTL. In applications, the specific device data sheets should be reviewed and interpreted for the worst-case conditions under which the circuit will be used.

**7.2 MICROCIRCUITS, LOW-POWER SCHOTTKY
TRANSISTOR-TRANSISTOR LOGIC**

7.2.6 Environmental considerations. Environmental considerations of microcircuit devices are primarily related to package style rather than device type. They are treated in the subsection 7.1 General, paragraph 7.1.7 Reliability considerations.

7.2.7 Reliability considerations. Reliability considerations of bipolar digital microelectronic devices are basically the same as those of other microelectronic devices. They are treated in the subsection 7.1 General, paragraph 7.1.7 Reliability considerations.

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7.3 MICROCIRCUITS, ADVANCED LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

7.3 Advanced low-power Schottky transistor-transistor logic.

7.3.1 Introduction. The advanced low-power Schottky transistor-transistor logic (ALSTTL or ALS) family is designed as an enhancement to the low-power Schottky logic family (LSTTL). In general, ALS devices are ideal for improving power efficiency at lower speeds. ALS devices provide reasonable propagation delay times, approximately 4 ns, which is approximately 60-percent improvement over LSTTL values. ALS devices provide a lower power dissipation, approximately 1 mW per gate which is approximately one-half the power dissipation of LSTTL. The general comments of subsection 7.2 Low-power Schottky TTL should be reviewed because they also apply to ALS microcircuit usage.

7.3.2 Usual applications. The ALS series is suitable for replacing all TTL families except in the very highest frequency applications. Replacement with ALS will result in lower power consumption, smaller power supply current spikes, and, in some cases, better noise immunity than other families. Some general guidelines will be presented here. The applications engineer should refer to device data sheets for specific guidance.

7.3.2.1 Available functional types. MIL-STD-975 (NASA), NASA Standard Electrical, Electronic, and Electromechanical Parts List, presents the available ALS part types that must be used for NASA microcircuit applications. This standard should be consulted for part selection. See subsection 7.1 General for discussion of Grades 1 and 2.

Table VIII presents the available functions for the ALS family that shall be used for NASA programs. This table summarizes information found in MIL-STD-975.

TABLE VIII. ALS functions available in MIL-STD-975

Gates	AND, NAND, OR, NOR, INV
Drivers	Buffer/line drivers
Counters	Binary
Flip-flop	D, J-K
Latches	D
Decoder	2-4 line
Multiplexers	2 to 1 universal

7.3 MICROCIRCUITS, ADVANCED LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

7.3.2.2 Comparative attributes specific to ALS microcircuits. Attributes specific to ALS microcircuits are specific values of:

- a. Interface parameters
- b. Noise margins
- c. Power dissipation
- d. Switching and propagation times.

A general discussion of these parameters was presented in subsection 7.2 Low-power Schottky TTL and the main points applicable to ALS microcircuits will be reviewed here. The application engineer should refer to the latest specific microcircuit data sheet for exact information.

ALS devices will interface properly with other technology families if the inequalities

$$V_{IH} < V_{OH} \text{ and } V_{OL} < V_{IL}$$

are satisfied (see paragraph 7.2.2.2 Comparative attributes specific to LSTTL microcircuits). The low-level noise margin (V_{NIL}) and the high-level noise margin (V_{NIH}) must also be satisfied by the ALS devices at worst-case operating conditions. Table IX gives the worst-case values of interface parameters and noise margins which must be satisfied under all conditions by ALS microcircuits.

TABLE IX. Worst-case values of primary interfacing parameters for ALS devices

Parameter	ALS Value
V_{IH} min	2 V
V_{IL} max	0.8
V_{OH} min	2.7
V_{OL} max	0.4
High-level noise margin $ V_{OH} - V_{IH} $	0.7
Low-level noise margin $ V_{OH} - V_{IH} $	0.4

7.3 MICROCIRCUITS, ADVANCED LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

Power dissipation is a function of device operating frequency; therefore, current data sheets should be checked for exact design values. Likewise, propagation delays (t_{PLH}) and (t_{PHL}) are a function of load capacitance and data sheets should be referenced. Similarly, other switching parameters such as f_{max} , t_w , t_r , and t_f (see paragraph 7.2.2.2 Comparative attributes specific to LSTTL microcircuits) should be extracted from the specific device data sheets. Approximate values for these parameters, for ALS microcircuits, are given in paragraph 7.3.5 Electrical characteristics.

7.3.2.3 Critical parameters. As discussed in paragraph 7.2.2.3 Critical parameters, proper interfacing, noise margins, power dissipation, and timing are critical in microcircuit applications. Absolute maximum ratings for supply voltage, input voltage, ambient operating temperature, and storage temperature should be adhered to. Data sheets must be consulted for acceptable design data.

7.3.2.4 Specific design precautions. All unused input pins of ALS devices should not be left floating, but rather should be tied to ground, power supply, or each other, depending on usage conditions. Resistors are recommended for certain applications. The application engineer should refer to the manufacturer's recommendations for each device connection. Ground planes are sometimes preferred over ground wires.

Interfacing ALS to another logic family should be in accordance with the data sheet parameters of both systems. The output and input requirements of both systems should be followed. Sometimes interface circuitry is required such as pull-up resistors or drivers. Drive capability of the ALS family is discussed in paragraph 7.3.5 Electrical characteristics.

The ALS family has internally designed electrostatic discharge protection to approximately 4000 V by incorporating npn Schottky transistors on input stages.

However, all microelectronic devices, including ALS, are electrostatic-discharge sensitive to some degree. Therefore, proper handling and manufacturing design precautions should be followed. Electrostatic-discharge sensitivity of microelectronic devices is discussed in subsection 7.1 General.

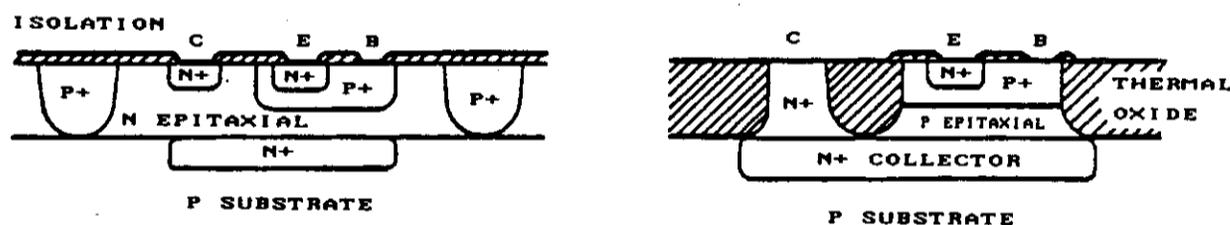
7.3.3 Physical construction. The ALS family is constructed with the planar process. Basic descriptions of fabrication, assembly, thermal characteristics, reliability, and other details are given in the General subsections for transistors (5.1) and microcircuits (7.1). The advanced low-power Schottky (ALS) process also uses transistors, and full Schottky clamping to achieve improved switching times at reduced speed-power products.

7.3.3.1 Oxide isolation process. This is a method to increase device densities through oxide isolation. This process is based on an approach to circuit isolation in which the active p-type diffusions that isolate conventional devices are replaced by passive insulator-oxide regions. With the isolation also serving as an insulator, there is no need to separate the isolation region from diffused areas. Hence, the oxide-isolated device achieves a considerable size reduction over its diode-isolated counterpart.

7.3 MICROCIRCUITS, ADVANCED LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

Implementation of the oxide isolation technique required development of a method to mask the active region of the transistor (where the buried collector and the base and emitter are located) during the oxide isolation step. The approach selected uses silicon nitride (Si_3N_4) layers to mask against thermal oxidation of the silicon in this region during the isolation step. Silicon nitride is ideal for this purpose because it remains practically inert during the oxidation step.

To understand the operation of these devices, consider the conventional diffused isolation npn transistor shown in cross-section in Figure 6. In this structure, the collector is buried beneath an epitaxial layer through which contact is made. A diffused p-type region isolates the collector region of one transistor from the collector region of an adjacent device. With oxide isolation, the p-regions are replaced by selectively grown thermal oxide regions formed to the depth of the buried collector, resulting in the cross-section shown in Figure 6. The electrical contact to the buried collector is surrounded by an additional oxide region located between the collector sink area and the base region. Analysis of the structure makes it apparent that good isolation is obtained with the oxide regions without having to separate the isolation from the transistor base. The region between the p^+ isolation and the base in Figure 6 can be eliminated entirely, bringing with it about a 40-percent savings in valuable die area.



Conventionally fabricated npn transistors use p-region for isolation, requiring space between isolation and base.

By using thermal oxide in place of the p-region, the base can abut the isolation region, saving 40% of die real estate.

FIGURE 6. Standard and oxide isolation structures.

The oxide isolation process provides the advantage of self-alignment. Because the nitride layers can be etched away selectively without harming oxide layers, the base and collector sinks can be fabricated out to the oxide, so that the oxide itself, as an insulator, limits the extent of these diffusions. Thus, mask alignment is far less critical. Resistors can be aligned in the same manner, indicating the possibility of preregistration of both passive and active components with respect to the isolation regions.

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7.3 MICROCIRCUITS, ADVANCED LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

Figure 7 shows a cross-section of a transistor fabricated with the ALS process with ion implantation, oxide isolation, and standard metal system. SBD refers to the Schottky barrier diode.

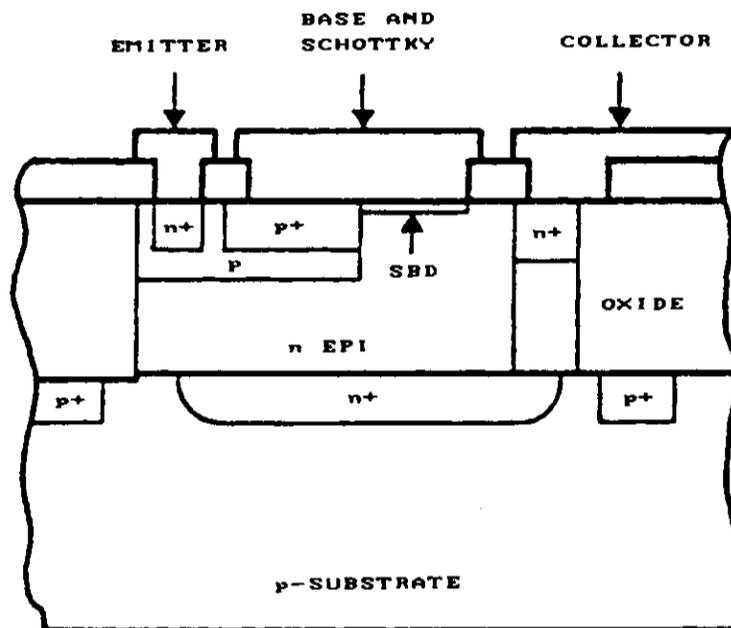


FIGURE 7. Cross section of a transistor fabricated with the ALS process. SBD is the Schottky barrier diode region.

7.3.3.2 Packaging. ALS devices are generally available in the four basic styles discussed in paragraph 7.2.3.2 Packaging, with variations in dimensions and number of leads. An additional description of these packages is given in subsection 7.1 General.

7.3.4 Military designation. Specifications and requirements for standard microcircuits used on military programs are called out in the General Specification for Microcircuits, MIL-M-38510, and associated detail specifications, MIL-M-38510 slash sheets. Descriptions of the designation for military microcircuits are given in subsection 7.1 General.

Military microcircuit parts which have the potential for being used in NASA programs shall have the MIL-M-38510 Class S or Class B qualification. NASA Grade 1 and Grade 2 parts are listed in MIL-STD-975 (NASA).

7.3 MICROCIRCUITS, ADVANCED LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

7.3.5 Electrical characteristics. The general comments discussed in paragraph 7.2.5 Electrical characteristics also apply to ALS device electrical characteristics and should be referenced. Table X gives typical values of power dissipation, propagation delay, maximum toggle frequency, and speed/power product for the ALS family.

TABLE X. Typical values of performance characteristics of ALS devices at 25 °C

Parameter	ALS
Power dissipation per gate (mW)	
Static	1
At 100 KHz	1
Propagation delay time (ns), C _L = 15 pF	4
Maximum toggle frequency (MHz), C _L = 15 pF	70
Speed/power product (pJ) at 100 KHz	4

Table XI gives the fan-out capability for the ALS family. The fan-out numbers given in Table XI are the number of loads of the various Schottky families which the ALS driver family can drive. For example, the ALS output high state can drive 20 LS devices in the high state, and the output low state can drive 10 LS devices in the low state. The lowest figure is taken so that the interconnection is reliable for both logic states. Therefore, one ALS gate can reliably drive 10 gates. Output and input current per gate at high and low logic levels are also given in Table XI.

The characteristics discussed here are meant to show only typical values of ALSTTL. In applications, the specific device data sheets should be reviewed and interpreted for the worst-case conditions under which the circuit will be used.

7.3 MICROCIRCUITS, ADVANCED LOW-POWER SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

TABLE XI. ALSTTL fan-out capability to selected Schottky families.

Driver Family	I/O	Output Current (mA)	Driven Family		
			LS	ALS	AS (F)
			H 0.02 $\frac{1}{\mu}$ L 0.4	0.02 $\frac{1}{\mu}$ 0.1	0.02 $\frac{1}{\mu}$ 0.5
ALS	H	0.4	20	20	20
	L	4	10	40	8

$\frac{1}{\mu}$ Maximum input currents (ma) for high and low logic states.

7.3.6 Environmental considerations. Environmental considerations of microcircuit devices are primarily related to package style rather than device type. They are treated in subsection 7.1 General, paragraph 7.1.7 General reliability considerations.

7.3.7 Reliability considerations. Reliability considerations of bipolar digital microelectronic devices are basically the same as those of other microelectronic devices. They are treated in paragraph 7.1.7 General reliability considerations.

7.4 MICROCIRCUITS, ADVANCED SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

7.4 Advanced Schottky transistor-transistor logic.

7.4.1 Introduction. The advanced Schottky transistor-transistor logic (ASTTL or AS) family is designed as an enhancement to the standard Schottky logic family (STTL). In general, the AS devices give an improved power efficiency at higher speeds. AS devices require approximately one-half of the supply current of the STTL family and have approximately twice the clocking frequency. AS devices provide minimal delay times, approximately 1.7 ns, at a power cost of approximately 8 mW per gate. AS circuits contain additional circuitry to provide a flatter power-frequency characteristic and their input configuration needs lower input current which translates into higher fanout. The general comments of subsection 7.2 also apply to AS microcircuit usage, the AS device family is also known as the "FAST" or "F" TTL device family, and these terms may be used interchangeably.

7.4.2 Usual applications. Advanced Schottky devices are electrically and pinout compatible with all existing TTL families. The AS devices are ideal for replacement of high-speed logic families to give lower power consumption. At frequencies higher than approximately 5 MHz, the power dissipation of AS devices is lower than that of the CMOS family.

7.4.2.1 Available functional types. MIL-STD-975 (NASA), NASA Standard Electrical, Electronic, and Electromechanical Parts List, presents the available AS part types that must be used for NASA microcircuit applications. This standard should be consulted for part selection. See subsection 7.1 General for discussion of Grades 1 and 2.

Table XII presents the available functions for the AS family that shall be used for NASA programs. This table summarizes information available in MIL-STD-975.

7.4.2.2 Comparative attributes specific to AS microcircuits. The AS family is electrically and pinout compatible with all existing TTL families. AS devices are ideal replacements in high frequency usage. These devices use approximately 40 percent of the power that STTL devices require and have one-fourth the speed-power product of STTL. All the specific attributes discussed in subsections 7.2 and 7.3, interface parameters, noise margins, power dissipation as well as switching and propagation times apply also to the AS family of devices. Table XIII gives the approximate worst case values of interface parameters and noise margins which must be satisfied under all conditions by AS microcircuits. Specific data sheets should be checked for more exact information. Data sheets should also be checked for exact design values for power dissipation, and switching and propagation times. Approximate values for these parameters for AS microcircuits are given in paragraph 7.4.5, Electrical characteristics.

7.4.2.3 Critical parameters. As discussed in subsection 7.2.2.3, proper interfacing, noise margins, power dissipation, and timing are critical in microcircuit applications. Absolute maximum ratings for supply voltage, input voltage, ambient operating temperature, and storage temperature should be followed. Specific data sheets must be followed for these design parameters.

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**7.4 MICROCIRCUITS, ADVANCED SCHOTTKY
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TABLE XII. AS or F functions available in MIL-STD-975

Gates	AND, NAND, OR, NOR, INV
Drivers	buffer/line
Counters	synchronous binary, BCD decade, up/down binary
Flip-flops	D, J-K
Latches	D
Registers	(2-port)
Adders	4-bit binary
Decoders	1 of 4, 1 of 8
Multiplexers	2-, 4-, 8-input
Transceivers	4-input, bus
Combinational	9-bit parity generator/checker

TABLE XIII. Worst-case values of primary inter-
facing parameters for AS devices

Parameter	AS(F) value
V_{IH} min	2V
V_{IL} max	0.8
V_{OH} min	2.7
V_{OL} max	0.4
High level noise margin $ V_{OH} - V_{IH} $	0.7
Low level noise margin $ V_{IL} - V_{OL} $	0.4

7.4 MICROCIRCUITS, ADVANCED SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

7.4.2.4 Specific design precautions. All unused input pins of AS devices should not be left floating but rather should be tied to ground, to power supply, or to each other, depending on usage conditions. Resistors are recommended for certain connections. The application engineer should refer to the manufacturer's recommendations for each device connection. Ground planes are sometimes preferred instead of ground wires.

Interfacing AS to another logic family should be in accordance with the data sheet parameters of both systems. The output and input requirements of both systems should be followed. Sometimes interface circuitry is required such as pull-up resistors or drivers. Drive capability of the AS family is discussed in subsection 7.4.5.

The AS family, like the ALS devices, has internally designed electrostatic-discharge protection to approximately 4000 volts. However, all microelectronic devices, including AS, are electrostatic-discharge sensitive in a somewhat unpredictable manner. Therefore proper handling and manufacturing process precautions should be followed. Electrostatic-discharge sensitivity of microelectronic devices is discussed in subsection 7.1 General.

7.4.3. Physical construction. The AS family is constructed with the planar process. Basic descriptions of fabrication, assembly, thermal characteristics, reliability, and other details are given in the General sections for transistors and microcircuits. The advanced Schottky (AS or F) and the advanced low-power Schottky (ALS) both use ion implantation of impurities instead of diffusion, oxide isolation of transistors, and full Schottky clamping to achieve improved switching times at reduced speed-power products.

7.4.3.1 Oxide isolation process. Details of the oxide isolation process, which is used to fabricate both AS and ALS families, are given in paragraph 7.3.3.1.

7.4.3.2 Packaging. AS devices are generally available in the four basic styles discussed in paragraph 7.2.3.2, with variations in dimensions and number of leads. An additional description of these packages is given in subsection 7.1 General.

7.4.4 Military designation. Specifications and requirements for standard microcircuits used on military programs are called out in the General Specification for Microcircuits (MIL-M-38510) and associated detail specifications (MIL-M-38510 slash sheets). Descriptions of the designation for military microcircuits are given in subsection 7.1 General.

Military microcircuit parts which have the potential for being used in NASA programs shall have the MIL-M-38510 Class S or Class B qualification. NASA Grade 1 and Grade 2 parts are listed in MIL-STD-975 (NASA).

7.4.5 Electrical characteristics. The general comments discussed in paragraph 7.2.5 also apply to AS device electrical characteristics and should be referenced. Table XIV gives typical values of power dissipation, propagation delay, maximum toggle frequency, and speed/power product for the AS family.

7.4 MICROCIRCUITS, ADVANCED SCHOTTKY TRANSISTOR-TRANSISTOR LOGIC

TABLE XIV. Typical values of performance characteristics of
AS devices at 25 °C

Parameter	AS (F)
Power dissipation per gate (mW)	
Static	8.5
At 100 kHz	8.5
Propagation delay time (ns), $C_L = 15$ pF	1.5
Maximum toggle frequency (MHz) $C_L = 15$ pF	200
Speed/power product (pJ) at 100 kHz	13

Table XV gives the fan-out capability for the AS family. The fan-out numbers given in Table XV are the number of loads of the various Schottky families which the driver family can drive. For example, the AS output high state can drive 100 ALS devices in the high state, and the output low state can drive 200 ALS devices in the low state. The lowest figure is taken so that the interconnection is reliable for both logic states. Therefore one AS gate can reliably drive 100 ALS gates. Output and input current per gate at high and low logic levels are also given in Table XV.

TABLE XV. ASTTL fanout capability to selected Schottky families

Driver Family	I/O	Output Current (mA)	Driven Family		
			LS	ALS	AS (F)
			H 0.02 $\frac{1}{}$ L 0.4 $\frac{1}{}$	0.02 $\frac{1}{}$ 0.1 $\frac{1}{}$	0.02 $\frac{1}{}$ 0.5 $\frac{1}{}$
AS(F)	H L	2 20	100 50	100 200	100 40

$\frac{1}{}$ Maximum input currents (ma) for high and low logic states.

The characteristics discussed here are meant to show only typical values of ASTTL. In applications, the specific device data sheets should be reviewed and interpreted for the worst-case conditions under which the circuit will be used.

**7.4 MICROCIRCUITS, ADVANCED SCHOTTKY
TRANSISTOR-TRANSISTOR LOGIC**

7.4.6 Environmental considerations. Environmental considerations of microcircuit devices are primarily related to package style rather than device type. They are treated in subsection 7.1.7 Reliability considerations.

7.4.7 Reliability considerations. Reliability considerations of bipolar digital microelectronic devices are basically the same as those of other microelectronic devices. They are treated in subsection 7.1.7 Reliability considerations.

7.5 MICROCIRCUITS, CMOS 4000B SERIES**7.5 CMOS 4000B series.**

7.5.1 Introduction. The 4000B CMOS series is a family of high-voltage small-scale integration and medium scale integration microcircuits consisting of functions from simple gates to complex counters, registers, and arithmetic circuits. This family gives the design features of CMOS technology of low-power consumption, high noise immunity, high speed, high fan-out, TTL logic compatibility, excellent temperature stability, and fully protected inputs and outputs. In addition, the 4000B series gives high-voltage operation, higher noise immunity, and standardized, symmetrical output characteristics.

7.5.2 Usual applications. The 4000B series CMOS microcircuits are very useful if very low quiescent power, low operating power, moderate frequency range, high operating voltage, and high noise immunity are required. In addition, some manufacturers offer a larger selection of this series in radiation hardened versions. Specific performance characteristics will be given in paragraph 7.5.5 Electrical characteristics.

7.5.2.1 Available functional types. MIL-STD-975 (NASA), NASA Standard Electrical, Electronic, and Electromechanical Parts List, presents the available 4000B part types that must be used for NASA microcircuit applications. This standard should be consulted for part selection. See subsection 7.1 General for discussion of Grades 1 and 2.

Table XVI presents the available functions for the 4000B family that shall be used for NASA programs. This table summarizes information available in MIL-STD-975.

TABLE XVI. 4000B series CMOS functions available in MIL-STD-975

AND, NAND, OR, NOR, INV	Gates
INV	Buffers
D, J-K, latch, multivibrator	Flip-flops
Binary, decade, divide-by-N	Counter
Synchronous, asynchronous, static	Shift registers

7.5.2.2 Comparative attributes specific to 4000B microcircuits. The CMOS 4000B series is compatible with existing MOS circuit families and existing TTL families either directly or with some interface circuitry. For example, interfacing TTL to 4000B CMOS will require pull-up resistors whereas this CMOS group may drive some TTL inputs directly. The applications engineer should consult the manufacturer's recommendations for specific family-family interfacing requirements.

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7.5 MICROCIRCUITS, CMOS 4000B SERIES

All the general attributes discussed in subsection 7.2 Low-power Schottky transistor-transistor logic regarding interface parameters, noise margins, power dissipation, and switching and propagation times also apply to the 4000B family of devices. Table XVII gives the approximate worst-case values of interface parameters and noise margins which must be satisfied under all conditions by 4000B microcircuits. Specific data sheets should be checked for more exact information and exact design values for power dissipation, as well as switching and propagation times. Approximate values for these parameters for 4000B microcircuits are given in paragraph 7.5.5 Electrical characteristics.

TABLE XVII. Worst-case values of primary interfacing parameters for 4000B CMOS devices

Parameter <u>1/</u>	4000B
V _{IH} min	3.5 V
V _{IL} max	1.5
V _{OH} min	4.95
V _{OL} max	0.05
High-level noise margin V _{OH} - V _{IH}	1 V
Low-level noise margin V _{IL} - V _{OL}	1

1/ Values at power supply of 5 V.

7.5.2.3 Critical parameters. As discussed in paragraph 7.2.2.3, proper interfacing, noise margins, power dissipation, and timing are critical in microcircuit applications. In addition, absolute maximum ratings should be observed. For the 4000B series, these absolute maximum ratings cover supply voltage, input voltage, power dissipation, operating temperature, storage temperature, and lead temperature during soldering processes. Specific data sheets must be followed for these design precautions.

7.5 MICROCIRCUITS, CMOS 4000B SERIES

7.5.2.4 Specific design precautions. All unused input pins of 4000B devices should be terminated to a power supply or to other used inputs depending on device type and usage conditions. Resistors are recommended for certain connections. The application engineer should refer to the manufacturer's recommendations for each device connection. The outputs of CMOS devices should not be subjected to voltages higher than power supply or lower than ground to prevent latch-up.

Interfacing 4000B devices to another logic family should be in accordance with the data sheet parameters of both systems. The output and input requirements of both systems should be followed. Sometimes interface circuitry is required such as pull-up resistors or drivers. Drive capability of CMOS families is discussed in paragraph 7.5.5 Electrical characteristics.

The 4000B family has internal electrostatic-discharge protection to approximately 4000 V, except for transmission gates, which are protected to approximately 800 V. However, all microelectronic devices, including the 4000B types, are electrostatic-discharge sensitive in a somewhat unpredictable manner. Therefore, proper handling and manufacturing process precautions should be followed. Electrostatic-discharge sensitivity of microelectronic devices is discussed in subsection 7.1 General.

7.5.3 Physical construction. The 4000B CMOS series is constructed with the planar process. Basic descriptions of fabrication, assembly, thermal characteristics, reliability, and other details are given in the General subsections for transistors (5.1) and microcircuits (7.1).

7.5.3.1 Complementary MOS process. Figure 8 shows the cross section of a 4000B CMOS inverter with an aluminum gate, bulk silicon technology. The 4000B family of CMOS digital devices from some manufacturers is representative of this technology.

7.5.3.2 Packaging. The 4000B CMOS family is generally available in the four basic package styles, discussed in subsection 7.2 Low-power Schottky transistor-transistor logic, paragraph 7.2.3.2 Packaging, with variations in dimensions and number of leads. An additional description of these packages is given in subsection 7.1 General.

7.5 MICROCIRCUITS, CMOS 4000B SERIES

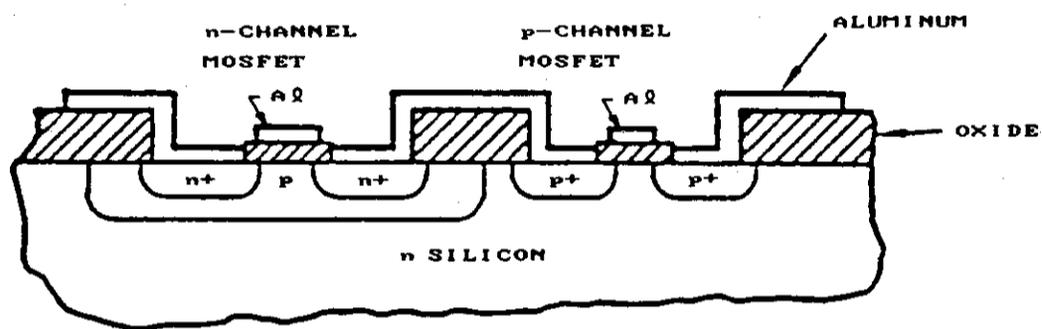


FIGURE 8. Cross section of a 4000B CMOS inverter with aluminum gate, bulk silicon technology.

7.5.4 Military designation. Specifications and requirements for standard microcircuits are called out in the General Specification for Microcircuits (MIL-M-38510) and associated detail specifications (MIL-M-38510 slash sheets). Descriptions of the designation for military microcircuits are given in subsection 7.1 General.

Military microcircuit parts which have the potential for being used in NASA programs shall have the MIL-M-38510 class S or class B qualification. NASA Grade 1 and Grade 2 parts are listed in MIL-STD-975 (NASA).

7.5.5 Electrical characteristics. The general comments discussed in subsection 7.2 Low-power Schottky transistor to transistor logic, paragraph 7.2.5 Electrical characteristics also apply to 4000B CMOS electrical characteristics and should be referenced. Table XVIII gives typical power dissipation, propagation delay, maximum toggle frequency, and speed/power product for the 4000B family. Although CMOS has low-power dissipation, this parameter increases with frequency and is comparable to Schottky dissipation values at higher frequencies. The recommended power supply voltage range is 3 to 18 V over the full ambient temperature range.

Table XIX shows fan-out capability for the 4000B CMOS family. The required input current of 0.001 mA illustrates the general rule that MOS devices need very little input current to be driven active. They are voltage-driven devices. The characteristics discussed here are meant to show only typical values of 4000B devices. In applications, the specific device data sheets should be reviewed and interpreted for the worst-case conditions under which the circuit will be used.

7.5 MICROCIRCUITS, CMOS 4000B SERIES

TABLE XVIII. Typical values of performance characteristics of 4000B devices at 25 °C

Parameter	4000B
Power dissipation per gate (mW)	
Static	0.001
At 1000 KHz	0.1
Propagation delay time (ns), C _L = 15 pF	105
Maximum toggle frequency (MHz)	
C _L = 15 pF	12
Speed/power product (pJ) at 100 KHz	11

TABLE XIX. Fan-out capability of 4000B series metal gate CMOS

Parameter	4000B
Maximum output drive (mA)	1.6
Maximum input current (mA)	0.001

7.5.6 Environmental considerations. Environmental considerations of microcircuit devices are primarily related to package style rather than device type. They are treated in subsection 7.1 General, paragraph 7.1.7 Reliability considerations.

7.5.7 Reliability considerations. Reliability considerations of CMOS digital microelectronic devices are basically the same as those of other microelectronic devices. They are treated in subsection 7.1 General, paragraph 7.1.7 Reliability considerations.

7.6 MICROCIRCUITS, HIGH-SPEED CMOS SERIES

7.6 High-speed CMOS series

7.6.1 Introduction. The high-speed CMOS or HC series of microcircuits includes an extensive line of devices that are pin compatible with many existing LSTTL and 4000B CMOS series digital logic circuits. Two versions of high-speed CMOS are available, HC and HCT. The HC series contains approximately 80 percent of the defined CMOS functions and is primarily intended for new designs. The HCT series consists of TTL-threshold compatible devices and was designed as a pin-for-pin replacement for LSTTL. The HCT series is primarily used for retrofitting.

The key features of the HC and HCT series devices are:

- a. Speeds equivalent to LSTTL types with typical gate delays of 10 ns
- b. Fan-out to 10 LSTTL loads
- c. Operating frequencies equivalent to LSTTL types, typically approximately 30 MHz
- d. High-voltage noise-immunity of CMOS, typically 45 percent of power supply, which is a two to three times improvement over LSTTL
- e. Wide range of power supply operating voltages, 2 to 6 V
- f. Low static power consumption, typically less than 1 mW.

7.6.2 Usual applications. The HC CMOS family gives all the advantages of standard 4000B CMOS devices, except HC may operate up to approximately 30 to 50 MHz. Many HC devices are pin-out compatible with 4000B CMOS devices. Because of the high density of HC designs, more complex functions may be implemented on a single chip with the HC technology than with the 4000B technology.

As discussed in paragraph 7.6.1 Introduction, the HCT series is designed primarily to replace LSTTL devices. In this application, the low power dissipation of HCT and subsequent decreased heat generation enhance system reliability by reducing or eliminating the need for heat sinks, fans, tightly regulated high current power supplies, and copper busses. For example, in a typical system in which LSTTL types have been replaced by equivalent HCT types, the system can be expected to operate with only one percent of the power, assuming an operating frequency of 10 KHz. Because of their high noise immunity, the HC and HCT series are ideal for use in noisy environments.

7.6.2.1 Available functional types. MIL-STD-975 (NASA), NASA Standard Electrical, Electronic, and Electromechanical Parts List, presents the available HC CMOS part types that must be used for NASA microcircuit applications. This standard should be consulted for part selection. See subsection 7.1 General for discussion of Grades 1 and 2.

7.6 MICROCIRCUITS, HIGH-SPEED CMOS SERIES

Table XX presents the available functions of the HC CMOS family that shall be used for NASA programs. This table summarizes information available in MIL-STD-975.

TABLE XX. HC series CMOS functions available in MIL-STD-975

Gates	(AND, NAND, OR, NOR)
Flip-flop	(D-type)

7.6.2.2 Comparative attributes specific to HC microcircuits. The HC series is compatible with existing CMOS families and the HCT series is compatible with, and is a replacement for, LSTTL devices. The application engineer should consult the manufacturer's recommendations for specific interfacing requirements.

All the general attributes discussed in subsection 7.2 Low-power Schottky transistor-transistor logic regarding interface parameters, noise margins, power dissipation, and switching and propagation times also apply to the HC family of devices. Table XXI gives the approximate worst-case values of interface parameters and noise margins that must be satisfied under all conditions by HC series CMOS microcircuits. Specific data sheets should be consulted for more exact information. Data sheets should also be checked for exact design values for power dissipation, as well as switching and propagation times. Approximate values for these parameters and for fan-out of HC series CMOS microcircuits are given in subsection 7.6.5 Electrical characteristics.

TABLE XXI. Worst-case values of primary interfacing parameters

Parameter ^{1/}	HC
V _{IH} min	3.5 V
V _{IL} max	1.0
V _{OH} min	4.9
V _{OL} max	0.1
High-level noise margin V _{OH} - V _{IH}	1.4
Low-level noise margin V _{IL} - V _{OL}	0.9

^{1/} Values at power supply of 5 V.

7.6 MICROCIRCUITS, HIGH-SPEED CMOS SERIES

7.6.2.3 Critical parameters. As discussed in paragraph 7.2.2.3, proper interfacing, noise margins, power dissipation, and timing are critical in HC series CMOS microcircuit applications. In addition, absolute maximum ratings should be observed. For the HC series, these absolute maximum ratings cover supply voltages, input and output voltages, clamp diode currents per pin, output and power supply currents, power dissipations, operating and storage temperatures, and lead temperatures during soldering processes. Specific data sheets must be followed for these maximum ratings.

7.6.2.4 Specific design precautions. All the design precautions discussed in subsection 7.5 CMOS 4000B series microcircuits, paragraph 7.5.2.4 Specific design precautions regarding 4000B series CMOS devices also apply to the HC series. Interfacing HC devices to another logic family should be done according to the input and output requirements of both families.

The HC series of CMOS microcircuits is guaranteed to be protected to 2000 V of electrostatic discharge. However, all microelectronic devices may be electrostatic-discharge sensitive in a somewhat unpredictable manner. Therefore, proper handling and manufacturing process precautions should be followed. Electrostatic-discharge sensitivity is discussed in the subsection 7.1 General.

7.6.3 Physical construction. The HC series CMOS microcircuits are constructed with the planar diffusion process. Basic descriptions of fabrication, assembly, thermal characteristics, reliability, and other details are given in the General subsections for transistors (5.1) and microcircuits (7.1).

7.6.3.1 HC process. Figure 9 shows a cross section of a CMOS inverter used to fabricate HC family devices. This technology uses a full ion implantation, oxide isolation, silicon gate process with projection photolithography, plasma etching, thin film chemical vapor deposition, and sputter metallization. Much higher densities and complexities in circuit design may be achieved compared with previous CMOS designs. The small geometries enable HC devices to operate easily at speeds in excess of 30 MHz.

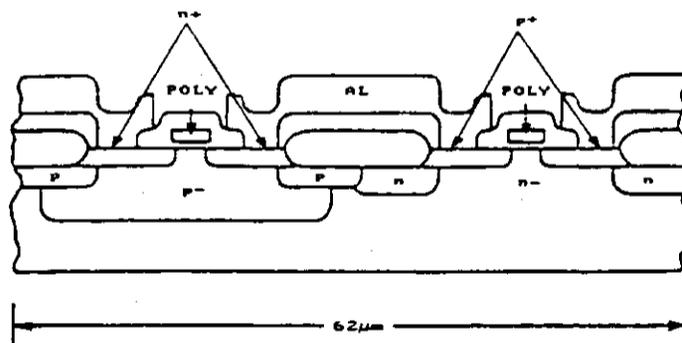


FIGURE 9. Cross section of HC series CMOS inverter with silicon gate, bulk silicon technology.

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7.6 MICROCIRCUITS, HIGH-SPEED CMOS SERIES

7.6.3.2 Packaging. The HC series is available in dual in-line and leadless chip carrier packages discussed in subsection 7.2 Low-power Schottky TTL, paragraph 7.2.3.2 Packaging, with variations in dimensions and number of leads. Additional description of these packages is given in the subsection 7.1 General.

7.6.4 Military designation. Specifications and requirements for standard microcircuits used on military programs are called out in the General Specification for Microcircuits, MIL-M-38510, and associated detail specifications, MIL-M-38510 slash sheets. Descriptions of the designation for military microcircuits are given in subsection 7.1 General.

Military microcircuit parts which have the potential for being used in NASA programs shall have the MIL-M-38510 class S or class B qualification. NASA Grade 1 and Grade 2 parts are listed in MIL-STD-975 (NASA).

7.6.5 Electrical characteristics. The general comments discussed in paragraph 7.2.5 Electrical characteristics also apply to the HC series CMOS electrical characteristics and should be referenced. Table XXII gives typical power dissipation, propagation delay, maximum toggle frequency, and speed/power product for the HC family. Although the HC series has extremely low static power dissipation, this parameter increases with frequency and is comparable to the Schottky dissipation values at high frequencies. Recommended power supply voltage range is 2 to 6 V over the full ambient temperature range.

TABLE XXII. Typical values of performance characteristics of HC CMOS devices at 25 °C

Parameter	HC
Power dissipation per gate (mW)	
Static	2.5×10^{-6}
At 100 kHz	0.17
Propagation delay time (ns), $C_L = 15 \text{ pF}$	8
Maximum toggle frequency (MHz)	
$C_L = 15 \text{ pF}$	40
Speed/power product (pJ) at 100 kHz	1.4

7.6 MICROCIRCUITS, HIGH-SPEED CMOS SERIES

Table XXIII shows fan-out capability for the HC CMOS family. The required input current of 0.001 mA illustrates the general rule that MOS devices need very little input current to be driven active. They are voltage-driven devices. The characteristics discussed here are meant to show only typical values of HC devices. In applications, the specific device data sheets should be reviewed and interpreted for the worst-case conditions under which the circuit will be used.

TABLE XXIII. Fan-out capability of HC series silicon gate CMOS

Parameter	HC Series
Minimum output drive (mA)	4
Maximum input current (mA)	0.001

7.6.6 Environmental considerations. Environmental considerations of microcircuit devices are primarily related to package style rather than device type. They are treated in paragraph 7.1.7 Reliability considerations.

7.6.7 Reliability considerations. Reliability considerations of bipolar digital microelectronic devices are basically the same as those of other microelectronic devices. They are treated in paragraph 7.1.7 Reliability considerations.

7.7 MICROCIRCUITS, INTERFACE

7.7 Interface.

7.7.1. Introduction. Use of interface microcircuits such as analog-to-digital (A/D) and digital-to-analog (D/A) converters makes it possible to use digital techniques in transmitting and processing analog data.

Analog data is converted by means of an A/D converter into a digital form, which can then be used in a computer, digital controller, data logger, or reconverted for analog controls through a D/A converter. The analog control function can be understood when placed in a data acquisition processing system such as shown in Figure 10. In this system, the physical variable measured in analog form by the sensor is amplified, filtered, then periodically sampled by the sample and hold circuit which holds the data voltage level at its output until the A/D converter performs its conversion operation.

The digital data derived from the A/D converter is applied to a line driver which drives an input-output buffer. The buffer, in turn, supplies a digital processor or computer providing digital commands to a D/A converter for eventual analog control or display. Similarly, encoding analog values into digital words (A/D conversion) can be achieved in many ways. These techniques will be covered later.

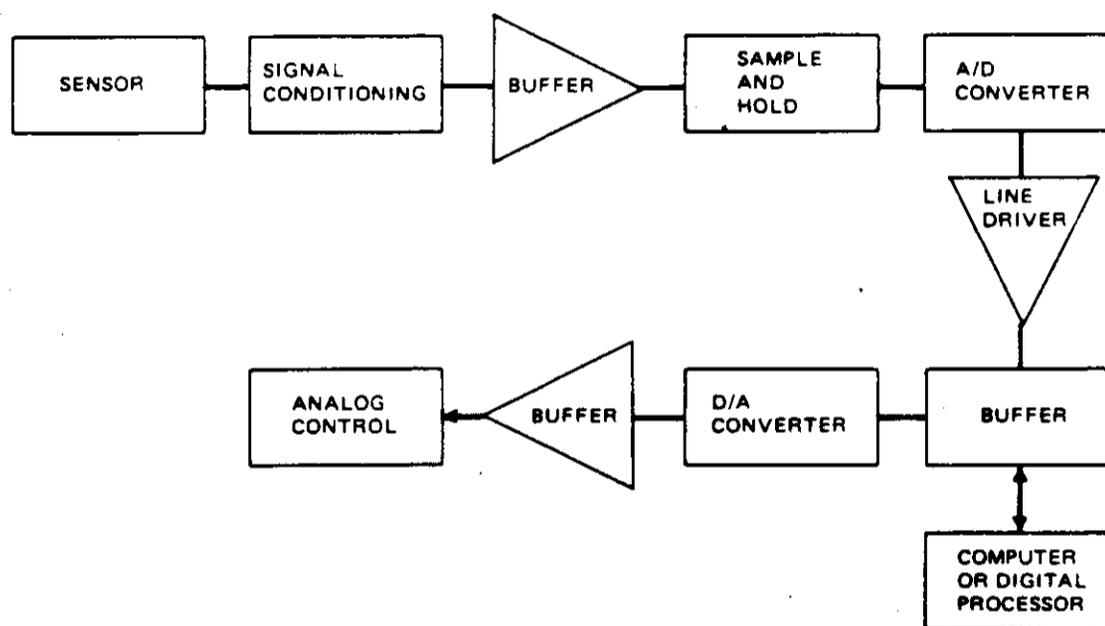


FIGURE 10. Typical data acquisition system.

7.7 MICROCIRCUITS, INTERFACE

7.7.2 Usual applications. Some examples of common applications are given below.

Analog control and monitoring. Analog outputs of strain gauges, thermocouple tachometers and other transducers are digitized by the A/D converter for analysis by a digital system that provides digital commands for conversion to analog control functions through a D/A converter. Monitoring of the analog signal also can be provided by the A/D converter digital system combination.

Data analysis. Single or multiple analog outputs of various transducers are digitized by the A/D converter and processed by a digital system to provide statistical or signal analysis.

Instruments. Measurement instrument outputs are digitized by an A/D converter and then subsequently processed in computation and control circuits. Examples of functions provided are: automatic zeroing, full scale calibration, linearization, computation, signal switching, and display control.

Loop controller. Analog data from 4 to 16 inputs are digitized by the A/D converter and processed by a digital system, which controls the loop through D/A converter analog outputs.

Status monitoring. Up to several hundred analog inputs are sampled by a sample and hold device and then digitized by the A/D converter. From this information, control and monitoring can be initiated by a digital system.

7.7.2.1 Available functional types and technology. Interface microcircuits are available in package types such as discrete component modules, hybrids (thick- or thin-film), and several monolithic package types. By function, they can be divided into a D/A class and an A/D class.

The D/A converters include the following types:

- a. Weighted current source
- b. R-2R
- c. Multiplying
- d. Corrected.

D/A converters presently available in the NASA Standard Parts Program are 8-bit multipliers.

The A/D converters are available in the following types:

- a. Counter
- b. Successive approximation

7.7 MICROCIRCUITS, INTERFACE

- c. Parallel (or flash)
- d. Integrating.

7.7.2.2 Comparative attributes specific to data converters. A/D and D/A converters interface with digital systems by means of a digital code. Although there are many possible codes to select, the most popular is natural binary (straight binary), which is used in its fractional form to represent a number.

$$N = a_1 2^{-1} + a_2 2^{-2} + a_3 2^{-3} + \dots + a_n 2^{-n}$$

Each coefficient "a" assumes a value of zero or one and "n" is the number of bits. The N has a value between zero and one. The leftmost bit has the most weight, 0.5 of full scale, and is called the most significant bit, or MSB. The rightmost bit has the least weight, 2^{-n} of full scale and is called the least significant bit (LSB). The analog value of the LSB is given by the expression

$$\text{LSB} = \frac{\text{FSR}}{2^n}$$

where FSR = the full scale range.

An important point is that the maximum value of the digital code, namely all 1's, does not correspond with analog full scale, but rather with one LSB less than full scale, or $\text{FS}(1 - 2^{-n})$.

Several other binary codes are used with A/D and D/A converters in addition to straight binary. These codes are offset binary, two's complement, binary coded decimal (BCD), and their complemented versions. Each code has a specific advantage in certain applications.

Not only are the digital codes standardized with data converters but so are the analog voltage ranges. Most converters use unipolar voltage ranges of 0 to +5 V and 0 to +10 V although some devices use the negative ranges 0 to -5 V and 0 to -10 V. The standard bipolar voltage ranges are ± 2.5 V, ± 5 V and ± 10 V. Many converters today are pin-programmable between these various ranges.

Table XXIV shows straight binary and complementary binary codes for a unipolar 8-bit converter with a 0- to +10-V analog full scale range. The maximum analog value of the converter is +9.961 V, or one LSB less than +10 V. The complementary binary coding used in some converters is simply the logic complement of the straight binary code.

When A/D and D/A converters are used in bipolar operation, the analog range is offset by half scale or by the MSB value. The result is an analog shift of the converter transfer function as shown in Figure 11. For this 3-bit A/D converter transfer function, the code 000 corresponds with -5 V, 100 with 0 V, and 111 with +3.75 V. Because the output coding is the same as before the analog shift, it is now appropriately called offset binary coding.

7.7 MICROCIRCUITS, INTERFACE

TABLE XXIV. Binary coding for 8 bit unipolar converters

Fraction $\frac{1}{}$ of FS	+10 V FS $\frac{1}{}$	Straight binary	Complementary binary
+FS - 1 LSB	+9.961	1111 1111	0000 0000
+0.750 FS	+7.500	1100 0000	0011 1111
+0.500 FS	+5.000	1000 0000	0111 1111
+0.250 FS	+2.500	0100 0000	1011 1111
+0.125 FS	+1.250	0010 0000	1101 1111
+1 LSB	+0.039	0000 0001	1111 1110
0	+0.000	0000 0000	1111 1111

$\frac{1}{}$ FS (full scale).

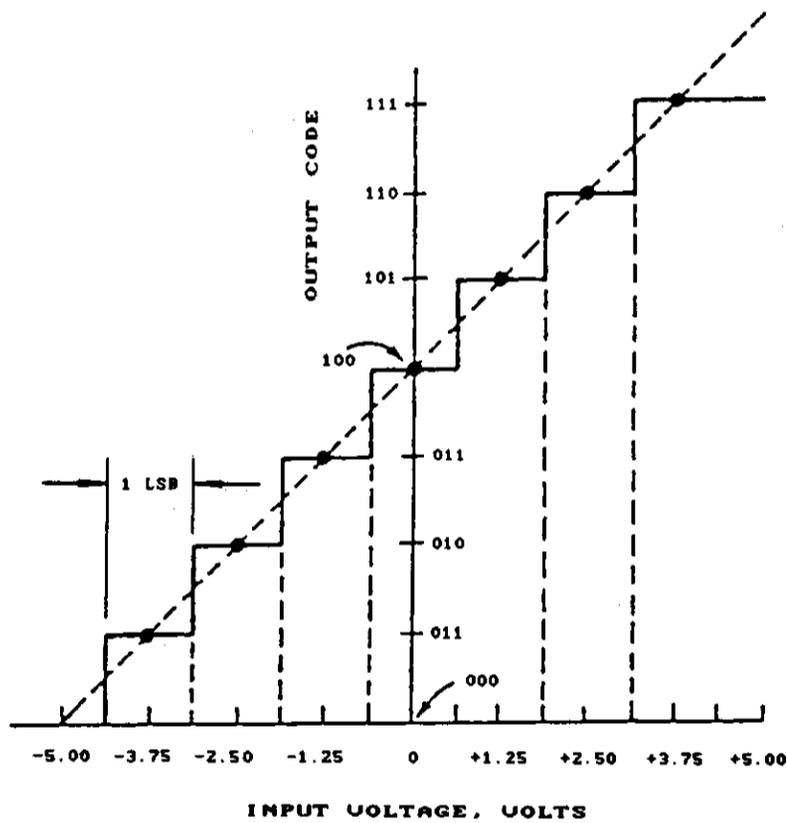


FIGURE 11. Transfer function for bipolar 3-bit A/D converters.

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Table XXV shows the offset binary code with complementary offset binary, two's complement and sign-magnitude binary codes. These are the most popular codes employed in bipolar data converters.

TABLE XXV. Popular bipolar codes used with data converters

Fraction <u>1/</u> of FS	± 5 V FS <u>1/</u>	Offset Binary	Comp. Off. Binary	Two's Complement	Sign-mag Binary
+FS -1 LSB	+4.9961	1111 1111	0000 0000	0111 1111	1111 1111
+0.750 FS	+3.7500	1110 0000	0001 1111	0110 0000	1110 0000
+0.500 FS	+2.5000	1100 0000	0011 1111	0100 0000	1100 0000
+0.250 FS	+1.2500	1010 0000	0101 1111	0010 0000	1010 0000
0	+0.0000	1000 0000	0111 1111	0000 0000	1000 0000 <u>2/</u>
-0.250 FS	-1.2500	0110 0000	1001 1111	1110 0000	0010 0000
-0.500 FS	-2.5000	0100 0000	1011 1111	1100 0000	0100 0000
-0.750 FS	-3.7500	0010 0000	1101 1111	1010 0000	0110 0000
-FS +1 LSB	-4.9961	0000 0001	1111 1110	1000 0001	0111 1111
-FS	-5.0000	0000 0000	1111 1111	1000 0000	--

1/ FS = full scale.

2/ Sign magnitude binary has two code words for zero as shown below.

	Sign-mag binary
0+	1000 0000 0000
0-	0000 0000 0000

The two's complement code has the characteristic that the sum of the positive and negative codes for the same analog magnitude always produces all zeros and a carry. This characteristic makes the two's complement code useful in arithmetic computations. The only difference between the two's complement and the offset binary codes is the complementing of the MSB. In bipolar coding, the MSB becomes the sign bit.

The sign-magnitude binary code, infrequently used, has identical code words for equal magnitude analog values except that the sign bit is different. As shown in Table II this code has two possible code words for zero: 1000 0000 or 0000 0000. The two are usually distinguished as 0⁺ and 0⁻, respectively. Because of this characteristic, the code has maximum analog values of full scale minus one LSB and reaches neither analog + FS nor - FS.

7.7 MICROCIRCUITS, INTERFACE

7.7.2.2.1 D/A converters. Very-high speed internal references and amplifiers are key features of the bipolar technology D/A converters. The CMOS technology D/A converters offer a much higher degree of logic interface function, while maintaining absolute minimums in power dissipation. A wide offering of microprocessor-interfaceable D/A converters simplify connection to 4-, 8-, and 16-bit microprocessor systems.

The reference may be specified as external or internal, fixed or variable, single-polarity or bipolar. If internal, it may be permanently connected or optionally connectible. If the D/A converter is a 4-quadrant multiplying type, the reference is external, variable, and bipolar. The user should check a converter's specifications to determine whether the full-scale accuracy specifications are overall or subdivided into a converter-gain specification and a reference specification.

There are a number of ways in which D/A converters differ in processing input data. The coding may be in any of the following forms: binary, offset-binary, two's complement, BCD, etc. The resolution ranges from 4-bits to 18-bits. The data levels accepted by the converter can be TTL, ECL, or CMOS. If buffer registers are desired, converters with an appropriate buffer configuration should be used.

7.7.2.2.2 A/D converters. A/D converters translate analog input voltages into an equivalent digital value. Manufacturers use both bipolar and CMOS technologies to optimize A/D converter designs. The bipolar technology lends itself to high speed A/D conversion and provides the complete A/D converter function including an internal reference. The CMOS technology offers A/D converter designs with very low power consumption and good interface versatility. At the digital output many interface data formats are available to match the wide variety of application needs. Although no A/D converters are listed in MIL-STD-975, NASA Standard (EEE) Parts List, a brief review will be given for completeness. A/D converters use one of the following conversion techniques.

- a. The counter A/D converter uses a digital counter to control the input of the converter. Clock pulses are applied to the counter and the output of the D/A converter is stepped up one LSB at a time. A comparator compares the D/A output with the analog input and stops the clock pulses when they are equal. The counter output is then the converted digital word.
- b. The successive-approximation converter compares the unknown input with sums of accurately-known binary fractions of full scale, starting with the largest (2^{-1}) and rejecting any that change the comparator's state. At the end of conversion (EOC), the output of the converter is a digital word, representing the ratio of the input to full scale by a fractional-binary code.

7.7 MICROCIRCUITS, INTERFACE

- c. Parallel (also flash, or simultaneous) A/D converters compare the analog signal against 2^n-1 graded voltage levels, using as many comparators, and the comparator output logic levels are processed by a priority encoder, which converts the output to a binary code. Because the whole conversion occurs essentially simultaneously, it is the fastest means of conversion, but it requires many accurate comparators and large numbers of gates.
- d. Integrating types count pulses for a period proportional to the input. Most frequently used are dual slope types, which count off the period required for the integral of the reference to become equal to the average value of the input over a fixed period. Integrating types can be made insensitive to drift by storing errors during an error-correcting cycle and subtracting them during the input-measuring cycle. This correction can be performed in analog fashion, using capacitance for storage, or digitally-using the information stored in a counter for correction.

Whatever the technique, A/D converters comprise several essential functions: an analog section, a digital data-generating section, data outputs, and digital controls.

The analog section requires a reference, one or more high-gain comparators, and either a D/A converter (successive approximations type) or a controllable integrator. The reference may be internal or external, fixed or variable, and of a specified polarity in relation to the analog input.

Successive-approximation A/D converters use the comparator in the current-summing mode; that is, the current output of the D/A converter is summed with the current developed in the D/A converter's feedback resistor by the input voltage of opposite polarity. The balancing action of the converter brings the summing junction towards a voltage null (much like that of an operational amplifier) at the end of conversion.

The typical D/A converter feedback options, when applied in an A/D converter, provide input-scaling choices. When the bipolar-offset connection is jumpered to the summing point, input signals of both polarities can be handled. The current-switching action of the D/A converter, at the typically fast clock rates used in successive-approximation converters, can disturb the output of the analog signal source, especially if it is a slow high-precision operational amplifier. In such cases, buffering may be necessary.

In successive-approximation converters, the digital data-generating section consists of a discrete or integrated successive-approximation register, its controls, and inputs from the comparator and clock. In integrating converters, this section consists of the clock-pulse generator, the counter, the input from the comparator, and the associated controls. Provisions are often made for the pulse train to be jumpered to the counter externally, so that the pulse train can be operated on externally, or can transmit its train of pulses to a remote counter. In a few types of converters, there are no on-board counters or

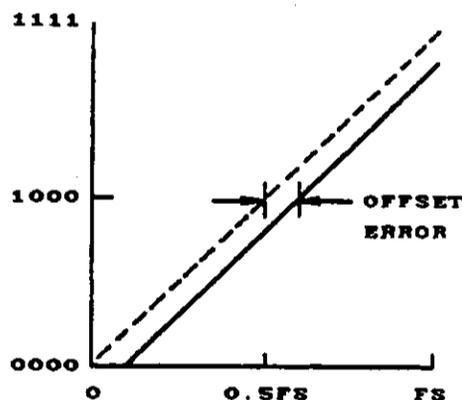
7.7 MICROCIRCUITS, INTERFACE

registers; the pulse train, magnitude, overrange, and control terminals are intended to communicate with external counters and registers.

Data output factors include coding, resolution, overrange information, levels, format, validity, and timing. Coding is usually binary, including jumper-connected offset-binary or two's complement for bipolar input signals. For some types of A/D converters, binary coded decimal (BCD) is available, with sign-magnitude for bipolar inputs. The resolution (number of output bits) ranges from 4 bits to 16 bits. The integrating types of A/D converters generally have no problems with missing codes (except sometimes at zero, with sign-magnitude coding); nevertheless, nonlinear integration can cause the conversion relationship to become nonlinear. Successive-approximation A/D converters have no way of determining overrange; they simply fill up. Some other types roll over and put out a carry flag to signal overrange. The data levels available at the converter output include TTL, ECL, or CMOS. Output formats are available in the following choices: parallel, serial, byte-serial, or pulse-train. A status (or busy) output changes state to indicate when the data becomes valid. The timing diagrams on specification sheets are usually accompanied by adequate descriptions of the conversion process and specifications of the critical interface parameters.

7.7.2.3 Critical parameters. Real A/D and D/A converters do not have the ideal transfer functions. There are three basic departures from the ideal: offset, gain, and linearity errors. These errors are all present at the same time in a converter; in addition, they change with both time and temperature.

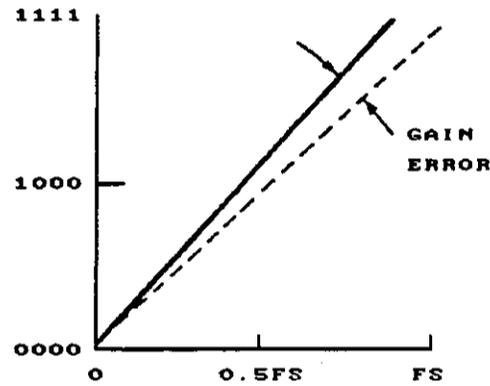
Figures 12, 13, and 14 show A/D converter transfer functions which illustrate the three error types. Figure 12 shows offset error, the analog error by which the transfer function fails to pass through zero. Figure 13 shows gain error, also called scale factor error; it is the difference in slope between the actual transfer function and the ideal, expressed as a percentage of analog magnitude.



FS = full scale

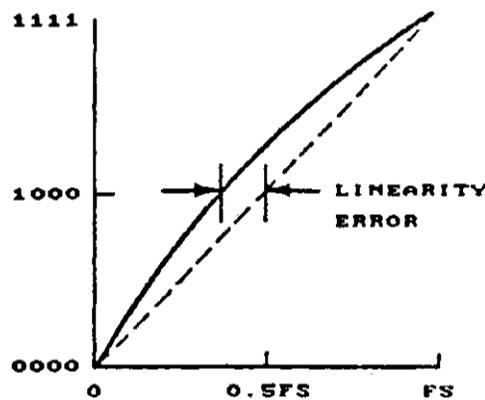
FIGURE 12. Offset error for an A/D converter.

7.7 MICROCIRCUITS, INTERFACE



FS = full scale

FIGURE 13. Gain error for an A/D converter.



NOTE:
1. FS = FULL SCALE

FS = full scale

FIGURE 14. Linearity error for an A/D converter.

In Figure 14 linearity error, or nonlinearity, is shown. This is defined as the maximum deviation of the actual transfer function from the ideal straight line at any point along the function. It is expressed as a percentage of full scale or in least significant bit (LSB) size, such as ± 0.5 LSB, and assumes that offset and gain errors have been adjusted to zero.

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Most A/D and D/A converters available today have provision for external trimming of offset and gain errors. By careful adjustment, these two errors can be reduced to zero, at least at ambient temperature. Linearity error, on the other hand, is the remaining error that cannot be adjusted out and is an inherent characteristic of the converter.

Basically there are only two ways to reduce linearity error in a given application. First, a better quality converter with smaller linearity error can be procured. Second, a computer or microprocessor can be programmed to perform error correction on the converter.

The linearity error discussed above is actually more precisely termed integral linearity error. Another important type of linearity error is known as differential linearity error. This is defined as the maximum amount of deviation of any LSB change in the entire transfer function from its ideal size $\frac{FSR}{2^n}$, where

FSR is the full scale range and n the number of bits. Figure 15 shows that the actual analog step size may be larger or smaller than the ideal. For example, a converter with a maximum differential linearity error of ± 0.5 LSB can have an analog step size between 0.5 LSB and 1.5 LSB anywhere in its transfer function. Integral and differential linearities can be thought of as macro- and micro-linearities, respectively.

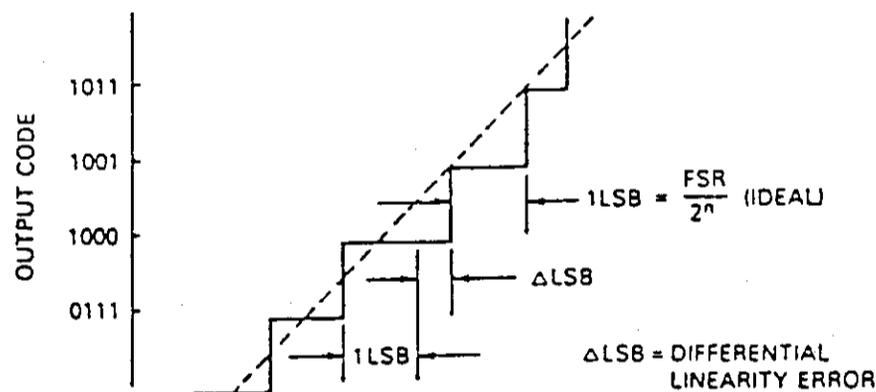
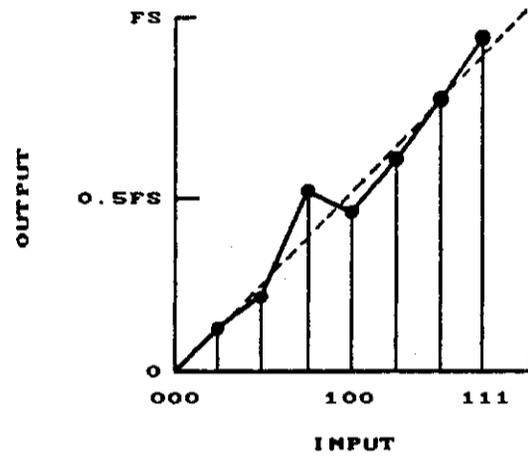


FIGURE 15. Defining differential linearity error.

Two important data converter characteristics closely related to the differential linearity specification are monotonicity and missing code. Monotonicity, which applies to D/A converters, is the characteristic whereby the output of a circuit is a continuously increasing function of the input. Figure 16 shows a nonmonotonic D/A converter output where, at one point, the output decreases as the input increases. A D/A converter may go nonmonotonic if its differential linearity error exceeds 1 LSB. If it is always less than 1 LSB, it assures that the device will be monotonic.

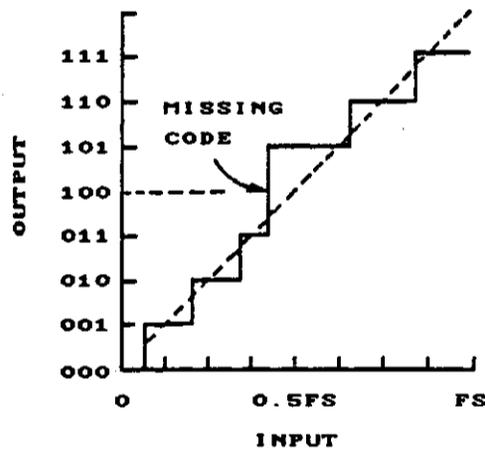
7.7 MICROCIRCUITS, INTERFACE



FS = full scale

FIGURE 16. Nonmonotonic D/A converter.

The term missing code applies to A/D converters. If the differential linearity error of an A/D converter exceeds 1 LSB, its output can miss a code as shown in Figure 17. If the differential linearity error is always less than 1 LSB, the converter will not miss any codes. Missing codes are the result of the A/D converter's internal D/A converter becoming nonmonotonic.



FS = full scale

FIGURE 17. A/D converter with a missing code.

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Ambient temperature change influences the offset gain and linearity errors of a data converter. These changes over temperature are normally specified in ppm of full scale range (FSR) per degree Celsius. When a converter is operated over a significant temperature change, the effect on accuracy must be carefully determined. Of key importance is whether the device remains monotonic, or has no missing codes, over the temperatures of concern. In many cases the total error change must be computed; i.e., the sum of offset, gain, and linearity errors due to temperature.

A number of considerations are important in selecting A/D or D/A converters. An organized approach to selection suggests drawing up a checklist of required characteristics. The checklist should include the following key items:

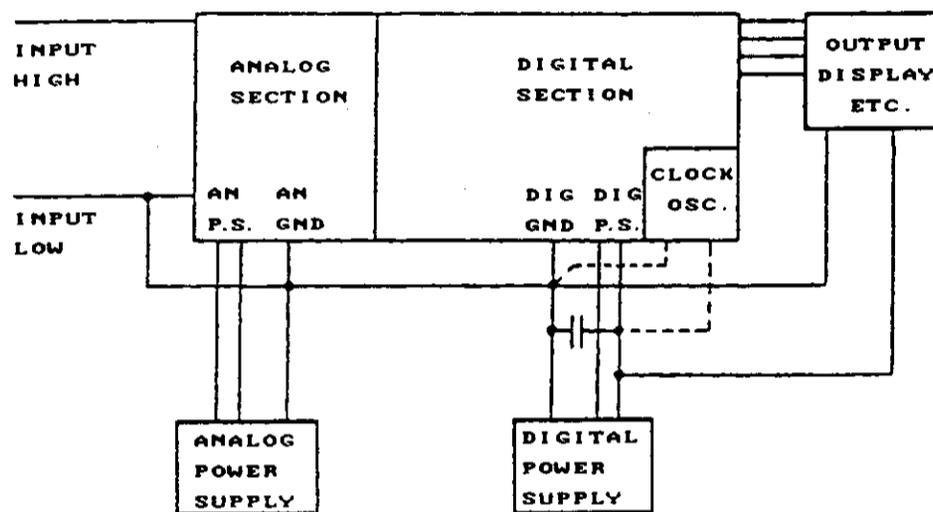
- a. converter type
- b. resolution
- c. speed
- d. temperature coefficient.

After the choice has been narrowed by these considerations, a number of other parameters should be considered. Among these are analog signal range, type of coding, input impedance, power supply requirements, digital interface required, linearity error, output current drive, type of start and status signals for an A/D converter, power supply rejection, size, and weight. These parameters should be listed in order of importance to efficiently organize the selection process.

7.7.2.4 Specific design precautions. The data converter can be considered as a component and, therefore, proper design procedures are necessary to obtain the optimum accuracy. For optimum performance from converters, care should be taken in the hook-up and external components being used. Test equipment used in system evaluation should be substantially more accurate and stable than the system needs to be. Following the precautions listed below will simplify the application of data converters.

- a. Do not introduce ground loop errors. Improper ground is a common source of error in analog or digital systems. Figure 18 shows that the digital and analog grounds are connected by a line carrying only the interface currents between sections, and the input section is also tied back by a low-current line. The display-current loop will not affect the analog section and the clock section is isolated by a decoupling capacitor. External reference return currents must also be returned carefully to analog ground.
- b. Do not couple digital signals into analog lines. For best results, keep analog and digital sections separated on PC boards.

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FIGURE 18. Proper hook-up of an analog-digital system.

- c. Use high quality components. For successive approximation converters, the resistors used should have excellent time and temperature stability to maintain accuracy. Adjustable potentiometers, etc., should be of compatible quality. For dual slope converters, the component selection is less critical. Long term drifts in the integrating resistor and the capacitor are not important. Resistive dividers used on the reference, especially if adjustable, should be of sufficient stability not to degrade system accuracy. Noisy components will lead to noisy performance.
- d. Use a good reference. No converter can be better than its voltage reference.
- e. Watch out for thermal effects. All integrated circuits have thermal time constants of a few milliseconds to dissipate temperature changes in the die. These can cause changes in such parameters as offset voltage and V_{BE} matching even with a carefully designed die. Thermal gradients between microcircuit packages and PC boards can lead to thermoelectric voltage errors in very sensitive systems.

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- f. Use the maximum input scale. To minimize all other sources of error, use the highest possible full scale input voltage. This is particularly important with successive approximation converters, where offset voltage errors can quickly get above 1 LSB, but even for integrating-type converters, noise and the various other sources discussed above will increase for lower than maximum full scale ranges. Preconverter gain is usually preferable for small original signals.
- g. Check these areas. Tie digital inputs low or high if they are not being used. This will avoid stray input spikes from affecting operation. Bypass all supplies with a large and a small capacitor close to the package. Limit input currents into any microcircuit pin to values within the maximum rating of the device (or a few mA if not specified) to avoid damaging the device. Ensure that power supplies do not reverse polarity or spike to high values when turned on or off. Many digital gates take higher-than-normal supply currents for inputs between defined logic levels.

7.7.3 Physical construction. Converters are available in three package types: discrete component modules, hybrids (thick- or thin-film), and monolithic. Each type has its advantages, as well as disadvantages.

The discrete component modules are mounted on printed wiring boards. These are usually encapsulated in an epoxy resin. Relatively low cost units are usually hard potted and are nonrepairable. However, many of the more expensive, high performance modules have a hard epoxy shell and a soft inner potting material. These can be repaired by the manufacturer. Because many of these modules can cost in excess of \$500 per device, being repairable is a definite asset.

The majority of modular devices incorporate epoxy encapsulated active elements (microcircuits, transistors). Hence, their operating temperature range is limited, usually 0 to +70 °C. For high reliability applications such as space flight the active elements are replaced with the hermetically sealed variety.

On an individual basis, the hermetically sealed active elements have an operating temperature of -55 to +125 °C whereas converter modules using them are usually only rated over -25 to +85 °C, or -55 to +100 °C. The primary reason for this difference is usually inability of the converter to meet tight electrical specifications at high and low ambient temperatures.

Due to the great demand for "built-in" high reliability devices as well as smaller physical configurations, many manufacturers have developed hybrid and monolithic versions of their discrete products. Modular devices, which are in the process of being phased out, are today being made for users who need parts to cover their needs for ongoing production and spares.

Power requirements are highest for modular converters and lowest for monolithic converters.

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Monolithic converters are a more recent development and many types are available today. Although they do not yet meet the performance characteristics of hybrids, monolithic data converters are highest in production volume because of their small size, standard packages, low cost, and potential for high reliability. An increasingly greater variety of these devices are being offered by manufacturers to give the designer a wide selection.

7.7.4 Military designation. Description of the NASA Standard Parts Program and the designations for military microcircuits are given in the subsections 1.1 Introduction and 7.1 General.

7.7.5 Electrical characteristics.

7.7.5.1 D/A conversion techniques. Because D/A converters are often used to manufacture A/D converters, the D/A conversion techniques will be discussed first.

D/A converters accept coded digital information at the input and generate an equivalent analog output signal.

The most popular D/A converter design in use today is the weighted current source circuit illustrated in Figure 19. An array of switched transistor current sources is used with binary weighted current. The binary weighting is achieved by using emitter resistors with binary related values of R , $2R$, $4R$, $8R$, ... $2^n R$. The resulting collector currents are then added together at the current summing line.

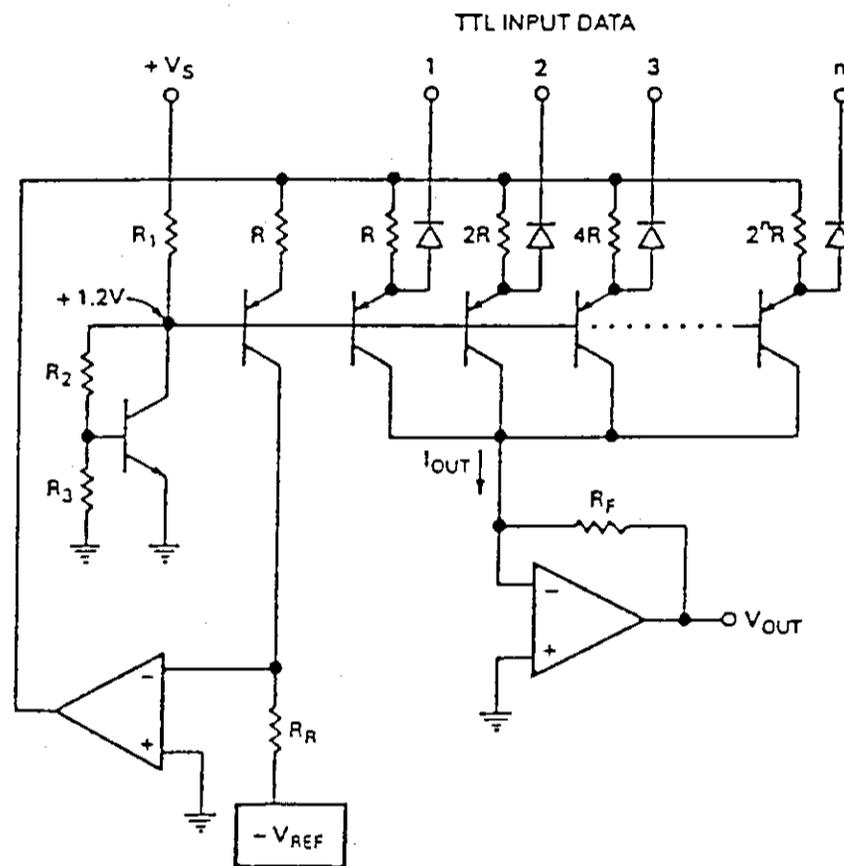
The current sources are switched on or off from standard TTL inputs by means of the control diodes connected to each emitter. When the TTL input is high the current source is on; when the input is low, it is off with the current flowing through the control diode. Fast switching speed is achieved because there is direct control of the transistor current and the current sources never go into saturation.

To interface with standard TTL levels, the current sources are biased to a base voltage of +1.2 V. The emitter currents are regulated to constant values by the control amplifier and a precision voltage reference circuit together with a bipolar transistor.

The summed output currents from all current sources that are on go to an operational amplifier summing junction; the amplifier converts this output current into an output voltage. In some D/A converters the output current is used to directly drive a resistor load for maximum speed, but the positive output voltage in this case is limited to about +1 V.

The weighted current source design has the advantage of simplicity and high speed. Both pnp and npn transistor current sources can be used with this technique although the TTL interfacing is more difficult with npn sources. This technique is used in most monolithic, hybrid, and modular D/A converters in use today.

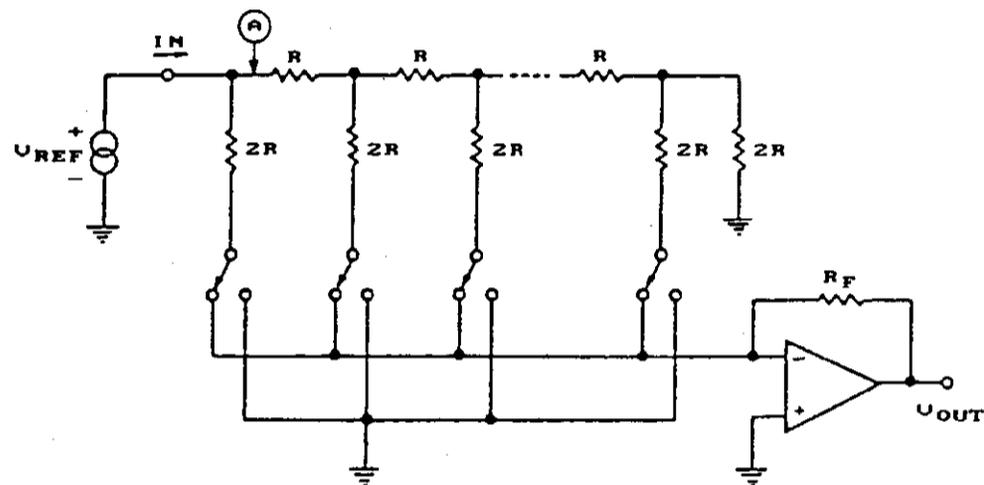
7.7 MICROCIRCUITS, INTERFACE

FIGURE 19. Weighted current source D/A converter.

A second popular technique for D/A conversion is the R-2R ladder method. As shown in Figure 20, the network consists of series resistors of value R and shunt resistors of value 2R. The bottom of each shunt resistor has a single-pole double-throw electronic switch which connects the resistor to either ground or to the output current summing line.

The operation of the R-2R ladder network is based on the binary division of current as it flows down the ladder. Examination of the ladder configuration reveals that at point A looking to the right, one measures a resistance of 2R; therefore, the reference input to the ladder has a resistance of R. At the reference input, the current splits into two equal parts because it sees equal resistances in either direction. Likewise, the current flowing down the ladder to the right continues to divide into two equal parts at each resistor junction.

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FIGURE 20. R-2R ladder D/A converter.

The result is binary weighted currents flowing down each shunt resistor in the ladder. The digitally controlled switches direct the currents to either the summing line or ground. Assuming all bits are on as shown in Figure 20, the output current is

$$I_{OUT} = \frac{V_{REF}}{R} \left[\frac{1}{2} + \frac{1}{4} + \frac{1}{8} + \dots + \frac{1}{2^n} \right]$$

which is a binary series. The sum of all currents is then

$$I_{OUT} = \frac{V_{REF}}{R} (1 - 2^{-n})$$

where the 2^{-n} term physically represents the portion of the input current flowing through the $2R$ terminating resistor to ground at the far right.

The advantage of the R-2R ladder technique is that only two values of resistors are required, with the resultant ease of matching or trimming and excellent temperature tracking. In addition, for high speed applications relatively low resistor values can be used. Excellent results can be obtained for high resolution D/A converters by using laser-trimmed thin film resistor networks.

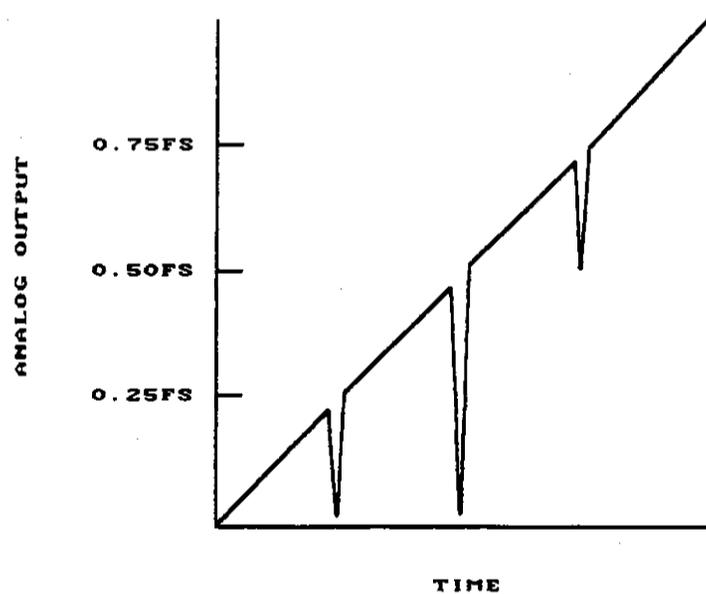
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The R-2R ladder method is specifically used for multiplying type D/A converters such as the DAC-08 listed in MIL-STD-975. With these converters, the reference voltage can be varied over the full range of $\pm V_{max}$ with the output the product of the reference voltage and the digital input word. Multiplication can be performed in one, two, or four algebraic quadrants.

If the reference voltage is unipolar, the circuit is a one-quadrant multiplying D/A converter; if it is bipolar, the circuit is a two-quadrant multiplying D/A converter. For four-quadrant operation the two current summing lines shown in Figure 20 must be subtracted from each other by operational amplifiers.

In multiplying D/A converters, the electronic switches are usually implemented with CMOS devices. Multiplying D/A converters are commonly used in automatic gain controls, cathode ray tube character generation, complex function generators, digital attenuators, and divider circuits.

One other specialized type of D/A converter used primarily in cathode ray tube display systems is the corrected or deglitched D/A converter. All D/A converters produce output spikes, or glitches, which are most serious at the major output transitions of 0.25 full scale (FS), 0.50 FS, and 0.75 FS as illustrated in Figure 21.



Note:
1. FS = full scale

FIGURE 21. Output glitches.

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Glitches are caused by small time differences between some current sources turning off and others turning on. The circuit shown in Figure 22 can virtually eliminate glitches. The digital input to a D/A converter is controlled by an input register while the converter output goes to a specially designed sample-and-hold amplifier. When the digital input is updated by the register, the sample-and-hold amplifier is switched into the hold mode. After the D/A converter has changed to its new output value and all glitches have settled out, the sample-and-hold amplifier is switched back into the sample mode. When this happens, the output changes smoothly from its previous value to the new value with no glitches present.

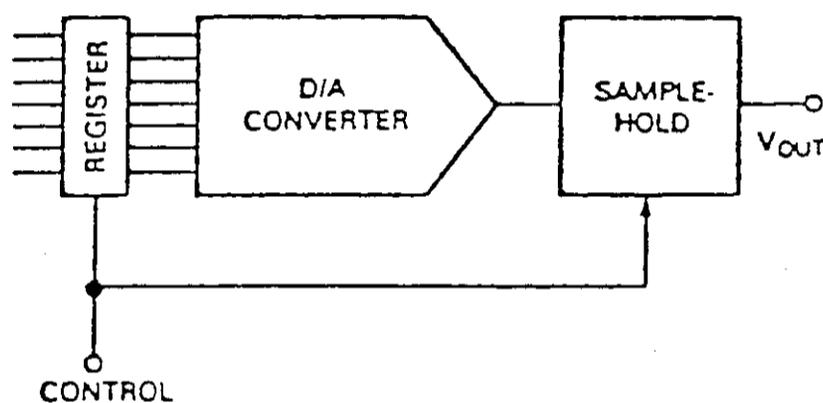


FIGURE 22. Corrected D/A converter.

7.7.5.2 A/D conversion techniques. Analog-to-digital conversion is somewhat more involved than is digital-to-analog conversion. Analog-to-digital converters employ a variety of different circuit techniques to implement the conversion function. Of the various techniques available, the choice depends on the resolution and speed required.

One of the simplest A/D converters is the counter, or servo, type. This circuit employs a digital counter to control the input of a D/A converter. Clock pulses are applied to the counter and the output of the D/A converter is stepped up one least significant bit (LSB) at a time. A comparator compares the D/A converter output with the analog input and stops the clock pulses when they are equal. The counter output is then the converter digital word.

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Although this converter is simple, it is also relatively slow. An improvement on this technique is shown in Figure 23 and is known as a tracking A/D converter, a device commonly used in control systems. Here an up-down counter controls the D/A converter and the clock pulses are directed to the pertinent counter input, depending on whether the D/A converter output must increase or decrease to reach the analog input voltage.

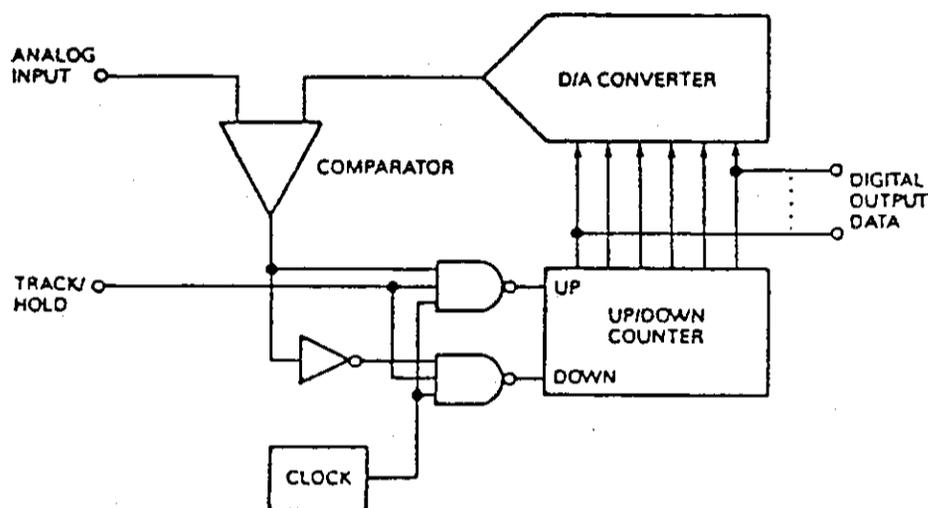


FIGURE 23. Tracking type A/D converter.

The obvious advantage of the tracking A/D converter is that it can continuously follow the input signal and give updated digital output data if the signal does not change too rapidly. Also, for small input changes, the conversion can be quite fast. The converter can be operated in either the track or hold modes by a digital input control.

A popular A/D conversion technique used for moderate to high speed applications is the successive-approximation type A/D converter. This method falls into a class of techniques known as feedback type A/D converters, to which the counter type also belongs. In both cases, a D/A converter is in the feedback loop of a digital control circuit which changes its output until it equals the analog input. In the case of the successive-approximation converter, the D/A converter is controlled in an optimum manner to complete a conversion in just n steps, where n is the resolution of the converter in bits.

In the successive-approximation A/D converter illustrated in Figure 24, a successive-approximation register (SAR) controls the D/A converter by implementing the weighing logic just described. The SAR first turns on the most significant bit of the D/A converter and the comparator tests this output against the analog input. A decision is made by the comparator to leave the bit on or

7.7 MICROCIRCUITS, INTERFACE

turn it off, after which bit 2 is turned on and a second comparison is made. After n comparisons, the digital output of the SAR indicates all those bits which remain on and produces the desired digital code. The clock circuit controls the timing of the SAR. Figure 25 shows the D/A converter output during a typical conversion.

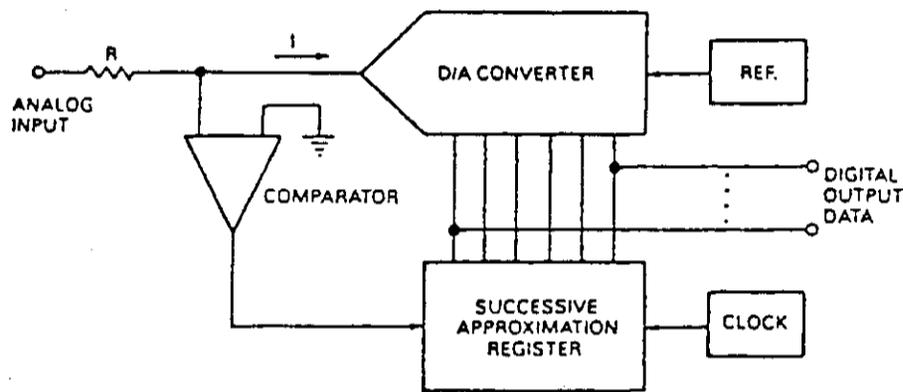
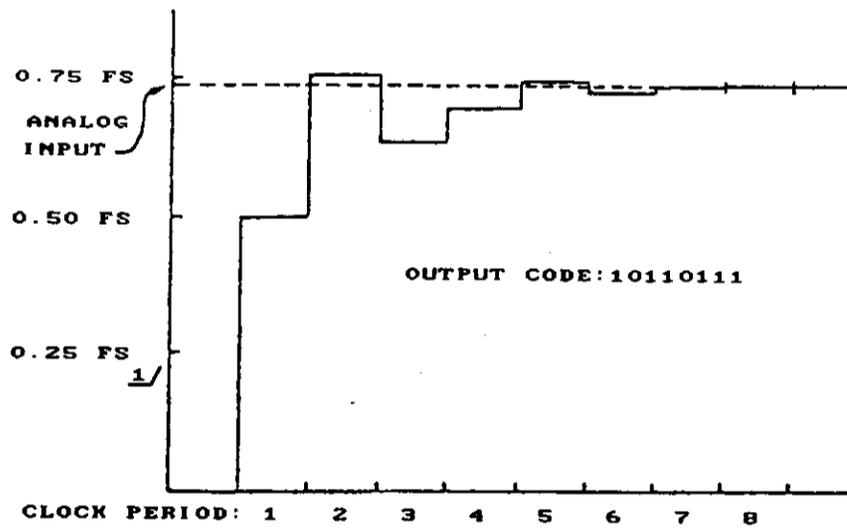


FIGURE 24. Successive approximation A/D converter.



Note:
1. FS = full scale

FIGURE 25. D/A output for 8-bit successive approximation conversion.

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The conversion efficiency of this technique means that high resolution conversions can be made in very short times. For example, it is possible to perform a 10-bit conversion in $1 \mu\text{s}$ or less and a 12-bit conversion in $2 \mu\text{s}$ or less. The speed of the internal circuitry, in particular the D/A and comparator, is critical for high-speed performance.

For ultra-fast conversions required in video signal processing and radar applications where up to eight bits of resolution is required, a technique known as the parallel (also flash, or simultaneous) method is used (see Figure 26).

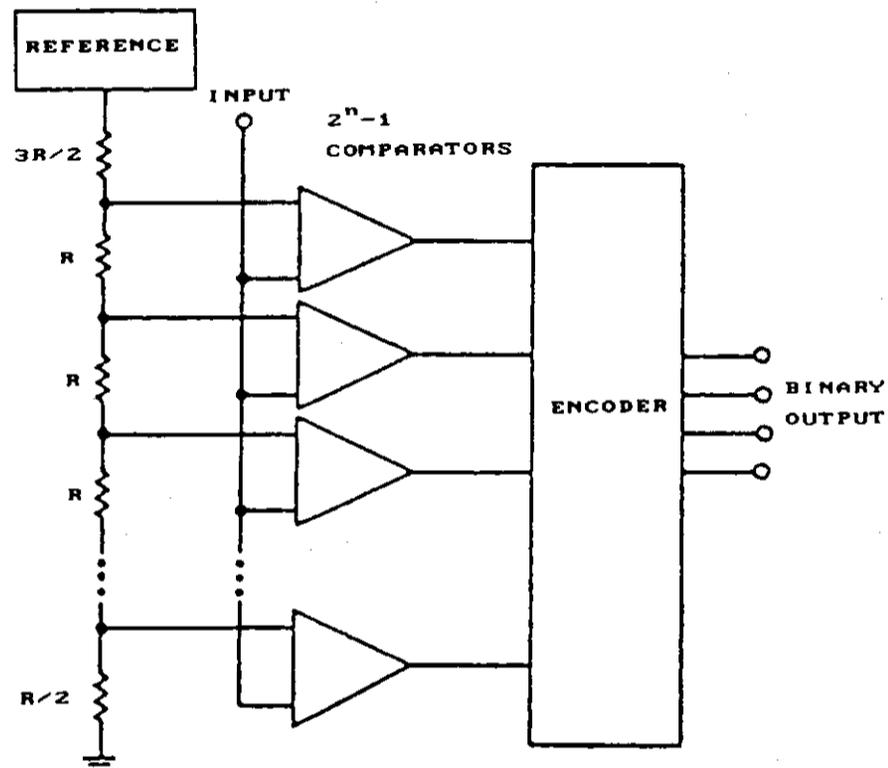


FIGURE 26. 4-bit parallel A/D converter.

This circuit uses $2^n - 1$ analog comparators to directly implement the transfer function of an A/D converter. The comparator trip-points are spaced 1 LSB apart by the series resistor chain and voltage reference. For a given analog input voltage all comparators biased below the voltage turn-on and all those biased above it remain off. Since all comparators change state simultaneously, the quantization process is a one-step operation.

A second step is required, however, since the logic output of the comparators is not in binary form. Therefore an ultra-fast encoder circuit is used to make the logic conversion to binary. The parallel technique reaches the ultimate in high speed because only two sequential operations are required to make the conversion.

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The limitation of the method, however, is in the large number of comparators required for even moderate resolutions. A 4-bit converter, for example, requires only 15 comparators, but an 8-bit converter needs 255. For this reason it is common practice to implement an 8-bit A/D converter with two 4-bit stages as shown in Figure 27.

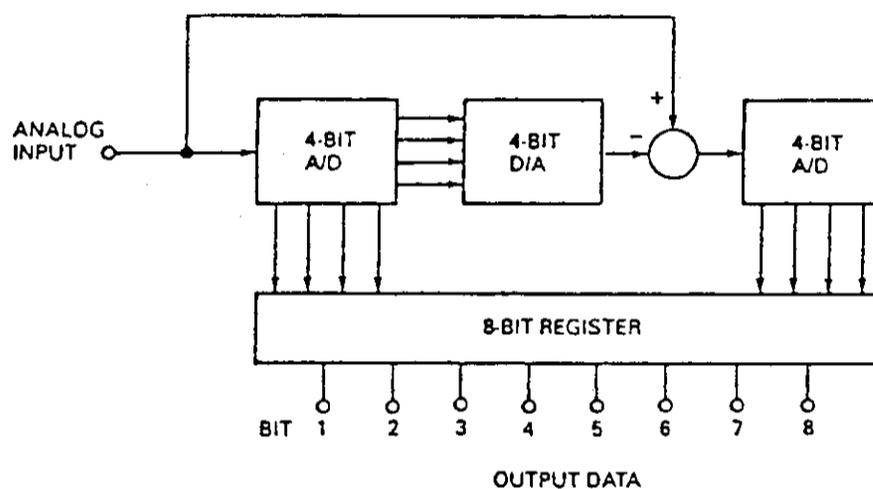


FIGURE 27. Two-stage parallel 8-bit A/D converter.

The result of the first 4-bit conversion is converted back to analog by means of an ultra-fast 4-bit D/A converter and then subtracted from the analog input. The resulting residue is then converted by the second 4-bit A/D converter, and the two sets of data are accumulated in the 8-bit output register. Converters of this type achieve 8-bit conversions at rates of 20 MHz and higher, whereas single stage 4-bit conversions can reach 50 to 100 MHz rates.

Another class of A/D converters known as the integrating type operates by an indirect conversion method. The unknown input voltage is converted into a time period which is then measured by a clock and counter. A number of variations exist on the basic principle, such as the single-slope, dual-slope, and triple-slope methods. In addition, there is another technique, completely different, which is known as the charge-balancing or quantized feedback method.

The most popular of these methods are dual-slope and charge balancing. Although both are slow, they have excellent linearity characteristics with the capability of rejecting input noise. Because of these characteristics, integrating type A/D converters are almost exclusively used in digital panel meters, digital multimeters, and other slow measurement applications.

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7.7.6 Environmental considerations. Environmental considerations are covered in subsection 7.1 General.

7.7.7 Reliability considerations. Reliability considerations for monolithic and hybrid converters are similar to those of other microcircuit types. They are covered in subsection 7.1 General, paragraph 7.1.7. Therefore this section shall be primarily devoted to reliability considerations for modular devices.

A factor to consider is that many converters are optimized at the listed power supply voltages. In many instances, deviations of more than a few percent will cause errors in the output of the converter.

The reliability of modular converters is, of course, dependent upon the reliability of the individual components within it. Therefore, upgrading the individual components enhances reliability considerably.

For NASA applications, the following degree of reliability is recommended.

At the component level:

- a. All transistors and diodes should be screened to JANS or JANTXV level per MIL-S-19500, depending upon the grade of the application.
- b. All microcircuits should be screened to class S or class B of MIL-STD-883, Method 5004, depending upon the grade of the application.
- c. All passive components should be established reliability (ER) types with a minimum failure rate level of "p."

At the module level, the following screens are effective in removing potential electrical/mechanical failure:

- a. High temperature storage for 48 hours at $T_A = 125\text{ }^\circ\text{C}$
- b. Temperature cycling for 10 cycles from -55 to $+125\text{ }^\circ\text{C}$
- c. Burn-in for 160 hours at elevated ambient temperature, with pre- and post burn-in electrical testing.

For hybrid converters, and depending upon the grade of the application, screening to class S or class B per Test Method 5008 of MIL-STD-883 is considered effective in removing potential electrical and mechanical failures. It should be used in NASA applications.

For monolithic converters screening per Test Method 5004 of MIL-STD-883 for class S or class B devices should be used.

7.8 MICROCIRCUITS, MEMORIES

7.8 Memories.

7.8.1 Introduction. This section describes semiconductor memory devices. It will cover several types of memory devices available, semiconductor technologies used in their fabrication, and various applications.

Early investigation of field effect transistors (FETs) was delayed for many years because of difficulties in controlling the surface states of the metal oxide semiconductor (MOS) devices. The discovery and understanding of minority carrier injection across pn junctions led to the development of bipolar transistor technology. As fabrication technologies improved, it became possible to integrate complete circuit functions within a single silicon chip. This resulted in the development of integrated circuit technology which has completely dominated the semiconductor industry which later expanded to large scale integration (LSI) and very large scale integration technologies for the complex digital circuits.

Implementing memory functions using this LSI technology is a logical progression. Because earlier MOS technologies were not refined enough, most semiconductor manufacturers concentrated on the bipolar process as the technology by which they would enter the memory marketplace.

As the demands for greater circuit complexity increased, the use of bipolar devices for complex circuit functions became less attractive. Power consumption became the limiting factor when the microcircuit manufacturers attempted to increase circuit complexity.

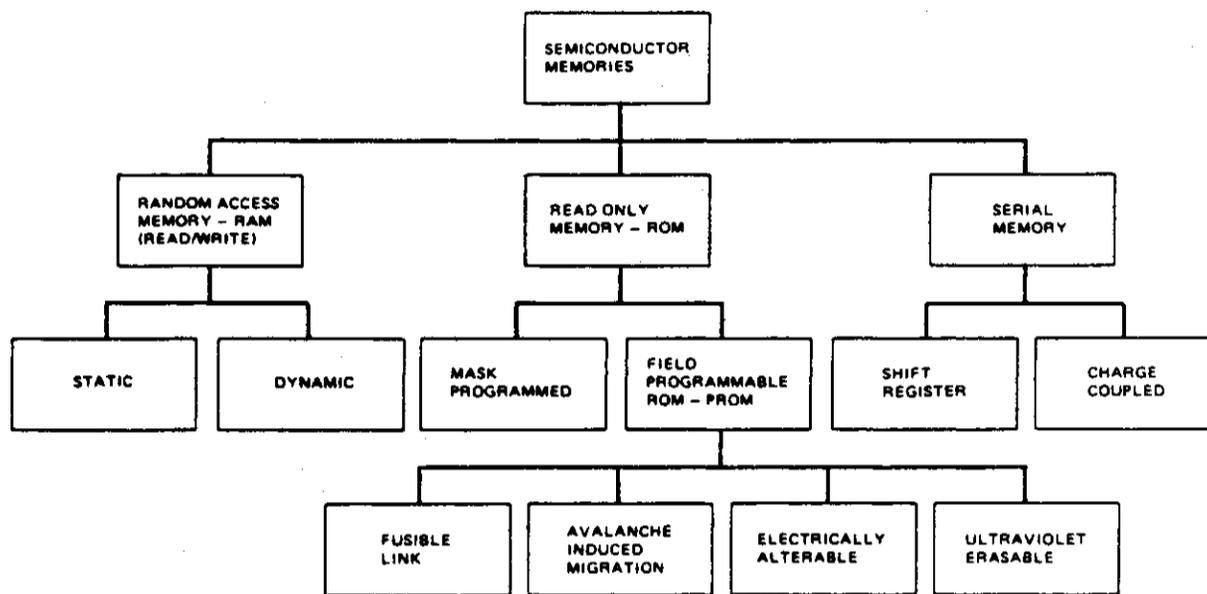
7.8.2 Usual applications.

7.8.2.1 Available functional types and technologies. The application of MOS technology to memory fabrication has triggered an electronic revolution. Although the MOS devices are inherently slower than bipolar devices, their advantages are smaller size, low power consumption, and a relatively simple fabrication process. These advantages have caused rapid proliferation of MOS memories in recent years.

The present semiconductor memories are all integrated circuits. Most are manufactured either in n-channel MOS (NMOS) and complementary MOS (CMOS) processes, or in transistor-transistor logic (TTL), integrated injection logic (I²L), emitter coupled logic (ECL) or advanced low power Schottky (ALS) processes of bipolar technologies.

Memory devices fall into three broad categories as shown in Figure 28. Except for charge coupled devices (CCD), electrically alterable programmable read only memory (EAROM) and erasable programmable read-only memories (EPROMs), which are developments within the MOS technology, all other types can be fabricated by either bipolar or MOS technologies.

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FIGURE 28. Semiconductor memory family.

The most common type of memory is the random access memory (RAM). It is also known as read-write memory where any address location can be chosen at random and the bits stored at that location arrive at the output in approximately the same time for any address used. In other words, this is a coordinate addressable type in which a word can be stored or read from any location independent of the location previously addressed. This is in contrast to the drum, shift register, or delay-line type of memories where the storage locations appear in fixed sequence and reading or writing must take into account the order of appearance of the addresses or storage locations. Although sequential memories are inexpensive, RAMs are much faster and more reliable. They are faster because their search mechanism is electronic instead of mechanical, and more reliable because there are no moving parts. Random access memories are used widely as general purpose data storage in all types of digital equipment. Their shortcoming is that their volatility allows data stored to disappear if power fails or is removed.

The two most popular types of semiconductor memory today are static and dynamic random access memories. As random access memory densities continue to increase, a growing number of memory applications will incorporate static devices. Static RAMs (SRAMs) are becoming more popular because memory systems built around these RAMs are fast, easy to design, and dense. Static memories are also suitable for applications that require high speeds.

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In a static cell, which is usually some form of a bistable latch, the data bit is retained indefinitely as long as power is applied. Dynamic cells, on the other hand, temporarily store charge on a capacitor; i.e., the gate capacitance of a MOSFET. Because there are leakage paths by which the charge representing the data can escape, the dynamic cell must be periodically clocked or refreshed to restore the charge. Because leakage current is a function of temperature, more frequent refreshes are required. In fact, refresh frequency doubles for every 8 to 10 °C rise in temperature.

Examples of typical static cells are shown in Figure 29. The most common bipolar cell is a multiple emitter, cross-coupled, bistable flip-flop with resistive loads. The voltage level of the work line determines the mode of operation. At the 3.0-V level, the standby current, which normally flows from V_{CC} to the word line, is diverted to the bit line of the transistor which is in the on state.

The bit line current is detected by a current sense amplifier during a read cycle. At a word line voltage of approximately 0.3 V and a bit line differential ($V_{BE} - V_{SAT}$), data can be written into the cell. There are many variations of this basic structure, such as the Schottky diode coupled cell which offers lower power consumption and a lower speed-power product.

Figure 29B depicts a six-transistor static MOS cell. $Q_1 - Q_4$ are enhancement mode n-channel devices, whereas $Q_5 - Q_6$ are depletion mode devices. Q_1 and Q_2 form the bistable latch, Q_5 and Q_6 form the depletion type load, and Q_3 and Q_4 connect the cell to the bit and word lines. A logic "0" level on the word line biases Q_3 and Q_4 "off", isolating the cell completely from all other circuitry. To gain access to the cell, the word line must be raised to logic "1" level.

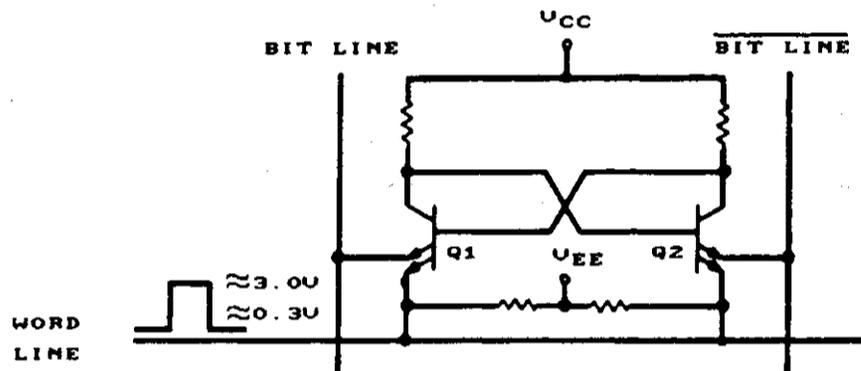
Either Q_1 or Q_2 will be biased "on" which determines the state of the cell. During a read operation, the FET which is "on" will lower the voltage on its bit line. This will be detected by a sense amplifier. During a write operation, the bit lines are driven by a differential amplifier which would bias either Q_1 or Q_2 "on." The state of the bistable latch is maintained by the constant application of V_{CC} .

Dynamic memory cells, in general, require less die area because they can be made with fewer transistors. The three-transistor design shown in Figure 30 has evolved as the standard dynamic cell. Charge is stored by the gate capacitance of Q_2 . The capacitor is shown in phantom in the schematic. This cell structure is currently being used on many 1 K and 4 K RAMs. A single transistor cell has also been developed and offers higher bit per die ratio and is common in 16 K and 64 K random access memory devices.

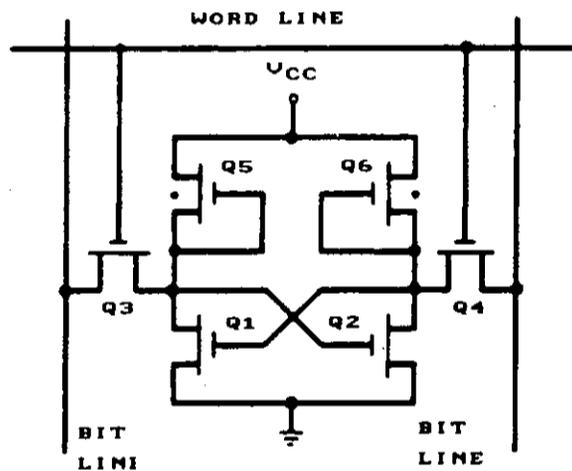
Another popular coordinate addressable variety, the read-only memory (ROM) is a memory that the computer is unable to write data into. In other words, it is a RAM without a write capability. This means that data must be stored into a read-only memory at the time the memory is manufactured or at least at the time it is installed in an application. Its reduced capability makes it cheaper than

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its RAM counterpart. It is still considered a random access device because the access time is independent of data location. ROMs are nonvolatile and stored data is not lost when power is removed.



A. Bipolar

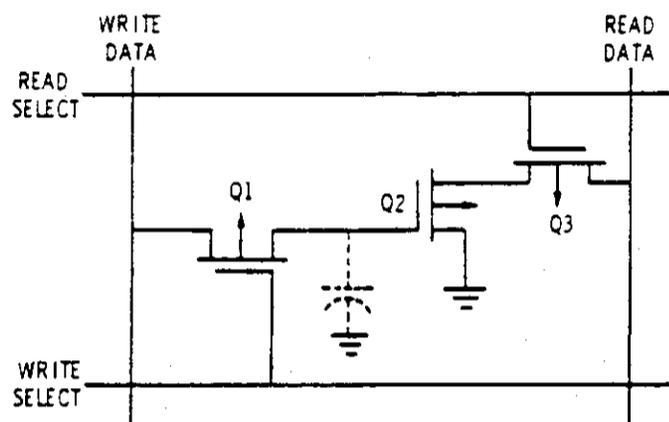


B. MOS

FIGURE 29. Static RAM cells.

Programmable ROM (PROM) devices are programmed by blowing minute fuses associated with the memory cells. The programming can be done by users. However, the memory content cannot be changed once programmed. This device is sometimes called a fuse-programmable ROM. Such PROMs are available in both bipolar and CMOS versions.

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FIGURE 30. Dynamic RAM cell.

The programming mechanism of the fuse-programmable ROM is as follows. A fuse of nichrome with a resistance on the order of a few hundred ohms, is deposited in series with each diode in the matrix. Thus, a conductive path between rows and columns exists for an unprogrammed fusible link PROM. Each bit, therefore, is at a logic "0" level. To program the bit to a logic "1" level, the nichrome fuse is blown to an open condition by a series of programming pulses. In general, the circuitry required between the data bit location and the output causes an inversion of that bit, thus:

<u>State</u>	<u>Bit logic status</u>	<u>Output logic status</u>
Unprogrammed	Low	High
Programmed	High	Low

Manufacturers recommend one to four short duration (100 to 500 μ s) programming pulses at current densities much greater than 2×10^7 A/cm². Under these conditions, the fuse melts rapidly and pulls back from the center, producing a cleaner gap. Problems of regrowth and programming failures can be avoided by using other suitable fuse materials, such as n-doped polycrystalline silicon (approximately 3500 Å thick). Because polysilicon is not a metal, electromigration is not observed.

During device fabrication, the surface passivation layer is removed from the active region of the fuse, permitting contact with the oxygen in the package cavity. This promotes oxidation of the silicon on the opened area, providing an additional dielectric barrier to inhibit regrowth. In addition, these fuses exhibit very well defined breaks. When programming current levels (20-80 mA) flow through the fuse, the polysilicon becomes molten and rolls back, exposing the oxide underneath.

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One problem with polysilicon fuses concerns the inadvertent programming of elements which are supposed to remain unprogrammed. During the programming operation the fuse being programmed sees 20 to 80 mA and up to 12 mA may flow through a nonselected fuse. On rare occasions, 12 mA can cause an element to be programmed.

The EPROM is basically a MOS structure using two polysilicon electrodes: the floating gate and the control electrode. The presence or absence of charge on the floating gate represents the data bits one or zero. It requires 12.5 to 20 V to program, which is higher than the standard 5-V operating voltage.

The stored data of EPROM can be erased by the application of UV light which has a wavelength of 2537 Å. This type of EPROM is called the UV erasable PROM (UVEPROM). It is equipped with a transparent quartz window placed over the packaged silicon chip to allow the user to erase after the chip has been packaged and to reprogram the device. With the UVEPROM, the data can be rewritten after erasure but the process is clumsy and slow. To erase, the chip has to be removed from the system and the entire memory is erased during the erase operation. Furthermore, an expensive package with a quartz window is required. However, this process gives users an opportunity to alter the program, and it is useful during the program development stage.

The structure and operating principles of the EEPROM are similar to those of the EPROM. The floating gate is used in both devices to store charge representing data bits. One important difference is that the EEPROM's charge is moved through the silicon dioxide insulator to the gate by tunneling. The gate of the EPROM is charged by charge injection from the silicon substrate. EEPROM's can be programmed, erased by an electrical pulse, then reprogrammed on a byte-by-byte basis.

The storage element used in UVEPROMs is shown in Figure 31. The silicon gate is floating in the thermal oxide. Operation of this structure as a memory cell is dependent on charge transport to the gate. This is accomplished by injection of electrons from either the source or the drain. Tunneling is not possible because the oxide under the gate is approximately 1000 Å thick. A negative source or drain potential with respect to the substrate is required to effect the charge transport.

Removal of the potential leaves the gate negatively charged. Because the thermal oxide is a very good insulator, the charge is held on the gate. The negatively charged gate induces the formation of a p-channel, turning the field-effect-transistor on.

In order to remove the charge from the gate, the electrons must be excited to a high enough energy level. Manufacturers' data sheets indicate exposure to UV light at an intensity of approximately 10W-s/cm^2 for about 20 minutes will erase the data bits. However, there is concern that these devices are susceptible to erasure by other sources of energy on other wavelengths. The electron activation

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is accelerated further by both operating and storage temperatures. Evaluation of potential problem areas must be accomplished before EPROMs can be considered for applications where they may be exposed to high intensity electromagnetic energy sources.

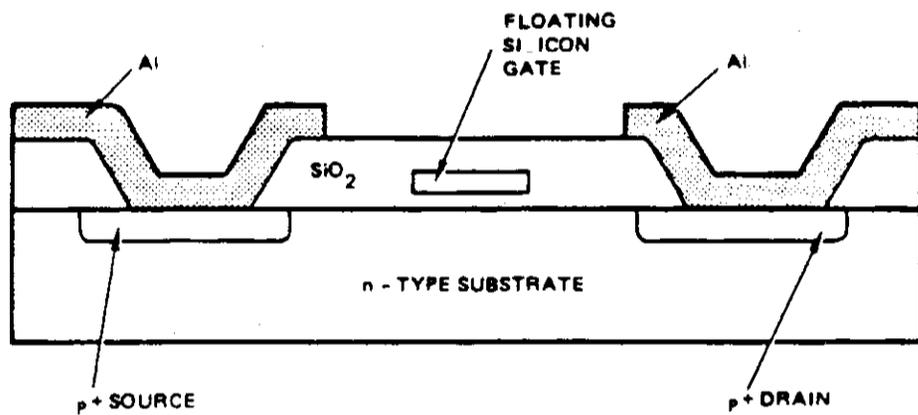


FIGURE 31. Floating gate avalanche MOS structure.

Another type of programmable ROM is the electrically alterable ROM (EAROM). It uses the split gate metal-nitride-oxide semiconductor (MNOS) transistor shown in Figure 32. The only physical peculiarity is that the gate oxide (SiO_2) is very thin (about 30 Å) on one portion of the device. By applying a negative potential to the gate electrode, positive charge can tunnel through the thin oxide to the oxide/nitride interface. When the voltage is removed, the charges are trapped at the interface, effectively altering the threshold level of the device.

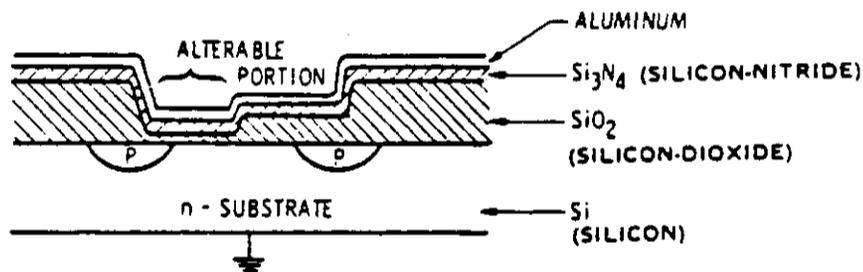


FIGURE 32. Typical cross section of EAROM device.

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Positive charge storage shifts the threshold in the negative direction. This is the mechanism by which data is written into the cell. Conversely, a positive voltage on the gate electrode causes negative charge tunneling, reversing or erasing the data bit.

EAROMs are useful in reduced power applications where data loss is intolerable (e.g., in severe noise environments or during recurring power interruptions). Such devices permit complete or selective writing of bits into "1" or "0" states. The memory can be electrically programmed while it is still in the circuit and alterations can be made without destroying the remaining stored information.

EAROMs are slow (μ s read times) and, for most program storage applications, are expensive. However, because they provide almost infinite storage time, they are being used as auxiliary memory in applications where remote systems are inaccessible for routine field changes and in satellite systems.

Despite greater complexity, CMOS devices are becoming easier to produce, and heat-dissipation considerations are driving the memory industry to use CMOS technology instead of the simpler NMOS structure.

7.8.2.2 Comparative attributes specific to memories. Each type of memory is based on a different operating principle and has its own merits. From a system viewpoint, static RAMs are easier to design in and are faster. They do not need the more complex read/write and refresh circuits required for dynamic RAMs. However, because each cell in a SRAM requires six components, compared with the two required in dynamic RAM (DRAM), the resultant die size and cost are typically three to four times greater. The DRAM is volatile, and all of the stored information is lost when the power is removed. Unlike the dynamic devices, static RAMs do not need to be refreshed, but they also lose memory content when power is removed. SRAMs are less likely to be used in large main memories; the preferred application is in a small portable system and in larger systems with CPU control store and cache memory.

Since its introduction, the EPROM family has rapidly outdistanced its ROM predecessors. EPROMs are a mainstay among nonvolatile memories. They were at one time ultraviolet-erasable programmable ROMs exclusively. However, because of the cost of the expensive package, the UVEEPROM was not as popular as the EEPROM.

EPROMs are keeping pace with DRAMs in the trend toward higher densities. While EEPROMs are competing with EPROMs and ROMs for program storage, lower density EEPROMs are finding applications in numerous devices, from satellite and cable TV decoders to automotive odometers. Unlike the UVEEPROMs that must be removed from a system for erasure and reprogramming in an EPROM programmer, the EEPROM can be erased and reprogrammed without being removed.

EAROMs utilizing MNOS technology are useful in applications where data loss is intolerable (for example, under severe noise environments or with recurring power interruptions), but such devices are expensive and too slow for most program storage applications and are not widely second-sourced.

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The speed-power relationship between bipolar and MOS integrated gates is shown in Figure 33.

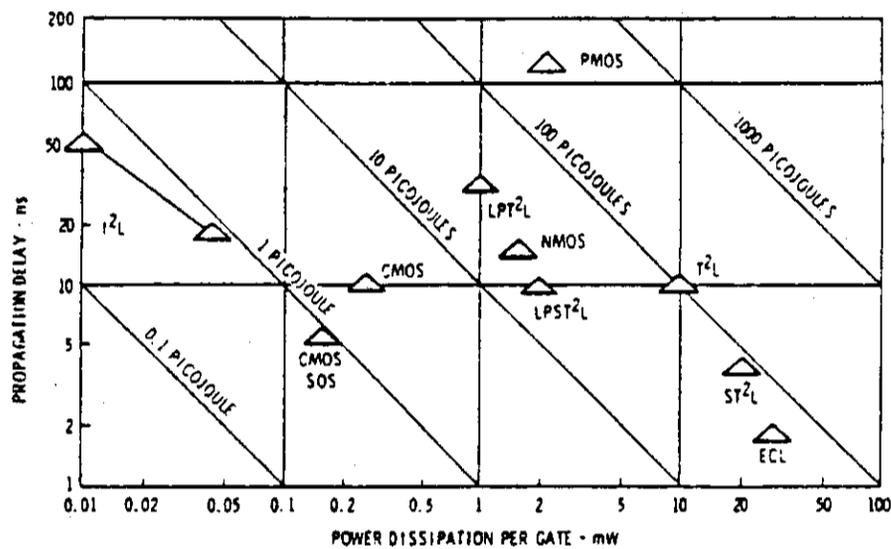


FIGURE 33. Speed-power product of various technologies.

In serial or sequential memories, data bits are not retained in one specific location to be retrieved later. Rather, bits of data are moved, in order, through each cell to the output. Most available shift registers feature bit serial inputs and outputs.

Most shift registers are fabricated using MOS technologies because of the many inherent advantages of MOSFETs such as:

- Extremely low cost per bit
- High input impedance of MOS devices
- Bilateral current flow.

The CCD memory is basically a shift register. It requires periodic refreshing and is used in image sensing and analog signal processing applications.

7.8.2.3 Critical parameters. The following is a list of critical parameters essential to selection of memories. Items marked with an asterisk apply to either RAM-type devices or to programmable ROM-type, but not to both.

MIL-HDBK-978-B (NASA)

7.8 MICROCIRCUITS, MEMORIES

<u>Critical parameters</u>	<u>Typical values</u>
Access time (ns)	50 to 200
Storage capacity (bits)	64 K to 1 M
Operating-temperature range (°C)	-55 to +125
Power consumption (mW/bit)	0.01
*Data retention (life)	10 years at room temperature
*Write/erase cycles for data endurance	1 x 10 ⁴ (min) 1 x 10 ⁶ (max)
ESD sensitivity (V)	1 K to 3 K
*Volatility	Power supply on or off
Read cycle	1 x 10 ¹² (min)
*Write/erase time (ms)	10 (min)
Radiation tolerance	10 ³ to 10 ⁷ RADS (Si)
Technology	MOS/bipolar

7.8.2.4 Specific design considerations. Memory selection depends mainly on the application and is a process of matching and compromising characteristics. A wide range of factors should be considered, such as the critical parameters listed in subsection 7.5, paragraph 7.5.2.3, when selecting memories. A single memory that is cost effective and suitable for all applications is unlikely. The choice of memory type is a tradeoff among system needs such as volatility, power consumption, reliability, capability to support upgrades, ease of manufacture, price, availability, speed, density, architecture, and board space. A major development in the ROM market and a reflection of what is happening throughout the semiconductor industry is a persistent trend toward CMOS design.

Aside from the one-time engineering cost involved in preparing the mask, ROMs are only cost effective when produced in quantity. When an error is found or a change in the code is required, costs are incurred for a new mask and for discarding all of the original parts. On the other hand, the latest generation of EPROMs incorporate features that make them easy to use and quickly programmable using today's programming hardware.

One feature that has contributed significantly to program efficiency is the electronic identifier silicon signature. The electronic identifier allows manufacturers to code their EPROMs so a programmer automatically can identify the device by manufacturer, density and programming voltage, and timing requirements.

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Appropriate cell dimensions and storage chip capacity of memories are shown in Table XXVIA and XXVIB as an example of technology trends.

TABLE XXVI. Memory cell dimensions and memory capacity

(A) Memory Cell Dimension	
Memory	Approximate Cell Dimension (Microns)
Mask programmable ROM	5
UVEPROM	7
DRAM	8
Pseudo SRAM	8
EEPROM	13
Fuse-programmable ROM (PROM)	14
SRAM	18

(B) Memory Capacity			
Memory	Year		
	1979	1983	1987
ROM, EPROM, DRAM	64 K	256 K	1 M
Fuse PROM, SRAM, pseudo SRAM	16 K	64 K	256 K
EEPROM	--	16 K to 64 K	256 K

7.8.3 Physical construction. The two basic types of transistors, bipolar and MOSFET, divide microelectronic circuits into two large families. The bipolar devices were the first to be developed. Although MOSFET was fabricated many years before the bipolar devices, because of the difficulty in controlling the surface state changes of the MOS device, MOSFET did not become practical until the early 1960s.

7.8.3.1 MOS technologies.

7.8.3.1.1 General. The MOS field effect transistor differs from bipolar transistor in that only one kind of carrier is active in a single device. A MOS transistor is also known as a majority carrier device. Devices that use electrons as carriers are called NMOS transistors and those that use holes are PMOS transistors.

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Figure 34 is a cross-section of PMOS and NMOS transistors in enhancement and depletion modes. In the case of the NMOS transistor, two islands of n-type material (one called the source and the other the drain) are diffused into the p-type silicon. A silicon dioxide (SiO_2) layer is present on the surface of the channel lying between the source and the drain, and a metallic layer called the gate lies above the SiO_2 .

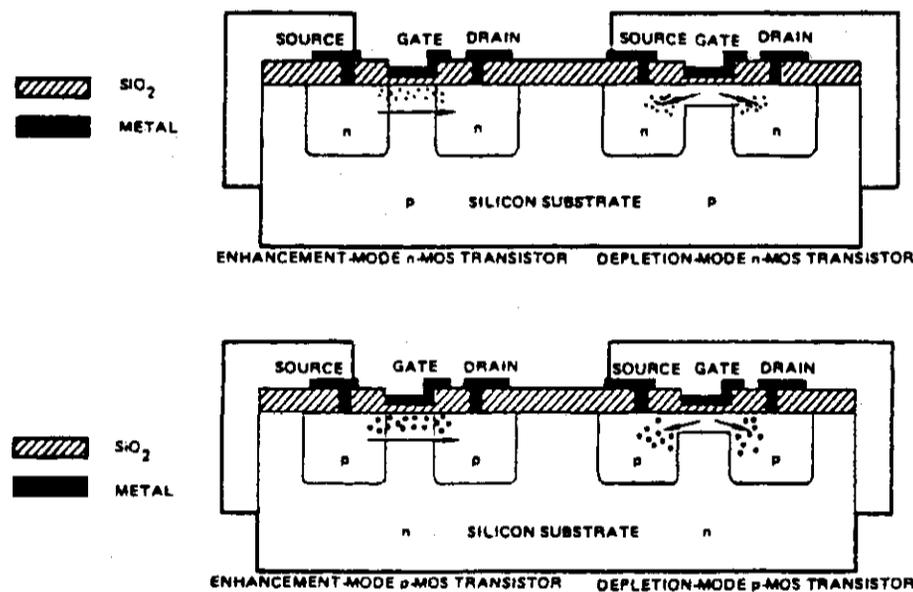


FIGURE 34 . Conventional metal-oxide-semiconductor field effect transistors (MOSFET).

7.8.3.1.2 Metal gate vs silicon gate. Aluminum has most of the desired properties in that it is a good electrical conductor and gives good ohmic contact to the P diffused areas. Aluminum adheres well to SiO_2 , has a small work function difference with doped silicon (around 0.3 V) and is obtainable in very pure form. Its bad features are that it scratches easily and cannot withstand the high temperatures required for boron diffusion. One way to obtain a low threshold voltage is to fabricate the gate structure from heavily doped silicon instead of metal.

Silicon gates are made of p-type polycrystalline silicon, whose work function is less than that of the aluminum used in ordinary MOS circuits. The difference between work functions of the gate and the semiconductor, therefore, is less and this influences the threshold voltage both directly and through a reduced surface state charge. In addition, silicon gates offer two advantages in fabrication: automatic gate alignment, and the possibility of mixing both bipolar and MOS circuits on the same substrate.

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The silicon-gate (Si-gate) process also has drawbacks. For example, the silicon deposition is an extra step as metal deposition is still required. External connections can't be made directly to the silicon. An additional metal layer must be deposited so that one can make external connections to the silicon. However, with the silicon layer isolated from the metal, the silicon achieves an extra level of intra-connection directly on the chip, thereby reducing the number of outside connections and the area of circuits.

7.8.3.1.3 Silicon nitrides vs silicon dioxide gate structure. The addition of silicon nitride to the silicon dioxide layer increases the dielectric constant by a factor of two, which in turn increases the gate capacitance, resulting in a lower threshold level. Silicon nitride is not deposited directly because it gives too high a Q_{SS} value (surface state charge), and large hysteresis effects are observed in the gate-voltage vs drain current characteristics. To avoid this, a thin layer of SiO_2 is grown before the Si_3N_4 is deposited so that we have a Si_3N_4/SiO_2 sandwich as a gate structure.

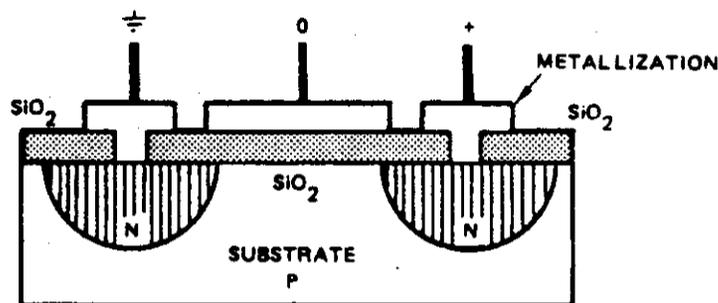
Another factor to consider is that at the same drive current (I_D), the silicon nitride device occupies only half the area. Sometimes Si_3N_4 passivation eliminates the instability caused by sodium ion contamination applied over a sodium-free oxide. The impervious nitride keeps the harmful sodium away from the chip. It is possible that a Si_3N_4/SiO_2 sandwiched gate structure is more likely to fail under voltage stress, because of partial pinholes. In addition, the silicon nitride process complicates fabrication; extra steps are required and the two dielectric materials may etch at different rates.

7.8.3.1.4 N-channel MOS transistor silicon gate. In an n-channel device, the conduction carriers are electrons, which have a mobility approximately three times that of holes. This results in faster devices. The higher mobility also means that channel current densities are higher. Therefore, n-channel devices can be made smaller and still carry the same current as p-channel devices.

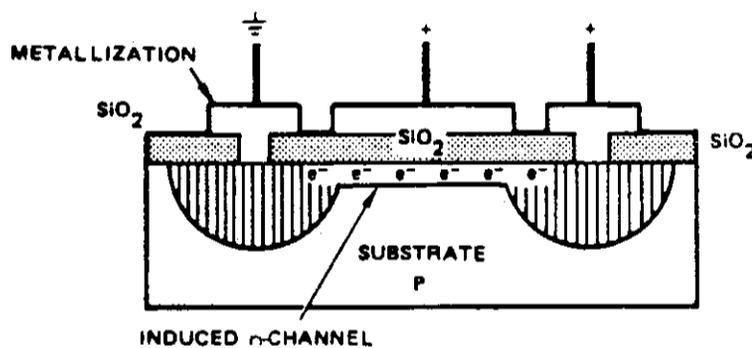
In the n-channel device as shown in Figure 35A, the drain and source are n-type islands in a p-type substrate. The enhancement mode MOSFET is off when a zero gate bias is applied. In Figure 35B when a positive gate bias is applied, electrostatic attraction brings electrons from the bulk of the silicon to the Si/SiO₂ interface. If enough of these electrons congregate along the interface, they invert the polarity of a layer of p-type silicon. This layer or channel provides a low-resistance path from drain to source, turning the FET on.

Ion implantation allows the construction of true enhancement mode NMOS transistors. This process permits more precise control of device geometry and doping levels than diffusion type processes. The result is that NMOS devices now compete successfully against bipolar in the race to decrease access times.

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- A. The enhancement-mode MOSFET is off when a zero gate bias is applied. In the n-channel device shown, the drain and source are n-type islands in a p-type substrate.

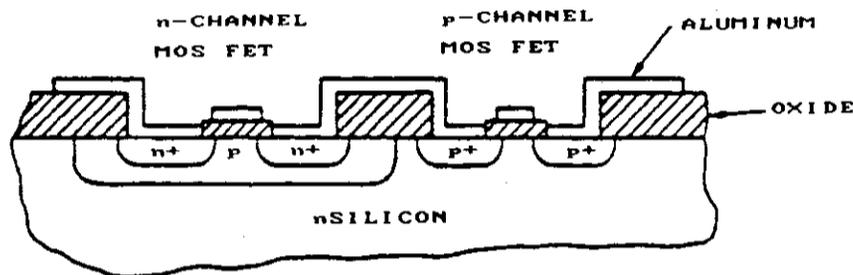


- B.

FIGURE 35. Cross section of n-channel MOS transistor.

7.8.3.1.5 Complementary MOS (CMOS). The CMOS is a version of the MOS technology that includes both n-type and p-type transistors. Figure 36 is a cross section of a CMOS device. When the device is fabricated in an n-type substrate as illustrated, a p-type channel transistor is made in the same manner as described for the MOSFET device, but note that an NMOS transistor requires an island of p-type material. This island requires an additional processing step during fabrication.

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FIGURE 36. Cross section of CMOS device.

Further illustrated in Figure 37 is a CMOS inverter device arranged to achieve low power consumption. This configuration has the gates of both transistors connected to a single input, since the two different type (n vs p) transistors require opposite polarity for conduction. They are never turned on at the same time and, therefore, little current flows from the power supply to ground and results in low power consumption. Both transistors are enhancement mode MOSFETs (normally off). CMOS is a mixture of n- and p-channel transistors on the same device. In conventional bipolar and p-channel MOS ICs, most of the power dissipation is quiescent power contributed by the load resistor. In the complementary-symmetry circuit, the transistor's load resistor is replaced by a transistor of opposite polarity. This results in considerably lower quiescent power dissipation.

The complementary inverter, shown in Figure 38, will be used to illustrate the basic operation of CMOS circuits. In this circuit, when V_{IN} is equal to V_{DD} , the n-channel device is turned on, the p-channel device is off and the output is near ground potential. Conversely, when the input voltage is at ground, the n-channel is off, the p-channel device is on and the output is near V_{DD} potential. Note that only during the actual switching period is there a direct connection between V_{DD} and ground. Thus, under static conditions, essentially no power is dissipated in the circuit.

The standby power dissipation of CMOS circuits is on the order of nanowatts. However, when a CMOS circuit switches, a considerable amount of current flows during the switching interval. Consequently, power consumption of CMOS circuits increases with increasing frequency and is, for example, comparable to Schottky device power dissipation at approximately 10 MHz or higher. Good operating speeds are expected from devices built with the complementary symmetry process because the output node capacitance is always charged and discharged through the "on" transistor.

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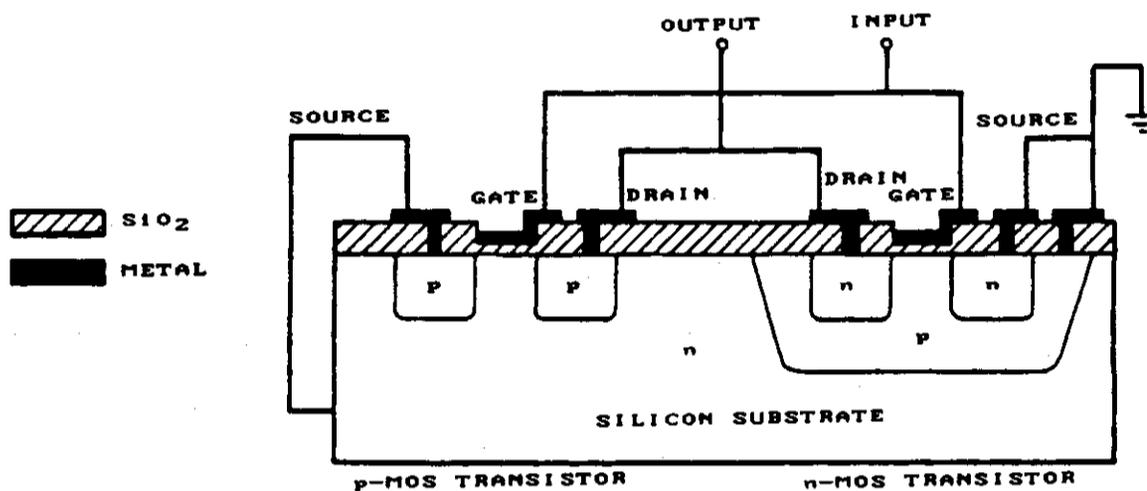


FIGURE 37. CMOS inverter.

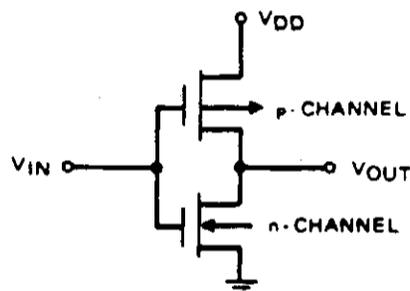


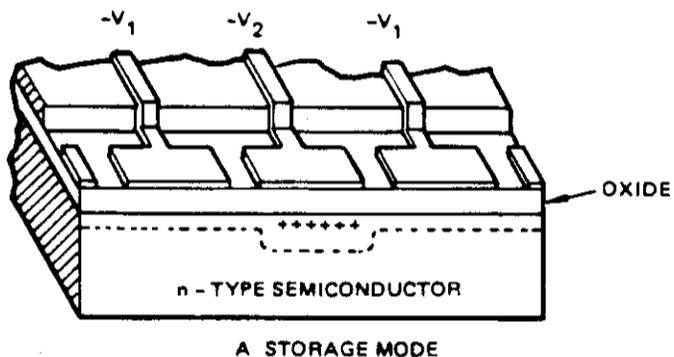
FIGURE 38. Complementary MOS inverter circuit.

7.8.3.1.6 Silicon-on-sapphire (SOS). The SOS technology is a drastic reduction in parasitic capacitance between elements of the field effect transistors, which translates to higher speed. The reduction in parasitic capacitance is a direct result of fabricating MOS structures on an insulating layer of sapphire. The elimination of an active reverse-biased junction to the substrate reduces the inter-element capacitance by a factor of 2 to 3.

7.8 MICROCIRCUITS, MEMORIES

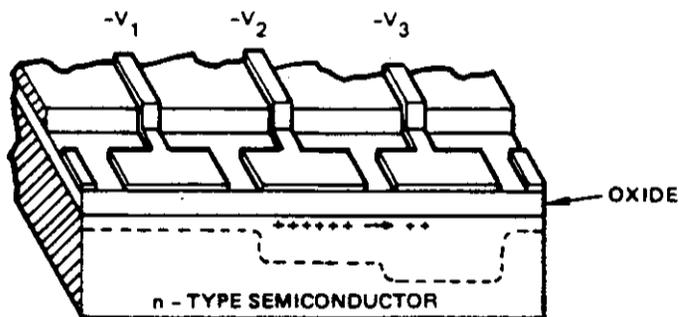
7.8.3.1.7 Charge coupled devices (CCD). CCDs are one of the latest developments within the MOS industry. The operational mechanism involves the transfer of discrete quantities of charge from one location to another. A basic charge storage element is a MOS capacitor. In a CCD, a large number of these capacitors are fabricated in close proximity. Charge packets are moved, in serial fashion, by applying multi-phase clock pulses to the storage elements in a prescribed order.

A typical cross section of a CCD showing the transfer of a charge packet is shown in Figure 39. Charge is stored in a potential well created by applying a negative voltage, $-V_1$, to the two outer elements and a more negative voltage, $-V_2$, to the center element.



A STORAGE MODE

A. Storage mode



B TRANSFER MODE

B. Transfer mode

FIGURE 39. Charge coupled device.

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To transfer the charge to the element on the right, $-V_1$ is reduced to a level $-V_3$ (more negative than $-V_2$). This creates a deeper potential well which causes the holes in the depletion region to drift so that they gather under the plate at $-V_3$. To complete the transfer cycle $-V_2$ is raised to $-V_1$ and $-V_3$ is raised to $-V_2$. The transfer of data has been explained in static terms. However, in reality, a three-phase clock would be used to transfer the data.

One parameter of greatest importance is transfer efficiency. That is, what percentage of the charge is successfully transferred. Because there is no amplification in a CCD, the transfer efficiency can never equal 100 percent. Devices currently being fabricated exhibit a transfer efficiency greater than 99.9 percent.

CCDs can be fabricated using two techniques, surface channel or buried channel. The primary difference between the two is that the buried channel device stores and transfers charge deeper than the substrate bulk due to additional doping. Generally speaking, surface channel devices are easier to fabricate and have a greater charge carrying capability while buried channel devices offer a better transfer efficiency, especially at higher transfer frequencies.

Although CCDs offer extremely good density and a low speed power product, their primary drawback is that they are a dynamic device, and thus, need periodic refreshing. The high refresh rate requirement increases the overall power dissipation of the device. Table XXVII compares some memory parameters between n-channel silicon gate and CCD devices.

TABLE XXVII. N-channel silicon gate and CCD memory parameters

Parameter	N-Channel Silicon Gate	CCD
Area/bit (square mils)	5 to 8	1.5 to 3
Power/bit (μ W)	70 to 100	5
Processing	Moderately difficult	Easy

7.8.3.2 New bipolar technologies.

7.8.3.2.1 General. Bipolar has been the leader in small-scale-integration (SSI) and medium-scale-integration (MSI). Improvements in the processing and design produced low power TTL, high speed TTL, Schottky TTL, low power Schottky TTL, advanced Schottky TTL, and advanced low power Schottky TTL. These variations give the application engineer great flexibility in selecting the most suited device for end requirements.

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7.8.3.2.2 Integrated injection logic (I²L). I²L, also known as merged-transistor logic (MTL), represents the most promising development in bipolar technology. I²Ls major strengths are density (more dense than MOS) and a speed power product on the order of 0.2 to 2 pico joules per gate.

An example of an I²L gate structure is shown in Figure 40. It consists of a lateral pnp transistor and a vertical npn transistor. In I²L, the vertical npn transistor is operated in an inverse mode. That is, what would normally be a multiple emitter structure is now a multiple collector structure. The pnp transistor acts as a constant current source, driving the npn device.

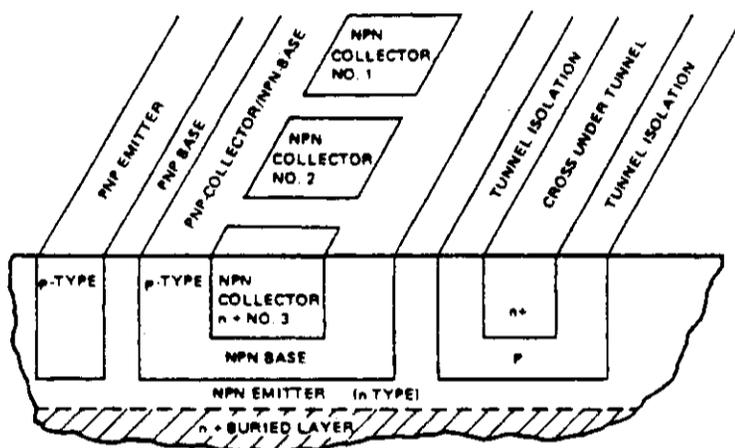


FIGURE 40. Integrated injection logic configuration.

The number of processing steps required to fabricate an I²L gate is less than that required for TTL, CMOS, or n-channel silicon gate. Only four mask steps and two diffusions are required, and the resulting gate is the smallest structure currently being produced. The entire gate area is less than 5 sq mils. However, as far as the speed is concerned, it still lags behind MOS and conventional bipolar devices.

7.8.3.3 CMOS vs bipolar technology. Presently, CMOS circuit densities are approximately the same as bipolar technologies. CMOS circuits draw considerably less power, approximately one-thousandth, than the bipolar equivalents at nominal operating frequencies. However at higher frequencies, of the order of 10 MHz and higher, CMOS power consumption is approximately the same as bipolar. CMOS has much lower driving capability compared to bipolar. For a comparison of CMOS and bipolar radiation hardness, see radiation considerations in subsection 7.1 General. Table XXVIII gives a broad comparison of the CMOS technology to NMOS, CCD, and I²L technologies.

7.8 MICROCIRCUITS, MEMORIES

7.8.3.4 Summary. Table XXVIII compares the processes which have been discussed on several items of general interest. All ratings are relative within the category of comparison.

TABLE XXVIII. Comparison of MOS and I²L technologies

Parameter	N-Channel Silicon Gate	Complementary (CMOS)	CMOS/SOS	CCD	I ² L
Processing	Very difficult	Difficult	Difficult	Easy	Easy
Size	Very small	Large	Large	Small	Very small
Speed	Very fast	Fast	Fast	Fast	Fast
Threshold	Low	Low	Low	High	--
Field inversion	Low	Low	Low	Low	--
Power consumption	Low	Very low	Very low	Very low	Very low
Temperature stability	Good	Good	Good	Not established	Good
Recurring cost	High	Average	Very high	Low	Average
Available LSI military temperature parts	Yes	Yes	Yes (few)	No	Yes
Comments	Flexible	Flexible; high noise immunity			

7.8.4 Military designation. Military designation of memory devices follows the same designation under MIL-M-38510 as other microcircuits and is described in the general section. Qualification to a specific slash sheet has been established for class B (NASA Grade 2).

7.8.5 Electrical characteristics. Device characteristics of MOS and bipolar memories are treated in subsection 7.5, paragraphs 7.5.2.1 Available functional types and 7.5.2.2 Comparative attributes specific to memories.

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7.8.6 Environmental considerations. A phenomenon that has recently surfaced regarding memory devices is their susceptibility to alpha particle radiation due to the low operating signal levels. Current 16 K dynamic MOS RAMs and CCDs have been found to exhibit a nonrecurring single bit failure mode that has been labeled a soft-error. These soft-errors are the result of an alpha particle passing through a memory data cell or any associated control circuitry (sense amplifiers, data lines, decoders, etc.). The prime source of the alpha-particle has been traced to the ceramic package. Further, the angle of incidence at which particles enter the die surface indicates that the package lid is the major offender. Improvements by the package manufacturers have provided materials with alpha particle flux densities of 0.04 to 0.37. The range indicates that various densities may be acceptable relative to the soft-error rate generated. Manufacturers have further dealt with the problems by increasing the storage cell capacitance, thus increasing the actual signal levels. Also overcoats on the die and the use of metal package lids have reduced the occurrence of soft errors. Finally, the user may increase the device supply voltages, thus increasing internal signal level and may further decrease the operating cycle time, thus reducing the time that data travels from storage cells to associated control circuitry. When these options are not tolerable, error-correction coding may be necessary in order to eliminate or reduce the soft error rate. Unfortunately the error-correcting code will use a portion of the actual memory array. Additional discussion of radiation is in subsection 7.1 General.

7.8.7 Reliability considerations. Reliability considerations for memory devices depend upon the technology used to fabricate the device and the package, and materials used for assembly. In addition to Reliability considerations of subsection 7.1 General, further cautions must be exercised.

In the case of using UVEPROMs, accidental erasure is a serious concern. After programming, avoid any exposure to light in general; an opaque cover over the package window is suggested. Further, storage temperatures greater than 125 °C will accelerate electron activation and must be avoided to preclude undesired data erasure. Memory cell data retention may be tested by burning in devices with a known data pattern. Duration may be acceptable from 24 to 160 hours at $T_A = 125$ °C.

Dynamic RAMs, as discussed earlier, if operated beyond 70 °C will require the refresh frequency to be doubled. Further, both memory devices fabricated with MOS technologies are sensitive to device surface and interface contaminants often found in ionic form. Static burn-in at $T_A = 125$ °C for 24 to 160 hours are effective in detecting the ionic contaminated devices. A caution must be raised in that the static burn-in alone is not effective in detecting other possible defects such as oxide pinholes, which may be recognized with a dynamic burn-in.

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7.8 MICROCIRCUITS, MEMORIES

Descriptions of some typical functional memory device failures are listed below (mostly related to RAMs):

- a. Cell opens and shorts. Opens and shorts are easily detected by simple tests. A cell may be stuck at a "0" or a "1." This may be due to faulty mask alignment, imperfect metallization, bad chip to pin connections, or possibly, if displayed by most or all cells, a faulty interface, power supply, or timing.
- b. Address non-uniqueness. This type of malfunction is usually due to either an address decoder failure or a multiple cell write. An address decoder failure will cause a part of the memory array to be unaddressable. Multiple cell write weaknesses are found in memory devices because of excessive capacitive coupling between cells.
- c. Cell/column/row disturb sensitivity. This type of problem can be caused by capacitive coupling between adjacent cells, cells in the same column, or cells in the same row. The coupled cells may not be affected until multiple disturbs occur.
- d. Sense amplifier interaction. Sense amplifier interaction is found when the level in one sense amp affects the resulting level in another. Sense amplifiers sometimes require an excessive recovery time to detect one logic state after detecting the opposite logic state for a long period of time. Sense amplifier inputs can also be overloaded by skewed cell decoders which temporarily select opposite data at the beginning of a cycle.
- e. Slow access time. This type of malfunction is typically caused by slow decoders, overloaded sense amplifiers, or too much capacitive charge on the output circuits, causing excessive time to discharge. These faults can cause the access time to be increased.
- f. Slow write recovery. Write recovery problems are caused by a saturated sense amp, during the write cycle, which is unable to recover in time to detect the correct level during the next read operation.
- g. Data sensitivity. This type of malfunction is shown by the varied response of the device as different data or address sequences are presented to the memory. This type of fault is not easily detected because of the number of possible input combinations. If some specific device weakness is suspected, the testing task is greatly reduced.
- h. Refresh sensitivity. Refresh sensitivity problems, aggravated by elevated temperatures, low supply voltage and disturb functions, may be found in DRAMs. After a specified time between refreshes, data is lost in cells because of excessive current leakage. The leakage of current reduces charge in the capacitive storage cell, causing the cell's logic level to change.

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- i. Static data losses. This type of malfunction may be found in SRAMs. After a long inactive period, some SRAMs lose data which should be retained as long as power is provided. The cause of this problem may be leakage current, open resistors, or open feedback loops within the chip.

Military internal visual requirements can not be economically or effectively implemented due to the LSI/VLSI die complexity and density. This has resulted in stress testing which supplements ineffective pre-seal inspection. The stress may be added burn-in, overvoltage, overcurrent, or combination of the same.

Due to the susceptibility of MOS devices to gate oxide rupture from static charges, special electrostatic protection precautions should be taken when handling these integrated circuits. See suggestions in handling electrostatic sensitive devices in subsection 7.1 General.

7.9 MICROCIRCUITS, MICROPROCESSORS7.9 Microprocessors.

7.9.1 Introduction. A microprocessor performs the storage of computation upon, and the processing of data sets. It can also monitor and control manufacturing processes, machine operations, and various other functions.

The microprocessor performs two major functions: logic execution and bus interface. All computational and logical decisions are made in the logic execution unit. Instructions and data are fed into the logic execution unit by the bus interface unit. All interaction with the system takes place through the bus interface unit. This unit fetches instructions and data as required from memory and passes and receives instructions and data to and from the logic execution unit when required. It is also capable of transferring data from a memory device or logic execution unit to another device, memory, display, or interfacing unit without use of the logic execution unit.

A variety of software aids are available for system support as shown in Table XXIX. Software support is a very important consideration when implementing a microprocessor-based system. The availability, cost, and level of software sophistication supporting a microprocessor are key factors in the selection process.

TABLE XXIX. Software and hardware aids for microprocessor implementation

Aid	Function
SOFTWARE	
Editor	Facilitates in developing source programs; allows adding, deleting, or replacing portions of a source program
Compiler	Translates a programming language into an "object" program for a particular microprocessor
Cross compiler	Translates one language "source" file into another language "source" file
Assembler	Translates the assembly level language "source" file for a given microprocessor into the "object" program for that microprocessor

7.9 MICROCIRCUITS, MICROPROCESSORS

TABLE XXIX. Software and hardware aids for microprocessor implementation (continued)

Aid	Function
SOFTWARE (continued)	
Cross assembler	Translates the assembly level language "source" program of a given microprocessor
Simulator	Used to test an "object" program when the microprocessor or hardware is not available
Loader	Transfer the "object" program to the microprocessor memory; directs and controls which memory locations the programs is loaded into
HARDWARE	
Emulator	Supports the development implementation of a microprocessor program; it includes the CPU and support devices
Input device	Allows loading of a program into memory
Video display and keyboard	Displays program text (source) as entered by the keyboard; allows visual display of text during "editing"
Disc operating system (DOS)	Controls and drives disc memories

7.9 MICROCIRCUITS, MICROPROCESSORS

7.9.2 Usual applications and characteristics.

7.9.2.1 Available technologies. Making the choice between 4- to 64-bit microprocessors depends upon system requirements. If high accuracy is needed and input data is at a relatively slow rate, an 8-bit microprocessor will probably meet the requirements. If high accuracy is not needed and input data rates are slow, 4-bit microprocessors may be adequate. Of course, high accuracy along with high data rates will require a 16-bit or 32-bit microprocessor. In some cases, the speeds associated with 16-bit, fixed instruction devices designed with the NMOS technology may not be fast enough. The faster bipolar technology utilizing bit slices may be necessary.

7.9.2.2 Applications. Table XXX lists a number of areas of application that benefit from 16-bit processing in terms of lower program requirements and improved system operating speed. Foremost are the application areas that require high computational accuracy or complex computations. Obviously, when the system is to be used as a computer such as minicomputers, personal computers, or "smart" terminals, this requirement may well be overriding.

TABLE XXX. Applications for 16-bit processors

Minicomputers
Personal computers
"Smart" terminals
Complex controllers
High-speed communications networks
Work processing systems
Diagnostic systems
Robotics
Artificial intelligence

7.9.2.3 Comparative attributes specific to microprocessors. The factors that influence microprocessor selection are word length, the speed at which the system performs the steps in the task solution, the instruction set, the timing and control signals available from the microprocessor for controlling other functional units, and the interrupt procedure.

7.9 MICROCIRCUITS, MICROPROCESSORS

Word length. The word length of the microprocessor refers to the number of data bits the microprocessor handles at one time. Typically, microprocessors are designed to handle data signals of 4 bits, 8 bits, 16 bits, or 32 bits at a time. For simple problems which require limited accuracy of the numbers it uses and a small number of different input codes, a 4-bit machine is more than adequate. For problems requiring higher number accuracy or where long strings of number or character codes are involved, a 16-bit or 32-bit microprocessor would be needed. For requirements in between, with a relatively limited number of different input and output codes, an 8-bit microprocessor can do the job. All microprocessors can be used to any accuracy desired by simply processing more bits but this can be achieved only by using more hardware and usually at the sacrifice of system speed. In addition, the greater the number of bits, the larger the memory directly accessible by the microprocessor.

The word length directly affects two important aspects of a digital logic system: information throughput and size. In general, the larger the word size the faster the throughput. For example, a 16-bit machine can add two 8-bit numbers simultaneously, as both numbers are processed in parallel. To do the same addition with an 8-bit machine requires processing each 8-bit number in sequence, a serial operation. Since more information can be contained in a 16-bit word as opposed to an 8-bit word, the 16-bit microprocessor has a higher information transfer rate. However, along with increased word size comes increased cost due to package size and package interconnections.

Speed. Speed of operation is important when sending and receiving information at a high rate and is very important when performing complex, many stepped computations. In both cases, a microprocessor that offers long-bit-length operation with high-speed clocks and an efficient instruction set is required. This is particularly true if the microprocessor has to respond to external conditions in real time.

Instruction set. A close check should be made to see that the instructions offered by a microprocessor are sufficient to handle the problem requirements. A calculator-type system needs efficient arithmetic operations including addition, subtraction, multiplication, division, absolute value, and so on. A communication system needs a wide choice of input and output instructions that can transfer information quickly. A logical system would need extensive logical, comparison, and decision-making instructions.

Timing and control features. The timing features of a microprocessor should be studied to see if its clock and instruction execution times are fast enough to solve the problem at hand in the time available. This should be carried through to include the other functional blocks to be connected to the microprocessor. In addition, the ease with which the microprocessor timing signals can be connected to outside units is very important. If the microprocessor control signals are such that they can be connected directly to the memory and input devices, the system timing is made easier because additional time delays through additional circuits are avoided.

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Interrupt structure. A microprocessor operating in a system performs one instruction after another--until there is a need to stop it at unexpected or random times for input or output information, or upon definite programming commands such as STOP or HALT. A control signal that interrupts the microprocessor in its sequence is called an interrupt signal.

Most microprocessors have a vectored interrupt structure that accepts both maskable and non-maskable interrupts. Maskable interrupts are only acted upon when software allows. Non-maskable interrupts are acted upon at the completion of the current instruction cycle. Then, depending upon the type of interrupt and its priority, the microprocessor accesses a specific memory address for instructions on how to handle the interrupt.

How fast a microprocessor can respond to an interrupt determines whether or not it is acceptable for many situations. If input information must be received over and over again into a system at unpredictable or random time, then whatever the microprocessor is doing is interrupted quite often. The microprocessor must be able to get its other jobs done despite the many interruptions. If the microprocessor were receiving information from satellite communications system and had to respond quickly, the interrupt structure would be critical and only a few microprocessors with that capability would be needed.

Table XXXI summarizes general characteristics associated with 8- to 16-bit microprocessors.

TABLE XXXI. Comparison of microprocessor power consumption and speed

Word size	Data move (μ s)	Add operation (μ s)	Multiply operation (8 x 8) (μ s)	Power Consumption
8-bit	3	4	1000 (by addition)	1.5 W
16-bit	2	1	72 (also 16 x 16)	2.3 W

The salient differences between 8- and 16-bit microprocessors are the accuracy and speed with which information may be processed. Most of the 8-bit microprocessors on the market today are classified as control devices. When compared to data processing minicomputers which process 16- and 32-bit numbers at high rates of speed, the 8-bit microprocessors are slow. Instruction cycle times for 8-bit microprocessors are normally 3 μ s.

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A 16-bit microprocessor processes two times as much data as does a 8-bit unit in a given instruction cycle. This means that the 16-bit unit will be at least two times more efficient in processing data and other information than the 8-bit unit. However, many applications do not require the efficiency and power the 16-bit unit instruction set features; they may only need the capabilities of an 8-bit unit. Thus, for certain applications, an 8-bit microprocessor may perform as well as a 16-bit unit.

A look at the instruction set of a microprocessor can give some insight into a machine's capability. Instruction sets indicate not only what basic computer operations can be performed, but at what speed (i.e., the number of clock cycles to perform). Instructions may be divided into categories such as arithmetic, (add, subtract, etc.), data transfer, input/output, logic, and branch.

7.9.2.4 Critical parameters. The following is a listing of design parameters considered essential to selection of a microprocessor. The reader should consult other material for further details and definitions.

- Address bus width
- ALU characteristics
- Architecture
- Bus structure
- Data bus width
- Hardware interface
- Instruction set needs
- Interrupt structures
- Memory refresh capability
- Operating temperature range
- Power consumption
- Price
- Reliability
- Second source (availability)

7.9 MICROCIRCUITS, MICROPROCESSORS

- Software support
- Speed of instruction execution
- Storage capacity (number of bits)
- Support devices
- System requirements
- Timer/control features.

7.9.2.5 Support devices. The availability of support devices is a very important consideration when selecting a microprocessor. For most applications, there are few cases where the microprocessor will be a "stand alone" device. Clocks, system control, input/output devices, and memory are usually needed. These functions could be fulfilled using discrete integrated circuits. However, this would not be taking full advantage of the packing density benefits available from microprocessors and the LSI/VLSI techniques. Most microprocessor manufacturers offer various support devices which will minimize the number of devices needed, although these support devices are not yet in MIL-STD-975.

Support devices may be divided into two general categories. First are those that provide timing, control, storage, and data movement and are tied directly with microprocessors such as digital processing support devices. Second are those that provide interface with the outside world.

Other support devices are the Memory Management unit (MMU), the Direct Access Memory Controller (DMA), the Counter Timer Controller (CTC), the Parallel Input/Output (PIO), and the Serial Input/Output (SIO).

7.9.2.6 Interface devices. Interface devices enable the microprocessor-based system to communicate with terminals, digital displays, and analog functions. These include the Universal Synchronous/Asynchronous Receiver/Transmitter (USART) which converts serial digital data to parallel digital data and vice versa, D/A and A/D converters, and display devices.

Figure 41 is a pictorial example of a microprocessor-based system configuration. Each box represents a device usually developed and offered by the microprocessor manufacturer. Most of those indicated were brought about by the microprocessor evolution. Some, although developed for one microprocessor, may be used with another microprocessor or even a non-microprocessor-based system. Second sources are available for most of these devices.

7.9 MICROCIRCUITS, MICROPROCESSORS

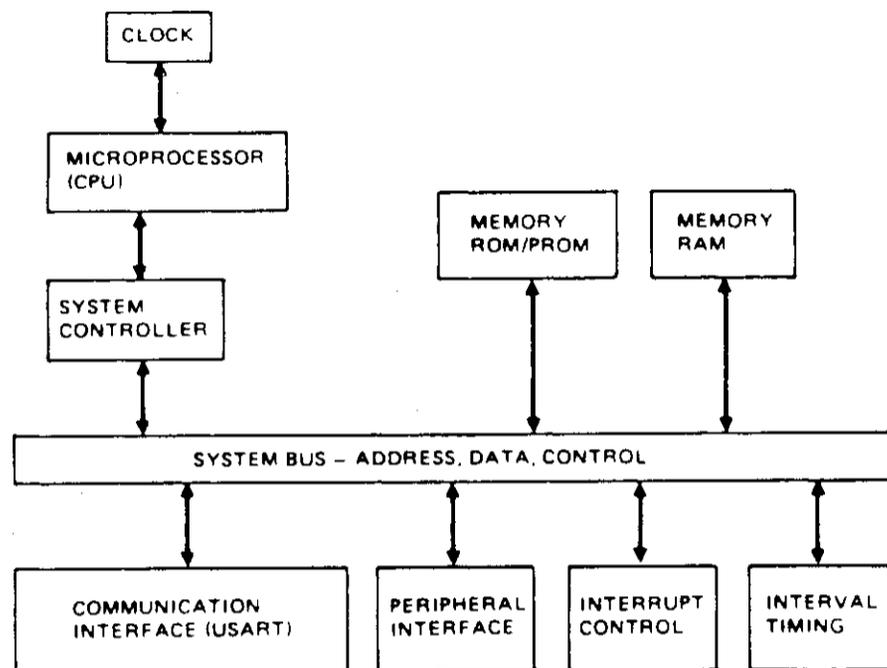
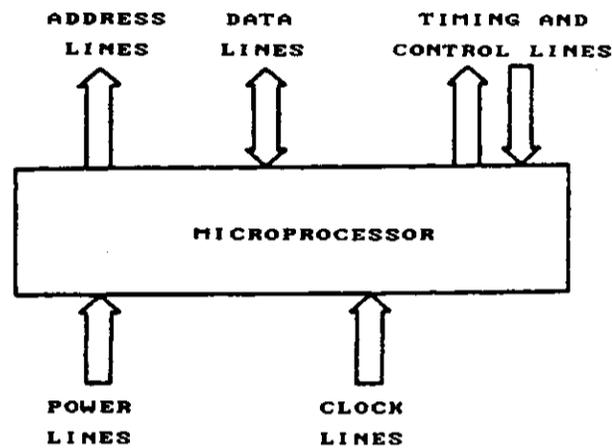


FIGURE 41. System block diagram.

7.9.2.7 Design considerations. The program and data to be used by the system must be stored in memory. Estimates should be made to determine how much memory will be required and what type of input and output circuits are needed. It may be difficult to accurately estimate the memory needed for program requirements until the program steps have been decided upon in some detail. Where available devices do not meet the system throughput needs, parallel processing schemes should be considered.

Most processors use signal lines as shown in Figure 42. These lines are:

- a. Address lines. The digital code that appears on these lines defines the location of instructions, data, or device to be used next by the microprocessor.
- b. Data lines. The instruction and data codes are sent to and from the processor on the data lines.
- c. Timing and control lines. All the timing and control signals sent to and from the microprocessor for the external functional blocks as well as the interrupt signals are included on these lines.
- d. Clock lines. For many microprocessors, the clock signals (the system's master timing signals) are formed externally and sent to the microprocessor. Other microprocessor and many microcomputers have circuits that generate these clock signals internally.

7.9 MICROCIRCUITS, MICROPROCESSORSFIGURE 42. Microprocessor signals.

Processing data in longer, 16-bit words provides for faster system operation, whether the system is oriented toward control, communications, or requires a great deal of computation. For any given operation, two to four times as much information can be processed by a 16-bit device and often instructions that accomplish more complete operations are available, resulting in shorter and more efficient programs.

7.9.3 Physical construction. Microprocessor dice are constructed using the basic semiconductor structures appropriate to the technology in which they are fabricated. Refer to the Microcircuit general section for details.

7.9.4 Military designation. The military part designation is described in the subsection 7.1 General.

7.9.5 Electrical characteristics. From the viewpoint of input and output as well as dc and ac electrical and timing characteristics, a microprocessor may be regarded as an assembly of electronic devices. For example, although there may be 16 output address pins on a microprocessor, many pins have the same characteristics regarding loading and timing. The same is true for the data pins, which are bidirectional (output pins when the device is placing data on the data bus, input pins when capturing data from the bus). Both the address pins and the data pins may be placed in a high impedance state, in effect disconnecting the data from the system address and data bus. This information need be specified only once for each pin type.

Other pins on the device are strictly input control functions (including power supply inputs) and output functions which indicate the state of the microprocessor (data in, data out, interrupt, status). For any given instant in time, all pins (except power supply pins) are either in a logic 1 or logic 0.

7.9 MICROCIRCUITS, MICROPROCESSORS

Exceptions sometimes occur. For example, the minimum "input high" voltage specified for one microprocessor is 3.3 V. A "high-output" for transistor-transistor logic (TTL) can be as little as 2.4 V. A TTL device providing a valid logic 1 level, say 2.9 V, would not be recognized as a valid logic 1 by that microprocessor. A close look at voltage level specification and loading currents is required before interfacing a microprocessor to other logic networks.

Device manufacturers supply timing diagrams which describe delay times and minimum setup times required for proper operation. For NMOS devices, the minimum and maximum clock period is also specified. This usually runs from about 250 ns to 2 μ s. Beyond this rate, the dynamic circuits within the microprocessor are not refreshed often enough.

7.9.6 Environmental considerations. Since most microprocessors are available in hermetically sealed ceramic packages, humidity and vibration are not critically environmental factors. The major limiting factor in microprocessors is temperature. Devices specified to operate over a wide temperature range such as -55 to +125 °C usually have reduced operating frequencies.

7.9.7 Reliability considerations. A major concern for device users is how long will the device operate under a given field environment. No matter how appealing a device may be in terms of function, utility, and costs, if it has a short life expectancy, it will remain undesirable. This is especially true with microprocessors, where a very complex semiconductor product will be used in extreme environments.

The reliability of a system is based on the sum of the individual device reliabilities. Reducing the number of devices should increase reliability, assuming that the devices implemented into the system are proportionately as reliable as previously used devices and software reliability is given adequate attention. From a system viewpoint, implementing a microprocessor should improve system reliability, assuming that a microprocessor substantially reduces the number of parts in a system. Fewer parts mean fewer wire bond connections (a susceptible failure area in any device) and fewer solder connections, cables, and connectors.

Potential failure modes are metallization and oxide defects, junction defects, and crystallization imperfections. Metallization defects include contamination under metal runs, electromigration, and microcracks caused by sharp oxide steps which prevent uniform metallization thickness. Oxide defects include contamination of the oxide layer with foreign material and pinholes and cracks in the oxide.

Each device manufacturer implements reliability tests to ensure that the aforementioned defects do not exist and hopefully, to catch any early failures. Microprocessor reliability data is available from both manufacturing experience and from field use. MIL-HDBK-217 contains empirical data that may be used to predict the device failure rate in a given application.

7.10 MICROCIRCUITS, LINEAR

7.10 Linear.

7.10.1 Introduction. Early development of integrated circuits was concerned almost entirely with logic devices for the computer market. Linear and special purpose devices are a more recent development for a considerably smaller but rapidly growing market in instrumentation, communications, signal processing, and other analog fields. These devices differ from logic devices in that their operation depends on biasing some or all of the transistors in their linear regions to reflect incremental variations as part of the circuit function rather than switching abruptly between high and low states.

In this section, the term linear device is used to designate a circuit in which function depends on operating some or all of the internal transistors in their linear region, such as amplifiers, comparators, line drivers, line receivers, timers, multiplexers, and voltage regulators. The term "special-purpose device" is used for a microcircuit which does not fit this definition of a linear device and is not a member of a logic family. The category includes interface circuits which drive or are driven by logic devices but are of linear design, such as voltage comparators and converters. Because of the different design and manufacturing techniques required, linear and special purpose devices are generally manufactured and marketed separately from logic devices.

7.10.2 Usual applications. Linear and special-purpose microelectronic devices are used in a wide variety of applications for which manufacturers are constantly creating new circuits. There is a fair degree of standardization with linear devices produced by semiconductor manufacturers. On the other hand, most major manufacturers are constantly generating new special-purpose circuits. So diverse is the spectrum of products that second sourcing of these can be a problem.

7.10.2.1 Available functional types and technologies. Linear microcircuits can perform a great variety of functions needed in circuit design for many applications in the instrumentation, computation, communication, and control fields. The important functional types of linear microcircuits that are most frequently used are available in a variety of processing technologies as shown in Table XXXII.

7.10.3 Analog switches.

7.10.3.1 Attributes of analog switches. Monolithic IC switches operate in video bandwidths and at high speed, minimize low-signal error and crosstalk, and can withstand high voltages. In some cases, manufacturers have added on-chip address latches to these switches and refined their overvoltage protection circuitry. To obtain the high speed necessary for video switching, manufacturers fabricate the chips using either a CMOS or DMOS (double-diffused NMOS) process. Lateral DMOS transistors are very fast because they have low on-resistance and low output capacitance.

7.10 MICROCIRCUITS, LINEAR

TABLE XXXII. Processing technologies for linear microcircuits

Functional Type	Processing Technology
Analog switches	CMOS
Multiplexers/demultiplexers	CMOS, Bipolar
Operational amplifiers	Bipolar, BIFET ^{1/} , CMOS, bipolar/MOS ^{2/}
Voltage regulators	Bipolar
Voltage comparators	Bipolar, BIFET
Line drivers and line receivers	Bipolar
Timers	Bipolar
Multipliers	Bipolar
Phase locked loops	Bipolar

^{1/} Bipolar operational amplifiers with JFET inputs

^{2/} Operational amplifiers with a combination of bipolar and CMOS technology.

To obtain fast switching, a video switch need not be used. A JFET model can be chosen, or in order to handle a wider analog-signal range, a CMOS switch developed for high-speed data acquisition can be selected. A good example would be a product like the 201 quad spst analog switch, which features 30 ns switching and 30 ohm on-resistance.

Figure 43 shows a schematic diagram of a simple analog switch.

7.10.3.2 Critical parameters of analog switches. The parameters which are considered to be important in analog switch applications are as follows:

On state resistance. On state resistance is the resistance of the channel when the analog switch is in the on state. FETs are available with on resistances of less than 2 ohms and some high-power bipolar transistors can have common emitter saturation resistances of less than 100 milliohms.

7.10 MICROCIRCUITS, LINEAR

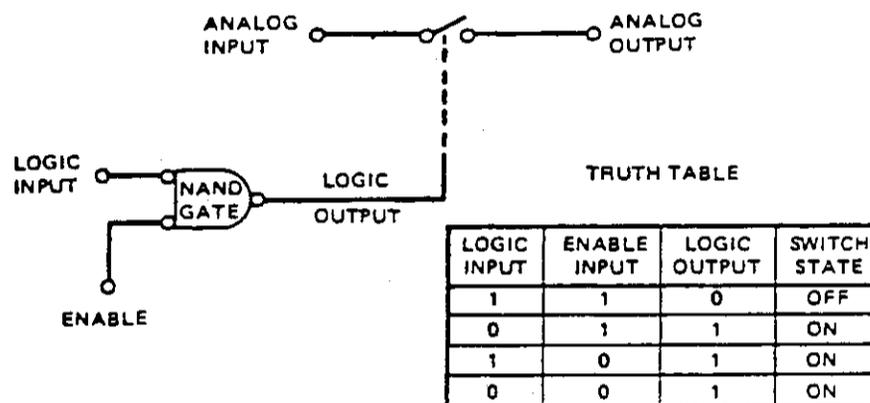


FIGURE 43. Simple analog gate.

Switching speed. Switching speed is the elapsed time from the application of the control signal to the appearance (or disappearance) of the analog signal at the output. Switching speed can be greatly affected by the load on the analog switch. Switching times of nanoseconds are easily attainable and maximum switching rates are often in excess of 10^6 operations per second.

Switch current. Switch current is the amount of current that can be fed through the switch channel. For example, the 201A can handle up to 70 mA of pulsed current or 20 mA of continuous current.

Break-before-make versus make-before-break. For most analog switch applications, break-before-make switching is desired because in most applications it is necessary to disconnect one signal source before connecting another to avoid crosstalk. However, to avoid opening the loop, make-before-break switching is critical in some control circuits such as the feedback resistor gain selector for programmable gain operational amplifiers.

Power supplies and power consumption. There are many possibilities for powering a particular analog switch. Bipolar supplies are the most common, but single supply operation is possible. Manufacturers' data sheets give application hints on power supply range versus analog signal range.

7.10 MICROCIRCUITS, LINEAR

Interfacing. Interfacing can be one of the most important parameters of an analog switch application because so many possibilities exist. The most important interface criteria is logic compatibility. The two most common logic levels are TTL and CMOS. Not all analog switches are compatible with both types of logic. Standard analog switches require a constant control signal present on the input to hold the switch in the desired position; this could tie up a control system unless external latches are added to control the switch.

7.10.3.3 Design precautions for analog switches. In selecting analog switches or multiplexers, attention must be paid to several key specifications. Break-before-make switching ensures that no two-channel inputs are simultaneously connected. Acquiring analog input signals within a specified time and error band are primary concerns affected by on resistance and output capacitance specifications. A low value of on resistance ensures minimum signal attenuation and maximum accuracy. The output capacitance forms an resistance-capacitance time constant with the on resistance placing fundamental limits on signal acquisition time. A low value of on resistance and output capacitance ensures minimum elapsed time between the channel select command and the acquisition of data to within a specified error band. High crosstalk and off isolation specifications prevent unsettled input signals from affecting the signal path.

7.10.4 Multiplexers.

7.10.4.1 Attributes of multiplexers. Multiplexers are a subset of analog switches which have many (4, 8, 16, or more) inputs with only one common output. They are used where it is necessary to transfer information from many signal channels at a transmitting point to a central or common receiving point, or vice versa. This is most often used when only one transmission line is available for all data transfer between points. The signals to be transmitted may be either analog or digital.

For high-voltage switching, multiplexers for analog signals in the ± 50 V range can be used. They offer a typical transition time of 1 μ s. These devices combine CMOS control circuitry with power-DMOS switch elements, providing high breakdown voltage and a low (30 to 300 milliohm) on-resistance.

Another variety of multiplexers is optimized for minimum low-signal errors. These devices can, for example, handle thermocouple signals without amplification. This error includes $R_{on} \times I_{D(on)}$ offset, and it also includes the thermocouple effects due to dissimilar-material junctions within the chip and package.

To further enhance their switches and multiplexers, some manufacturers add overvoltage protection circuitry. Overvoltage can appear at any terminal, and can range from transients of many hundreds of volts to a supply potential just a little higher than the manufacturer recommends.

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7.10.4.2 Critical parameters of multiplexers. In a multiplexing application, the following factors should be considered:

- a. System attenuation. Includes loss in analog signal due to the multiplexing and demultiplexing devices and the transmission path. This is a frequency dependent factor.
- b. Channel isolation. At low frequencies, this is principally a function of channel off leakage currents and at high frequencies is a function of device and system capacitances.
- c. Crosstalk. There are several sources of crosstalk, the main ones being overlap between switching channels due to imperfect break-before-make switching, switch leakages, off switch capacitances, inter-switch capacitance, and stray circuit capacitances.
- d. Noise. There are several sources of noise, including thermal or Johnson noise generated in any resistive components, crosstalk, leakages, switching transients, and thermal EMF.
- e. Switching rates. These are important in sampling systems where they determine the maximum analog signal handling frequency of the multiplexer.

7.10.4.3 Design precautions for multiplexers. Multiplexers are available in two output configurations: single ended or differential. Figures 44 and 45 demonstrate these options. Single-ended multiplexing, as shown in Figure 44, applies to systems that have signal sources that are referenced to a common point (usually ground).

Differential multiplexing, as shown in Figure 45, is utilized when all signal sources are not referenced to the same voltage level. Major considerations are switch matching (on resistance, leakage current, and capacitance), common-mode rejection, and the system's tolerance to switching transients introduced by the break-before-make switching sequence.

7.10.5 Operational amplifiers (op amp).

7.10.5.1 Attributes of op amp. The op amp is classified as a linear device. This means that its output voltage V_O tends to proportionately follow changes in the applied differential input V_{id} . Within limits, the changes in the output voltage V_O are larger than the changes in the input V_{id} by the open-loop gain A_{VOL} of the op amp. The amount that the output voltage can change (swing), however, is limited by the dc supply voltages and the load resistance R_L . Generally, the output voltage swing is restricted to values between the $+V$ and $-V$ supply voltages.

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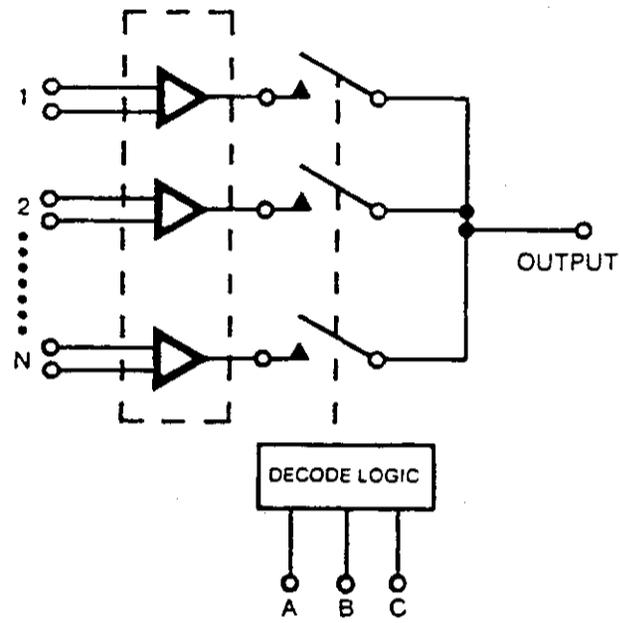


FIGURE 44. Single-ended multiplexing.

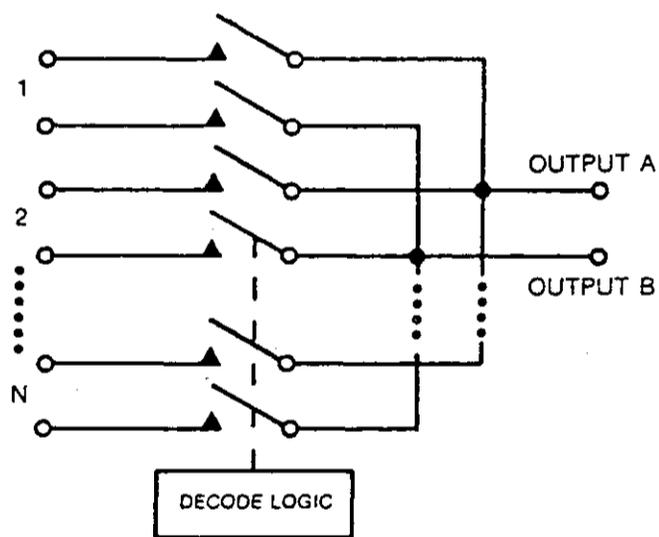


FIGURE 45. Differential multiplexing.

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Op amps produced by bipolar processes offer advantages in low offset voltage and drift, high common-mode and power-supply rejection ratios, high open loop gain, low noise, and high speed. If a commonly used precision op amp had to be chosen against which to measure others, it would be the OP-07 developed several years ago. The amplifier's 25- μ V maximum offset voltage, 0.6- μ V/ $^{\circ}$ C maximum offset drift, and 300,000 V/V open-loop gain put the device in a class by itself for a while. Virtually every op amp manufacturer now lists the OP-07A family in its catalog. Several manufacturers now offer devices that are improvements of the OP-07A in offset, drift, open-loop gain, or all three.

In some applications, an op amp's noise performance is just as important as its dc parameters. In terms of input noise-voltage density, bipolar op amps out perform all other types. Voltage-noise figures lower than 10 nV/kHz are exceptionally good for monolithic op amps, and several high-precision units meet this criterion.

An op amp's speed can be characterized in several ways: by slew rate, settling time, or bandwidth. An op amp's slew rate is limited by the currents available for charging and discharging the unit's internal capacitances. Settling time (i.e., the time required for the amplifier's output to settle to within a certain error band around the final value) is an important parameter in such precision applications as driving high-resolution data converters. Finally, most data sheets specify an op amp's bandwidth in terms of the unit's gain-bandwidth product.

Bipolar op amps offer some speed advantages over JFET-input types. For example, there are bipolar devices whose slew rates are much greater than those of JFET input amplifiers. The tradeoff that must be made is in settling-time precision. Many of the JFET units specify settling times to within $\pm 0.01\%$ error band, whereas almost all the fast bipolar devices use a $\pm 0.1\%$ error band. Many high-slew rate bipolar amplifiers do not possess high open-loop gain. They must keep the number of gain stages low in order to minimize phase shifts and the slew limiting, capacitances.

Certain needs in analog-signal processing cannot be met by bipolar op amps and are spurring the development of devices that combine JFET inputs and bipolar circuitry. Bipolar op amps cannot meet the low bias current figures that JFET-input amplifiers specify. Low input bias current, impedance of course, implies high input impedance. The high impedance results from the use of a bootstrap input stage. The bootstrapping technique ensures that the input bias current is independent of the common-mode voltage. JFET-input op amps meet fast settling times which bipolar units cannot.

The $\pm 0.1\%$ error band of most bipolar devices is about four least significant bits (LSBs) in 12-bit systems and therefore the bipolar op amps cannot be used with any assurance of $\pm 1/2$ LSB settling. The specifications for almost all

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JFET-input op amps, on the other hand, guarantee settling times to $\pm 0.01\%$, about 1/2 LSB at 12-bit resolution. Another significant area where JFET-input amplifiers are superior to their bipolar counterparts is power consumption. Some devices take minimal quiescent current from their power supplies, yet display remarkable dc accuracy.

CMOS-processed op amps do not compare favorably with bipolar units in terms of offset voltages and noise performance. Op amps that use CMOS or a combination of MOS and bipolar processing can provide very high input impedances and, in some cases, operate from very low supply voltages.

Some manufacturers are marketing op amps whose characteristics can be varied somewhat to suit special applications. These are called programmable op amps and they have an additional input bias terminal. By controlling the current in this terminal by external means, some of the op amp's parameters can be adjusted to values that will optimize its performance in any given application.

7.10.5.2 Critical parameters of op amps. The detailed and specific performance characteristics of a particular op amp should be obtained from the appropriate specification. Major op amp characteristics are: input offset current and drift, input bias current and drift, input off-set voltage and drift, output voltage swing, voltage gain, common-mode rejection ratio, power supply rejection ratio, power supply current, unity-gain bandwidth, slew rate, rise time, overshoot, settling time, open-loop voltage gain, input resistance, and output resistance.

An ideal op amp has zero input offset voltage and no drift. However, because of the mismatch of input transistors and resistors on the monolithic circuit, typical op amps have a low but definite offset voltage. Many have provisions for connecting an external resistor so that the input offset can be adjusted to zero. The exact method used and total resistance of the null adjustment is dependent upon the type of op amp circuit selected. A general purpose internally compensated op amp such as a 741 may require 10 k Ω . A JFET-input or externally compensated op amp may require 100 k Ω . Recommended input offset voltage null adjustment circuits are usually shown in the data sheet.

Input offset voltage temperature coefficient is specified in volts per degree Celsius. The amount of drift that occurs with temperature variations is directly related to how closely matched the input characteristics are when the device is manufactured. Bipolar and JFET-input op amps usually have 10 to 15 $\mu\text{V}/^\circ\text{C}$. The CMOS op amp family has from 0.7 to 8 $\mu\text{V}/^\circ\text{C}$, depending upon the bias mode selected.

The input common-mode range may be defined as the maximum range of the input voltage that can be simultaneously applied to both inputs.

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Unlike the ideal op amp whose characteristics include the absence of internal noise, typical op amp performance is degraded by added noise components. Noise components determine the ultimate lower limit of signal handling capability. Noise is usually specified on the data sheet as equivalent input noise and is increased by the gain of the amplifier. There are several potential sources of noise in an op amp. The most common are thermal noise caused by resistances, noise current, and noise-voltage generators. Thermal noise increases with higher temperatures, larger resistances, and wider bandwidths. In specifications, these two parameters are detailed separately. Noise voltage is specified at a low source resistance. Noise current is specified at a high source resistance. Both parameters are measured with a narrow bandwidth filter at a series of points across a useful spectrum of the amplifier. Data is usually given in terms of noise voltage versus frequency. Typically a frequency and source resistance will be given in the test conditions included in the device data sheet.

In general, JFET-input or low-input-bias-current bipolar op amp will have lower noise current and tend to be quieter at source impedances above 10 k Ω . Below this value, the advantage swings to bipolar op amp which have lower input voltage noise. The noninverting op amp configuration has less noise gain than the inverting configuration for low signal gains and thus a higher signal-to-noise ratio. At high gains, however, this advantage diminishes.

Two commonly used criteria for determining the stability of an op amp are gain margin and phase margin. The gain margin is defined as the amount of loop gain in decibels at the frequency at which the phase angle of the loop gain is 180 degrees. If the gain margin is negative, this gives the decibel rise in open-loop gain, which is theoretically permissible without oscillation. If the gain margin is positive, the amplifier is potentially unstable. The phase margin is 180 °C minus the magnitude of the phase angle of the loop gain at the frequency at which the absolute value of the loop gain is unity (zero decibels). For a linear amplifier of good stability gain and phase margins of at least 20 dB and 45 °C, respectively, are required.

7.10.5.3 Design precautions for op amp. The choice of circuit configuration and of associated component values is related to the choice of an op amp for a given application. Op amps are specified as open-loop devices, but in a circuit application they generally have feedback applied. The designer must predict the closed-loop circuit performance as determined by his choice of an op amp and circuit configuration. Detailed literature is available on circuit configurations to accomplish particular analog circuit functions using op amps. The following gives recommended guidelines and a design strategy to best meet the designer's needs:

- a. The first step is to completely define the design objectives, including the frequency of the input signal, the accuracy or allowable distortion, the output load, the operating temperature range, power supply characteristic, and noise environment.

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- b. Op amps can be divided into four basic categories as shown in Table XXXIII.

TABLE XXXIII. Selection process for op amps

Category	Primary Characteristics
General purpose	"741A types," M38510/101 types
High accuracy	Low input offsets ($V_{i0} < 1$ mV), high dc gain, high CMRR, and low noise. Leading part types are OP-07A and OP-27A. M38510/135 types
High speed	Optimized for high slew rate, high gain-bandwidth, and fast settling time; 118 type M38510/10107 and /122 types
Low power, wide supply range	Low supply drain (< 1 mA), wide input and output voltage range. Includes micropower (< 100 μ A) units for battery operation. 124 type M38510/11005

There can be overlap between some categories, whereas others are mutually exclusive. For example, some devices cover both high accuracy and low power. However, high speed and low power tend to be mutually exclusive. It is difficult to simultaneously optimize both speed and power.

Economics is another important dimension of the selection process. The general purpose category is generally lowest in cost, but a high accuracy op amp with low input offset voltage may be more cost effective if it eliminates the need for external trimming components. The high-accuracy OP-07 is often used in place of general-purpose 741 types because of its low input-offset-voltage and high gain.

- c. Consideration of ac requirements for an application is a good starting point in the op amp selection process. If high frequency (slew rate > 10 V/ μ s) is the primary concern, then the choice quickly narrows down to the high speed category. Two factors will generally dictate the op amp choice: (1) the loop gain (excess of open-loop gain over closed-loop gain) must be sufficient at the highest frequency of interest, and (2) the slew rate must be high enough to follow the fastest signal input without causing distortion or other anomalies.

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- d. If the frequency requirements are relatively modest (slew rate $< 10 \text{ V}/\mu\text{s}$) and the circuit requires closed-loop gain above unity, a high accuracy (see Table XI) op amp should be chosen. Selection of a specific op amp type within the high accuracy category is generally determined by impedance levels of the input signal and feedback elements. High impedances (above $10 \text{ k}\Omega$) imply a need for an op amp with low input bias currents. This need for low bias current can be met through use of JFET-input op amps, or by using bipolar-input op amps specifically designed for low input-bias-current.

- e. Low noise, always desirable, is sometimes the primary consideration. In many high-gain active-filter or amplifier applications, low noise can be more important than dc offset. These are the three basic rules for obtaining low noise:

Using low impedances minimizes the effect of current noise flowing through the source impedance and reduces resistor thermal noise.

Noise outside the frequency range of interest can usually be attenuated by filtering. Block high-frequency power-supply noise from the signal path by use of decoupling capacitors at the op-amp supply inputs.

Some op amps such as the bipolar-input OP-27A are designed for minimum noise. The input stage current is set to a relatively high value which reduces input noise ($5.5 \text{ nV}/\text{Hz}$ max at 10 Hz). Output swing is increased to $\pm 10 \text{ V}$ into 600Ω to allow the use of low impedance, low noise feedback elements.

- f. The op amp power supply requirements are the next factors to consider. If the circuit is to be operated from a battery, such as in spacecraft, selections should be made from the low-power, wide supply range category. Low-power op amps are designed for minimum quiescent supply current. Speed is traded off for lower power consumption and output drive is generally reduced. The input and output stages are designed for linear operation over a wide voltage range which is very helpful for single-power-supply operation.

The low-power family includes programmable micropower op amps that offer the designer another dimension in circuit design. The quiescent supply current is set by an external resistor which allows the circuit designer to trade off between quiescent supply current and speed. Since the quiescent current directly controls slew rate and gain-bandwidth product, these programmable op amps are easily frequency compensated in such circuits as active filters, oscillators, or multistage instrumentation amplifiers.

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- g. In an op amp, as in any other feedback amplifier, the phase of the feedback must be controlled to assure that the design is stable with frequency and that the desired gain-frequency response is obtained. Op amp stability can be assured by proper circuit design which takes into account the causes of instability.

7.10.6 Voltage regulators.

7.10.6.1 Attributes of voltage regulators. Various types of voltage regulator microcircuits are available. The type of regulator used depends primarily upon the designer's needs and trade-offs in performance and cost.

Voltage regulators can be classified by the polarity of their regulated output voltage, by whether their output voltage is fixed or adjustable, and by their control element. A positive regulator is used to regulate a positive voltage and a negative regulator is used to regulate a negative voltage.

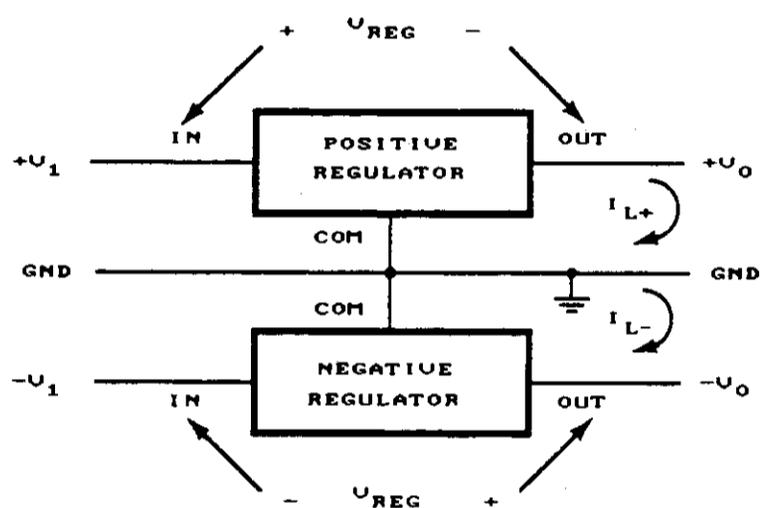
Figure 46 illustrates conventional positive and negative voltage regulator applications using a continuous and common ground. For systems operating on a single supply, the positive and negative regulators may be interchanged by floating the ground reference to the load or input. This design approach is recommended only where ground isolation serves as an advantage to overall system performance or when negative regulators with the desired characteristics are not available. Figures 47 and 48 illustrate a positive regulator in a negative configuration and a negative regulator in a positive configuration, respectively.

Many fixed three-terminal voltage regulators are available in various current ranges from most major microcircuit manufacturers. These regulators offer the designer a simple, inexpensive method to establish a regulated voltage source. Their particular advantages are ease of use, few external components required, reliable performance, internal thermal protection, and short-circuit protection.

There are disadvantages. The fixed three-terminal voltage regulators cannot be adjusted because their output voltage sampling elements are internal. The initial accuracy of these devices may vary as much as $\pm 5\%$ from the normal value. Also, the output voltages available are limited. Current limits are based on the voltage regulator's applicable current range and are not adjustable. Extended range operation requires external circuitry.

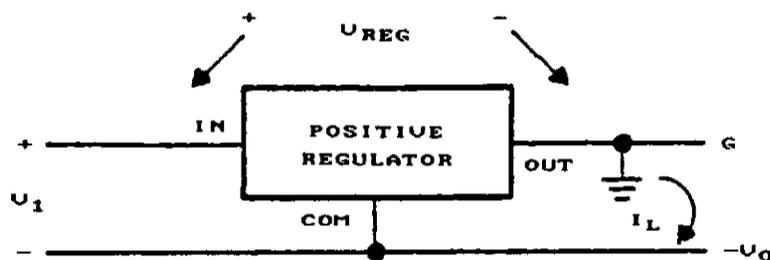
The adjustable regulator may be well suited for those applications requiring voltages not available in fixed regulators. Additionally, all adjustable regulators use external feedback, which allows the designer a precise and infinite voltage selection. Note that most fixed three-terminal regulators may be used as variable regulators above the fixed output voltage.

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FIGURE 46. Conventional positive/negative regulator.

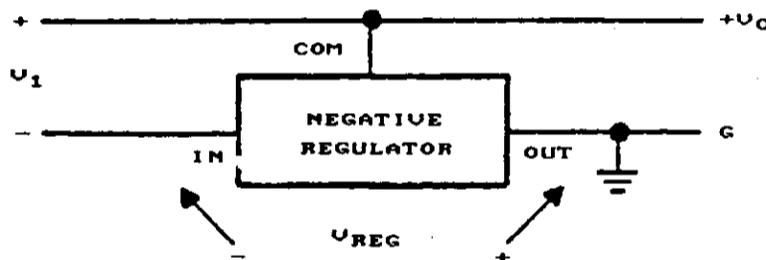
The output sense may also be referred to a remote point. This allows the designer to not only extend the range of the regulator with minimal external circuitry, but also to compensate for losses in a distributed load or external pass elements. An additional feature found on many adjustable voltage regulators is access to the voltage reference element and shutdown circuitry.

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NOTE: U_1 MUST FLOAT.

FIGURE 47. Positive regulator in negative configuration.



NOTE: U_1 MUST FLOAT

FIGURE 48. Negative regulator in positive configuration.

The series regulator is well suited for medium current applications with nominal voltage differential requirements. Modulation of a series pass control element to maintain a well regulated, prescribed output voltage is a straight forward design technique. Safe-operating-area protection circuits such as overvoltage, fold-back current limiting, and short-circuit protection are additional functions that series regulators can supply. The primary disadvantage of the series regulator is its power consumption. The amount of power

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a series regulator will consume depends on the load current being drawn from the regulator and is proportional to the input-to-output voltage differential. The amount of power consumed becomes considerable with increasing load or differential voltage requirements. This power loss limits the amount of power that can be delivered to the load because the amount of power that can be dissipated by the series regulator is limited.

7.10.6.2 Critical parameters of voltage regulators. The safe operating area (SOA) is a term used to define the input and output voltage range and load current range within which a device is considered to operate reliably. Exceeding these limits may result in a catastrophic failure or will render the device temporarily inoperative (zero output), depending upon the device and its performance characteristics. Integrated circuit voltage regulators with internal current limiting, thermal and short circuit protection will merely shut down. On the other hand, external components, such as pass transistors, may respond by failing catastrophically.

Although particular design equations depend upon the types of voltage regulator microcircuit used and its application, there are several parameters that apply to all regulator circuits for safe, reliable performance:

The limits on the input voltage are derived from two considerations: absolute maximum rated input and input-to-output differential voltage.

Load current is the maximum load current deliverable from the regulator microcircuit. The power that can be dissipated within the regulator is the product of the input-to-output differential voltage and the load current, and is normally specified at or below a given case temperature. This rating is usually based on a 150 °C junction temperature limit. The power rating is an SOA limit unless the integrated circuit regulator provides an internal thermal protection.

7.10.6.3 Design precautions for voltage regulators. Selection of the proper voltage regulator microcircuits and external components will result in a reliable design in which all devices can operate well within their respective safe operating areas. Fault conditions such as a short circuit or excessive load may cause components in the regulator circuit to exceed their safe operating area (SOA). Because of this situation, as well as protection for the load, certain protection circuits should be considered.

A potentially dangerous condition may occur when a voltage regulator becomes reverse biased. For example, if the input supply were crowbarred to protect either the supply itself or additional circuitry, the filter capacitor at the output of the regulator circuit would maintain the regulator's output voltage and the regulator circuit would be reverse biased. If the regulated voltage is large enough, the regulator circuit may be damaged. To protect against this, a diode can be connected from the input to the output such that the capacitor will be discharged by passing the regulator under low input voltage conditions.

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Current limiting techniques are commonly used in conjunction with protection of voltage regulators against fault conditions. The two basic current limiting techniques constant current and fold-back current limiting.

Three-terminal voltage regulator microcircuits have been especially useful to the designer of small, regulated power supplies or on-card regulators. Three-terminal regulators are popular because they are small and require a minimum number of external components.

Mounting and using three-terminal regulators usually presents no problems. However, there are several precautions that should be observed. It is good practice to use bypass capacitors at all times on the input and output of the regulator.

Like any semiconductor circuit, lower operating temperature greatly improves reliability of a voltage regulator. It is good practice to make the input-to-output drop across a three-terminal regulator as low as possible while maintaining good regulation. Larger voltage drops mean more power dissipated in the regulator. Although most regulators are rated to withstand junction temperatures as high as 150 °C, heat sinking should be provided to maintain the lowest possible temperature.

Circuit lead lengths should be held at a minimum. Lead lengths associated with external compensation or pass transistor elements are of primary concern. Components should be located as close as possible to the regulator control circuit.

Improper placement of the input capacitor can induce unwanted ripple on the output voltage.

The voltage regulator should be located as close as possible to the load. This is especially true if the output voltage sense circuitry is internal to the regulator. Excessive lead length will result in an error voltage developed across the line resistance. If the voltage sense is available externally, the effect of the line resistance can be minimized. By referencing the low current external voltage sense input close to the load, losses in the output line are compensated for.

7.10.7 Voltage comparators.

7.10.7.1 Attributes of voltage comparators. The comparator function demands high gain and wide bandwidth, properties not easily combined on one microcircuit. The manufacturer's option to trade one for the other is restricted because the typical application requires both. Comparators fall into three categories determined by speed. At the top end are devices that are based on bipolar ECL and feature propagation delays as short as 2 ns. Next are the middle performance comparators that attempt to provide as much speed and accuracy as possible in one device. In the third category most of the comparators are found with response times from 200 ns to several microseconds. These devices include low supply-voltage versions, low-power versions, FET-input versions, and dual and quad versions.

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A basic comparator is similar to a differential amplifier operating in the open-loop mode. Because of high gain, the output is normally saturated in either the high state or the low state depending upon the relative amplitudes of the two input voltages. With these conditions, the comparator provides a logic-state output which is indicative of the amplitude relationship between two analog input signals. Initially, op amps were used in the open-loop mode to perform comparator functions. However, devices designed specifically for this operation resulted in improvements in recovery time, switching speed, and output levels. The output logic-state levels of a comparator amplifier stage normally match those required by the following stage (e.g., a TTL logic stage).

7.10.7.2 Critical parameters. Because comparators are normally used to drive logic circuits, the output must change states as rapidly as possible. High open-loop gain, wide bandwidth and slew rate are key factors in comparator speed. Operation in the open-loop mode (no feedback), with minimum or no frequency compensation, results in maximum gain-bandwidth product for best performance. Most comparators operate in this manner.

7.10.7.3 Design precautions for voltage comparators. When selecting a comparator, certain device parameters must be considered for proper design and application. These parameters are: input offset voltage, response time, slew rate, power supply rejection ratio, input bias current, common-mode voltage range, output configuration, and voltage gain.

The input offset voltage for a comparator should be as small as possible because in a high gain circuit it is the dominating factor that determines the exact threshold level. For this reason, comparators should be nulled or a precision comparator used so that the input differential voltage is as close to zero as practical when the output is at the logic switching threshold.

The voltage gain determines the sensitivity and threshold accuracy of a comparator. For the ideal comparator, the gain could be considered infinite. An extremely small voltage applied between the two inputs will cause a change in the output. In practice, some minimum voltage variation will be required at the input to effect a change in the output state. This minimum sensitivity will be determined from the voltage gain of the comparator. The relationship is as follows:

where:

$$\Delta V_{IN(MIN)} = \frac{\Delta V_O}{A_V}$$

A_V = voltage gain
 ΔV_O = difference between the high and low state of the output

Applications in which the input signal varies slowly can cause the output to change proportionally within the hysteresis bond. This becomes a problem when the comparator is used to trigger a logic stage requiring fast rise and fall inputs. This problem can be solved by the introduction of positive feedback. This causes a fast or Schmitt trigger action. This action is accomplished by

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feeding a portion of the output signal back to the noninverting input. Depending upon the amount of positive feedback, a new trip-level will be introduced after each transition. These are called the upper threshold point (UTP) and lower threshold point (LTP). The difference between the two points is the hysteresis. A comparator with positive feedback is shown in Figure 49. In this figure, R_3 equals the product of R_1 and R_2 divided by the sum of R_1 and R_2 . The positive feedback reduces the width of the hysteresis band.

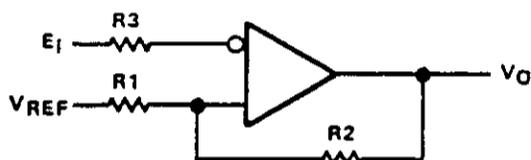


FIGURE 49. Comparator with positive feedback.

A typical hysteresis loop diagram for this type of circuit is shown in Figure 50.

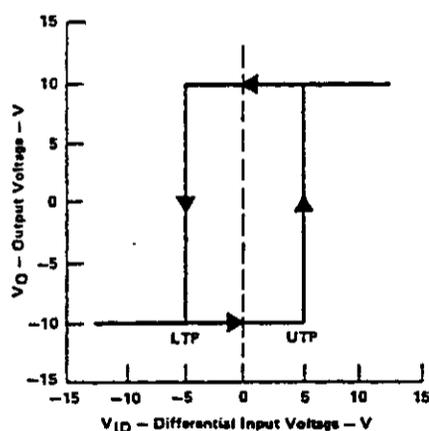


FIGURE 50. Typical comparator hysteresis loop.

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7.10.8 Line drivers and receivers.

7.10.8.1 Attributes of line drivers/receivers. Interfacing subsystems and transmitting data over a distance, whether it is a few inches on a circuit board or many feet to another unit in the system, can present a problem. The quality of the signal reproduced is dependent on the line driver and receiver and the transmission line between them. Receiver characteristics such as input impedance, hysteresis and signal input threshold, and frequency response will effect the quality of the signal.

7.10.8.2 Critical parameters of line drivers/receivers. Devices currently available are very versatile. The following parameters are those that are necessary for the stated reasons:

Line drivers. Inputs should be compatible with the logic and supply voltage levels of the system. Outputs are complementary in differential line drivers. For WIRED-OR applications, open collector output devices should be used. Figure 51 shows a logic diagram for a line driver.

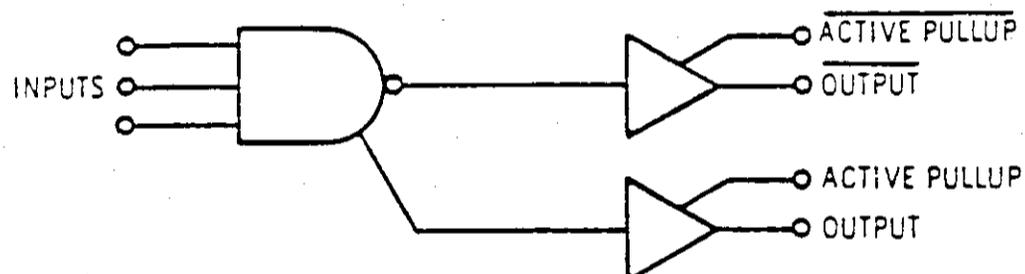
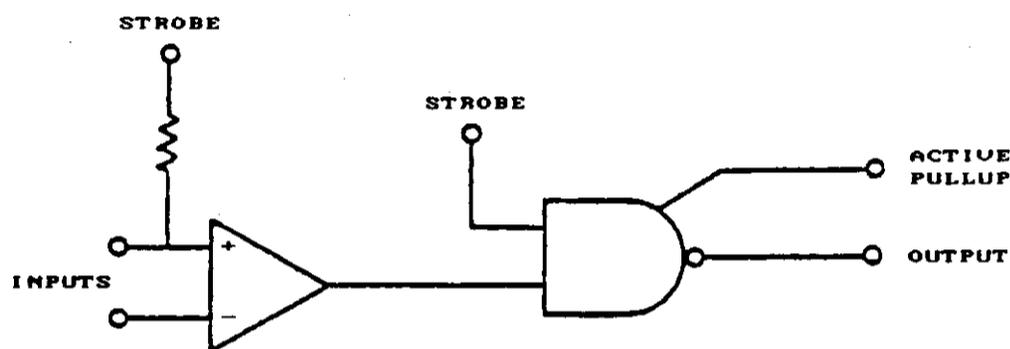


FIGURE 51. Line driver logic diagram.

Line receivers. Output should be compatible with the logic and supply voltage levels of the system. For WIRED-OR applications, open collector output devices should be selected. To reduce the noise to an acceptable level at the input terminals, a high value of common-mode rejection ratio is necessary. Figure 52 shows a logic diagram for a line receiver.

7.10.8.3 Design precautions for line drivers/receivers. Impedance matching the line driver and the line receiver to a transmission line is a necessity in most systems because reflections due to mismatched impedances can introduce errors. The two most common methods for impedance matching are parallel termination and series termination (back matching).

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FIGURE 52. Line receiver logic diagram.

Parallel termination is best suited for applications where there are many receivers for one driver, and the data is fairly symmetrical in nature. The terminating resistor, equal in value to the transmission line impedance, is placed across the transmission line at the furthest point from the driver.

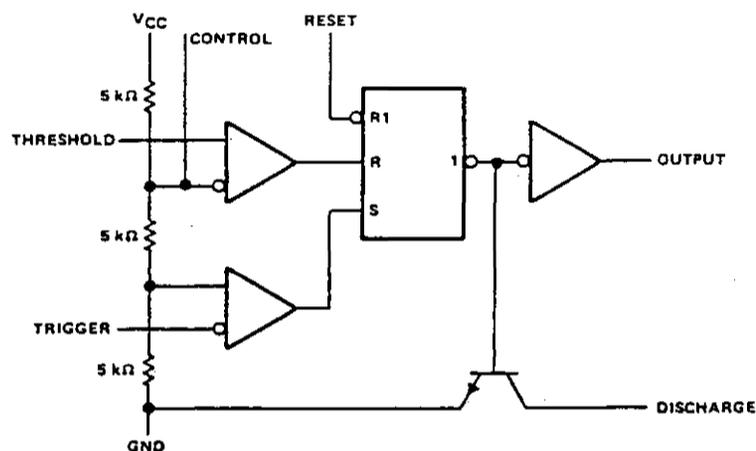
Series terminator can only be used when there is one receiver for every driver. The series resistor has such a value that its resistance plus the output resistance of the driver equal the transmission line impedance. This applies to the single-ended system. For the balanced differential system, each output of the driver has an external series resistance, which when added to its own output resistance, equals one half the impedance of the transmission line.

7.10.9 Timers.

7.10.9.1 Attributes of timers. Monolithic timing circuits are highly stable controllers capable of producing accurate time delays or oscillation. In the time delay mode of operation, the time is precisely controlled by one external resistor and capacitor. For a stable operation as an oscillator, the free running frequency and the duty cycle are both accurately controlled with two external resistors and one capacitor. The functional block diagram of a typical timer is shown in Figure 53.

Today different types of timers are available in single or dual versions, and in bipolar or CMOS technology. Applications for timers include missing pulse detectors, oscilloscope calibrators, darkroom enlarger timers, touch switches, basic square wave oscillators, alternating LED flashers, voltage-controlled oscillators, programmable voltage-controlled timers, frequency synthesizers and many others.

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FIGURE 53. Timer functional block diagram.

7.10.9.2 Critical parameters of timers. General types of monolithic timers are available for use in linear and digital signal processing applications. The important parameters are as follows:

- a. Timing control range nanoseconds to hours
- b. Astable or monostable operating modes
- c. Adjustable duty cycle
- d. High output current
- e. Output logic and supply voltage system compatibility
- f. Temperature stability over the full operating temperature range.

7.10.9.3 Design precautions for timers. Several precautions should be taken with respect to the power supply. The most important is good power supply filtering and bypassing. Voltage ripple on the supply line can cause loss of timing accuracy. A capacitor from the power supply to ground, ideally directly across the device, is necessary. The capacitance value will depend on the specific application. Values of from 0.01 μF to 10 μF are not uncommon. The capacitor should be as close to the device as physically possible.

If timing accuracy is to be maintained, stable external components are necessary. Most of the initial timing error is due to the inaccuracies of the external components. The timing resistors should be the metal film type if accuracy and repeatability are important design criteria. If the timing is critical, an adjustable timing resistor is necessary. A good quality multi-turn pot might be used in series with a metal film resistor to make up the resistive portion of the RC network.

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The timing capacitor should also be high quality, with very low leakage. Ceramic disc capacitors should not be used in the timing network under any circumstances. Several acceptable capacitor types are silver mica, Mylar, polystyrene, and tantalum.

If timing accuracy is critical over temperature, timing components with a small positive temperature coefficient should be chosen. The most important characteristic of the capacitor is low leakage. Obviously any leakage will subtract from the charge count causing the actual time to be longer than the calculated value.

The power dissipation of the package should never be exceeded. With extremely large timing capacitor values, a maximum duty cycle which allows some cooling time for the discharge transistor may be necessary.

7.10.10 Multipliers.

7.10.10.1 Attributes of multipliers. A four-quadrant multiplier with normal XY transfer function can also divide in two quadrants with a $10 V Z/X$ transfer function, and also perform square root and squaring functions. Selection of function in most cases may be accomplished with external passive components only.

In theory, a multiplier has an output which is ideally the product of two input variables, X and Y , divided by the 10-volt scaling voltage. However, the practical multiplier is subject to various offset errors and nonlinearities, which must be accounted for in its application.

As shown in Figure 54, the multiplier may be considered as having two parts; one contains the input circuitry and the multiplying cell and the other is the gain-conditioning operational amplifier, A .

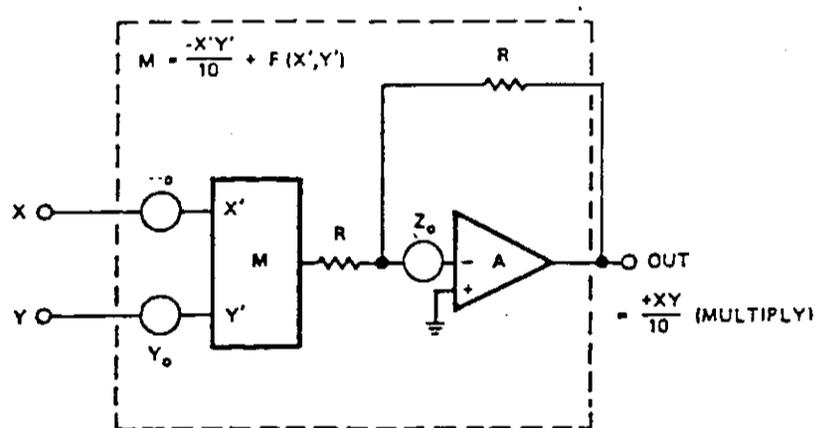


FIGURE 54. Functional block diagram of typical multiplier.

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The "total error" specification includes the effects of all errors. Although a guide to performance, it may produce an excessively conservative design in some applications. For example, output offset is not important if the output is to be capacitively coupled or the initial offset is nulled. Gain error is not important if system gain is to be adjusted elsewhere in the system or if gain is not a critical factor in system performance. If frequent calibration of offset is possible and scale-factor errors are allowed, nonlinearity becomes the limiting parameter. In such cases, improvements in predicted error can be achieved by using the approximate linearity equation:

$$f(X, Y) \cong |V_x| \epsilon_x + |V_y| \epsilon_y$$

where ϵ_x and ϵ_y are the specified fractional linearity errors and V_x and V_y are the input signals.

Multipliers are used for modulation and demodulation, remote gain adjustment, power measurement, and mathematical operations in analog computing, curve fitting, and linearizing.

X or Y feedthrough is the output signal for any value of X or Y when no volts are applied to the other input, the value depends on test conditions, including frequency, temperature and applied signal amplitude.

7.10.10.2 Critical parameters of multipliers. The critical parameters of a multiplier are total error, feedthrough error, input common mode rejection ratio, and output offset voltage.

Total error: The sum of the effects of input and output dc offsets, nonlinearity and feedthrough. Feedthrough error is the sum of the effects of input offsets and nonlinearity.

Input common mode rejection ratio is the ratio of the input common mode voltage to the output error voltage. Larger values mean better rejection of common-mode signals and are therefore more desirable in applications where noise is a problem.

Output offset voltage is the offset voltage at the output amplifier stage. It is usually minimized during manufacture but varies with temperature.

7.10.10.3 Design precautions for multipliers. When selecting a multiplier, device parameters important in a given application must be considered. Attention must be given to parameters which vary with temperature and power supply voltage. This information can be obtained from the data sheet.

All possible causes of output error must be taken into consideration in order to assure the required accuracy of the output.

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7.10.11 Phase locked loops (PLL).

7.10.11.1 Attributes of phase locked loops. Monolithic integration has changed the phase locked loop from a specialized design technique to a general-purpose building block. Today over a dozen different integrated PLL products are available from a number of manufacturers. Some of these are designed as general-purpose circuits, suitable for a multitude of uses. Others are intended or optimized for special applications such as tone detection, stereo decoding, am detection, frequency synthesis, fm demodulation, frequency-shift keyed demodulation, signal conditioning, data synchronization, and motor speed control.

As a functional block, the primary purpose of a PLL is to compare an incoming signal with an internally generated reference signal in both the frequency and phase domain, and to compensate for any differences between the two by adjusting its internal references to match the input signal.

A simple block diagram of a PLL is shown in Figure 55. As can be seen from the figure, the basic elements of a PLL are a phase comparator, a low pass filter, and a voltage-controlled oscillator. The basic principle of operation of a PLL can briefly be explained as follows:

With no input signal applied to the system, the error voltage V_d is equal to zero. The voltage-controlled oscillator (VCO) operates at a set frequency (f_0) which is known as the free-running frequency. If an input signal is applied to the system, the phase comparator compares the phase and frequency of the input signal with the VCO frequency and generates an error voltage, $V_e(t)$, that is related to the phase and frequency difference between the two signals. This error voltage is then filtered and applied to the control terminal of the VCO. If the input frequency, f_s , is sufficiently close to f_0 , the feedback nature of the PLL causes the VCO to synchronize, or lock, with the incoming signal. Once in lock, the VCO frequency is identical to the input signal, except for a finite phase difference.

7.10.11.2 Critical parameters of phase locked loops. Two key parameters of a PLL system are its lock and capture ranges. They are defined as follows:

Lock range: The range of frequencies around the f_0 over which the PLL can maintain lock with an input signal. It is also known as the tracking or holding range. Lock range increases as the overall gain of the PLL is increased.

Capture range: The band of frequencies in the vicinity of f_0 where the PLL can establish or acquire lock with an input signal. It is also known as the acquisition range. It is always smaller than the lock range and is related to the low-pass filter bandwidth. It decreases as the bandwidth is reduced.

7.10.11.3 Design precautions for phase locked loops. The following is a discussion of the key performance parameters associated with various applications.

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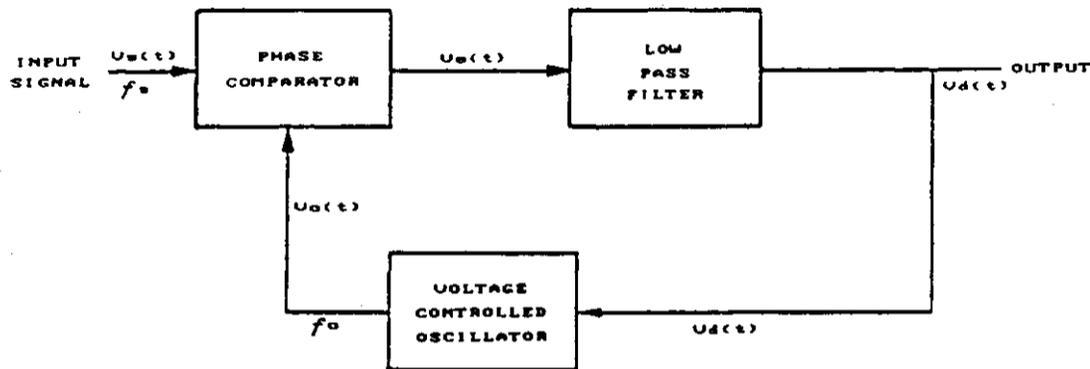


FIGURE 55. Basic diagram of a phase locked loop.

FM demodulation. There are three key parameters which should be examined:

- a. Quality of demodulated output. This is normally measured in terms of the output level, distortion, and signal/noise ratio for a given fm deviation.
- b. Voltage-controlled oscillator (VCO) frequency range and frequency stability. For reliable operation, the VCO upper frequency limit should be at least 20 percent above the fm carrier frequency. VCO frequency stability is important, especially if a narrow-band filter is used in front of the PLL, or multiple input channels are present. If the VCO exhibits excessive drift, the PLL can drift out of the input signal band as the ambient temperature range varies.
- c. Detection threshold: This parameter determines the minimum signal level necessary for the PLL to lock and demodulate an fm signal of given elevation.

In most fm demodulation applications, it is also desirable to control the amplitude of the demodulated output.

Frequency-shift keyed (FSK) decoding. Frequency-shift keying used in digital communications is very similar to analog fm demodulation. Therefore, any PLL integrated circuit can be used for FSK decoding, provided that its input sensitivity and the tracking range are sufficient for a given FSK signal deviation. Some of the basic requirements and desirable features for a PLL used in FSK decoding are center frequency stability, logic compatible output, and control of VCO conversion gain. Center frequency stability is essential to ensure that the VCO frequency range stays within the signal band over the operating temperature range. A logic compatible output is desirable to avoid the need for an external voltage comparator to square the output pulses.

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Frequency synthesis. This application requires a PLL circuit with the loop opened between the VCO output and the phase comparator input, so that an external frequency divider can be inserted into the feedback loop of the PLL.

Signal conditioning. Most signal conditioning applications require very narrow-band operation of the PLL. This in turn may require the use of active filters within the loop, i.e., between the phase detector and the VCO.

7.10.12 Physical construction and mechanical considerations. Construction and mechanical considerations of linear and special purpose circuits are generally the same as for other microcircuits. A vast majority of monolithic microcircuits are constructed with the planar diffusion process. Basic descriptions of fabrication, assembly, thermal characteristics, reliability and other details of semiconductor devices are given in the general subsection on transistors (5.1) and microcircuits (7.1).

7.10.13 Military designation. A description of the designation for military microcircuits are given in subsection 7.1 General, paragraph 7.1.32.

7.10.14 Environmental considerations. Environmental considerations of microcircuits are included in subsection 7.1 General, paragraph 7.1.7 General reliability considerations.

7.10.15 Reliability considerations. Reliability considerations of linear and special purpose microcircuits are generally similar to other microcircuits. They are treated in subsection 7.1 General, paragraph 7.1.7 General reliability considerations. Derating is an important technique of increasing linear microcircuit reliability. MIL-STD-975 should be consulted for applicable derating factors for the various types of linear microcircuits.

7.11 MICROCIRCUITS, HYBRID

7.11 Hybrid.

7.11.1 Introduction. A hybrid microcircuit is a microelectronic device that uses a combination of active and passive circuit elements from more than one technology. Circuit functions can be implemented in hybrid form that offers an economic and technical advantage over the same function implemented in monolithic form. Hybrid microcircuits are divided into three general categories.

- a. Multichip hybrid microcircuits. The multichip hybrid microcircuit is defined as a device that contains several active elements mounted on a header with no thin- or thick-film associated circuitry.
- b. Simple hybrid microcircuits. The simple hybrid microcircuit is more complex than the multichip version and consists of one or more active and passive elements mounted and interconnected by deposited metallization on a single layer substrate. It may also have thin- or thick-film resistors deposited on the substrate. Military specification MIL-M-38510 defines the package inner-seal perimeter of simple hybrid microcircuits as not exceeding 2 inches.
- c. Complex hybrid microcircuits. Complex hybrid microcircuits contain the same basic combination of active and passive elements as the simple hybrid microcircuits except that the number and complexity of the elements is greater and the substrate may consist of several conducting layers. The package inner seal perimeter of the complex hybrid is defined as greater than 2 inches.

Compared with monolithic devices, the design costs of hybrids are relatively small. Minor changes readily can be made to an existing design to optimize a hybrid device for an application. Although some standard circuits are marketed, most hybrid devices are at least partially custom designed. Custom designing provides hybrid microcircuits with a performance advantage, but at some expense of reliability assurance because high volume production of any one design is seldom achieved.

Hybrid microcircuits will continue to be a major factor in new system designs even though monolithic devices continue to expand to higher levels of complexity and encompass functions previously available only in hybrid form. Continued use of hybrid microcircuits will be accomplished through the use of the new, more complex monolithic dice into even more complex hybrid microcircuits, thus allowing hybrid circuits to remain at the forefront of microcircuit technology.

7.11.2 Usual applications. Although nearly any integrated circuit can be made in hybrid form, those that provide characteristics or functions beyond the present capabilities of monolithic devices are of great interest for aerospace applications. Monolithic devices are generally preferred for applications within their capabilities because of their greater anticipated reliability and economy and, sometimes, smaller size.

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Some microcircuits would be impossible to implement in monolithic form, but may be easily realized with a hybrid design. For example, circuits requiring large capacitances or resistor networks are easily fabricated in hybrid form. Hybrid microcircuits are often called for when power, high frequency, or close tolerance circuit elements are required. Some specialized hybrid circuits use the thermal coupling between semiconductors in close proximity to provide controls not found in discrete circuits, such as a corrective feedback to prevent thermal runaway.

Typical systems applications of hybrid microcircuits include control, display, power, transmitter, receiver, and signal processor circuits. Hybrid micro-wave integrated circuits are normally used as receivers, mixers, transmitters, phase comparators, and levelers. A hybrid microcircuit can be designed and prototype hardware delivered in four to six months. By comparison, development of a custom monolithic circuit may require 12 months for delivery of prototype samples. Hybrid packages are ideally suited to consolidate SSI and MSI chips into small physical volumes with high electronic capability.

7.11.2.1 Available functional types. A large variety of hybrids are available off-the-shelf. Some of the functional applications of standard hybrids are in digital signal processing, operational amplifiers, voltage references, analog multiplexers, sample-and-hold amplifiers, D/A and A/D converters, flash converters, and data acquisition systems. Custom hybrids may be obtained to these products and to such applications as voltage to frequency converters, instrumentation amplifiers, modulators, filters, analog switches, bus interface, analog and digital interface, and thin-film resistor networks.

7.11.2.2 Comparative attributes of packages. A large variety of hybrid packages are used in aerospace applications. The major package types are: leadless, plug-in bathtub or uniwall, the all-metal power-package, beryllia-base power-package, platform, and flatpack. Figure 56 shows four typical package types. The standard flatpack is the most common. Figure 57 shows the general features of a flatpack package and Figure 58 displays details of the package cross-section.

All package types are available in a wide range of sizes and lead count. For example, the flatpack comes in sizes from approximately 0.6 x 0.6 inch to 2 x 2 inch length and width, and lead counts of 20 to approximately 120. Table XXXIV is a comparison of advantages and disadvantages of six major hybrid package types. The choice of package will depend on the design constraints imposed on the hybrid such as high-wattage elements, minimum board area, and high lead count, etc.

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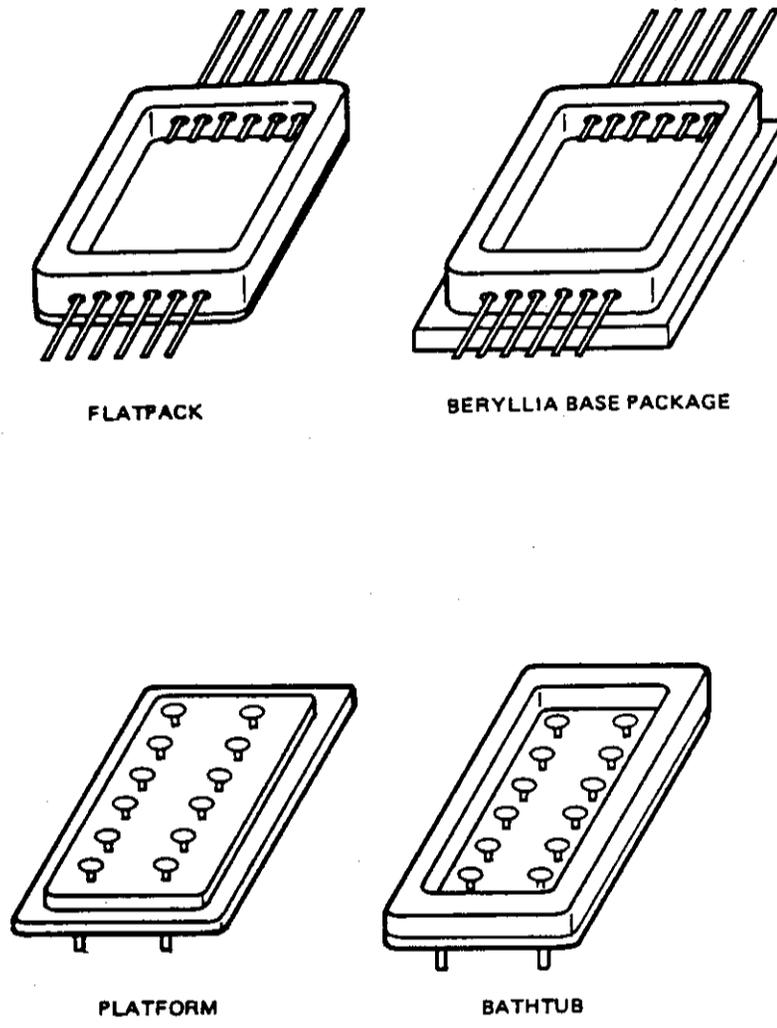


FIGURE 56. Four common hybrid package types without covers.

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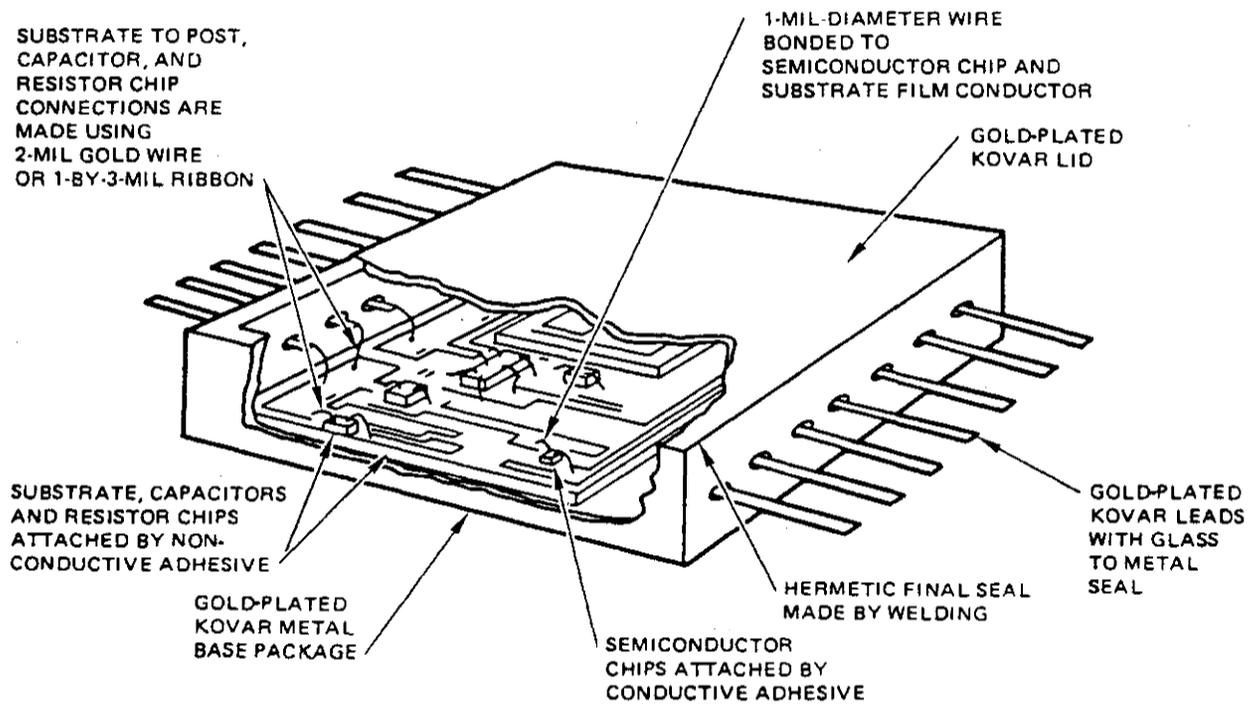


FIGURE 57. General features of a flatpack hybrid microcircuit.

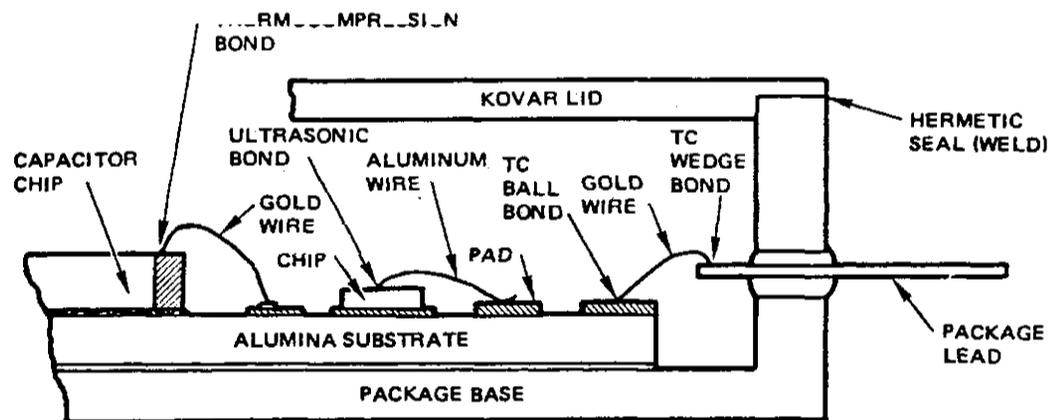


FIGURE 58. Cross-section of a hybrid substrate packaged in a flatpack.

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7.11 MICROCIRCUITS, HYBRID

TABLE XXXIV. Comparison of characteristics of common hybrid packages

Type	Characteristics
<p>Leadless packages</p> <p>Advantages</p> <p>Disadvantages</p>	<p>Ceramic with metal ring frame for weldable covers and castellated leads</p> <p>No glass-to-metal seal, uses less PWB area</p> <p>Because of thermal expansion these packages can only be used with ceramic PWBs. Connections can only be made using solder. It is difficult to inspect the connection and to remove the case from the PWB.</p>
<p>Plug-in bathtub</p> <p>Advantages</p> <p>Disadvantages</p>	<p>Uniwall metal base with sidewalls and pins on two, three, or four sides, preferred to platform packaging for new designs.</p> <p>Packages can be welded using automated techniques. Parts can be opened and resealed.</p> <p>Some attention must be given to device positioning, especially for active elements that are to be wedge- or aluminum-wire bonded so that bonding tool tip clearances are satisfied. Domed lids must usually be special-ordered. Packages cannot be repaired if the pins are fractured near the package walls or if the glass beads crack. When plug-in packages are removed from module boards, package hermeticity is lost if glass beads around the package pins are damaged during removal; the hybrid then needs repackaging.</p>
<p>Metal power packages</p> <p>Advantages</p> <p>Disadvantages</p>	<p>Copper, Kovar, or molybdenum power packages with weldable lids.</p> <p>Best for heat dissipation in power circuits. (NOTE: Kovar is not as good as beryllia, described elsewhere).</p> <p>Require special process and equipment. They are heavy and usually are custom designed. Also, they are expensive and difficult to seal.</p>

7.11 MICROCIRCUITS, HYBRID

TABLE XXXIV. Comparison of characteristics of common hybrid packages (continued)

Type	Characteristics
Beryllia-base power-package	Metal ring frame soldered to a beryllia base
Advantages	Used when the heat generated by high-wattage elements in a hybrid must be more efficiently dissipated than is possible using alumina or all-metal packages. A number of different beryllia-base thicknesses are available. Packages can be sealed using flat or domed lids. Sealing can be done by welding or soldering.
Disadvantages	Though sealed packages can be opened and resealed, the operation is not as repeatable, and is therefore more expensive than when using all-metal packages. This is because the glass beads in the ring frame of the beryllia package are less likely to withstand repeated thermal and mechanical stresses. Beryllia is toxic and requires special handling and processing.
Plug-in platform	Flat metal base with pins on two, three, or four sides. This is not recommended for new designs.
Advantages	Because there are no sidewalls, wire bonding near the edge of the substrate is not a problem. Domed lids of different heights are available. Sealed packages can be opened and resealed.
Disadvantages	The configuration is sealed by hand soldering. There is some possibility of solder-ball formation, especially with long (>1.5-inch) packages. If pins are fractured or if the glass beads around the pins crack, the package cannot be repaired. Package hermeticity is lost if glass seals around the package pins are damaged when the package is removed from module boards. The hybrid then needs repackaging.
Flatpack	Metal ring frame brazed to a metal base or uniwall construction. Radial leads on two, three, or four sides.
Advantages	Packages can be sealed by welding. Step or flat lids are normally used, but domed lids are available so that larger elements can be incorporated. These packages can be opened and resealed.
Disadvantages	Leads fractured near package walls or cracked glass beads around these leads cannot be repaired.

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7.11.3 Physical construction. Construction of hybrid devices begins with a substrate on which the circuit is built. This substrate is usually an alumina- Al_2O_3 ceramic which provides a good combination of electrical insulation and heat conduction properties.

Selection criteria for various substrates are summarized in Table XXXV.

TABLE XXXV. Substrate selection criteria

Substrate	Comment
Unglazed alumina	Normally preferred for thick- and thin-film circuits.
Polished alumina, sapphire	Preferred for microwave circuits.
Ground alumina	Preferred for multilayer and large-area thick-film circuits.
Beryllia	High-power dissipation; can be used as a mother board for alumina substrates.

In applications where superior heat transfer properties are required, (e.g., due to the use of power transistor dice), a suitable substrate material is beryllia [beryllium oxide (BeO)]. However, a word of caution is necessary concerning the use of beryllia. This substrate is extremely poisonous under certain conditions. It should not be ground, filed, machined, or otherwise abraded. Exposure to high temperature (above 250°C) is also to be avoided. Airborne particles and fumes which are released under the above conditions are lethal.

Microwave hybrid microcircuit designs are optimized by the use of sapphire substrates. Although this may be more expensive, its use can be justified by the superior performance which is achieved. Parameters which are significantly improved through the use of sapphire are intermodulation distortion (crosstalk) and noise figure.

7.11 MICROCIRCUITS, HYBRID

Conductors, insulating layers, and passive elements are deposited on the substrate using thin-film or thick-film techniques. Semiconductor dice are generally attached to gold pads by means of silicon-gold eutectic bonding, although other techniques, such as epoxy and Teflon as well as beam lead bonding are sometimes used. Interconnection of semiconductor bonding pads, passive elements, and external leads is usually accomplished by gold wire using ball bonding (see Figure 59 for a typical construction).

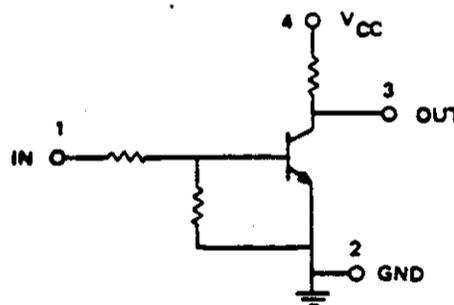
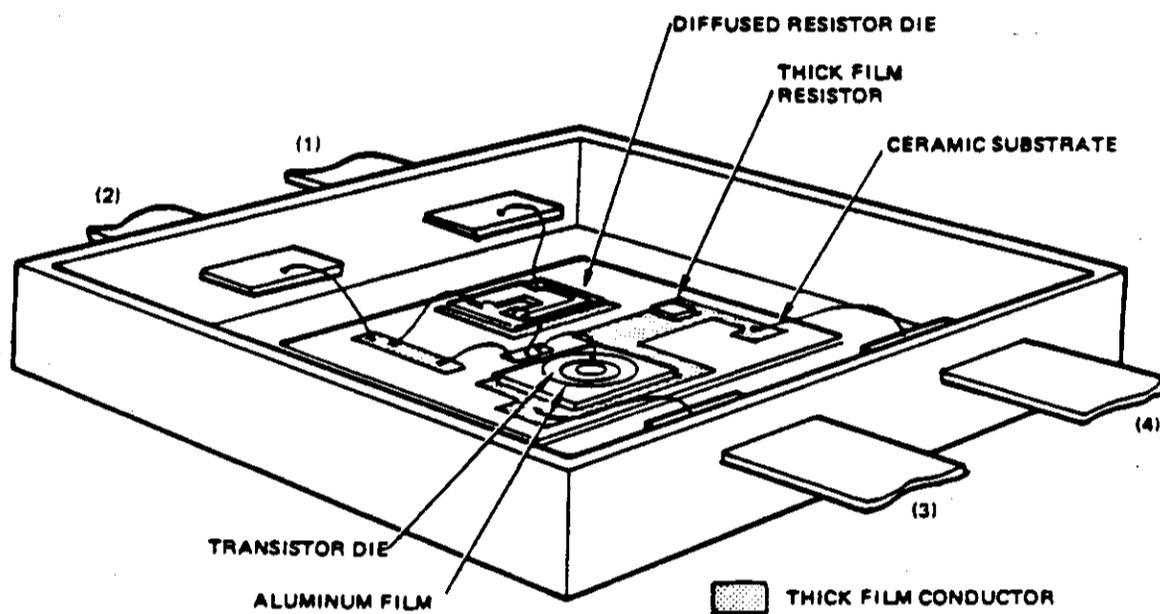


FIGURE 59. Comparison of schematic diagram and typical construction of a hybrid microcircuit.

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Although hybrid microcircuits allow use of circuit elements with much closer tolerances than those provided by monolithic circuits, close tolerances of the completed circuit are not always easily achieved. As with diffused resistors, the film resistors as deposited provide a better matching of resistances and temperature coefficients than tolerances on absolute values. They may, however, be trimmed to much closer tolerances by means of power pulses, abrasion, or laser trimming. More complex circuits which use many semiconductor dice may suffer degradation of parameters as a result of long exposure at the high temperatures encountered in die bonding and wire bonding.

The development of the leadless chip carrier and the increasing availability of circuit elements in this form is of significant value to the manufacturer of complex hybrid microcircuits. Utilization of circuit elements packaged in the chip carriers allows 100-percent reliability screening prior to installation in a hybrid microcircuit, a distinct advantage over unpackaged dice which historically have caused electrical and/or environmental yields as low as 60 percent in the finished hybrid microcircuit. Chip carriers are hermetically sealed and also lend themselves to automated assembly techniques, thus simplifying hybrid microcircuit assembly and processing.

7.11.3.1 Thin-film. Thin-film circuits are made by vapor deposition of conductors, resistors, and dielectrics through a metal mask in a vacuum chamber (selective etching may be used as an alternative). Conductor patterns are generally gold, although aluminum may be used. Nichrome resistors with sheet resistance up to 300 Ω /square and cermet resistors with sheet resistance up to 5000 Ω /square are the most widely used types, although other metals such as tantalum and titanium may be used. Thin-film capacitors using silicon monoxide or silicon dioxide as the dielectric may be used for smaller values, but they require too much area for values greater than about 100 pF. For larger values, discrete types such as ceramic chip capacitors are used.

See summary of thin-film element characteristics in Table XXXVI.

7.11.3.2 Thick-film. Thick-film circuits are most generally produced by applying metallic and insulating pastes to the substrate through screening masks. Each application is followed by a pass through an oven to set the paste and to bond it to the substrate. Gold paste is generally used for the conductor pattern. Resistor pastes, which are generally proprietary formulations, come in a wide range of sheet resistivities. Capacitors may be made to values as high as 10,000 pF by thick-film techniques or discrete capacitors may be used. A protective glass coating is generally applied to the entire circuit as a final step. For additional circuit characteristics, see summary of thick-film element characteristics in Table XXXVI.

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TABLE XXXVI. Thin-film and thick-film characteristics

Material	Comment
Thin-film conductors	These are used when high-density, single-layer packaging is required. They allow narrower lines and spacing than are possible with thick-film conductors.
Thin-film resistors	These can be etched from thin-film conductor substrates. They are used for values up to 20 k Ω (and in some instances up to 100 k Ω) and for dynamic trimming to tight tolerances. These resistors are limited to the lower values because only relatively low sheet resistivities are available.
Thick-film conductors	They are less expensive than thin-film conductors, especially when many single-layer, conductor-only substrates are involved. Conductor lines and spaces are not as narrow as those possible with thin-film, so some packaging density is lost. Packaging density can sometimes be improved by redesigning a dense, single-layer thin-film circuit to take advantage of the thick-film multilayer concept.
Thick-film resistors	A wide range of resistor values can be obtained by selecting from a range of sheet resistivities that vary from 1 Ω to more than 1 M Ω per square. Often, chip resistors are used in hybrids to take advantage of their high resistivities and relatively low demands for space on the substrate. These resistors can be dynamically trimmed but it is not recommended.

7.11.3.3 Die requirements. Microcircuits and semiconductor die used for construction of class S or class B hybrids shall meet the screening and quality conformance provisions of Method 5008 of MIL-STD-883 for the appropriate reliability class.

7.11.3.4 Specification of passive elements and packages. Passive elements and packages used in the construction of class S and class B hybrids shall meet all quality and reliability requirements listed under class S or class B, respectively, as specified in Appendix G of MIL-M-38510 and further detailed in Method 5008 of MIL-STD-883.

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7.11 MICROCIRCUITS, HYBRID

7.11.4 Military designation. Quality and reliability assurance requirements for class S and class B hybrid microcircuits are called out in Appendix G in the General Specification for Microcircuits (MIL-M-38510). NASA quality levels of Grade 1 and Grade 2 are satisfied by class S and class B, respectively. Implementation of the requirements of Appendix G is detailed in Method 5008 of MIL-STD-883, Test Methods and Procedures for Microelectronics. Method 5008 establishes the detailed screening and quality conformance procedures for achieving class S and class B levels for microcircuits, semiconductors, passive elements, and packages used for hybrid construction and should be strictly satisfied for NASA hybrid applications. A further description of NASA Standard Parts Program and the designation for military microcircuits are given in subsection 7.1 General.

7.11.5 Electrical characteristics. Most hybrid microelectronic devices are custom made for their particular applications. However, the part operation should be precisely defined by a table of electrical characteristics which includes all necessary parameters with test conditions, parameter minima and maxima, parameter typical values, and associated symbols with units. Electrical characteristics should also include areas such as a functional diagram, a block diagram, a table of pin assignments, a mechanical outline, and a table of absolute maxima ratings. A worst-case circuit design analysis should be performed following the requirements of Appendix G of MIL-M-38510.

All active and passive elements of the hybrid should have electrical characteristics derated according to the appropriate derating factors listed in MIL-STD-975 for NASA standard parts. Derating will increase the safety margin between the operating stress level and the actual failure level for the constituent hybrid parts and will provide added protection for system anomalies unforeseen by the applications engineer.

7.11.5.1 Special testability considerations. Special care should be taken in hybrid applications to ensure that the device is testable. Ideally, built-in test procedures should be incorporated to set the state of all elements of the hybrid and to observe such states. The hybrid designer should anticipate and allow for test equipment and system test limitations when designing built-in test features. Test time should be minimized, because test machine time may sometimes cost more than the part itself. Extra test pins are sometimes needed to adequately test a hybrid part.

7.11.5.2 Thermal design considerations. A thermal design analysis should be performed on the hybrid following the conditions of Appendix G of MIL-M-38510. All active and passive elements of the hybrid should be derated to worst-case conditions.

All active and passive elements in hybrids should be, as a minimum, derated to the requirements of MIL-STD-975.

Hybrid manufacturers shall comply with the facilities and manufacturing line requirements of MIL-STD-1772.

7.11 MICROCIRCUITS, HYBRID

7.11.6 Environmental considerations. NASA hybrids typically have demanding environmental requirements. Extremes in environmental tolerance arise from maximum operating limits of temperature-sensitive chip elements, electrical performance, power dissipation, vibration and shock, and radiation. Environmental test requirements are satisfied, except for radiation, by the class S requirements of Table VII of Method 5008, MIL-STD-883. Radiation testing of the constituent microcircuit and semiconductor die of class S hybrids should be done following the requirements of Table III of Method 5008.

7.11.7 Reliability consideration. Hybrid microcircuits for NASA Grade 1 or 2 applications require class S or B quality and reliability assurance levels. Appendix G of MIL-M-38510 details the procedures for documenting the requirements for such hybrids. Appendix G references method 5008 of MIL-STD-883 to describe the hybrid device evaluation procedures to implement these requirements. The hybrid device evaluation has four general requirements: element evaluation, process control, device screening, and quality conformance evaluation. These requirements should be satisfied at the class S quality level for NASA Grade 1-type applications. For Grade 2 NASA applications, the hybrid device evaluation must satisfy the class B quality level.

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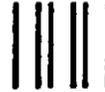
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15 MARCH 1984

MILITARY HANDBOOK

NASA PARTS APPLICATION HANDBOOK

(VOLUME 4 OF 5)

CRYSTALS, FILTERS, TRANSFORMERS,
INDUCTORS, DELAY LINES, MOTORS



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NASA Parts Application Handbook

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FOREWORD

This handbook provides a technological baseline for parts used throughout NASA programs. The information included will improve the utilization of the NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List (MIL-STD-975) and provide technical information to improve the selection of parts and their application, and failure analysis on all NASA projects. This handbook consists of five volumes and includes information on all parts presently included in MIL-STD-975.

This handbook (Revision B) succeeds the initial release. Revision A was not released. The content in Revision B has been extensively changed from that in the initial release.

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8. CRYSTALS

8.1 Introduction. Crystals are not listed in MIL-STD-975, NASA Standard Electrical, Electronic, and Electromechanical (EEE) Part List, but are included in this handbook. Because of the complexity of quartz crystal technology and the variety of application techniques, the information contained herein is restricted to brief reviews of basic design considerations and application factors of crystal units and crystal oscillators.

Both natural and cultured quartz are used for frequency control. Most natural quartz comes from Brazil. Cultured (or synthetic) quartz is grown at several domestic facilities. Natural quartz is variable in size and shape with numerous flaws and inclusions whereas with cultured quartz, the major axis of growth is predetermined and the quality is essentially uniform throughout the stone. Processing natural quartz into resonators of critical dimensions and crystallographic angular orientation is much more labor intensive and wasteful of material than cultured quartz.

Quartz is useful in the electronics field because it is not only piezoelectric, but also has very stable physical properties and useful behavior over wide temperature ranges. The piezoelectric effect causes an electrical potential difference to appear between two opposite faces of a quartz plate when the plate is mechanically stressed and, conversely, causes mechanical stress and movement to be generated in the quartz plate when a potential difference is applied to two opposite sides.

Quartz plates cut to proper dimensions and appropriately mounted will vibrate in mechanical resonance at the frequency of an alternating voltage applied to opposite faces of the plates. The unusually high stiffness and elasticity of this very hard crystalline material make it possible to produce electro-mechanical resonators that will operate over a wide band of frequencies extending to 175 MHz; experimental work is being done to 1 GHz.

During the past few years, the quartz crystal art has advanced by what might be considered two orders of magnitude. Part per million accuracy has advanced to part per hundred million and what was part per hundred million is today part per 10 billion. As more and more basic cause and effect relationships are proven, better predictions of expected performance under widely varying environments may be expected.

8.1.1 General definitions

Accuracy. The degree of precision with respect to a referenced, specific, or nominal value. Accuracy and stability are not directly related, yet they are commonly used incorrectly and interchangeably.

Coupled modes. Also known as activity dips, are vibrational modes which generally manifest themselves over a very narrow temperature range, often as little

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as a fraction of a degree. Coupled modes cause the crystal frequency to decrease, with a simultaneous increase in equivalent resistance; thus, the term "activity dip." Their magnitude, width, and quantity are proportional and highly sensitive to the drive level. They may actually be made to disappear at a low enough drive level. Coupled modes are of concern in crystals used in phase-locked loops and in temperature-compensated oscillators.

Crystal cut. Quartz crystals have three mutually perpendicular axes. The angle of rotation with respect to these axes at which the resonator is cut from the stone determines its electrical parameters, frequency range of feasibility, and temperature characteristics. Most crystal cut designators consist of two letters such as AT and SC.

Crystal holder. The sealed enclosure in which a quartz resonator is mounted; it includes a cover, a base or other means of closure, and suitably insulated pins or leads. It does not include the resonator mounting structure.

Crystal unit. An assembly consisting of a quartz resonator suitably mounted in a crystal holder.

Drive levels.

- a. Minimum. The lowest value at which the crystal will oscillate. It is used to minimize internal temperature rise when determining frequency change from before to after a test such as vibration or aging.
- b. Rated. The power dissipation level at which the crystal unit is designed to operate within specified tolerances of various electrical parameters. This drive level is generally established in a specific test-set at the maximum allowable equivalent series resistance.
- c. Reduced. The name of a test used to determine the ability of a crystal to begin oscillating at a much lower level than the rated drive level (typically less than 10 μ W). It is indicative of the cleanliness and surface condition of the quartz resonator. This test generally applies to overtone units.

Equivalent resistance. The equivalent resistance of a crystal unit is defined as follows:

- a. For crystal units designed to operate at series resonance, equivalent resistance is defined as the equivalent ohmic resistance of the unit when operating in the specified crystal impedance meter (or other test-set) adjusted for the rated drive level and tuned to the specified crystal unit frequency.
- b. For crystal units designed to operate at parallel or antiresonance, equivalent resistance is defined as the equivalent ohmic resistance of the unit and a series capacitor of the specified load value when operating in the specified crystal impedance meter (or other test-set) adjusted for rated drive level and tuned to the specified crystal unit frequency.

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Operable temperature range. The temperature range over which a unit designed for controlled temperature conditions is to continue to oscillate (not necessarily within specified performance limits).

Operating temperature range. The temperature range over which a unit is to operate within specified performance limits.

Oscillators. A device which has a specific output frequency and accuracy in relation to specific input power. The three types of oscillators are:

- a. OCXO. Oven temperature controlled crystal oscillator (ovenized)
- b. TCXO. Temperature compensated crystal oscillator
- c. XO. Crystal oscillator.

Reference temperature. The ambient temperature at which certain crystal parameter measurements are made. For controlled temperature units, the reference is the midpoint of the controlled operating temperature range. For noncontrolled temperature units, the reference temperature selected is normally room temperature (23 ± 1 °C).

Specified frequency. The nominal frequency at which the crystal unit is designed to operate under specified conditions.

Stability. The amount of change due to a specific event, and it may be transient or permanent. Stability and accuracy are not directly related, yet they are commonly used incorrectly and interchangeably.

Unwanted modes. Also known as spurious modes or spurs. Vibrational modes near the desired mode of oscillation that are intrinsic in many crystallographic designs. They will be similar from unit to unit. Because they are predictable, they can be dealt with in oscillator designs by trapping or filtering, and they are often the limiting factor in filter performance. They are not particularly drive-level sensitive.

8.1.2 Symbols. Some of the symbols commonly used in quartz crystal terminology and associated circuitry are listed below.

C_d	Total distributed capacitance across leads or terminals of a crystal unit
C_e	Electrostatic capacitance across the quartz resonator
C_1 or C_M	Motional arm (series-arm) capacitance of a crystal equivalent circuit.

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C_0	Total electrostatic shunt capacitance of a crystal unit
C_L or C_x	Load capacitance in series with a parallel-mode crystal unit
C_T	Total capacitance ($C_0 + C_x$) shunting series arm of a crystal equivalent circuit
DRAT	Doubly-rotated AT-cut crystal unit
f_a	Antiresonant frequency of a crystal unit
f_o	Oscillator output frequency, frequency of a crystal unit
f_p	Parallel resonant frequency of a crystal unit
f_r	Resonant frequency of a crystal unit
f_s	Series resonant frequency of a series arm
Δf	Any change in frequency, often used to express difference between f_a and f_r ; i.e., bandwidth
L_M	Motional arm (series arm) inductance of a crystal unit
Q	Quality factor $\frac{1}{\pi F C_M R_S} = \frac{2\pi F L}{R_S}$
r	Ratio of total electrostatic shunt capacitance (C_0) to motional-arm capacitance (C_M) of crystal equivalent circuit
R_S or ESR	Equivalent series resistance of crystal unit

8.2 Usual applications.

8.2.1 Crystal units. Crystal units are the basic crystal device. They, in turn, are used in crystal oscillators and crystal filters.

8.2.2 Crystal oscillators. Crystal oscillators are made in a large variety of types to suit a large number of widely differing needs. They range from the 100-ppm accuracy class to 10^{-12} /day stability class, from 2 grams to 3 pounds, and from 0.02 cubic inch to over 40 cubic inches. They consume from a few microwatts to several watts and cost from a few dollars to more than \$10,000. Designs range from a simple clock output to oscillators with a great array of features (some of which are listed below) and are realizable in many combinations.

This section is intended as an overview of crystal oscillators and not as a catalogue of available devices. Refer to MIL-O-55310 and the parts specialist for further details.

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In the basic quartz crystal-controlled oscillator (XO), the crystal serves to stabilize the output frequency of an oscillator circuit.

The largest single source of oscillator frequency instability is the response of the crystal to temperature change. To minimize this effect, oscillators may be temperature-compensated (TCXO) or "ovenized" (OXCO); i.e., incorporating an oven as an integral part of the device construction to provide temperature control for optimum stability. The difference in performance between these oscillator types is shown in Table I.

TCXO. Temperature compensation is achieved by circuits using thermistors, varactors, and resistors. Currently evolving digital compensation techniques will add EPROMs, A/D and D/A converters, and comparators. This circuitry creates a temperature characteristic (TC) equal to, but of opposite slope, of the crystal TC, which maintains frequency stability over temperature by acting as a variable load on the crystal oscillator circuit. The rate of change of ambient temperature (ramp rate) that can be tolerated while maintaining desired frequency stability is rather limited (e.g., 1 °C per minute for $\pm 2 \times 10^{-6}$ stability).

OXCO. Temperature control is achieved by various combinations of devices: heater windings, heater blankets, thermostatic control, proportional control, power transistors as the heat source, self-limiting ovens for just the crystal, single and double ovens, booster heaters for fast warm-up, or Dewar flasks of glass, steel, or titanium. Safety devices to prevent thermal runaway may be provided.

The sophistication of temperature control ranges from coarse (several degrees) to very fine (better than 0.01 °C), over a wide ambient temperature range. In some applications only the crystal itself is temperature controlled.

8.2.2.1 Oscillator features. The following list of oscillator features, although not complete, is indicative of the many options available to the designer and it highlights the inadequacy of speaking of a "standard oscillator."

- a. Cased crystals
- b. Uncased crystals
- c. Single and multiple output frequencies
- d. TTL, CMOS, ECL, sinewave outputs
- e. Reverse polarity protection
- f. Supply voltage conditioning
- g. Electrical and mechanical adjustment of f_0 and E_0 using a potentiometer, fixed resistor, or variable capacitor

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- h. Remote frequency update in 1×10^{-11} steps
- i. Voltage control of f_0 with linearity adjustment
- j. Integral crystal filter (for phase noise control)
- k. Integral output buffer
- l. Integral wave shaper
- m. Integral amplifier
- n. Integral impedance matcher
- o. Integral multipliers and dividers
- p. Test points (for oven voltages)
- q. Output disable (for more rapid on and off)
- r. Any number of physical configurations, mounting means, connectors, and terminals
- s. Discrete components or hybridized construction (perhaps in combination) with various criteria for selection and screening.

8.2.3 Crystal filters. Crystals are used in band-pass (and occasionally band-reject) filters. Narrow bandwidths and high attenuation are attainable in a small package because of the high Q of the crystals. Filters with only AT-cut crystals are discussed here. The number of crystals per filter varies from 2 to more than 16. In size and weight, filters range from those with discrete components in boxes of 6 cubic inches at 6 ounces, to monolithic filters in TO-8 or flatpack packages at 1 cubic inch and 3 grams. Bandwidths of 0.01 to 3 percent of center frequency are practical for fundamental frequency AT-cut crystals. On the third overtone, bandwidths of up to only a few tenths of one percent are realizable. The upper bandwidth limit is determined by the spurious responses of the crystals. The lower limit is determined by the need for the bandwidth to be sufficiently wider than the desired signal spectrum to allow for temperature and aging effects.

Attainable stopband attenuation ranges up to 100 dB with the most common applications being in the range of 40 to 70 dB. The ultimate attenuation is limited by the grounding and shielding techniques applied externally to the filter.

Not all bandwidths are available at all frequencies. Some typical values are shown in Table II.

TABLE I. Typical crystal oscillator categories

Type of Oscillator	Frequency Stability	Temperature Range	Power Consumption	Configuration, Volume, Weight	Comments	Relative Cost
Crystal controlled, with TTL, CMOS, ECL, or sine-wave output	±50 ppm ±20 ppm ±10 ppm	-55 to +125 °C -20 to +70 °C 0 to +70 °C	<100 mW	Cold-weld or seam-sealed TO-5, TO-8, DIP, or LCC; 0.02 cubic in; 3 grams	Components may be discrete or hybridized. Crystals may be uncased. Power depends on frequency and type of output.	\$X
Temperature-compensated crystal oscillator (TXCO)	±1 ppm ±2 ppm	-10 to +70 °C -55 to +85 °C	10 to 100 mW	1-3 cubic in, 2 oz	There is a warm-up time of several seconds. There is a permissible ambient temperature ramp of 1 °C/minute max.	\$10X
Ovenized crystal oscillators (OCXO)	±1 x 10 ⁻⁷ per day to ±1 x 10 ⁻¹² per day	-40 to +85 °C	4 W warm-up; 1 W sustaining; 0.5 W oscillator	40 cubic in, 3 lb	Warm-up time may be hours or days.	\$100X

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TABLE II. Typical filter bandwidths

Center Frequency	Mode of Oscillation	Bandwidth
30 MHz	Fundamental	500 kHz 1.6%
60 MHz	3rd Overtone	60 kHz 0.1%

8.2.4 Other applications. Quartz crystals are also used in discriminators. The output of a discriminator is a dc voltage, the polarity and amplitude of which is directly relatable to the frequency of the rf input.

Certain crystal cuts have a reasonably linear temperature characteristic of 200 to 300 ppm/°C, making them ideal as high resolution thermometers. Because the frequency of crystals is sensitive to pressure and weight, they are used as indirect indicators in industrial processes such as measuring plating thickness, detecting liquid flow, and determining the presence of atmospheric impurities.

8.3 Materials and physical construction. Crystals are typically edge-clamped in a two- or three-ribbon mounting system. Electrodes are plated on the opposite crystal surfaces with the plating extending to the opposite edges to provide attachment to the mounts. The crystal faces are only partially plated so that the effective electrode area is confined to a small circular region at the center of the blank. Thus, the capacitance is kept to a minimum and the principle resonating activity is confined to the middle where the crystal is most likely to be of uniform thickness. Both of these factors contribute to frequency stability. Final plating is done to fine tune the crystal frequency response. Figure 1 shows a typical crystal unit construction.

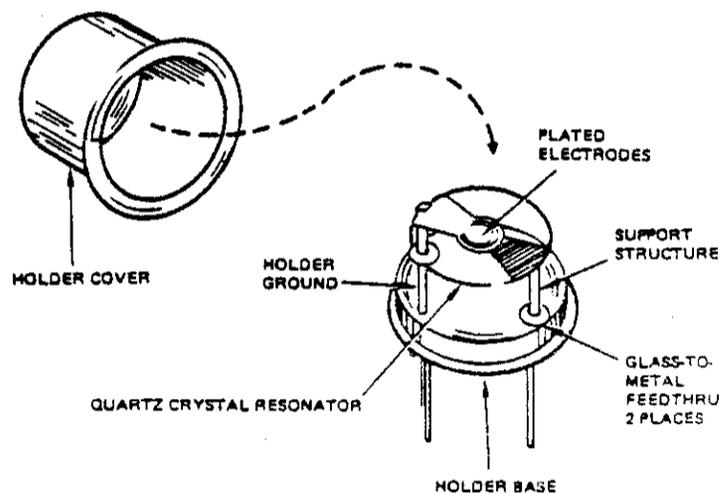


FIGURE 1. Typical construction of a quartz crystal unit in a transistor style holder.

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8.3.1 Quartz material. The quartz resonator should be pure Z-growth cultured quartz. The crystal should be swept (impurities removed by an electrodiffusion process) unless it can be clearly demonstrated to be unnecessary for the application. The fundamental frequency should not be less than 1 MHz and preferably greater than 3 MHz. Minimum thickness of the crystal should be 0.0025 inch with 0.003 inch preferred. The base plate should be gold. Gold applied in a dry plating process should be used for final frequency adjustment.

The crystal should be of a high-frequency crystallographic cut, either AT or doubly-rotated AT (DRAT) cut.

8.3.2 Crystal cut. The crystal cut is the angle in respect to the three crystallographic perpendicular axes to which the crystal is ground or cut. It is the determining factor in crystal performance. The AT and double-rotated AT cuts are preferred. Lower frequency cuts should be avoided.

The advantages and disadvantages of the DRAT versus AT crystallographic cuts are stated in generalities and approximations because the specific values are dependent upon frequency, crystallographic angle, surface finish, plating configuration, support structure, holder, and the precision of manufacture.

The DRAT-cut crystals have lower acceleration sensitivity, can operate at a high drive level to minimize phase noise without the appearance of activity dips, display rapid thermal equilibrium during warmup, and exhibit higher fundamental frequencies and better aging. The DRAT crystals also have a much wider inflection point in temperature characteristics which reduce frequency to temperature changes. Figures 2, 3, 4, and 5 demonstrate this temperature-frequency relationship for one AT and three double-rotated cuts. For reference, Figure 6 shows the relationship of the zero angle cut for each.

The disadvantages of DRAT-cut crystals compared with AT cut are that they have a relatively large and steep negative frequency deviation at temperatures below 0 °C which limits DRAT-cut crystals to ovenized constructions, they are very frequency-sensitive to impressed dc voltages, and they require more complex circuitry.

The DRAT-cut crystals are more expensive due the greater manufacturing difficulties associated with maintaining two crystallographic angles simultaneously rather than a single angle.

For these reasons, the AT-cut crystal is by far the more popular cut. The DRAT cut is reserved for precision applications.

8.3.3 Support structure. The support structure can be a 2-or 3-ribbon structure. The resonator crystal should be bonded to the support structure by electrically conductive cement or by thermocompression bonding.

8.3.4 Holder. Holders or enclosures should be metal, metal-glass, or all glass constructions. The holders should be evacuated and cold-welded. If resistance welding is unavoidable, PIND testing should be performed. Do not back-fill the enclosure or use solder sealing methods to complete the seal. Figure 7 shows crystal unit holders and Figure 8 shows oscillators.

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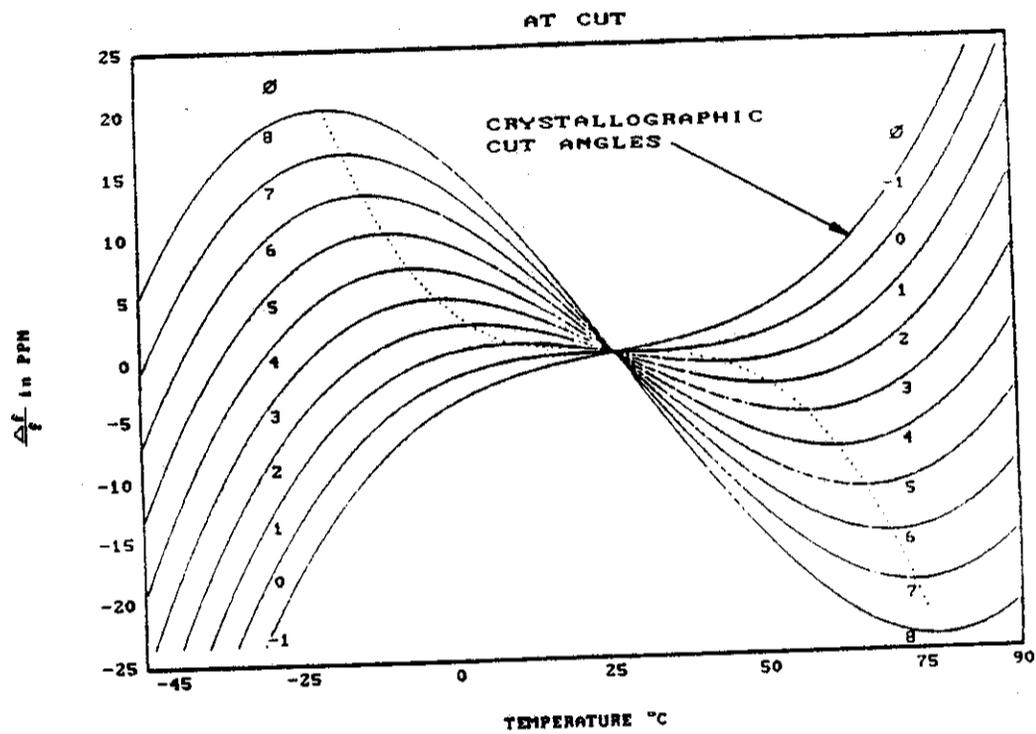


FIGURE 2. Generalized temperature/frequency characteristics for several AT crystals.

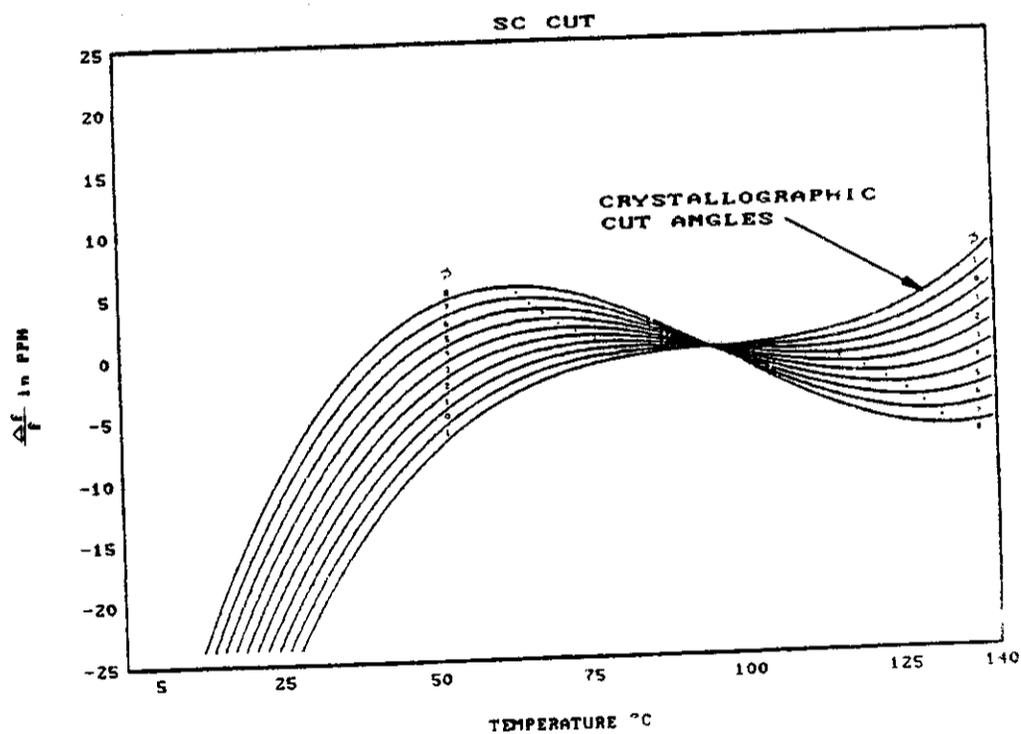


FIGURE 3. Generalized temperature/frequency characteristics for several double-rotated SC crystallographic cuts.

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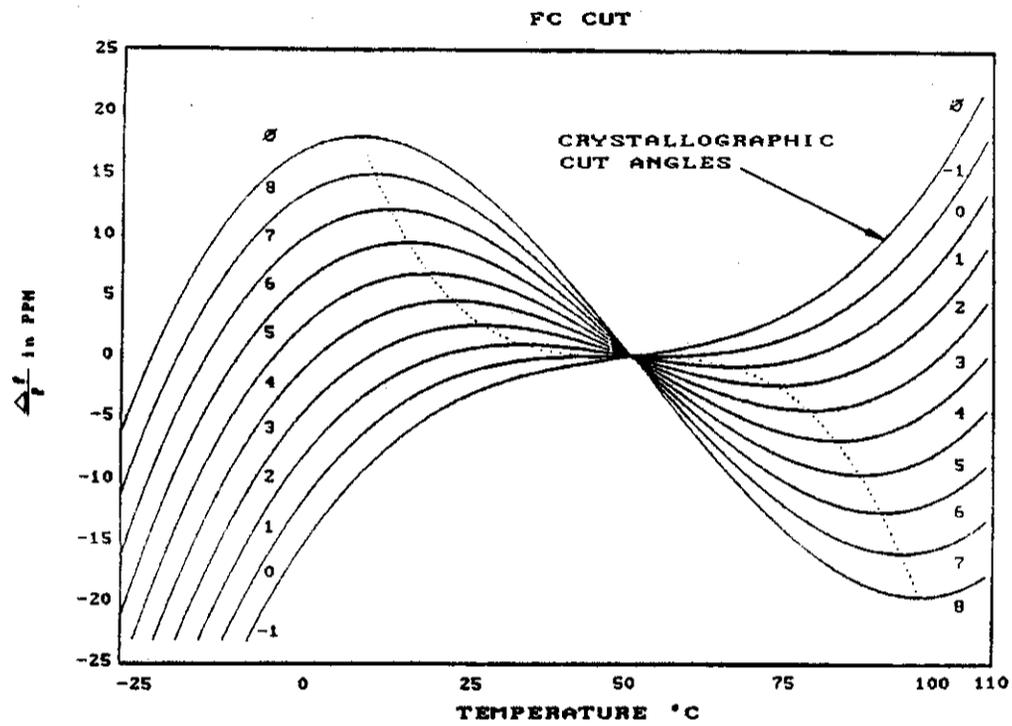


FIGURE 4. Generalized temperature/frequency characteristics for several AC double-rotated crystallographic cuts.

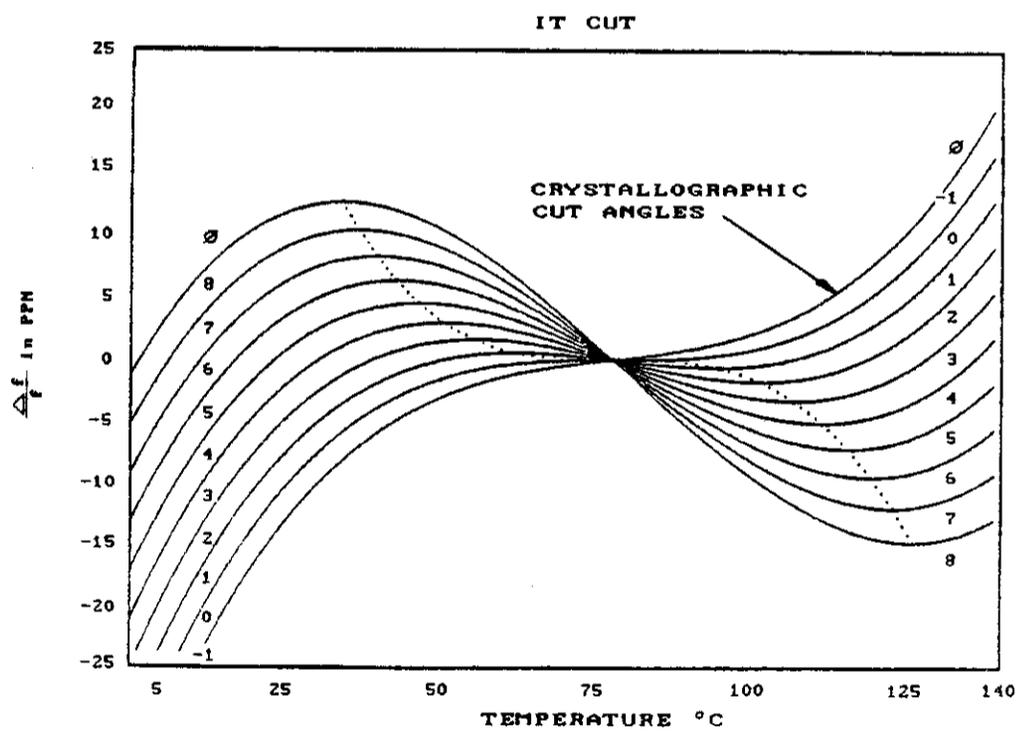


FIGURE 5. Generalized temperature/frequency characteristics for several IT double-rotated crystallographic cuts.

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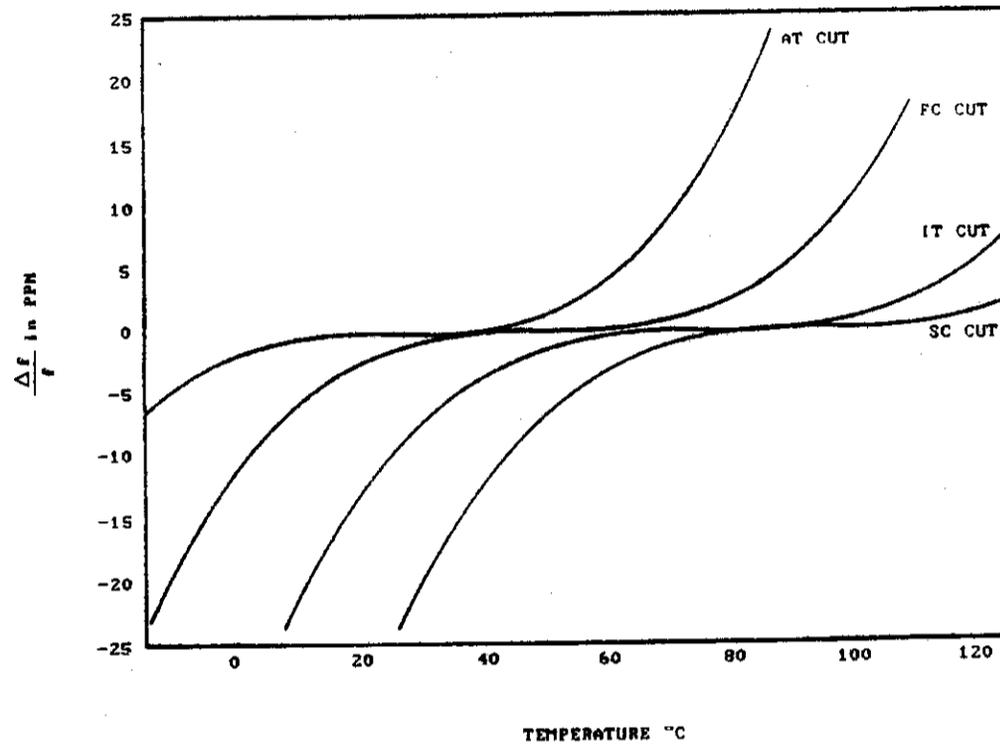


FIGURE 6. Temperature/frequency characteristics for zero angle crystallographic cuts.

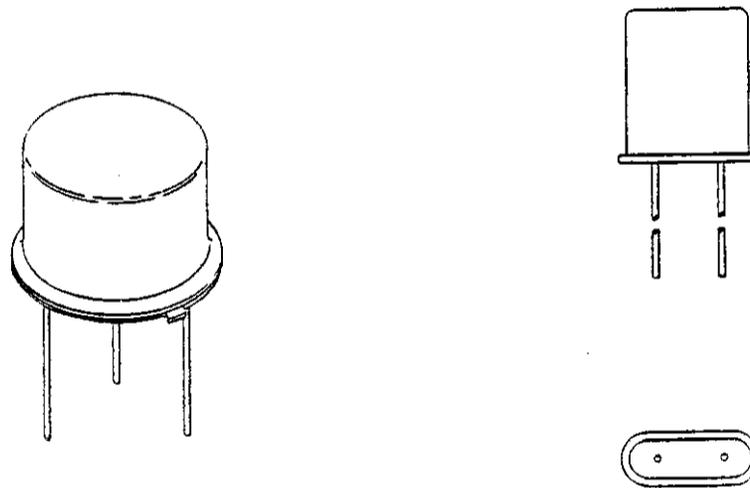
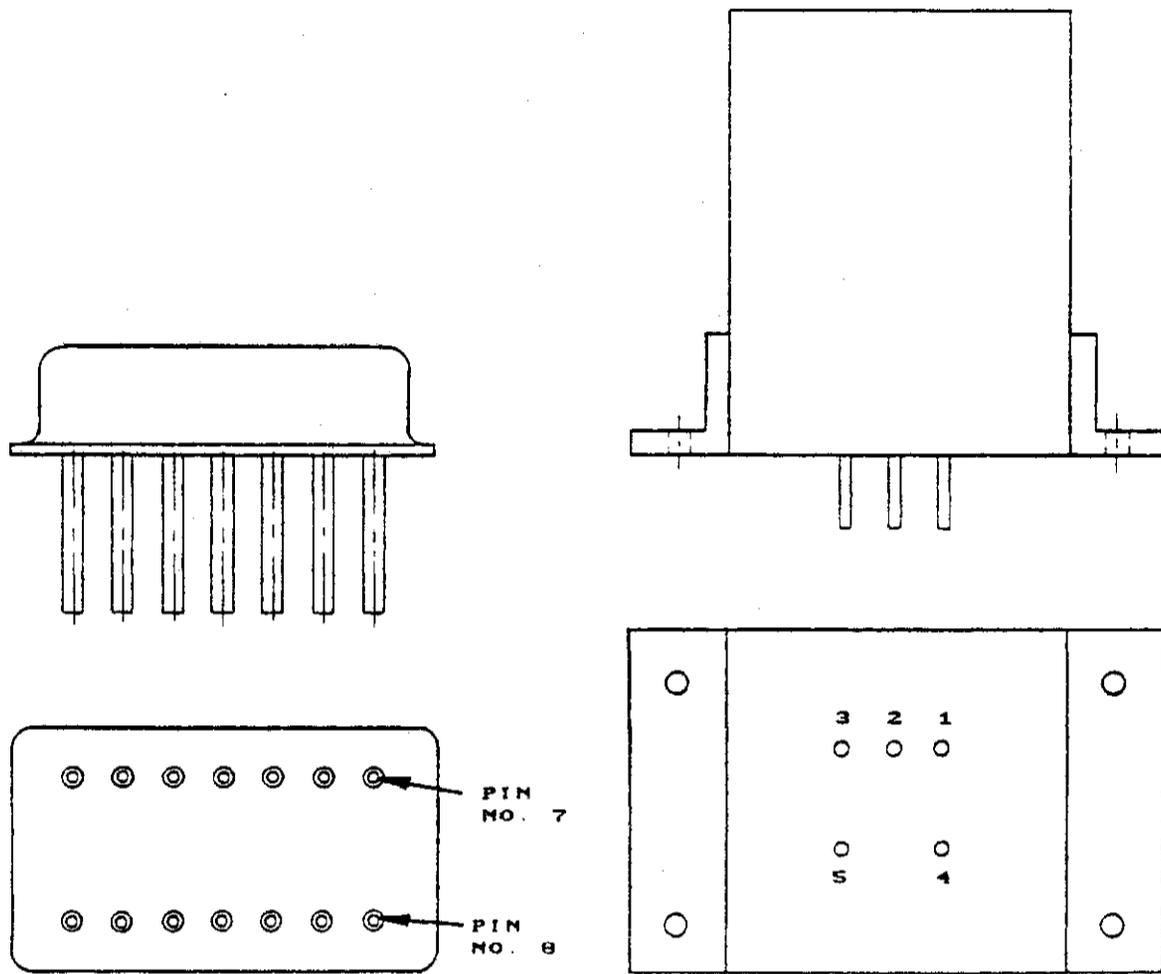


FIGURE 7. Outline drawings for typical crystal units.

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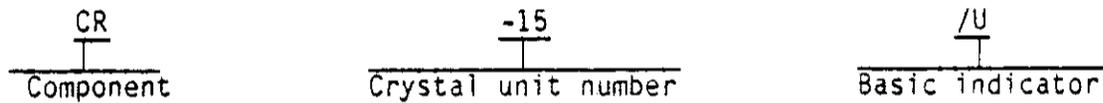


A. Oscillator (XO)

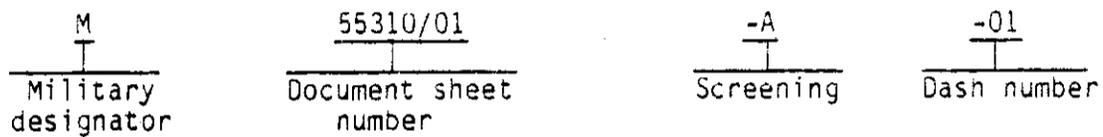
B. Oven controlled oscillator

FIGURE 8. Outline drawings for typical oscillators.

8.4 Military designation. The military specification for quartz crystal units is MIL-C-3098. The designation is:



The military specification for crystal oscillators is MIL-O-55310. The designation is:



No crystals are listed in MIL-STD-975.

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8.5 Electrical characteristics.

8.5.1 Equivalent electrical circuit of a quartz resonator. For electrical design considerations, the schematic representation of a crystal is shown in Figure 9 and the lumped constant equivalent circuit is shown in Figure 10.

As is characteristic of most mechanical resonators, the motional inductance (L_M) resulting from the mechanical mass in motion is generally relative to that obtained from coils and their magnetic fields. The L_M ranges from thousands of henries for the low frequency cuts to millihenries for frequencies over 100 MHz. The extreme stiffness of this glass-like crystalline material allows very small values of the motional capacitance (C_M), whereas the very high order of elasticity allows the motional resistance (R_M) to be relatively low. The L, C, R equivalent circuit, therefore, is one having a high L/C ratio, and a very high (L/C)/R, or Q ratio. Representative values in frequency regions of interest are shown in Table III.



FIGURE 9. Schematic diagram of crystal unit.

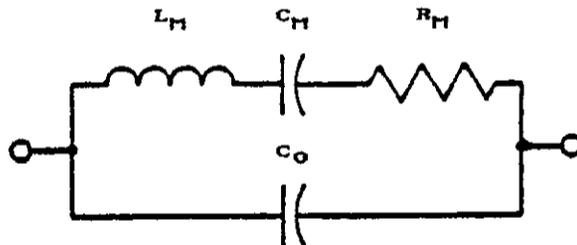


FIGURE 10. Lumped-constant equivalent circuit of crystal.

TABLE III. Representative crystal unit values

Frequency	Near 1 MHz	High frequency overtones
L_M	10 H	mH
C_M	0.0X pF	0.000X pF
Q	2.5×10^6	$<100 \times 10^3$

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The shunt capacitance (C_0) is the electrostatic capacitance existing between the crystal electrodes with the quartz plate as a dielectric. It is present whether or not the resonator is vibrating. The value of C_0 varies with the area and thickness of the quartz plate. Electrode size is part of the crystallographic design and is selected as a compromise value primarily for natural bandwidth (Δf), R_s , unwanted modes, or pulling linearity.

The values of C_0 are large with respect to the values of C_M . The ratio C_0/C_M varies from a low of 125, to a high of over 10,000. The ratio, C_0/C_M , sometimes referred to by the symbol "r," is a very important performance parameter for crystal resonator design and application considerations. If a dc voltage is applied across the terminals of the equivalent circuit shown in Figure 10, it can be seen that the reactive energy stored in the C_M is actually stored as strain or deformation of the resonator, whereas that stored in C_0 has nothing to do with mechanically activating the resonator. Thus, crystal resonator designs having higher C_0/C_M ratios exhibit higher equivalent electrical impedances.

8.5.2 Reactance curve of a quartz resonator. The other important effect of the C_0/C_M ratio has to do with the natural bandwidth (Δf) of the crystal unit. Figure 11 shows a curve of reactance versus frequency of a crystal unit.

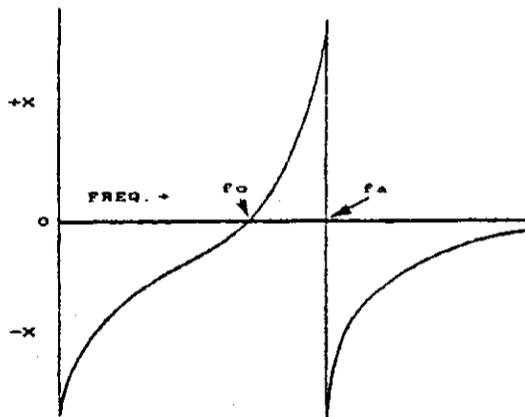


FIGURE 11. Reactance vs frequency of a crystal unit.

As can be seen, this curve is what would be expected if the reactance C_0 alone were plotted, except in the vicinity of mechanical resonance. Mechanical resonance occurs where the curve goes through zero at f_0 , and proceeds at a much higher rate to a very high positive value and then plunges rapidly downward through zero reactance to a very large negative value, returning to the expected C_0 curve only slightly higher than f_a . The frequency of zero reactance f_0 occurs very close to the frequency where X_{LM} is equal to X_{CM} and is given by the equation:

$$f \cong \frac{1}{2\pi\sqrt{L_M C_M}}$$

where f_0 is the resonant frequency of the crystal unit.

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The frequency f_a occurs when $X_{LM} - X_{CM} = X_{C_0}$ and is given by the equation:

$$f_a \approx \frac{1}{2\pi \left[\frac{(L_M C_M C_0)}{(C_0 + C_M)} \right]^{1/2}}$$

where f_a is the antiresonant frequency of the crystal unit.

The natural bandwidth Δf is $f_a - f_0$. If the equations above are substituted, Δf can be shown to be equal to:

$$\Delta f = \frac{f_0}{2 C_0/C_M} = \frac{f_0}{2r}$$

8.5.3 The crystal unit as an electronic resonator. An oscillator may be defined as an electronic device for converting dc into ac. Such devices are inherently closed loop systems and are composed of an amplifier, a resonator, and some means of limiting the amplitude of oscillation. A block diagram of such a system is shown in Figure 12. In this system, the amplifier provides its own limiting.

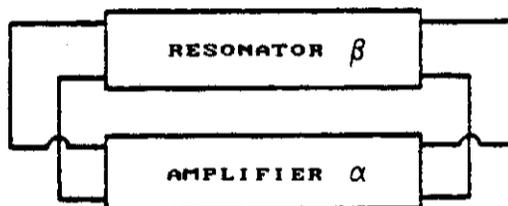


FIGURE 12. Block diagram of an oscillator.

The amplifier has a gain of α , and the resonator has a transmission coefficient of β . If the product of $\alpha\beta$ is greater than unity and has a zero phase angle at a particular frequency, oscillation will normally occur at this frequency. The amplitude of oscillation will increase at this frequency until limiting, as provided by the amplifier or an external means, satisfies the condition of $\alpha\beta = 1$, known as Barkhausen's condition for oscillation.

If a phase shift occurs in the amplifier α , there must be an equal and opposite shift in phase in the transmission characteristic of β .

This phase-shift correction by the resonator is accomplished by the system experiencing a change in frequency of oscillation. It is the purpose of a good resonator to provide phase correction with minimum frequency shifts. If the

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resonator is assumed to be working at or very near a frequency of resonance, and because we are concerned with only very small changes in phase so that the change in phase angle can be assumed to be equal to the change in the tangent of the phase angle, then:

$$\Delta\phi\alpha = -\Delta\phi\beta$$

$$\Delta\phi\beta = \frac{\Delta X}{R} = \frac{4\pi L\Delta f}{R}$$

and

$$\frac{\Delta f}{f} = \frac{\Delta\phi}{(4\pi L\Delta f/R)} = \frac{\Delta\phi/2}{2\pi\Delta fCR} = \frac{\Delta\phi}{2Q}$$

Where

L = effective series inductance

C = total effective series capacitance

R = total effective series resistance of the loop

The fractional change in frequency for a phase shift change of ϕ will be small if the effective Q of the resonator is large. The Q of typical quartz crystals is in the range of 10,000 to 100,000 and runs above 1,000,000 for special precision units. These high Q values, coupled with the high degree of performance of physical and electrical properties, make the quartz crystal outstanding as a resonator.

8.5.4 Series resonance. In the circuit shown in Figure 13, at the frequency of series resonance,

$$f = \frac{1}{2\pi\sqrt{L_M C_M}}$$

Because the magnitudes of X_{CM} and X_{LM} are equal, one cancels the other, resulting in a minimum impedance equal to R_M and essentially zero phase shift. In most series resonant applications below VHF, C_0 may be neglected.

8.5.5 Parallel resonance. At frequencies slightly higher than series resonance, X_{LM} will increase and X_{CM} will decrease, resulting in a net inductive reactance, X_L .

When this net $X_L = X_{C_0}$, then $f = \frac{1}{2\pi\sqrt{L_M (C_M C_0)/(C_M + C_0)}}$

At this point, maximum impedance will appear at the terminals, indicating parallel resonance of the crystal unit. The frequency separation between series and parallel resonance is determined by the capacitance ratio, C_0/C_M .

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In practical oscillators not operating at the series resonant frequency of the crystal, the circuit adds its input capacitance (C_x) to C_0 , resulting in a total C_T .

The formula for frequency then becomes $f = \frac{1}{2\pi\sqrt{\frac{(C_T C_M) L_M}{C_T + C_M}}}$

C_x is the oscillator input capacitance at the crystal terminals and its value must be known in order to adjust the crystal to the desired frequency, just as the capacitance must be known for an inductor to be preadjusted to resonate at a given frequency.

Parallel resonant crystals can be excited by a high impedance oscillator (Figure 14) or one with low impedance (Figure 15). It can be seen from Figure 14 that if C_x is zero, the crystal will be at its own parallel self-resonant frequency. As C_x is increased, the frequency is lowered, with the net inductance of the series arm resonating with C_T . As C_x increases, the voltage to the crystal decreases until infinite C_x results in no coupling to the crystal.

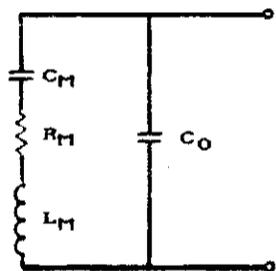


FIGURE 13. Simplified equivalent circuit.

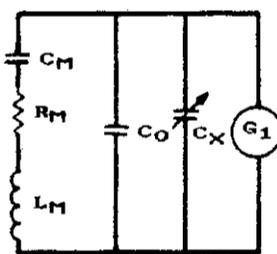


FIGURE 14. High Z generator.

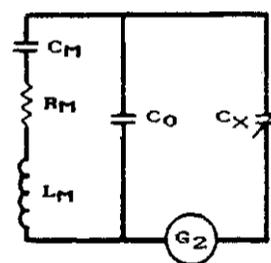


FIGURE 15. Low Z generator.

In the circuit of Figure 15, if C_x is zero, there is no coupling to the crystal. As C_x increases, C_T increases, gradually lowering the frequency of operation, just as in Figure 14. Thus a given capacitance, C_x , will result in essentially the same frequency in Figures 14 and 15 for a given crystal. If C_x reaches infinity in Figure 15, the crystal will be at its series resonant frequency.

Thus, it can be seen that the only difference between crystals designed for series resonance and those for parallel resonance is the oscillator input reactance for which they are calibrated. In fact, a crystal calibrated for parallel resonance will operate at its calibrated frequency in a series resonant circuit with the addition of an appropriate value of series capacitance.

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From the above, it is obvious that a crystal cannot be properly specified without stating the reactance with which it is to be calibrated.

8.5.6 Consideration of frequency stability as a function of the crystal resonator in an oscillator. The equation for the fractional frequency change of the resonator to offset incremental changes in phase angle in the associated circuitry has been previously discussed, and has been stated as

$$\frac{\Delta f}{f} = \frac{\Delta \phi}{2Q}$$

Oscillators are generally nonlinear devices, but will be assumed to be linear in operation for this discussion. Small changes in phase angle in the oscillator circuitry associated with the resonator may be due to many causes. One important universal cause, for almost all conceivable oscillator designs, are small changes in capacitance to ground on either side of the amplifier. The small change in capacitance across this resistance will result in a change in phase angle, $\Delta \phi$, which may be expressed as $R\pi\Delta C$. The fractional frequency change of the circuit will then be:

$$\frac{\Delta f}{f} = \frac{\Delta \phi}{2Q} = \frac{R\pi\Delta C}{2Q} = \frac{R2\pi f\Delta C}{2Q}$$

The conclusion so reached is important, because it states that as the frequency of operation increases, the Q of the resonator must increase proportionally, to obtain the same fractional frequency stability, when phase shift changes of this type are considered.

8.6 Environmental considerations. Quartz crystals are uniquely useful as an electronic component not only because of the high Q factors available but also because of the relatively small and generally predictable effects induced by exposure to environmental stress.

8.6.1 Frequency stability versus environmental changes. The reactance and resistance of the equivalent circuit of the crystal are changed in value when the crystal is subjected to the following:

- a. Change in temperature
- b. Change in excitation or crystal drive
- c. Change in mechanical stress
- d. Passage of time
- e. Change in the load reactance.

8.6.2 Frequency stability versus temperature. Frequency versus temperature characteristics are fairly well established for many crystal designs covering

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the frequency range of from approximately 1 KHz to over 200 MHz. The general behavior of the preferred type crystal cuts versus temperature were shown in Figures 2 through 6.

8.6.3 Frequency stability versus crystal excitation. The driving excitation current flowing through the crystal affects the frequency of operation. This is the result of temperature gradient strains due to localized heating of the quartz. Because the power dissipated in the crystal is proportional to the square of the current, the slope of the frequency change versus crystal current increases as the current is increased. It is highly advisable to operate the crystal at a low level of drive current and ensure that the oscillator will start and maintain a most constant crystal drive current, regardless of small changes in the voltage applied to the oscillator by some means of AGC action. Lightly driven (less than 10 μ W) precision AT-cut crystals will change less than a couple of parts per billion per decibel change in drive level. If the same crystals are driven at 5-mW levels in a manually adjusted gain test oscillator, it is practically impossible to repeat frequency measurements to 1×10^{-8} accuracy, even when a voltage regulator is used with the oscillator. Higher drive levels are useful when spectral purity of the oscillator output signal outweighs all other considerations for frequency stability, aging, and activity dips. At some point, drive level can cause incipient or catastrophic fractures.

8.6.4 Frequency stability versus shock and vibration. Environmental changes which mechanically induce strains in the quartz cause a change in the frequency of operation, and mechanically induced strains in the mount will have an additional effect upon the frequency.

The tip-over test measures the change in frequency resulting from a 2 G; i.e., the pull of gravity first from one side of the crystal then from the other. From this test, the fractional frequency change per G may be determined for various orientations of the crystal. The change in frequency resulting from a larger, but constant, stress may be obtained by subjecting the crystal to the centrifugal force of a centrifuge.

If the crystal resonator and its mounting system are rigid enough to remain free of mechanical resonance over the frequency range of the mechanical disturbance, then the frequency deviation per G of disturbance, as determined from the tip-over or centrifuge test, may be used to predict the expected frequency deviation for applied sinusoidal vibration up to the level where the the elastic limits of the design are exceeded.

Frequency changes of three general types are encountered in conducting vibration and shock tests. First, there are those which dynamically follow the mechanical excitation and disappear completely and instantly upon cessation of the excitation.

Second, during the course of the vibration run, or during initial subjection to shock, the reference frequency may undergo a temporary change from that observed before the test, but returns within a few moments to its original value. The

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phenomena will repeat and is probably associated with the plasticity and flow-back nature of the solder or cement used to attach the resonator to its support.

Finally, there are permanent changes that may result from the chipping of the resonator, permanent deformation or fracture of the mounting system, or damage to the attachment points of the mount to the resonator.

The stiffer and more rigid the mounting system, the higher the natural resonance of the system but the lower the capability to withstand high impact shock. If the design is to undergo no permanent damage from high mechanical shock, either the deceleration time must be short enough for the natural compliance of the system to prevent the system from exceeding its elastic limits, or shock buffers must be designed into the system to limit the excursion of the resonator to a displacement safely within the elastic limits of the regular mounting system. Designs incorporating shock buffers cannot be expected to provide frequency stability during shock, but if carefully designed, can provide minimal frequency change on a before-and-after exposure basis.

The AT-cut crystal, with its relatively inactive peripheral mounting points, provides opportunity for designs capable of withstanding much higher shock and vibration than designs with the resonator soldered to fine wire leads (low frequency cuts). The periphery of AT-cut crystals operating on overtone modes is even more quiescent than those operating on the fundamental mode, and such designs are generally preferred for maximum stability under shock and vibration.

8.6.5 Frequency change versus time (aging). In most applications, the crystal is relied upon to control or minimize the effects of changes in value with the passage of time on all other circuit elements of the frequency generating (or controlling) system. Therefore, it is most important that the crystal itself change as little as possible with time, and that whatever change that does occur is predictable.

There is still much to be learned concerning the basic cause and effect relationship associated with the aging process of crystal units. It is generally agreed that the principal cause of change in frequency with time of AT-cut crystals is due to a gradual transfer of mass to or from the crystal resonator. The material (solid, liquid, or gas in nature), which after final processing remains within the enclosure (where it can transfer to or from the resonator) is considered to be a contaminant. This contaminant may be set in motion by thermal agitation and the direction of transfer is from the hotter to the cooler surface.

Stress and a change of stress imparted to the crystal by its mounting structure is another major contributor to aging. In the application, the drive level must be minimized to limit its effect on aging. Storage and operating temperatures should be kept to their lowest applicable values to diminish the aging process.

Processing techniques are continually being improved to reduce contamination as a cause of aging. The behavior generally observed in precision-made crystals is an increase in frequency with an exponential reduction in rate with the passage of time as the free material gradually assumes a quiescent state.

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The major amount of change takes place in the first few weeks after manufacture and may be accelerated by high-temperature storage or temperature cycling, thus attaining the minimal and decreasing rate upon application.

The thermal transfer of free contaminant theory reasonably explains the observation that crystals once stabilized may not retrace to the same frequency when cooled off and reheated. This effect can be minimized if the heating pattern is consistent, a condition which is approached if the crystal is used in an oven with a highly repeatable warm-up characteristic. Further, the more care taken in the processing to minimize the possibility of free contaminants, the better the retrace behavior becomes.

The use of modern clean room processing techniques and ion pump high-vacuum systems, the continual improvement in measurement techniques, and the availability of the atomic standard reference frequency have made possible precision AT-cut crystal resonators that exhibit aging stabilities approach one part in 10^8 per year.

It is important to realize that even though the same careful control of contaminants can be applied in the processing of other types of crystals, there are other basic reasons why this high order of fractional stability is generally limited to the AT-cut that operates in the 1- to 5-MHz range. As previously pointed out, the effects of the mounting system must be negligible before the intrinsically stable properties of the quartz plate and its gold electrodes can be used to full advantage. This becomes increasingly difficult, if not impossible, at frequencies much lower than 1 MHz, regardless of the type of quartz cut used.

The frequency-determining thickness dimensions of an AT-cut crystal varies inversely with frequency; therefore, so does the fractional frequency stability for a given amount of free contaminant within the enclosure. It is possible to use higher orders of overtones to offset this situation, but even so, the best Q realizable will be inversely proportional to the frequency. Accordingly, it is generally recognized that the use of 1- to 5-MHz frequency precision crystals in oscillators, followed by well designed multipliers, will provide higher stability at the higher frequencies than that possible with the best frequency crystals.

8.6.6. Short Term Stability. If the output of an oscillator is frequency or phase modulated, the sideband energy frequencies will mix with frequencies contained in the noise or phase jitter of subsequent multipliers, thereby degrading the spectral purity of the output frequency. The degradation increases exponentially with the multiplication. For signal-to-noise ratios generally encountered with broad band frequency multipliers, the ratio of power in the side bands of the multiplied signal to that in the carrier of the multiplied signal will N^2 times the same power ratio of the signal before multiplication, where N represents the number of times the signal is multiplied.

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The noise, or sideband energy, associated with the signal from the crystal oscillator, is attributable primarily to the amplification required to bring the crystal resonator energy to a useable level. Operating the crystal resonator at very low-power levels with high gain in the following amplifier stages is most desirable for better long-term stability, but tends to degrade the spectral purity of the oscillator output signal.

The problem of separating the contribution of the measurement system from the inherent short-term stability of the oscillator is a difficult one. However, good precision reference oscillators in the 1- to 5-MHz range can be produced which exhibit a frequency stability in the region of 1×10^{-10} rms for 1-s periods and also maintain long-term stabilities in the order of a few parts in 10^8 per week.

8.6.7 Effect of over-specifying crystals. It is not uncommon when establishing performance requirements to make the operating temperature range wider than what actually will be encountered. This is to establish a safety margin. Because of the nonlinear nature of the frequency versus temperature characteristic of high-frequency crystals (a cubic curve in the case of AT-cut crystals), this practice can be counter-productive.

In the case of AT-cut crystals, the temperature characteristic is forced to become steeper in the region symmetrical near room temperature as a more radical crystallographic angle is chosen to accommodate specified requirements. This could necessitate the use of temperature compensation or temperature control to achieve the desired frequency stability.

In the case of DRAT-cut crystals, lowering the cold temperature end unnecessarily below 0° or -20°C may obviate the use of a DRAT crystal or, at least, force the addition of a heater for temperature control.

Thus, it is apparent that great care is warranted in establishing the frequency stability versus the operating temperature range requirement.

8.6.8 Effects of radiation. When AT- and DRAT-cut crystals are exposed to nuclear radiation, there is a measureable, often significant change in frequency and equivalent resistance (low-frequency cut crystals may cease to oscillate completely). Design criteria to minimize this effect include the use of swept quartz and gold electrodes. It may not be necessary to observe these precautions on all crystal-controlled devices such as wideband filters and low-stability oscillators.

When crystal-controlled oscillators are irradiated, there is a measureable (often significant) change in the output frequency whereas the output waveshape may not be affected. With the current oscillator designs and component technology, most of the output frequency change can be attributed to the crystal, whereas the contribution of the associated circuitry can be made negligible. An improper oscillator circuit design may result in burn-out.

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No single number can fully describe this behavior of a crystal or oscillator because it is a function of the following:

- a. The design of the specific device
- b. The type of radiation (such as CO^{60} , FXR, and N) and whether the exposures are in series or combination
- c. The pulse spectrum and duration of impingement
- d. The exposure history of the device
- e. The fact that some of the effects are permanent whereas some are transient
- f. Indirectly, the technique and timing of the very precise frequency measurements which must be made under the most difficult of circumstances.

Tests performed on the best of radiation-hardened crystal-controlled oscillators at the kilo-Rad and mega-Rad levels have indicated an average decrease in frequency ($\Delta f/f$) in the range of 10^{-13} to 10^{-15} per Rad. A conservative working estimate appears to be 10^{-12} per Rad. At low exposure levels of only a few Rad, the change may be even more severe. Positive frequency shifts in the 10^{-10} to 10^{-11} region preceding the negative shift have been observed.

As stated before and emphasized here, the actual performance obtained is a function of the specific device and the specific radiation conditions. Therefore, the values stated above are intended to convey only a general ideal of crystal behavior under nuclear radiation.

8.7 Reliability considerations. Quartz crystals, properly used, are among the most reliable of electronic components.

8.7.1 Failure modes and mechanisms. For all practical purposes, there is no wearout mechanism associated with the operation of a quartz crystal when it is operated within its specifications at or below rated drive level. As described in paragraph 8.6.5, there is a continuous aging process, the rate of which reduces exponentially with time. Operation above the rated drive level will result in deterioration of stability, even though the dissipation is well below the level at which the crystal may shatter. A high drive level will also cause degradation of the aging characteristics.

There are certain failure mechanisms associated with manufacturing defects which, if not detected prior to equipment installation, can result in either catastrophic failure or excessive drift after some period of operation. Some of the more common defects are listed below.

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- a. Poor seal
- b. Excessive contamination
- c. Excessive internal mechanical stresses introduced by the resonator support structure
- d. Poor electrical connections
- e. Poor plating (electrode) adherence
- f. Incipient fractures of the resonator
- g. Support structure fracture.

A poor seal will result in a gradual downward frequency drift and an increase in crystal resistance due to the entrance of atmosphere and moisture and other surface contaminants. The crystal may fail either by drifting out of tolerance or by an increase in resistance to the point that it may no longer oscillate.

The effect of excessive contamination is the same as that of a poor seal, except that the contaminants are present from the time of manufacture. This defect is relatively rare among quality manufacturers and the use of cold-welded cases reduces the problem of contaminants (such as solder flux) being introduced during the sealing operation. Occasionally, the design of the internal mounting structure and/or the assembly techniques used on a batch of crystals will result in the resonator being under mechanical stress in its mounted condition. Such stresses can also be introduced by high shock loads introduced by improper handling. This condition increases the effective resistance of the crystal and can cause it to be temperature sensitive, either failing to operate or operating out of frequency tolerance in the operating temperature range. Units with this type of defect, when it is sufficient to cause the crystal to exceed the frequency tolerance limits, will usually be out of tolerance on the low side.

Poor electrical connections may result in intermittent operation or total failure depending on the nature of the joint.

8.7.2 Screening. One hundred percent screening techniques are effective in eliminating most of the manufacturing defects described above. Recommended screening techniques are listed in Table IV in the sequence in which they should normally be conducted.

8.7.3 Reliability derating. No derating of quartz crystals is required. As emphasized previously, the drive level must not exceed the maximum rated level to preclude incipient damage and the crystal must be at the rated drive level for specified tolerances to apply.

8. CRYSTALSTABLE IV. Recommended screening techniques for crystals

Screening Test	Purpose
Thermal shock	To stabilize the crystal and to aggravate any imperfection in the seal or electrical connections
Reduced drive level	To check quartz resonator surface quality
Random vibration	To check for undesirable mechanical resonance of the mount and manufacturing defects
Gross and fine leak test	To ensure seal integrity
Aging	To stabilize performance
Performance over temperature range	To verify performance

8.7.4 Failure rate. Because quartz crystals are low-population parts, significant statistical data on failure rate is lacking. MIL-HDBK-217 lists the average failure rate as 0.02%/1000 hours. Pending further accumulation of data, a figure of this order of magnitude may be used for predictions.

MIL-HDBK-978-B (NASA)

9. FILTERS, ELECTRICAL

9.1 Introduction. Electromagnetic interference (EMI) control requires suppression of unwanted electrical noises, transients, harmonics, and other spurious signals which may prevent proper operation of either the equipment being designed or other electronic systems within the associated environment. Anticipation and disposition of noise is best accomplished as early in the design stage as possible, so that optimum techniques of lead dress, shielding, decoupling and mechanical design can be used to the maximum extent possible, thus minimizing problems associated with the addition of EMI filters as the design approaches completion.

An interference filter may be a single inductor or capacitor or it may combine inductive and capacitive elements in various combinations. If space and cost permit, almost any desired performance can be obtained. For the purposes of this section, discussion is limited to standard electromagnetic interference filter sections, with emphasis on standardization of types.

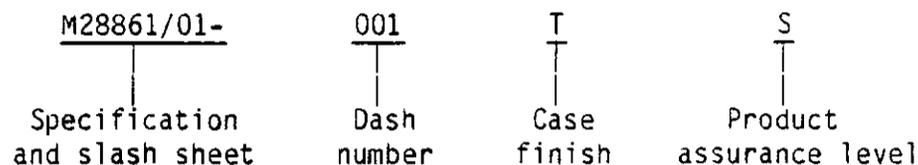
9.2 Usual applications. EMI filters used in dc lines are usually single capacitors or pi networks or sections and are usually used in 28 Vdc applications. EMI filters used in ac lines are usually L or T networks. The ac line filters are usually used in higher voltage applications of 120 Vrms. The current applications of both dc and ac line filters vary and devices are available in current ratings of up to 20 A. Both dc and ac filters use toroidal inductors and ceramic discoidal capacitors in their construction and are intended for low frequency applications.

Another type of ceramic filter uses a ferrite bead as the inductive element. These are also available in the three basic electrical configurations, but are effective only at higher frequencies (10 MHz and up). Here, the ferrite bead functions not as a true inductor, but as an absorber of rf energy.

9.3 Physical construction. EMI filters are made in many sizes, shapes, and combination of sections, depending on their electrical ratings and characteristics. Standard filters, however, tend toward single-section units of tubular and discoidal configuration with axial terminations and a threaded bushing mount. An example of typical physical construction is shown in Figure 1.

9.4 Military designation. The military specification for EMI filters is MIL-F-28861. This specification includes a number of individual specification sheets (slash sheets) for various styles.

The part number is in the following format:



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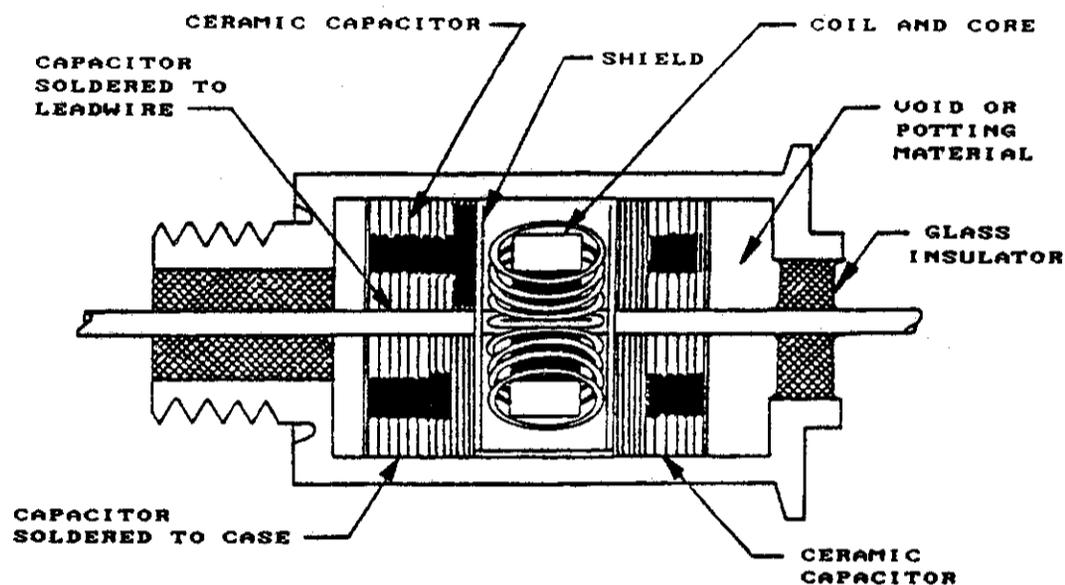


FIGURE 1. Typical construction of EMI filters.

9.5 Application and design considerations.

9.5.1 Electrical considerations.

9.5.1.1 Definition of requirements. In selecting a filter for a particular application, the following parameters should be taken into consideration if the filter is to be effective in its intended application.

- a. Insertion loss. The insertion loss is affected by load current, temperature, and voltage and is dependent on source and load impedance
- b. Voltage and current ratings
- c. Maximum voltage drop, if applicable
- d. Maximum and minimum inductance and capacitance
- e. Peak transient voltage and its duration
- f. Minimum filter life under rated conditions at maximum temperature
- g. Temperature range of operation.

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9.5.1.2 Selection of filter circuits. Although selection of an optimum filter to accomplish the required interference attenuation can be quite complex, the following general rules can be utilized for the selection of the filter parameters.

9.5.1.2.1 Bypass capacitor. A bypass capacitor can be used as a filter when both source and load impedances (Z_S and Z_L) are large with respect to capacitor reactance (X_C).

Under these conditions,

$$\text{Attenuation (db)} \approx 20 \log \frac{1}{X_C} \left[\frac{Z_S Z_L}{Z_S + Z_L} \right]$$

Thus, it can be seen that as $\frac{Z_S Z_L}{Z_S + Z_L}$ increases with respect to X_C , attenuation increases.

However, if either Z_S or Z_L approaches zero, the bypass capacitor is not an effective filter.

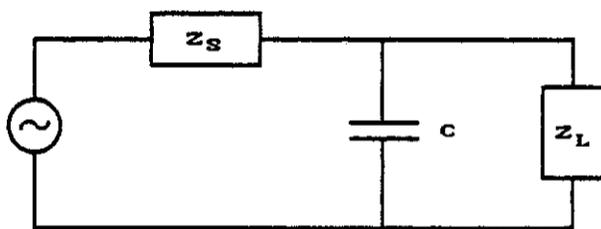


FIGURE 2. Bypass capacitor filter.

9.5.1.2.2 Inductor. A series inductor can be used as a filter when both Z_S and Z_L are low with respect to inductor reactance (X_L).

Under these conditions,

$$\text{Attenuation (db)} \approx 20 \log \left[\frac{\omega L}{Z_S + Z_L} \right]$$

Thus attenuation is significant when Z_S and Z_L are both small with respect to X_L .

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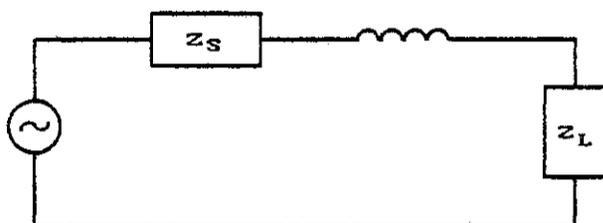


FIGURE 3. Inductor filter.

9.5.1.2.3 Inductive input "L" section. This circuit should be used whenever Z_S is much smaller than Z_L .

Under this condition, above cutoff,

$$\text{Attenuation} \approx 20 \log \left[\frac{\omega L}{Z_L} + LC\omega^2 \right]$$

and if $\omega L \approx Z_L$

$$\text{Attenuation} \approx 20 \log [LC\omega^2]$$

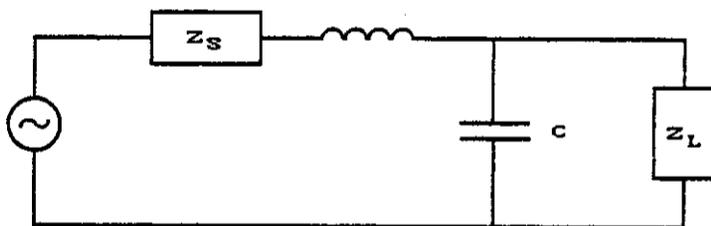


FIGURE 4. Inductive input "L" section filter.

9.5.1.2.4 Inductive output "L" section. This circuit should be used whenever Z_L is much smaller than Z_S .

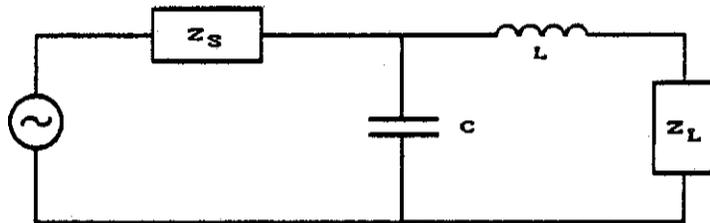
Under this condition, above cutoff,

$$\text{Attenuation} \approx 20 \log \left[\frac{\omega L}{Z_S} + LC\omega^2 \right]$$

and if $\omega L \approx Z_L$

$$\text{Attenuation} \approx 20 \log [LC\omega^2]$$

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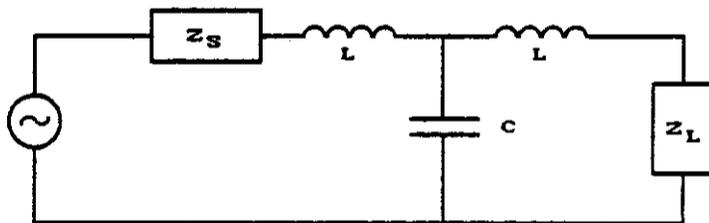
FIGURE 5. Inductive output "L" section filter.

9.5.1.2.5 "T" section. A "T" circuit should be used whenever Z_S and Z_L are small in value or are capacitive.

If Z_S and Z_L are both small:

$$\text{Attenuation} \approx 20 \log \left[\omega^2 LC + \frac{L^2 C \omega^2}{Z_S + Z_L} + \frac{2\omega L}{Z_S + Z_L} \right]$$

It can be seen that, as Z_S and Z_L approach zero, the last two terms become very large.

FIGURE 6. "T" section filter.

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9.5.1.2.6 Pi section. This circuit should be used whenever Z_S and Z_L are large in value or are inductive.

If Z_S and Z_L are both large,

$$\text{Attenuation} = 20 \log \left[\omega^2 LC + \frac{LC^2 \omega^3 Z_S Z_L}{Z_S + Z_L} + \frac{2\omega C Z_S Z_L}{Z_S + Z_L} \right]$$

It can be seen that, as Z_S and Z_L increases, the attenuation increases for Pi filters.

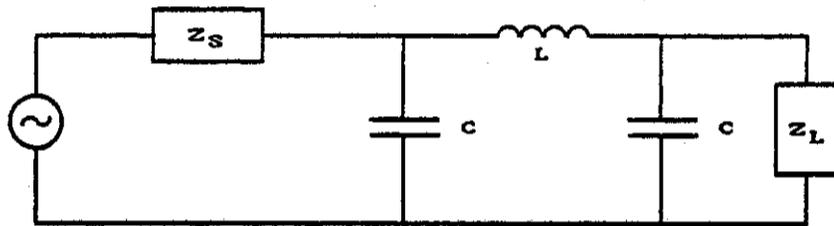


FIGURE 7. Pi section filter.

9.5.1.3 Selection of materials. Magnetic materials generally used for these filters are molypermalloy powder, iron powder and ferrite beads. Molypermalloy powder and iron powder cores are stable with temperature and load current, but ferrite beads become unstable when load current and temperature increase. Ceramic materials generally used for these filters are Z5U, X7R, and BX dielectrics. However, for space use, X7R/BX dielectrics should be considered with limitations on voltage stress and temperature coefficients. BX and X7R (lower range of dielectric constant) are more stable with higher voltages and temperatures. Like ferrites, Z5U becomes unstable when load voltage and temperature increases.

The following guidelines should be considered in the filter construction.

- a. Plating and marking should be resistant to solvents.
- b. The plating composition and solder composition should be compatible. The melting point of the solder should not be lowered after mixing with the plating material.
- c. Encapsulant used in the filter construction should have a low moisture absorption coefficient, good adhesion to metal, good thermal shock resistance, and a low expansion coefficient.

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9.5.2 Mechanical considerations. Once a filter has been selected for a particular application, its contribution to EMI control should not be degraded through faulty installation. Quite often, filter insertion loss characteristics are compromised by improper application. The following guidelines are suggested for proper application.

- a. Filters should be installed close to the problem areas they are to correct.
- b. The mounting area should be free of foreign particles and nonconducting finishes and surfaces to maintain a zero impedance current path through the metal-to-metal contact. This will provide good grounding and bonding, which will enable the design objectives of the filters to be achieved.
- c. Filters should not be held with pliers or other gripping tools. Pressure exerted on the filter case may crack the ceramic element.
- d. The specified mounting torques should not be exceeded.
- e. Isolation should be provided between the input and output terminals of a filter by the use of an rf-tight bulk head.

9.5.3 Thermal considerations. A filter's reliability may be degraded by the improper application of heat. The following guidelines are suggested.

- a. A heat sink should be used when attaching leads. Methods such as preheat, slow heating, and uniform heating should be used. The same precautions apply when cooling down the devices.
- b. Filter leads or terminals should not be heated for long periods (5 to 10 s maximum).
- c. Soldering iron tip temperatures should not exceed 450 °F.
- d. Filters should be allowed to cool to room temperature before cleaning.

9.5.4 Equipment EMI requirements. Equipment EMI requirements vary, therefore it is important that the filter parameters be adequately specified on the specification control drawing. Various combinations of inductance and capacitance, as supplied by various vendors, might meet the specified insertion loss requirements as measured per MIL-STD-220, but system EMI characteristics could vary drastically as measured with various combinations of different vendors' filters.

Probably the simplest method of insuring continuing compatibility of filter and system requirements is to specify the filter circuit, minimum inductance, capacitance values, and minimum insertion loss characteristics over the required frequency range.

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9.6 Environmental considerations. Filter construction should be such that it can meet the shock and vibration requirements of the intended application. Materials used in the construction should be capable of withstanding thermal shock and be capable of resisting solvents and soldering heat. The materials should also be capable of meeting the outgassing and radiation requirements of space application.

9.7 Reliability considerations.

9.7.1 Failure modes and mechanisms. The possible failure modes of interference filters are determined by the number and types of parts making up the filter. The most common failure modes are:

- a. Shorts
- b. Opens
- c. Low insertion loss.

Shorts. Catastrophic failure by shorting of a capacitor is probably the most common failure mode, and the frequency of occurrence is a function of the quality of the capacitors used and the stress to which they are subjected. Many EMI filters are exposed directly to the hazards of power line variations, and the combined effects of ac heating of the capacitors and power line transients can result in early life failure of marginal designs.

Opens. Opens are usually the result of poor solder joints. In addition to marginal joints which are sometimes made by the filter manufacturer, careless assembly techniques by the user can result in remelting of the internal joints and subsequent opens during equipment operation. This problem is particularly important with power line filters, because of the heavy lead wires which are often required for such applications. To minimize this problem, the user should specify that high temperature solder be used by the manufacturer on terminations which will be subjected to solder heat during subsequent user assembly operation.

Low insertion loss. Low insertion loss can be caused by an open capacitor, a shorted inductor, poor solder joint, improper grounding of internal shielding, or a bad capacitor. Insertion loss may also be low at full rated current because of saturation of the inductor, or may be low at temperature extremes due to capacitance temperature coefficient. These latter two faults, however, are design deficiencies rather than true failure modes.

9.7.2 Screening. Screening techniques for EMI filters are essentially the same as those normally applied to the type of capacitor used in the construction of the filter. Thermal shock cycling and operating burn-in have been found to be effective in stressing the filters. For effective screening, limits should be imposed on capacitance change and dissipation factor (at some specified high frequency), after the thermal shock and burn-in tests. Ideally, the burn-in

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would be conducted at full rated current and full rated voltage, but application of rated current is usually only marginally effective from a cost standpoint. Unless the I^2R losses in the inductors are high, which would be poor design practice, simple voltage burn-in will screen most defects. Most failures in normal operation are the result of capacitor breakdowns. Inductor failures are usually the result of external overloads, rather than degradation of the inductor itself. However, for applications where high reliability is the prime consideration, application of rated current during burn-in is recommended.

9.7.3 Reliability derating. Interference filters should be voltage derated in the same manner as is recommended for the capacitors used. Refer to derating criteria in MIL-STD-975. This is particularly important on filters which are directly across system power lines. Such devices must be capable of withstanding all voltage variations and transient conditions encountered in its application. Derating under normal conditions will insure that transient peaks do not exceed the voltage rating of the capacitors used.

Severe current derating is normally not required, provided that the device is designed such that its internal temperature rise is low. Filters designed to meet the requirements of MIL-F-28861 have an allowable temperature rise of 25 °C maximum, as measured at the case hot spot.

9.7.4 Failure rate. The failure rate of an interference filter is usually taken as the sum of the failure rates of its internal components. For prediction purposes, the latest available data for the individual parts, such as is available in MIL-STD-217, should be consulted.

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10. TRANSFORMERS AND INDUCTORS

10.1 General.

10.1.1 Introduction. This section contains information on transformers, inductors, and radio frequency coils. MIL-STD-975 (NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List) specifies inductor coils but contains no specific transformers. Instead, it specifies that all transformers be procured to the requirements of MIL-STD-981 (Design, Manufacturing and Quality Standards for Custom Electromagnetic Devices for Space Applications). This section will describe transformers and inductors in terms of the appropriate military specifications and MIL-STD-981.

In its most elementary form, a transformer consists of two inductively coupled wire wound coils. When alternating current at a given frequency flows in either coil an alternating voltage of the same frequency is induced in the other coil. The value of this voltage depends on the degree of coupling and the flux linkages in the two coils. The coil connected to a source of alternating voltage is usually called the primary coil and the voltage across this coil is the primary voltage. Voltage induced in the secondary coil may be greater or less than the primary voltage, depending on the ratio of the primary to secondary turns. Accordingly, a transformer is termed a step-up or a step-down transformer.

Most transformers have stationary iron cores around which the primary and secondary coils are wound. Because of the high permeability of iron, most of the flux is confined to the core, and a greater degree of coupling between the coils is thereby obtained. So tight is the coupling between the coils in some transformers that the primary and secondary voltages bear almost exactly the same ratio to each other as the turns in the respective coils or windings. Thus, the turns ratio of a transformer is a common index of its function in raising or lowering voltage.

Inductors are used in electronic equipment to smooth out ripple voltage in direct current (dc) supplies so they carry dc in the coils. It is common practice to build such inductors with air gaps in the core to prevent dc saturation. The air gap, size of the core, and number of turns depend upon three interrelated factors: inductance desired, current in the winding, and ac voltage across the winding. The number of turns and the air gap determine the dc flux density, whereas the number of turns, applied voltage, applied frequency, and core size determine the ac flux density. If the sum of the two flux densities exceeds the saturation value, then noise, low inductance, and nonlinearity result.

Radio frequency transformers and coils without iron cores or with small slugs of powdered iron are commonly used in electronic circuits. In air core transformers all the current is exciting current and induces a secondary voltage proportional to the mutual inductance. Radio frequency coils are used to pass dc and present high impedance to ac signals of radio frequency.

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**10.1 TRANSFORMERS AND INDUCTORS,
GENERAL**

10.1.1.1 Applicable military specifications. Applicable military specifications are listed below:

<u>Mil Specs</u>	<u>Title</u>
MIL-T-27	Transformer and Inductors (Audio, Power, and High-Power Pulse)
MIL-C-39010	Coils, Fixed, Radio Frequency, Molded, Established Reliability
MIL-T-21038	Transformers, Pulse, Low Power
MIL-C-83446	Coils, Radio Frequency, Chip, Fixed and Variable

10.1.2 General definitions.

Air gap. A nonmagnetic discontinuity in a ferromagnetic circuit. The term is used even when the space is filled with such nonmagnetic materials as wood or brass.

Audio frequency. Any frequency audible to the normal human ear. Audio frequencies range from 15 to 20,000 Hz.

Choke. A coil which conducts dc but impedes the flow of ac due to its inductance.

Coil. A conductor wound in helical or spiral shape to form an inductor. Also, a number of turns of wire wrapped around a rod or tube of either ferromagnetic or insulating material. Coils offer considerable opposition to the passage of ac but very little opposition to dc.

Core. A magnetizable portion of a device or component such as an inductor or transformer.

Corona. An electrical discharge into the atmosphere from a high voltage circuit. It can usually be heard as a hiss and seen in the dark as a purplish light near sharp points carrying high voltage.

Degauss. Demagnetize.

Differential transformer. A transformer that connects two or more signal sources to a single transmission line, keeping them isolated from each other and from the line, and with an output proportional to the difference between the signals.

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10.1 TRANSFORMERS AND INDUCTORS,
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Efficiency. Ratio of the useful power output obtainable from a device to the total input power required by the device.

Ferromagnetic. Capable of being highly magnetized.

Harmonic distortion. The distortion of the shape of the sine wave of the fundamental frequency due to the addition of harmonics of the fundamental frequency.

Hotspot temperature. The highest temperature within the coil. It is important because it is necessary to restrict the temperature of the insulation to a safe operating value.

Inductor. A coil inserted in a circuit to supply inductance. It may have a magnetic core, nonmagnetic core, or air core.

Insertion loss. At a given frequency, the insertion loss of an element connected into a system is defined as the ratio of voltage appearing across the line immediately beyond the point of insertion, before and after insertion.

Leakage. Electrical loss due to imperfect insulation. It is the undesired flow of electricity through or over an insulator.

Magnet wire. Insulated copper wire used in the winding of coils. Usually it is solid and enameled and/or covered with cotton.

Permeability. A measurement of the ability of a material to conduct the lines of force of a magnetic field. The permeability of air is considered to be unity and the permeability of other materials is measured in relation to air.

Primary. The current-carrying transformer winding which, through induction, generates a current in the secondary winding.

Q. The ratio of energy stored to energy dissipated in a device. In an inductor, the ratio of reactance to effective series resistance at a given frequency. A measure of frequency selectivity or the sharpness of resonance.

Radio frequency. Any frequency from 10 kHz to 10,000 MHz.

Resonance. The resonant frequency of a circuit is the frequency at which the inductive reactance is equal to the capacitive reactance.

Ripple. The ac component in the output of a dc power supply.

Saturation. The point in magnetic field beyond which the magnetic material cannot be further magnetized.

Secondary. The transformer winding in which voltage is electromagnetically induced by the primary.

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Shield. A conducting material that provides an electromagnetic barrier between portions of circuits.

Steady state. Pertains to the condition of a circuit or component in which all values remain constant after all transient effects have disappeared.

Toroid coil. A coil of wire wound around a doughnut-shaped core.

Transformer. An electromagnetic device used to increase or decrease ac voltage. If the voltage is increased the current is decreased and vice versa, so that the power is unchanged, except for the losses in the transformer.

Waveform. The shape of an electromagnetic wave or its graphic representation showing the variations in amplitude with time.

Winding. A continuous conducting path, usually wire, generally formed into an electromagnetic coil.

10.1.3 NASA standard parts. NASA standard parts are as listed in MIL-STD-975, section 6 for inductors, and section 12 for transformers. The transformer section contains no specific parts but requires that procurement of transformers be made to the requirements of MIL-STD-981.

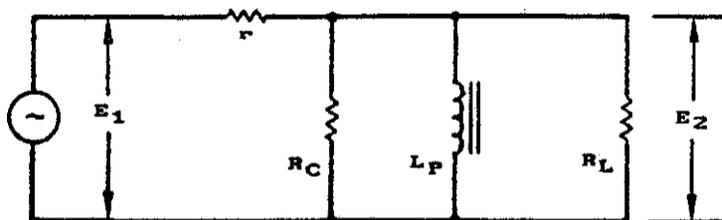
10.1.3.1 MIL-STD-981. MIL-STD-981 is the design manufacturing and quality standard for custom electromagnetic devices for space applications. All transformers, and MIL-C-83446 chip inductors if used in Grade 1 applications, shall be procured to the requirements of MIL-STD-981. MIL-STD-981 establishes the requirements for acceptable design, manufacturing, and quality criteria for custom electro-magnetic devices for space applications. It includes provisions for Class S parts intended for critical flight and mission-essential ground support applications as well as for Class B parts for noncritical applications.

10.1.4 General device characteristics.

10.1.4.1 Audio, power, and high-power pulse transformers and inductors. These types of components include transformers and inductors weighing 300 pounds or less, having rms test voltage ratings of 50,000 V or less, or high-power pulse transformers where the peak pulse power is greater than 300 W and the average pulse power is greater than 5 W.

10.1.4.2 Low-power and pulse transformers. Low-power pulse transformers have peak pulse power of 300 W or less and an average pulse power of 5 W or less.

10.1.4.3 Low-frequency transformers. Transformers used in power supplies generally operate at a single frequency that is usually in the lower frequency range of 25 Hz to 400 Hz. In low-frequency applications, the leakage impedance due to L_p is small with respect to the winding resistance (r) and the effect of the capacitive shunt impedance can be neglected when compared with the primary inductance. The equivalent circuit for the low-frequency case can be simplified as shown in Figure 1.

10.1 TRANSFORMERS AND INDUCTORS,
GENERALFIGURE 1. Low-frequency transformer equivalent circuit.

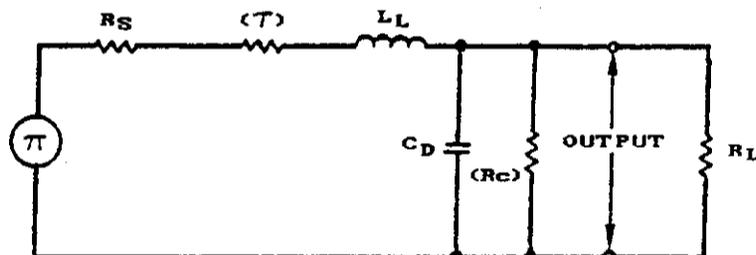
The primary current consists of the load current E_2/R_L and an excitation current. The excitation current is actually two out-of-phase currents, a resistive current due to the core losses and a magnetizing current controlled by the shunt inductance. In case of a resistive load, the voltage drop across the series resistor (r) subtracts almost directly from the primary voltage E_1 (the excitation current is generally small compared with the load current). At low frequencies, the phase angle between input and output voltage is mainly affected by the resistive losses (r) and primary inductance (L_p).

$$\tan \beta = \frac{r}{2\pi f L_p}$$

10.1.4.4 High-frequency transformers. Transformers used in pulse applications generally operate at much higher frequencies than power transformers. In order to transmit the applied pulse shape with as much fidelity as possible, the transformers must have a relatively wide frequency band response. The low frequency response of the transformer affects the pulse width, whereas the upper frequency response affects the pulse rise and fall time. An acceptable method of estimating the rise time from the upper cutoff frequency (f_{c2} - 3 db point) of the transformer is:

$$t_r = \frac{0.35}{f_{c2}}$$

The equivalent circuit of a high-frequency transformer is shown in Figure 2.



R_S - SOURCE (GENERATOR) IMPEDANCE,
 $T \ll R_S, R_C \gg R_L$

FIGURE 2. High-frequency transformer equivalent circuit.

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10.1.4.5 Coils. The rf coil is used to pass direct current and to present a high impedance to radio frequency currents. Applications of rf coils include discriminator circuits, oscillator and frequency multiplier tank circuits, and filters.

10.1.4.6 Saturable reactors. Transformers used as saturable reactors in magnetic amplifier circuits generally use cores with square B-H loop characteristics. The magnetic characteristics of the pair of cores, which are usually of toroidal shape, are matched in sets. Each core is individually wound with one or more gate windings. The coils are then stacked together and control windings are wound over both cores. There are many variations of windings and core shapes--single cores, E cores with gate windings on the outside legs, and control windings on the center leg, or other core types and shapes--but the principle of operation is the same. The rule of equal magnetomotive forces requires that the primary and secondary ampere-turn products be equal. The varying ac impedance of the gate winding controls the amount of signal that the load is receiving. The dc current through the control windings saturates or desaturates the core to various degrees and thus controls the impedance of the gate windings.

10.1.5 General parameter information. The following general terms and equations are helpful to understand and evaluate general characteristics of transformers and inductors.

10.1.5.1 DC resistance. Coil dc resistance tolerances vary greatly with the wire size used for winding. On fine wire coils, it is impractical to maintain tight dc resistance tolerances. When specifying Q and dc resistance of the inductor, the values must be compatible or the dc resistance specification should be omitted.

10.1.5.2 Inductance and related parameters. When the flow of electric current through a coil is varied, the resulting change in the magnetic field surrounding the coil causes a voltage to be induced in the coil, which opposes the supply voltage. This results in the coil having "self inductance" or simply "inductance." Inductance can be defined as that property of an electric circuit which opposes any changes in the current flowing in the circuit.

Inductance (L) represents a factor by which the rate of change of current is multiplied to obtain the induced electromotive force (EMF):

$$e = (-L) (di/dt)$$

The constant is called the coefficient of self-induction and is measured in henries (H). One henry equals an induced EMF of 1 V for 1 A per second rate of change of current. The minus sign indicates that the self-induced voltage is opposite in polarity to the supply voltage.

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The amount of inductance is determined by the amount of flux linking a given coil, which depends on the number, size, and arrangement of the turns forming the coil and the presence or absence of magnetic substances in the core of the coil.

Flux linkages represent energy stored in the form of magnetic flux. The amount of such energy depends on the inductance and current.

$$W = LI^2/2$$

Thus, an inductor can be also considered an energy storing device. The term LI^2 establishes the size of the inductor.

The flux (Φ) is defined as the total number of lines of force and is measured in lines or maxwells (Mx). Flux density (B), the number of lines per unit area, is measured in gauss and represents the number of maxwells per square centimeter. The flux density is proportional to the magnetizing force (H), the proportionality factor being the permeability (μ) of the medium, or:

$$B = \mu H$$

The magnetizing force (H) is a measure of the work required to move a unit pole one cm against the field and is measured in oersteds (Oe). The work required to move the unit pole around the total magnetic path is defined as the magnetomotive force and is expressed in Gilberts (Gb). The magnetomotive force is proportional to the product of amperes and turns and does not require that the turns be distributed evenly over the entire magnetic path.

In order to have an inductance, a core of magnetic material is not essential. It is often omitted at high frequencies. At medium and low frequencies, however, a magnetic core is essential for all but the lowest values of inductance.

The ratio of the number of lines in a given medium to the number of lines which the same magnetizing force would produce in air is termed the "permeability" of the medium. In an iron core, the flux density (B) is not a linear function of the magnetic intensity (H). Therefore, the permeability (μ), representing the slope of the B-H curve, is not a constant. Furthermore, the permeability also depends on the "past history" of the iron core, a phenomenon known as core "hysteresis".

Permeability can be further complicated by direct current flowing in the coil. In this case, the incremental permeability is of prime importance in establishing the inductance, i.e., the permeability of magnetic material to alternating currents superimposed on direct current. This is defined as the permeability of the material to small increments of alternating magnetomotive force. Permeability in the concept of flux density and field intensity is analogous to permittivity of dielectric substances in electric fields. While permittivity of dielectrics is usually independent of the magnitude of the electric field intensity, permeabilities of ferromagnetic substances are critically dependent on magnetic field conditions.

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An inductor usually has ohmic losses which can be represented as a resistance (R) in series with the inductance. When a voltage (V) is applied across that inductor, the current rises gradually to its steady value (V/R), following the logarithmic curve:

$$I_d = (V/R) (1 - e^{-t/T}), \text{ where } t = L/R$$

The term L/R, called the time constant, represents the time in seconds required for the current to reach 63.2 percent of its final value.

The decay of current in an inductor will also follow a logarithmic curve given by the formula:

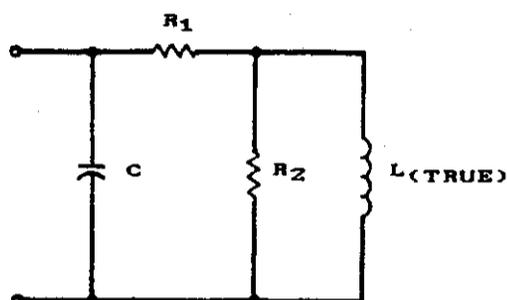
$$I_d = (V/R) (e^{-t/T})$$

The L/R ratio for a given inductor within the winding space factor variation is a constant. For example, doubling the number of winding turns increases the inductance by a factor of 4. However, to fit in the core window space, twice the turns requires half the wire size, which results in a resistance increase by a factor of four.

Inductors are widely used in electronic and electrical equipment. Such coils are generally designed for a specific application whenever the end use is known. The performance of a coil, besides being affected by its shape, size, and core material, can be drastically influenced by the mode of operation. In order to establish the performance of an inductor under actual field conditions, measurements should simulate, as closely as possible, the operating conditions of that coil.

10.1.5.3 Quality factor (Q). In addition to the inductance of a coil, the most important parameter describing the efficiency or quality of the inductor is the quality factor. When alternating current flows through a coil, energy is stored in the coil during a portion of the cycle. Most of the stored energy is fed back into the circuit later in the cycle. The difference between the stored energy and the returned energy is the energy dissipated in the coil. A perfect coil with no losses would return all the energy stored. The Q is the ratio of stored to dissipated energy per cycle. Q can be expressed as the quotient of the inductive reactance of the coil and the resistive losses. For a series representation, $Q = 2\pi fL/R$ (series) while for parallel, $Q = R$ (parallel)/ $2\pi fL$.

10.1.5.4 Self-resonant frequency. When an inductor is placed in a circuit or across the terminals of a bridge, it represents a complex network which includes its inductance, resistance, and capacitance. If we neglect flux leakage between turns of the winding, which in most cases is insignificant, we can simplify the coil equivalent circuit to that shown in Figure 3.

10.1 TRANSFORMERS AND INDUCTORS,
GENERALFIGURE 3. Equivalent circuit of a coil.

R_1 represents the copper winding resistance and is generally independent of frequency. Only at extremely high frequencies would R_1 appear to rise in value due to the "skin effect." R_2 , representing core losses, is a combination of three frequency-dependent losses: eddy current, hysteresis, and residual losses. These losses increase with frequency and flux level. Capacitor C represents the total shunt capacitance effect of the "between turns" capacitance and the capacitance from each turn to the core or ground. This capacitance forms a parallel circuit with inductance L , which resonates at the self-resonant frequency of the coil.

Due to this self-resonant effect, only those inductance measurements made far below the self-resonance of the coil will give the true inductance value of the coil. As the bridge frequency approaches the self-resonant frequency of the inductor, the apparent inductance measured differs (increases) from the true inductance according to the relationship

$$L \text{ (Apparent)} = L \text{ (True)} / (1 - (f/f_0)^2)$$

where f is the bridge frequency and f_0 is the frequency of the coil self-resonance.

The self-resonant frequency affects the apparent inductance of the coil and adds dielectric losses to it. In addition, it limits the useful frequency range of the coil.

10.1.5.5 Distributed capacitance. The net effective distributed capacitance (CD) of the coil has a direct effect on the self-resonant frequency. This capacitance can be measured directly on some commercial bridges, such as the Boonton "Q" meter. In general, the measuring frequency should be selected far above the self-resonant frequency of the coil.

10.1.5.6 Temperature rise. One important parameter often overlooked in high-current-carrying inductors is the temperature rise of the unit. This is directly related to winding resistance, core losses, and type and size of coil enclosure. Ambient operating temperature plus the self-generated temperature rise determine the operating temperature, life expectancy, and reliability of the inductor.

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10.1.6 General guides and charts.

10.1.6.1 Considerations when specifying inductors. Most transformers and inductors are designed for specific applications. When ordering these devices and using the military specifications as a baseline, the following information should be specified:

- a. Inductance value, tolerance, applied voltage and frequency and dc load current at which it is measured
- b. DC resistance and tolerance
- c. Q value, tolerance, applied voltage and frequency and dc load current at which it is measured
- d. Minimum coil self-resonant frequency
- e. Permissible changes in coil parameters over the frequency range, voltage and current ranges, and temperature range
- f. Dielectric strength and insulation resistance requirements
- g. Static and magnetic shielding requirements
- h. Description of inductor application and the associated circuitry
- i. Environmental conditions: shock and vibration and maximum altitude
- j. Permissible size and shape of inductor package: type of enclosure, (metal case, molded unit), type of termination, mounting method, outside coating (paint) and marking, and the total permissible weight.

10.1.6.2 Trade-off considerations. An existing transformer is frequently required for an application where some conditions are changed. Often the transformer can be used as is. However, at times the new application may require a modification of the design, even to the extent of changing the size and shape.

As the parameters and conditions subject to change are numerous, it is only possible to give some general guidelines for trade-off considerations. The guidelines and formulas referenced in this section still apply, however.

A power transformer, whose input voltage level is lowered, will have a correspondingly lower output power, but will generally not require a change in transformer design, provided the operating frequency is not changed and a slight improvement in regulation (increase of loaded output voltage) is acceptable. However, increasing the power output of a transformer will generally require a corresponding increase in the transformer size and weight. Note that transformer core sizes come in distinct steps. This may cause a slightly larger size increase than the power level would indicate.

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Similarly, a reduction in the operating frequency may increase the transformer size. This may occur if the number of turns in the winding or the core cross-sectional area is appreciably increased in order to keep the flux level from going into saturation. An increase of the inductance requirement of winding will generally require an increase in the winding turns, although in some cases a change of core material or the core size can also increase the inductance.

Any dc resistance changes require a change of wire size, number of turns, or the mean turn length (winding location or core size). Interwinding capacitance and leakage inductance are to a great extent dependent on the physical winding configuration (spacing), number of turns, winding traverse and insulation materials. Increasing the spacing between windings reduces the capacitance but increases the leakage inductance between the windings. There are various methods of winding construction to minimize the leakage inductance and the dynamic capacitance.

For a pulse transformer to maintain a flat pulse with a minimum droop requires a given winding pulse inductance (large with respect to the winding and the source resistance). Increasing the pulse width results in a larger pulse droop. The droop can also increase when the pulse amplitude or the pulse repetition rate is increased. This results in an equivalent rise of the dc offset affecting the core permeability, and can even cause core saturation during the end portion of the pulse width.

10.1.7 General reliability considerations. The military-type electronic transformer is potentially one of the most reliable components utilized in electronic equipment. Because of similarity of construction, these general reliability considerations described herein are applicable to transformers, inductors and RF coils, although reference is made only to transformers.

From a dielectric viewpoint transformers have stress levels that are normally low when compared to other component categories. Electronic transformers have completely static parts well suited to adequate structural design techniques. In addition, relatively large thermal capacity generally exists in transformers to absorb unusual conditions of equipment under end use. However, because of many variables involved from the conceptual design stage through the final product stage, transformers sometimes become chronic reliability "offenders".

When trying to obtain maximum reliability of a magnetic component, one should recognize the vital effect of each link in the total chain from concept to use. Some of these links are under the ultimate user's control, some are the responsibility of the equipment manufacturer, and, certainly, a major responsibility rests with the component manufacturer.

In order to improve the reliability of electronic transformers, it is necessary to understand what causes failures before one can act to help prevent them. The following are general considerations of failure patterns and the principal stresses involved:

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10.1.7.1 Stress. Design safety factors and manufacturing controls should take into account the stresses a transformer will be subjected to in its lifetime. These stresses have an effect on the four basic parts of a transformer; the dielectric system, the conductive system, the structural system, and the magnetic system. Stresses of significance in electronic transformers are:

- a. Temperature
- b. Vibration
- c. Mechanical shock
- d. Humidity
- e. Dielectric
- f. Thermal shock.

In addition to their magnitudes, all of these stresses except dielectric have a significant interaction with time or number of cycles applied. The most important effects of the stresses are shown in Table I.

TABLE I. Effects of primary stress situations on electronic transformers

Stress	Effects
Temperature (long duration, steady state)	Chemically deteriorates insulation materials; weakens solder joints and connections; weakens structural parts; contributes to dielectric breakdown of insulation
Dielectric stress	"Plating" effect deposits conducting ions between parts that should be electrically far apart; deterioration effects as function of time are believed to be negligible in transformers
Vibration and mechanical shock	Contributes to failure of mechanical parts if limits are exceeded; detects weak weld joints; will cause loose electrical parts to move, abrade, and wear out
Thermal shock	Breaks wire, seams, etc., by fatiguing or simple stressing to breaking point; erodes insulation

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In summary these stress effects are:

- a. Temperature, vibration, mechanical shock, and thermal shock, directly age or degrade the transformer as a function of time; these four stresses interact significantly and are the sole causes of failures in all four transformer parts
- b. Dielectric stress accelerates aging effects in the dielectric system.

Because a transformer is a composite of materials with various expansion coefficients, temperature changes establish mechanical stresses. These stresses vary widely as the transformer temperature changes. Therefore, it is important to minimize the mechanical stresses by proper selection of the core, windings, insulation, and impregnation materials, as well as the transformer construction.

Another cause of transformer failure is excessive voltage stress. High voltage stresses between windings, turns, terminals, or to ground (or core) may, if they exceed the dielectric withstanding ability of the insulation barrier, cause a breakdown of the insulation, which will result in transformer failure. Winding configurations, the shape of winding edges, and terminations (electrodes) can intensify electric fields, thus increasing electrical stresses considerably.

10.1.7.2 Failure modes. Failure modes are shown in Table II.

TABLE II. Failure modes in electronic transformers

System	Detection and Interaction
Dielectric	Turn-to-turn or layer-to-layer failures (within a winding) cause radical changes in transformer electrical performance. Heating by excessive current flow may be severe enough to cause failure in the conductive system. Turn-to-turn (between windings) and turn-to-ground failures also cause radical changes in the transformer electrical performance.
Conductive	Failure of current-conducting parts may cause catastrophic or parametric failure.
Structural	Failure of an external structural part may be detected visually, but may also induce failures in the dielectric and conductive systems. Internal structural part failures induce dielectric and conductive system failures.
Magnetic	Failure in the magnetic system is usually characterized by slow deterioration of the magnetic properties.

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10.2 Audio, power, and high-power pulse.

10.2.1 Introduction. This section covers audio, power, and high-power pulse transformers and inductors for use in electronic and communications equipment. These components are in accordance with MIL-T-27. They have rms test-voltage ratings of 50,000 V or less, and include high-power pulse transformers where the peak pulse power is greater than 300 W and the average pulse power is greater than 5 W.

10.2.2 Usual applications. Transformers are used in applications to optimally couple a load to a source, provide isolation, transform voltages and currents, match impedances, control and shape signals, and to shield and filter signals.

Transformers used to deliver power, whether rectified or not, generally operate at lower frequencies or at a single frequency, and are predominantly governed by different transformer elements than pulse transformers, which operate at much higher and wider frequency bands.

10.2.3 Physical construction.

10.2.3.1 Inductors.

Core materials. The two basic types of cores used in inductors are solid magnetic steel alloys and powder cores.

Magnetic steel cores are obtained in the form of laminations, C-cores, bobbins, sleeves, or toroids. Most magnetic steel alloys are extremely sensitive to mechanical stress from handling, winding pressure, or bending and must be protected to preclude distortion of magnetic properties. For this reason, many designs use cores placed in rigid aluminum or plastic boxes filled with a silicon compound for core cushioning.

Powder cores include molybdenum permalloy (moly-perm), ferrite, and powdered iron cores. Moly-perm cores are made by reducing the magnetic alloy material (nickel/iron/molybdenum) to a very fine powder and are then formed under high pressure and temperature to the desired shape. Ferrites are combinations of various metallic oxides formed into cubic polycrystalline structure by solid state reaction and then pressed and sintered into toroidal cups and other shapes. Powdered iron cores are made in a similar manner. Powdered iron cores are superior to moly-perm cores at frequencies above 100kHz and are considerable cheaper than either moly-perm or ferrites.

Wire and winding. The most important element in an inductor is its winding. High-frequency toroidal inductors have been wound on "air cores" (that is, nonmagnetic cores) which were machined from wood or plastic merely to serve as a support for the winding.

The winding consists mainly of round insulated copper wire, although copper or aluminum sheet strips and square wires may be used.

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Due to the high abrasion experienced by the wire during winding, toroids are generally wound with a double (or multiple) insulated wire. Heavy nyleze- and heavy polythermaleze-type insulated wire are used extensively. Nyleze is a wire coating with very high abrasion resistance and is rated for operating temperatures up to 130 °C. Nyleze-coated wire is solderable and has an advantage in production in that it can be soldered directly at the termination. Polythermaleze is a high-temperature wire rated up to 200 °C and it also has a high abrasion resistance. However, it cannot be soldered directly. The insulation must be stripped off, either chemically or mechanically.

Several types of toroidal winding methods are employed. Continuous winding is employed for inductors used at low frequencies. This winding puts the most wire on the core and results in the highest L/R ratio. The wire turns are applied parallel, traversing a multiple of 360 degrees on the core in one direction.

For medium- and high-frequency applications, the effect of distributed capacitance in the coil must be taken into consideration. Consequently, a winding scheme (either bank or progressive) to minimize coil capacitance may be used. They result in somewhat higher winding resistance, as wire one size finer often must be used (compared with a continuous winding) to accommodate the turns required for the inductance. A bank winding consists of several distinct winding sections or segments. Progressive winding is a bank winding with a large (continuous) number of banks; that is, the entire winding is applied in one 360 sweep.

Effect of coil embedment on coil capacitance. The capacitance of the coil is affected not only by the wire insulation and type of winding but also by the coil impregnating and potting compounds. Generally, distributed capacitance is increased by the impregnant, as most of the impregnants have a dielectric constant greater than unity. For low distributed capacitance in a coil, dry air is the best dielectric. Coils operating at high frequencies are often embedded in tiny glass-sealed air bubbles or just sealed in a dry nitrogen atmosphere. Potting compounds vary from waxes and tars to a wide variety of epoxies. The coils are often coated with silicone rubber, which cushions the coil from pressures exerted by the potting compound.

Enclosures. Three types of enclosures (open coil, molded coil, and metal-encased coil) are used for inductors.

Open coils have the least environmental protection. They consist mainly of a winding wound on a core and terminated with plastic insulated leads. Open coils are frequently coated on the outside with plastic, which serves less for environmental protection than for mechanical protection against scraping or breakage of the winding wires during handling. Open coils are normally not recommended for NASA applications.

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For environmental protection, coils must either be molded (encapsulated) or hermetically sealed (metal-encased). Encapsulated units meet stringent requirements of Grade 5 of MIL-T-27 and are generally smaller in size and less expensive than comparable metal-encased coils. Termination of the windings is also simpler and is provided by a variety of solder terminals, printed-circuit pins, or flexible insulated leads. In hermetically sealed coils, termination must be made through a sealed insulated terminal which is soldered into the case.

Generally, no mounting provision is made for open coils. Molded units, in addition to solidly embedded terminal pins, can be provided with a variety of mounting hardware, such as brackets, threaded metal inserts, or studs, thus offering inductors that can withstand high shock and vibration levels. Of course, metal-encased units can be supplied with the same mounting provisions and will withstand the most severe environmental stresses. Also, additional magnetic and static shielding can easily be provided for metal-encased units.

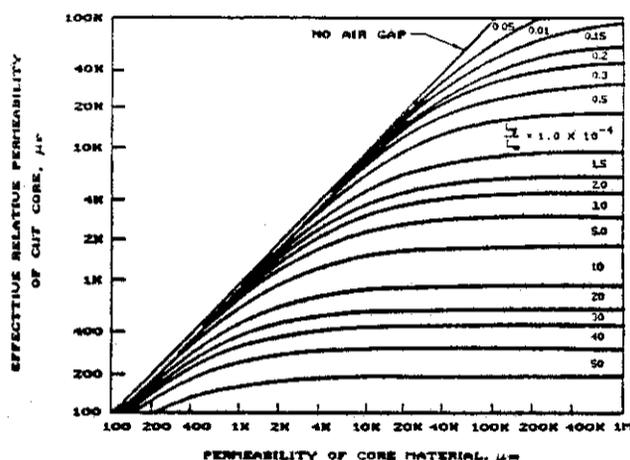
10.2.3.2 Transformers.

Cores. Most transformers have stationary cores in various shapes of thin iron alloy sheets, pressed powders, or ferrites. The high permeability of these cores confine the bulk of the flux to the core area and help achieve close coupling between the coils of a transformer.

In selecting the core material, consideration must be given to many core characteristics, such as core losses, saturation flux, magnetizing force, incremental permeability, operating temperature, and others. Some core characteristic information can be obtained from the core B-H (hysteresis) loop. The flux density is proportional to the magnetizing force (H), the proportionality factor being the permeability (μ) of the core material. Thus, the permeability is defined as $\mu = B/H$. The core data indicate that for most core materials the flux density (B) is not a linear function of the magnetic intensity (H). It also depends on the frequency, temperature, and level of operation. Therefore, the permeability (μ) representing the slope of the B-H curve is not a constant. Additionally, the permeability depends also on the core hysteresis. Permeability can be further complicated by a direct current in the coils. In this case, it is the incremental permeability (ie the permeability of the core material to small increments of alternating current superimposed on direct current) which is of importance. The introduction of an air (nonmagnetic) gap in the flux path of a core lowers the effective relative permeability. Figure 4 shows the relation of effective permeability to the core material permeability for various gap ratios.

The ratio of residual flux (B_r) to the saturation flux (B_s) indicates the "squareness" of the core, a characteristic that has a substantial effect on transformer operation, particularly in unidirectional pulse applications or with dc-biased windings (e.g., pulse transformers, saturable reactors, and magnetic amplifiers).

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$$\mu_r = \frac{\mu_m}{1 + \frac{L_g}{L_m} \mu_m}$$

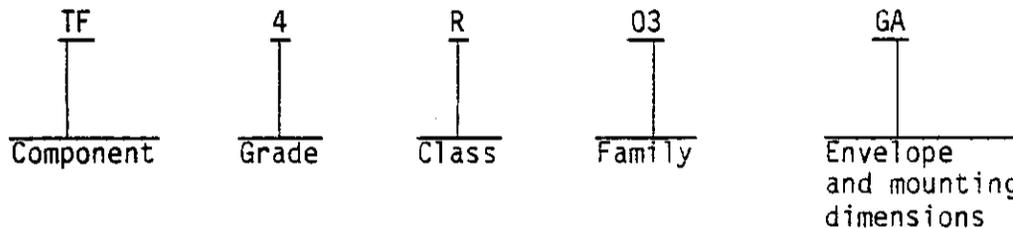
WHERE: L_g = EFFECTIVE AIR GAP
 L_m = MEAN CORE LENGTH

NOTE: CURVES NEGLECT STACKING FACTOR AND FRINGING FLUX. AIR GAP AND MATERIAL CROSS-SECTIONS ASSUMED TO BE EQUAL.

FIGURE 4. Effective permeability of cut core vs permeability of the material.

Cores are available in soft magnetic steels, ferrites, and powdered iron materials. The most significant advantage of ferrite over laminated and powdered iron cores is its high resistivity, which affords a dense homogeneous magnetic medium with high permeability, stable with respect to both temperature and time, but without the high eddy current losses inherent in conventional core materials.

10.2.4 Military designation. The designation of MIL-T-27 units is in the form depicted below. Note that MIL-STD-975 contains no transformers, all transformers must be procured to the requirements of MIL-STD-981.



- a. Component. Transformers and inductors are identified by the two-letter symbol "TF."

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- b. Grade. The grade is identified by a single digit.

Grade 4. These units are metal encased, including such constructions as those having a metal shell, with separately fabricated headers which provide a hermetic seal.

Grade 5. These units are encapsulated, including molded or embedded constructions, or are units with a metal shell, open at one or both ends and filled with encapsulant material.

Grade 6. These units are open type and are generally intended for subsequent potting, molding, or embedment in an assembly with or without component parts.

- c. Class. The class is identified by a single letter denoting the maximum operating temperature (temperature rise plus maximum ambient temperature).

- d. Envelope and mounting dimensions. The envelope and mounting dimensions are identified by a two-letter symbol in accordance with MIL-T-27.

10.2.5 Electrical characteristics. Transformers and inductors are usually intended for custom applications and as such, MIL-T-27 provides provisions for specifying parameters. The following are typical ratings which should be specified:

10.2.5.1 Power transformers.

- a. Nominal primary voltage and possible variation (taps on winding clearly defined)
- b. Operating frequency range
- c. Secondary rms load voltages with allowable tolerance at nominal input voltage and rated loads
- d. Secondary rated rms and dc load currents and possible variations
- e. Allowable regulation the basis for which shall be clearly stated; e.g., 5 to 100 percent load, over temperature range, etc.
- f. Electrostatic shielding
- g. Polarity of windings
- h. Surge conditions and transient peaks
- i. Corona limits, if absolutely necessary

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- j. Capacitive or inductive input if used in a rectifier or filter circuit.
- k. Allowable dc resistance of each winding
- l. Self-resonant frequency
- m. Dielectric withstanding voltage limit.

10.2.5.2 Inductors.

- a. Rated inductance and required limits at nominal rms voltage and frequency, and dc current
- b. Allowable dc resistance
- c. Quality factor (Q) at the specified voltage and frequency.

10.2.5.3 Audio transformers.

- a. Source and load impedances
- b. Allowable variations in primary impedance when operating at rated load on secondaries and at the specified frequency
- c. Primary and secondary dc currents
- d. Frequency response at the specified power level
- e. Harmonic distortion
- f. Insertion loss at the specified frequency
- g. Self-resonant frequency
- h. Electrostatic shielding
- i. Magnetic shielding
- j. Polarity
- k. Phase shift
- l. Resistive, inductive, and capacitive unbalances, if applicable
- m. Allowable insulation resistance
- n. Dielectric withstanding voltage limit.

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10.2.5.4 Winding inductance. The inductance of a winding is proportional to the square of the number of turns and the effective permeability of the magnetic medium, and inversely proportional to the total magnetic path length. With an air gap present in the magnetic path, the inductance (L) of a coil can be calculated as:

$$L = \frac{0.4 \pi N^2 \times 10^{-8}}{\frac{L_m}{\mu A_m} + \frac{L_g}{A_g}}$$

where

L = coil inductance in henries
 N = number of turns in the coil
 L_m = magnetic length of the core in cm
 L_g = length of the air gap in cm
 A_m = core cross-sectional area in cm²
 A_g = effective gap area in cm²
 μ = core permeability.

10.2.5.5 Induced voltage. Most transformers have stationary cores in various shapes of laminations, pressed powders, or ferrites. The high permeability of these cores confine the bulk of the magnetic flux to the core area and help achieve close coupling between the coils of a transformer. The close coupling enables the primary and secondary voltages to have almost the same voltage per turn. Thus, an alternating voltage applied to the primary winding of a transformer induces in the secondary winding a voltage directly proportional to the winding turns ratio. The induced voltage is proportional to the rate of flux change and the number of secondary turns:

$$E = - \frac{Nd\phi}{dt} \times 10^{-8}$$

10.2.5.6 Flux density. Flux density (the number of flux lines per unit area of the core cross-section) is an important parameter that characterizes the ability to contain the flux within the magnetic core. For a sinusoidal voltage applied to a winding, the flux density is:

$$B = \frac{E \times 10^8}{4.44 \times f \times N \times A}$$

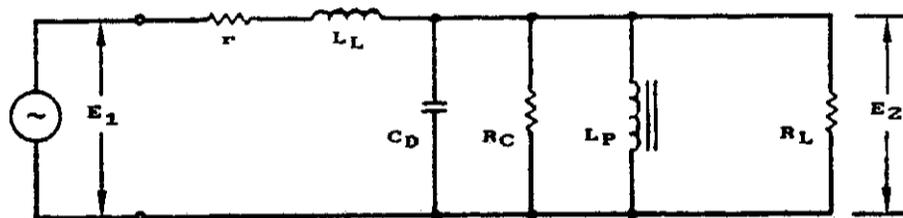
where

B = flux density (gauss)
 E = voltage (rms)
 f = frequency (Hz)
 N = number of turns in winding
 A = effective core cross-section area in square centimeters.

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10.2.5.7 Equivalent circuit representation. Most transformers are not ideal. They have finite inductance (permeability), winding resistance, core losses, leakage inductance, and dielectric losses, and the core flux capacity is limited. Therefore, the induced voltages and the output voltages do not exactly correspond to the winding turn ratios and the regulation of a transformer usually cannot be neglected.

A transformer can be represented in an equivalent circuit (shown in Figure 5) with all important transformer elements represented as lumped parameters and reflected to one side (e.g. primary), including the load impedance (R_L). All impedances are reflected by the square of the turns ratio.



where

- r = primary and reflected secondary winding resistances
- L_L = primary and reflected secondary leakage inductances
- C_D = distributed capacitance represented as a total shunt capacitor
- R_C = shunt resistance equivalent to the core losses
- L_p = primary open circuit inductance
- R_L = load impedance.

FIGURE 5. Transformer equivalent circuit.

10.2.6 Environmental considerations. The major environmental conditions affecting transformer and inductor life and performance and influencing the methods of construction and protection of such devices are as follows:

10.2.6.1 Temperature. Temperature has the following effects on transformer construction.

- a. Magnetic materials. These materials must be selected so that their variations with temperature will not result in appreciable variations in circuit performance.
- b. Insulation. Class of insulation material and allowable winding temperature rise must be selected to give the required life at maximum operating temperature.

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- c. Sealing. Materials and type of construction must be such that wide temperature range and rapid cycling over this range will not degrade the moisture seal.
- d. Regulation. When regulation is critical, design techniques must be applied to minimize copper losses and to deliver normal secondary voltage at rated load and at design center temperature.

10.2.6.2 Altitude. High-altitude, low-pressure operation affects the dielectric properties of insulation and the temperature rise. Because voltage breakdown strength and the corona-inception voltage of air are affected by the air density, internal insulation and spacing must be designed for the lower pressure at higher altitudes.

Another factor in high-altitude operation is the temperature rise in the transformer where significant power is dissipated. At ground-level altitudes under 10,000 feet, much of the heat dissipated in the transformer can be carried away by conduction, convection and radiation. As the altitude is increased and air density decreased, the rates of conduction and convection are reduced and the heat must be almost entirely removed by radiation. As a result, the temperature rise of a transformer with a given power loss increases substantially with altitude. Care must be taken in locating the transformer within a package, considering the effect its temperature rise may have on other components.

10.2.6.3 Vibration, shock, and acceleration. Power transformers and inductors usually have a higher mass to size ratio and are unusually susceptible to vibration and mechanical shock. Mountings may be damaged as may winding and lead wires if they are not properly secured.

10.2.6.4 Explosive atmosphere. Explosive atmospheres generally do not cause deterioration of transformer parts. However, a spark which could be the result of a failure from some other cause could be disastrous. For this reason, transformers used in explosive atmospheres should be hermetically sealed.

10.2.6.5 Radiation. Some insulation materials used in transformers and inductors may deteriorate when exposed to radiation. Magnetic properties, however, are not affected.

10.2.7 Reliability considerations. The following failure mechanisms are common causes of transformer and inductor failures.

10.2.7.1 Excessive primary voltage. Excessive primary voltage induces excessive voltage in the secondary windings. The overvoltage may lead to immediate puncture of the insulation or to premature breakdown. The overvoltage on the primary will saturate the core and increase the input current, which increases the core and copper heat losses. This results in excessive temperature and rapid deterioration of the insulation system.

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10.2.7.2 Excessive secondary currents. Excessive secondary currents will result in a corresponding increase in primary current which will cause over-heating in both windings. This will cause a rapid deterioration of the insulation system, open or short circuited windings, or misshapen or broken containers, as a result of rapid expansion of the potting or filling materials.

10.2.7.3 Input frequency fluctuations. Operating frequencies below the lower design range limit will result in low reactance due to core saturation and higher than rated input current. This will result in overheating of the transformer and rapid deterioration of the insulation system.

10.2.7.4 Corona. This phenomenon occurs at points of high voltage stress and causes accelerated aging of the insulation by the liberation of ozone and by increasing temperatures. It creates weak spots in the insulation that eventually lead to insulation breakdown.

10.2.7.5 Other causes. Most transformers and inductors fail as a result of insulation breakdown resulting from insulation embrittlement and degradation of insulation resistance due to exposure to excessive hotspot temperatures. Open circuits occur in some instances because of poor wire terminations and mechanical structures that are inadequate to support the unit. Poor workmanship in coil winding and inadequate location of insulation contribute to failures because of opens resulting from broken wires and short circuits between uninsulated current-carrying parts. Poor workmanship in soldering connections and incorrect techniques of making wire joints are also common causes of failures.

10.2.7.6 MIL-STD-981. MIL-STD-981 establishes the requirements for acceptable design, manufacturing, and quality criteria. Transformers and inductors must meet the criteria and quality assurance provisions identified in that Military Standard.

10.3 TRANSFORMERS AND INDUCTORS, PULSE TRANSFORMERS

10.3 Pulse transformers.

10.3.1 Introduction. This section covers low-power pulse transformers used in electronic and communications equipment. These components are manufactured to meet the requirements of MIL-T-21038 and are transformers where the peak pulse power is 300 W or less and the average pulse power is 5 W or less.

10.3.2 Usual applications. Low-power pulse transformer applications fall into two categories:

- a. Those applications which may be characterized as coupling or impedance matching
- b. Those applications in which the transformer acts in conjunction with some nonlinear element such as a transistor to form a pulse-generating circuit. Most applications in this category are blocking oscillator circuits.

All applications which can be discussed in terms of generator constants, load impedances and desired pulse responses fall into the first category defined above. All others are in the second. This distinction between the two categories is emphasized because in a pulse generating circuit, the dynamic characteristics of other circuit elements have as much to do with the pulse shapes as the parameters of the transformer itself.

Some common applications of low-power pulse transformers are described below.

10.3.2.1 SCR circuits. Pulse transformers find broad application in silicon-controlled rectifier (SCR) speed and load control circuits. Commonly, they supply between 30 mA and 1 amp of current to an SCR gate. Although 1:1 is the most common transformer turns ratio in SCR circuits, often it is more convenient and economical to use a step-up transformer. Its voltage gain is especially useful in servo controlled motors and in speed-control applications where a transducer signal must be applied.

10.3.2.2 Blocking oscillator circuits. Another valuable use of pulse transformers is in blocking oscillator circuits. The transformer should have an in-circuit saturation time at least double the required pulse width. Then, by carefully selecting the values of components, the pulse width of the circuit may be reduced. The resulting circuit will be relatively unaffected by changes in transistor parameters or temperature.

10.3.2.3 Low power data bus coupling pulse transformers. Data bus coupling transformers provide signal coupling and fault isolation between the stub and the main data bus. Terminal interfaces must be designed to guarantee proper signal level and minimum distortion to achieve the desired bit and word error rates.

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**10.3 TRANSFORMERS AND INDUCTORS,
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10.3.3 Physical construction. Low-power pulse transformers are generally small in size because of the low-power requirements. The outer surface of the transformer may be a metal shell for magnetic shielding or it may be the potting material in its normally molded shape or conformally coated configuration. These small parts are usually mounted by their lead wires. Many designs are used in printed circuit construction, where they are mounted by their lead wire pins. Figure 6 demonstrates a typical outline drawing.

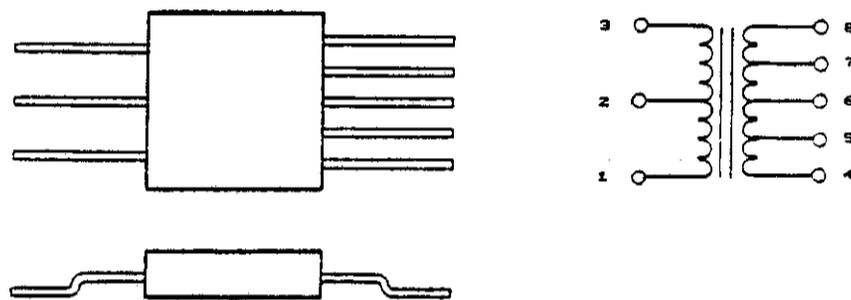
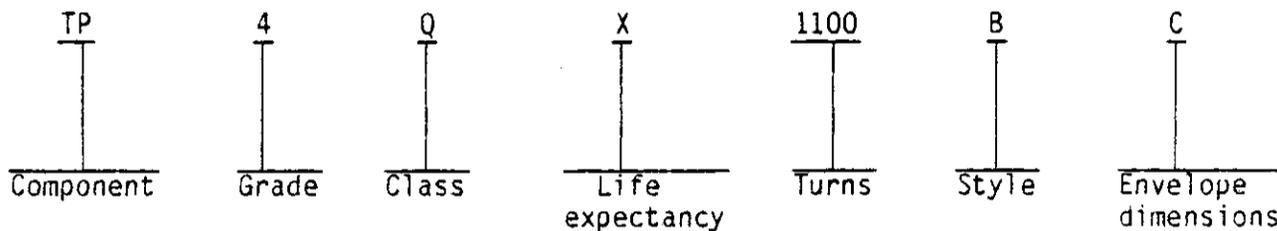


FIGURE 6. Typical outline drawing and schematic.

The internal construction of a low-power pulse transformer consists of a magnetic core with two, three, or four coil windings wound on the core which are insulated from each other. The ends of the coil windings are soldered to lead wires which serve as external electrical connections to the transformer. The core and coil windings are encapsulated with an insulation potting material, which adds additional insulation, serves as a protection from moisture, and provides mechanical support for the core and coils.

10.3.4 Military designation. The type designation specified in MIL-T-21038 is in the form depicted below. Note that MIL-STD-975 contains no transformers. All transformers must be procured to the requirements of MIL-STD-981.



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Component. Low-power pulse transformers are identified by the two-letter symbol "TP."

Grade. The grade is identified by a single digit.

Class. The class is identified by a single letter denoting the maximum operating temperature (temperature rise plus maximum ambient temperature).

Life expectancy. The life expectancy is identified by a single letter.

Turns ratio. The turns ratio for the first four windings is identified by a four-digit number.

Style. The style is identified by a single letter, in accordance with MIL-T-21038.

Envelope dimensions. The envelope dimensions are identified by either a single letter or two letters, in accordance with MIL-T-21038.

10.3.5 Electrical characteristics. Pulse transformers are usually intended for custom applications in which parameters are tailored to the circuit. MIL-T-21038 has provisions for specifying these parameters. Following are typical ratings which should be specified.

- a. ET constant
E is pulse amplitude (in volts)
T is pulse width (in microseconds)
- b. Power level
- c. Pulse waveform characteristics
- d. Number of windings
- e. Winding dc resistance
- f. Turns ratio
- g. Winding-to-winding capacitance
- h. Open circuit inductance
- i. Leakage inductance
- j. Core material description
- k. Magnetizing current
- l. Winding structure

**10.3 TRANSFORMERS AND INDUCTORS,
PULSE TRANSFORMERS**

- m. Dielectric strength test voltage
- n. Induced test voltage
- o. Insulation resistance test voltage
- p. Environmental conditions
- q. For coupling transformers, specify the source and load impedance and describe the input pulse and the desired output pulse.

10.3.6 Environmental considerations. The environmental considerations for low-power pulse transformers are essentially the same as those described for power devices (section 10.2.6). The low-power pulse transformers are usually small because of their low power requirements and thus are usually adequately designed to meet the requirements of shock and vibration. Units which are mounted by their terminal lead wires should be provided with additional mechanical support for the body.

Temperature cycling and thermal shock environments are the most severe environments for these parts. When adjacent materials have different rates of expansion with temperature changes, there are often stresses applied to the fine wire in the coil windings. If materials separate or crack during temperature change, the coil wires may break and cause an open coil winding.

10.3.7 Reliability considerations.

10.3.7.1 Usual failure mechanisms. The usual failure mechanisms for low-power pulse transformers fall into one of the following categories:

- a. Broken coil wire
- b. Insulation failure
- c. Change in core characteristics.

Broken coil wire occurs most often where small diameter wire is used because of the low power requirement, and where the mechanical construction of the part is not adequate to withstand the mechanical stresses encountered in normal handling, or when exposed to normal environments. Circuit discontinuities also occur at failed solder joints where the wire is soldered to the terminal. These failures are due to faulty solder joints or excessive mechanical stress.

Insulation failures may occur between coil windings or between the current-carrying coils and the core, mounting structure or metal cases. These failures are caused by overvoltage stress on the insulation or the use of inadequate insulation, and when insulation is damaged by excessive mechanical stress on the insulation materials. Generally, the dielectric withstanding voltage and induced voltage tests are adequate to assure that insulation failures will not occur during normal operation.

**10.3 TRANSFORMERS AND INDUCTORS,
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Change in core characteristics can be caused by mechanical forces on the core, high ambient temperature, exposure to magnetic fields and dc saturation of the core during dc resistance measurements of the coil windings. A change in the core characteristics will cause a detrimental change in the transformer pulse waveform parameters. Special consideration of the design, construction, handling, and testing of pulse transformers is essential for the preservation of the waveform parameters.

10.3.7.2 MIL-STD-981. MIL-STD-981 establishes the requirements for acceptable design, manufacture, and quality criteria. Pulse transformers must meet the criteria and quality assurance provisions identified in that military standard.

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**10.4 TRANSFORMERS AND INDUCTORS,
COILS AND CHIP INDUCTORS**

10.4 Coils and chip inductors.

10.4.1 Introduction. This section covers coils and chip inductors for use as inductive elements in radio frequency circuits. They are low profile, miniature devices suitable for use on printed wiring boards and in hybrid circuits.

10.4.2 Usual applications.

10.4.2.1 Coils. A radio frequency coil is used to pass direct current and present a high impedance to radio frequencies. The coil may have high radio frequency voltage across it. Typical applications are in oscillators, amplifiers, converters, and exciters. High stability and high Q characterize these inductors, which operate up to 200 MHz. Inductance values from 0.1 microhenries through 100 millihenries are available. Typical Q ranges from 50 to 80. These devices meet the requirements of MIL-C-39010 for established reliability coils.

10.4.2.2 Chip inductors. Fixed and variable chip inductors are primarily intended for use in hybrid microelectronic circuits and have termination systems suitable for solder attachment to alumina substrates. These devices are manufactured to meet the requirements of MIL-C-83446.

10.4.3 Typical construction.

10.4.3.1 Coils. Fixed radio frequency coils consist of a coil of insulated copper wire wound on a coil form. Two terminal wire leads are anchored in the coil form and the wire coil ends are soldered to the terminal lead wires. This internal assembly is encapsulated with a molded jacket of plastic insulating material or conformally coated with a protective coating such as varnish.

RF coils may be enclosed in housings of magnetic or conducting material to form electromagnetic shielding. Considerations of economy and good practice indicate that shielded coils should be used only in applications where the environmental and service conditions are severe and where space is at a premium. Advantages of shielding are the exclusion of unwanted pickup from random internal and external sources, the prevention of feedback which may cause oscillation, and the confinement of the magnetic and electrostatic fields generated by the coil itself.

Coil forms may be powdered iron, ferrite, ceramic or phenolic. Powdered iron and ferrite coil forms are used when high inductance values are required. The ceramic and phenolic forms are used for lower inductance values. The phenolic form is used where environmental conditions are not so severe as to require ceramic forms. Figure 7 shows a typical coil.

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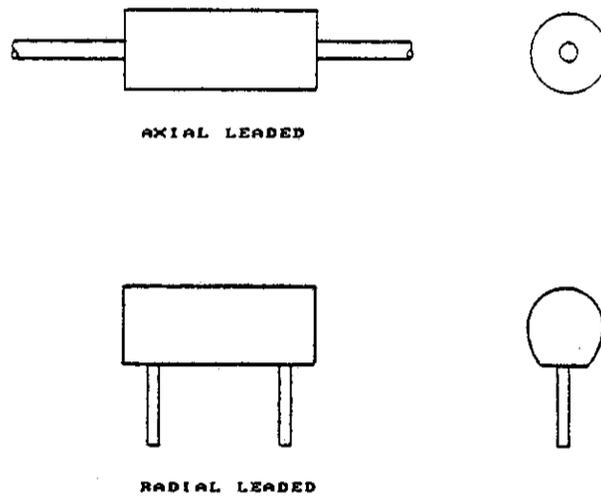


FIGURE 7. Typical outline drawing for RF coils.

10.4.3.2 Chip inductors. Chip inductors are miniature coils molded onto alumina substrates having metallized termination pads compatible with standard hybrid attachment techniques, as shown in Figure 8.

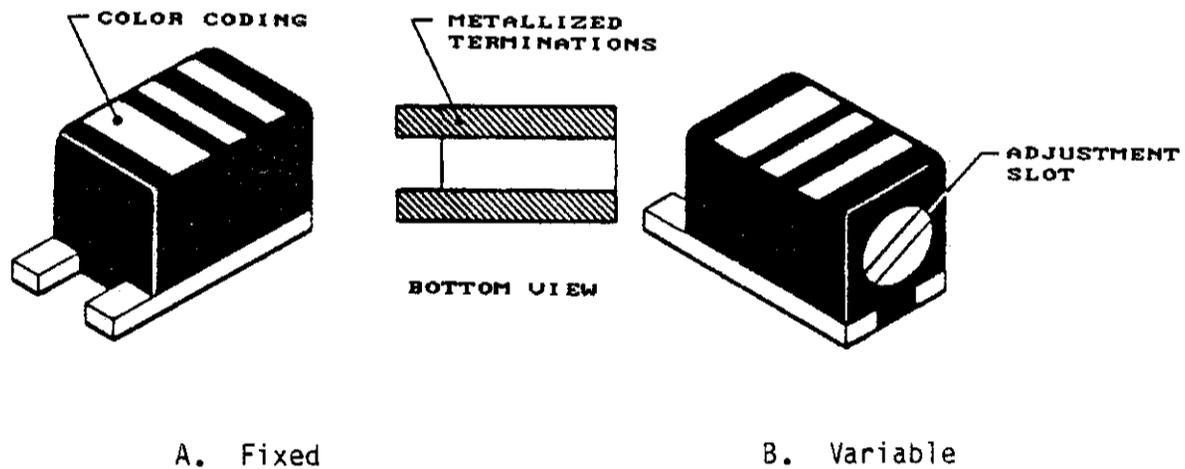


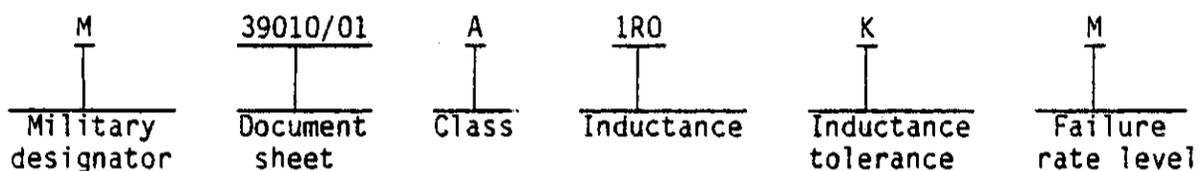
FIGURE 8. Typical outline drawing for chip inductors.

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**10.4 TRANSFORMERS AND INDUCTORS,
COILS AND CHIP INDUCTORS**

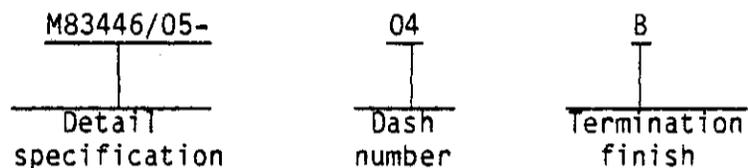
10.4.4 Military designation.

Coils per MIL-C-39010. MIL-C-39010 is an established reliability specification. It designates parts as follows:



- Class. A single letter denoting the maximum operating temperatures
- Inductance. nominal inductance expressed in microhenries
- Tolerance. single letter denoting tolerance
- Failure rate level. letter denoting failure rate level.

Chip inductors per MIL-C-83446. The designation specified in MIL-C-83446 is in the following form:



- Detail specification. The specification slash sheet
- Dash number. Sequentially assigned number
- Termination finish. A single letter denoting finish.

10.4.5 Electrical characteristics. Due to their small physical size, radio-frequency coils have low inductance values, ranging from 0.1 to 33 microhenries for all but the largest size (M39010/03). This device has inductance range up to 100 millihenry.

Chip coils have inductance ranges up to 1 millihenry for fixed inductors and 0.1 millihenry for variable inductors.

10.4.6 Environmental considerations. The environmental considerations for radio frequency coils are essentially the same as those described in the section "Transformers, power."

10.4 TRANSFORMERS AND INDUCTORS, COILS AND CHIP INDUCTORS

Radio frequency coils are small in size and, as a result, are usually adequately designed for shock and vibration requirements without special considerations. Generally, the voltage level of radio frequency coils is low and insulation problems do not exist, except where the voltage level above ground requires special insulation between the coil and ground.

Molded jacket coils are well protected from moisture and contaminants. Open type coils with only a varnish coat should be used in more protected applications.

10.4.7 Reliability considerations. Molded rf coils tend to fail because of degradation of Q, shorted turns, and open terminations. In the instances where Q degrades, there will be a higher incidence of failures in coils with powdered iron cores because the powdered iron slug, if overheated, will degrade. The maximum operating temperature of coils with powdered iron cores should be limited to 105 °C by suitable derating and adequate cooling. Compression molding of insulating jackets can produce coil failures due to pressures which crush the coil or push it to one side of the molded jacket. This condition produces open circuits, short circuits, core characteristic changes, broken cores, or dielectric strength failures. Shrinkage of epoxy as it cures can place mechanical stress on the coil winding and the wire terminations and can cause an open circuit, which is the most predominant failure mode in units with encapsulated coils.

Failure mechanisms for variable rf coils are similar to those for fixed rf coils. An inability of the tunable slug to stay in position or to adjust as required is caused by mechanical failure. Tunable slugs are frequently broken by excessive force applied during adjustment.

11.1 DELAY LINES, GENERAL

11. DELAY LINES

11.1 General.

11.1.1 Introduction. This section discusses delay lines for NASA applications. These devices are not included in MIL-STD-975. The delay line is a versatile device, the basic function of which is to introduce a time delay in the transmission of an electrical signal. This delay phenomenon can be achieved by electromagnetic devices (e.g., lumped constant and distributed constant delay lines) or by electromechanical devices (e.g., metal and quartz media delay lines). Some of the applications of a delay line are:

- a. Phase shifting
- b. Triggering
- c. Time interval measurement
- d. Event synchronization
- e. Phase angle measurement
- f. Pulse train positioning
- g. Pulse storage.

An ideal delay line would perform the above functions without distorting or modifying the information contained in the signal. However, depending upon the design and the application, delay lines induce various types of losses.

11.1.2 Definitions. The large variety of delay lines technologies has given rise to an equally large vocabulary of delay definitions. However, all delay lines regardless of their physical construction share certain characteristics. These shared characteristics and their corresponding definitions are listed below and are depicted in Figure 1.

Characteristic impedance (Z_0). The value of terminating impedance that will produce a minimum reflection to the input.

Crosstalk (C_S). The amount of input pulse reflected directly into the output pulse.

Delay time (T_0). The time duration between the 50 percent point on the leading edge of the input pulse and the 50 percent point on the leading edge of the output pulse.

Dispersion. The variation of delay as a function of frequency.

Feed through. A spurious signal which arrives at the output by electrical coupling.

Frequency response. The insertion loss or voltage attenuation as a function of frequency normalized to a maximum of 0 dB.

Input fall time (F_{t1}). The time duration between the 90 and 10 percent points on the decreasing edge of the input pulse.

Input rise time (T_{r1}). The time duration between the 10 and 90 percent points on the increasing edge of the input pulse.

11.1 DELAY LINES, GENERAL

Output fall time (F_{t2}). The time duration between the 90 and 10 percent points on the decreasing edge of the output pulse.

Output rise time (T_{r2}). The time duration between the 10 and 90 percent points on the increasing edge of the output pulse.

Output voltage (E_{out}). Amplitude of the output voltage.

Postpulse spurious (P_{ps2}). The output pulse excursions following the main pulse.

Prepulse spurious (P_{ps1}). The output pulse excursions prior to the main pulse.

Pulse attenuation (a). The difference in voltage of the input and output pulses.

Pulse distortion (S). The magnitude of the largest peak amplitude of all spurious responses in either a positive or negative direction relative to pulse amplitude.

Pulse overshoot (P_{os}). The amplitude of the overshoot on the leading edge of the output pulse.

Pulse top ripple (P_r). A measure of the deviation from the average amplitude in the output pulse top.

Pulsewidth (P_w). The time duration on the input pulse between the 50 percent point on the increasing edge and the 50 percent point on the decreasing edge.

Sonic delay line. A device that uses electroacoustic transducers and the propagation of an elastic wave through a medium to achieve a signal delay.

Spurious level. The ratio of a spurious signal to the desired signal, measured at a specific frequency over a band of frequencies.

Tapped line. A delay line having more than one terminal pair associated with a single delay channel.

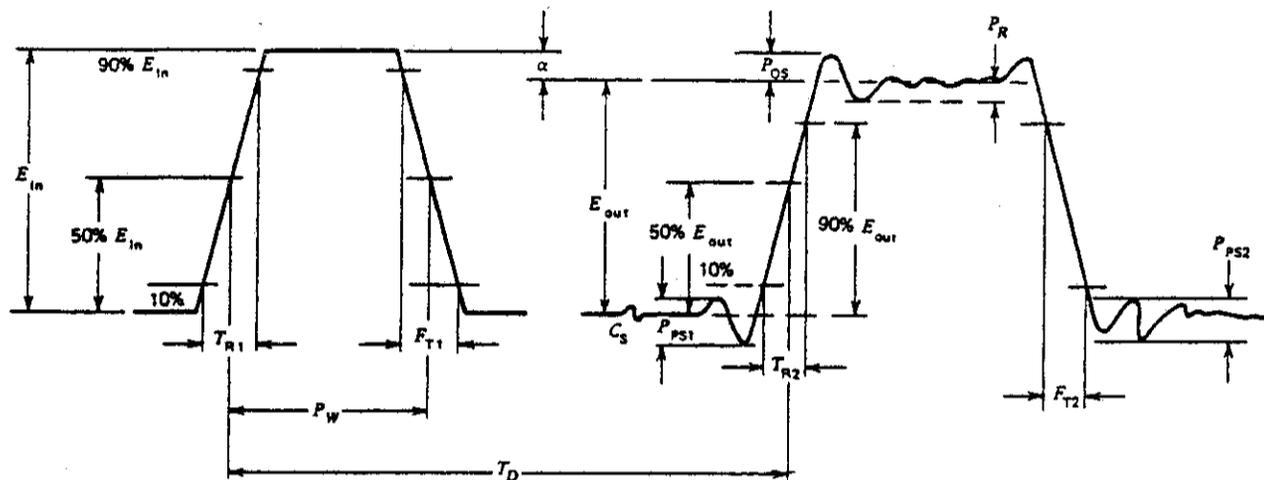
Temperature coefficient of delay. The variation of delay with temperature measured in parts per million per degree centigrade (PPM/°C.), microseconds per microsecond per degree centigrade ($\mu s/\mu s/^\circ C.$) or percent per degree centigrade ($\%/^\circ C.$).

Terminal impedance. The impedance at specific terminals and specific frequencies under conditions of negligible incident acoustic energy.

Terminal inductance. The imaginary part of the terminal impedance at a specified frequency divided by twice that frequency.

Terminal resistance. The reciprocal of the real part of the effective terminal admittance.

11.1 DELAY LINES, GENERAL

FIGURE 1. Pulse characteristics.

11.1.3 NASA standard parts. No delay lines appear in MIL-STD-975 NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List.

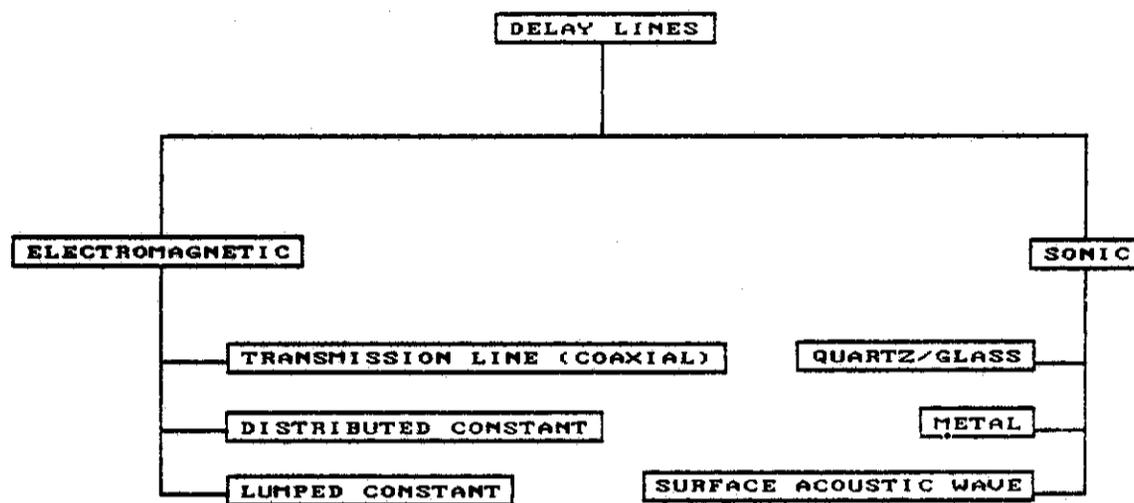
11.1.4 General device characteristics. Delay lines are essentially of two types: electromagnetic and sonic.

An electromagnetic delay line is a device that uses the inherent characteristics of electrical components to provide precise pulse delays. This is accomplished by designs that use relatively high values of inductance and capacitance in miniature packages.

Electromagnetic line designs can take the form of coaxial cables, distributed capacitance-inductance circuits, and lumped capacitance-inductance circuits. Lumped devices provide the best overall electrical characteristics of the electromagnetic delay lines. However, they are also the most expensive.

The second class of delay lines is designated as sonic because it converts electrical signals into mechanical energy in the form of acoustic waves. These acoustic waves are injected into a metal or quartz/glass medium and propagate at a velocity much slower than electrical energy. The acoustic signal is then reconverted into an electrical signal. This design has the advantage of affording very long delay times. Electromagnetic and sonic delay lines with their subsequent derivations are depicted in Figure 2.

11.1 DELAY LINES, GENERAL

FIGURE 2. Types of delay lines.

11.1.5 General parameter information. Delay line parameters consist of the following. However, not all delay lines utilize the same parameters and specifications because there are many diverse delay line technologies.

- a. Delay time
- b. Delay tolerances
- c. Tap delay times
- d. Tap delay tolerance
- e. Impedance
- f. Maximum rise time (or bandwidth)
- g. Attenuation
- h. Output loading
- i. Temperature coefficient of delay
- j. Distortion
- k. Operating temperature range
- l. Storage temperature range
- m. Maximum operating voltage
- n. Package size.

11.1.6 General guides and charts. Table I outlines the various delay line technologies and their typical performance characteristics. This table should be used as a first step in device application. It should also be noted that these are typical characteristic values and individual vendors may offer delay lines whose performance can exceed those given in Table I. This is because delay lines are essentially custom devices and devices that appear to be similar can exhibit wide variations in internal design.

11.1 DELAY LINES, GENERAL

TABLE I. Guide to delay line application

Electrical Requirement	Limits	Select
Delay time (T_d)	≤ 10 ns < 6 μ s 0.1μ s-200 ms Custom Custom Custom	Coaxial Distributed Lumped Quartz Metal SAW
Delay to rise time (T_d/T_r)	< 20 < 10 < 150 Custom Custom Custom	Coaxial Distributed Lumped Quartz Metal SAW
Characteristic impedance (Z_0)	$25 - 1$ K Ω $100 - 2$ K Ω $50 - 10$ K Ω Custom Custom Custom	Coaxial Distributed Lumped Quartz Metal SAW
Temperature coefficient of delay	>100 PPM/ $^{\circ}$ C. >150 PPM/ $^{\circ}$ C. > 50 PPM/ $^{\circ}$ C. Custom Custom Custom	Coaxial Distributed Lumped Quartz Metal SAW

11.1.7 General reliability considerations. Delay lines are not true components but are assemblies of components that are more electrically complex (and less reliable) than stand-alone components such as resistors and capacitors. They also tend to have complex responses to environmental conditions. In addition, delay lines are made using a variety of diverse technologies to achieve desirable delays with acceptable levels of distortion.

Delay lines are also labor intensive devices and must have numerous in-process checks and controls imposed during their fabrication to ensure acceptable levels of quality and reliability. Subsequently, delay lines require rigorous reliability screening prior to installation in the end application.

11.1 DELAY LINES, GENERAL

Continuously variable delay lines should be avoided because their mechanically complex construction is often susceptible to environmental hazards such as shock and vibration. Tapped delay lines such as LC devices tend to be relatively immune to their environment in comparison to continuously variable designs.

All delay lines regardless of technology or construction should undergo screening prior to assembly, packaging, and installation into a high reliability application. The nature and scope of such screening will depend upon contractual requirements and the component engineering function.

11.2 DELAY LINES, COAXIAL

11.2 Coaxial.

11.2.1 Introduction. The passage of an electromagnetic signal through any medium other than free space introduces a delay phenomenon. The simplest device using this effect is the coaxial delay line. Several attributes that make coaxial cable delay lines attractive to noncritical applications are their wide bandwidth, fast rise times, low cost, and ease of replacement.

11.2.2 Usual applications. The two most common applications of coaxial cable delay lines are video pulse processing and radio-frequency (rf) pulse envelope processing. However, coaxial delay lines in video applications have drawbacks such as dribble-up. In an rf application, there can be a tendency for the attenuation of the cable to vary at the band limits. Dribble-up and attenuation variation are discussed further in section 11.2.5.

Another consideration in the application of coaxial delay lines is availability. Diameters less than 0.013 in. may not be available in more than 8-ft lengths and diameters less than 0.034 in. in more than 10-ft lengths. Cables with diameters of 0.034 in. are available in 15-ft lengths and larger diameters are available in 20-ft lengths. Consequently, splicing is sometimes necessary to meet the application requirements. The implications of splicing coaxial cable are discussed further in paragraph 11.2.7, Reliability considerations.

11.2.3 Physical construction. Typical flexible coaxial cable construction is shown in Figure 3. The center conductor is usually copper, copper-plated high-resistivity base material, or silver-plated high-resistivity base material. A sheath of dielectric material (e.g., polyethylene, Teflon TFE, or Teflon FEP) separates the center conductor from the outer conductor. The outer conductor typically consists of a braid of copper or silver-plated copper. This braid is then sheathed in a protective cover of vinyl or fiberglass.

In addition to the typical construction, several variations are common. These variations include:

- a. The use of a secondary layer of conductive braid added either underneath or above the protective cover to provide more shielding
- b. A layer of protective armor added above the protective cover
- c. The inner and outer surfaces of the dielectric made conductive to reduce electrical noise caused by mechanical motion of the cable.

A form of high impedance coaxial cable is shown in Figure 4. This construction is the same as flexible cable construction described above except a nylon insulating thread is wound around the center conductor which results in air gaps in the nylon winding. The air gaps are used to lessen the overall dielectric constant of the sheath. Consequently, the capacitance per unit length is also lessened. Although the resultant higher impedance is often attractive in delay line applications, the propagation velocity also increases and greater cable length is required to maintain the same delay.

11.2 DELAY LINES, COAXIAL

The effective inductance of the center conductor is increased as shown in Figure 5 by winding the center conductor in the form of a coil. This inductive effect can be further enhanced by making the core of the coil magnetically loaded. A dielectric sheath insulates the coiled center conductor from a continuous wrap of alternately insulated and bare wires. This continuous wrap of wire is used instead of conventional coaxial cable braid because it yields an increase in line impedance and line delay. A conventional braid for the outer conductor would place a shorted turn adjacent to the inductor, thus decreasing the inductance and increasing overall cable losses. The outer wires are connected together at only one end to prevent ground loops also acting as shorted turns.

The lack of magnetic shielding on this type of cable makes it more susceptible to stray fields than any of the more conventional cables. Part of these stray fields are self-generated and it is advisable not to attempt to crowd the cable or place it in the vicinity of similar cables unless the application can tolerate considerable feedthrough and spurious signals.

Semirigid cable construction is shown in Figure 6. The center conductor consists of either solid or stranded material. A sheath of dielectric material (polyethylene, Teflon TFE, or Teflon FEP) separates the center conductor from the outer conductor which consists of seamless tubing. Semirigid cable is available in smaller diameters than typical flexible cables. However, small cable sizes usually mean greater losses over frequency.

Continuously variable coaxial delay lines are also available. Two common configurations are the "line stretcher" and the "trombone" in which sections of the inner and outer conductors slide into and out of corresponding sections of an adjacent assembly.

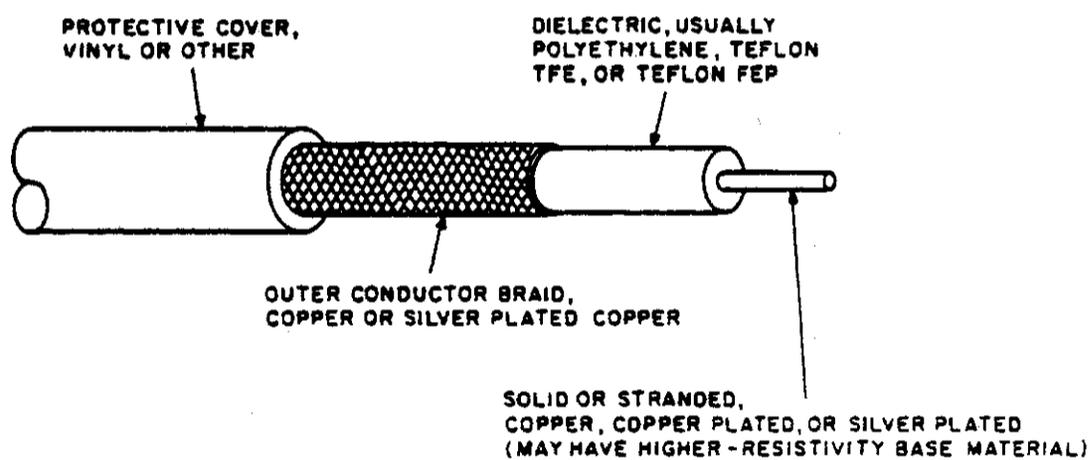


FIGURE 3. Flexible coaxial cable construction.

11.2 DELAY LINES, COAXIAL

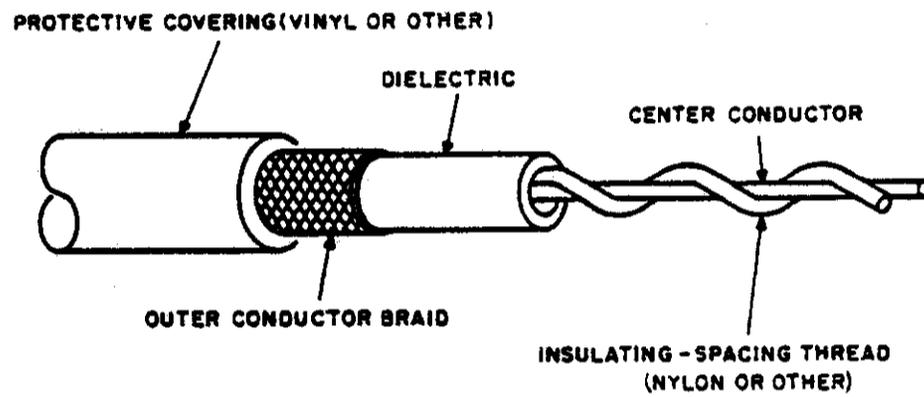


FIGURE 4. Low-capacitance/high-impedance cable construction.

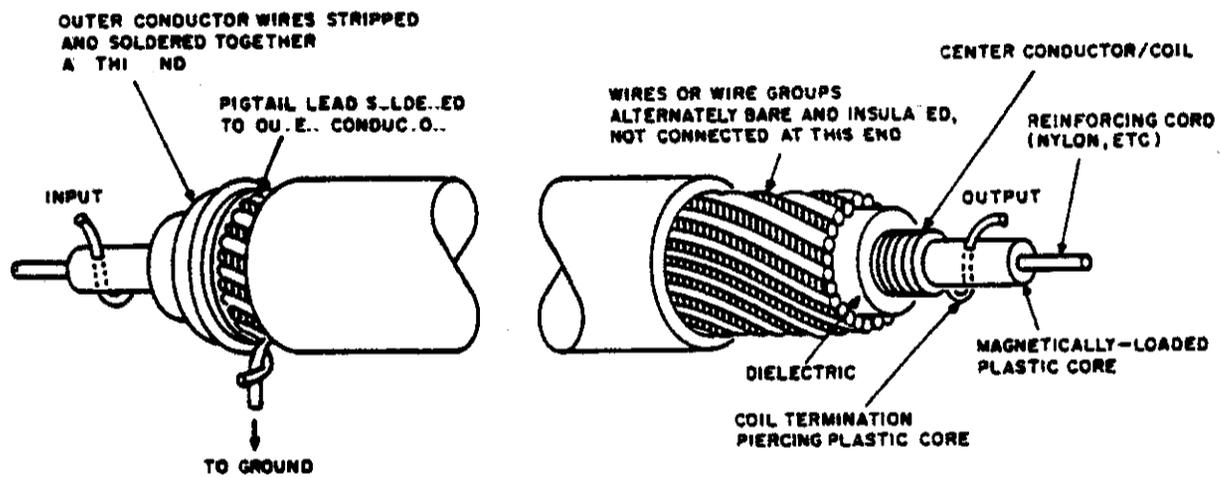


FIGURE 5. High-inductance/high-impedance cable construction.

11.2 DELAY LINES, COAXIAL

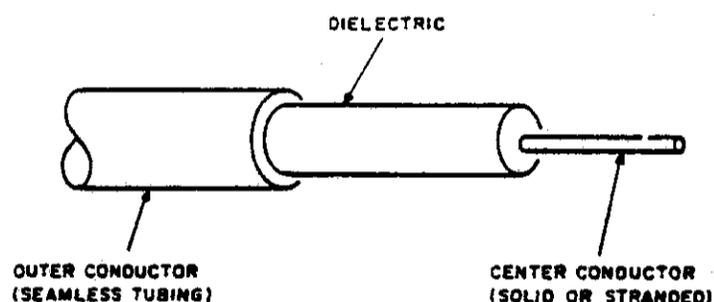


FIGURE 6. Semirigid cable construction.

11.2.4 Military designation. There are no military specifications directly applicable to the application of coaxial cable delay lines. However, there are several military specifications for coaxial cable in general. These are:

MIL-C-17	Cable, Radio Frequency, Flexible and Semirigid, General Specification for
MIL-C-22931	Cable, Radio Frequency, Semirigid, Coaxial, Semi-Air-Dielectric, General Specification for
MIL-C-23806	Cable, Radio Frequency, Coaxial, Semirigid, Foam Dielectric, General Specification for
MIL-C-28751	Cable, Radio Frequency, RG-373/U

There are no delay lines included in MIL-STD-975.

11.2.5 Electrical characteristics. One of the most common applications for coaxial cable delay lines is for the processing of video pulses. In this application however, long lines can exhibit what is called dribble-up. Dribble-up occurs when the rise time from the 10 to 50 percent point may be fast but the rise time from the 50 percent point to the full pulse height may be several times as long.

Rf pulse envelope processing, the second common application of coaxial cable delay lines, can exhibit an undesirable variation of cable attenuation as a function of frequency. This is not a particularly serious effect if phase stays proportional to frequency. This effect can be explained by the equation for the characteristic impedance of a transmission line shown below.

$$Z_0 = \sqrt{(R + j2\pi fL)/(G + j2\pi fC)}$$

11.2 DELAY LINES, COAXIAL

where

R and L = the equivalent series elements
G = equivalent shunt conductance
C = capacitance

Not only does the ratio of L/C have to remain constant, but it must also equal the ratio R/G if the line characteristics are not to change as a function of frequency. When cable attenuation varies as a function of frequency, the ratios do change but the condition is so gradual that changes of amplitude and phase may not be significant over wide bandwidths. It is advisable to check this factor, especially at the higher frequencies.

Some types of lines, such as the high-inductance and high-impedance design shown in Figure 5, exhibit the best characteristics if not bent, or if bent, done so on a large radius. This is due in part to changes in cable inductance resulting from handling and in part to feed-through if the bend approaches or exceeds 360 degrees. The semirigid cable of Figure 6 is recommended as probably more constant in characteristics after strenuous handling.

Flexible coaxial cable lines are susceptible to degradation from mounting in addition to bending. Tight mechanical clamps can cause spurious reflections that can result in signal degradation.

11.2.6 Environmental considerations. The environmental considerations associated with coaxial delay lines are:

- a. Electromagnetic environment
- b. Reduced barometric pressure
- c. Humidity
- d. Operational temperature range
- e. Shock and vibration.

In the case of high leakage coaxial designs, such as shown in Figure 3, the delay line should be dressed well away from susceptible circuits or circuits that generate high electrical or magnetic fields. Even other coaxial cables can present a hazard due to rf leakage through the outer conductor braid of the cable.

Altitude also affects delay line performance. Certain materials have the tendency to outgas at reduced barometric pressure. This outgassing can eventually lead to degradation of the part. The degradation can be physical or electrical in nature, or a combination of both.

11.2 DELAY LINES, COAXIAL

Flexible cables are not particularly suitable for exposure to conditions of high humidity. Braid and other forms of multiple wire conductors can act as moisture traps which cause corrosion. Condensation occurring in the construction of devices similar to Figures 4 and 5 can cause immediate (although sometimes temporary) changes in performance characteristics.

At low temperatures flexible cables tend to become stiff. For this reason, it is advisable to provide mechanical support for coaxial cable delay lines. Coaxial designs tend to be relatively immune to the high end of the operating temperature range (-55 to +125 °C.)

Properly dressed and mounted coaxial cables are usually insensitive to the effects of shock and vibration. However, improperly installed cables that undergo shock and vibration can generate spurious signals and sustain mechanical fatigue.

11.2.7 Reliability considerations. Because cable splicing is inevitable due to limited cable selection and demands for longer delay times, it is advisable to have the cable manufacturer make the assembly. If some adjustment is necessary on the part of the user, a portion of the outer layer of the cable assembly can be left unbound or unpotted for trimming after delivery.

After a coaxial delay line has been assembled, the number of reliability screening options is limited. Radiographic inspection is useful only in locating gross mechanical defects in the assembly. For electrical integrity, radiographic inspection is usually not an advisable screen.

The electrical integrity of a coaxial cable delay line can be further assured by the use of time domain reflectometry (TDR). The TDR screening process injects a pulse down the line under inspection and the resultant type and magnitude of any reflections can be recorded and examined. This is especially important when the cable has been spliced or trimmed.

Final trimming of the cable should not be made until after the assembly has been subjected to several cycles of thermal shock. Thermal shock has a tendency to promote dimensional changes in the cable. These dimensional changes can alter electrical performance characteristics.

11.3 DELAY LINES, DISTRIBUTED CONSTANT

11.3 Distributed constant.

11.3.1 Introduction. Distributed constant (dc) delay lines were developed to provide larger induced signal delay times than are obtainable from coaxial delay lines. The delay lines are either inductive designs with incidental capacitance or capacitive designs with incidental inductance. The capacitance and inductance arise from magnetic and electric fields set up in the structures and, therefore, the effect is distributed over a considerable part of the structure of the device.

11.3.2 Usual applications. DC delay lines can be utilized for a conventional delay function. In addition, they can be used to invert the polarity of a pulse so the output of a delay line may be shorted and the inverted pulse will emerge from the input terminals after a period equal to twice the delay of the line.

Another application is to derive a train of pulses from a single pulse with a line having several taps. A single input pulse will appear, in turn, at each successive tap. The pulses are then summed to form a binary word. It should be noted that each tap will reflect some energy back to the input and likewise deduct some energy from the single traveling pulse. There has to be some compromise between the amount of tap loading and characteristics of the output waveform.

11.3.3 Physical construction. The most common form of the distributed delay line is a coil wound on an insulating mandril with conductive longitudinal stripes (Figure 7). The longitudinal stripes are joined with a broken circle at one end to provide the ground connection. The circle is broken to keep the ground connection from looking like a shorted turn in the magnetic field of the coil.

The above method of construction can be reversed. The coil can be placed on the inside and wound on a magnetic core (Figure 8). The ground plane of the capacitor is a sheath of insulated wires wrapped around the outside of the coil on a relatively long pitch. In some configurations only every other wire is insulated and the alternate wires are bare.

The advantage of this form of construction is that higher characteristic impedances can be obtained. The disadvantages are that external electrical field problems tend to increase and the device is a harder to protect from a hostile electromagnetic environment.

DC delay lines can also be made in the form of a modified capacitor (Figure 9). The inductor is formed by winding a copper conductor on a continuous strip of magnetically permeable material. The distributed capacitance is achieved by placing conductive foil strips between two layers of dielectric sheathing. This insulated conductive screen (ground plane) provides one plate of the distributed

11.3 DELAY LINES, DISTRIBUTED CONSTANT

capacitor whereas the individual turns of the inductor strip form the other. When combined with the inductor the ground plane also provides shielding between successive turns of the spiral when wound.

Variable dc delay lines are also available. A common design uses a slider that runs along a coil wire from which the insulation has been removed. A second slider electrically connected to the first transfers the signal from the tap connections to a rail (Figure 10).

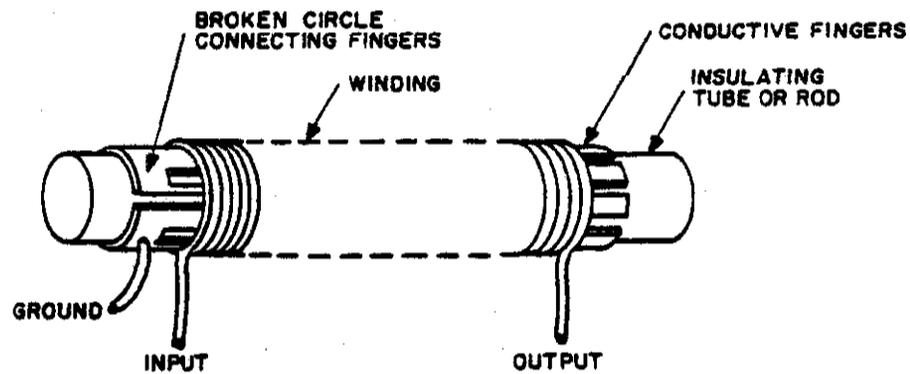


FIGURE 7. Distributed constant delay line mandril construction.

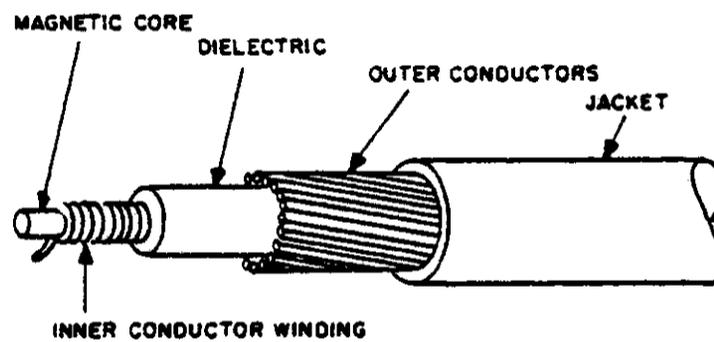


FIGURE 8. Distributed constant delay line construction.

11.3 DELAY LINES, DISTRIBUTED CONSTANT

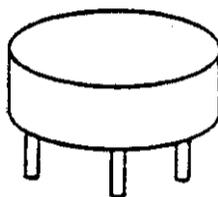
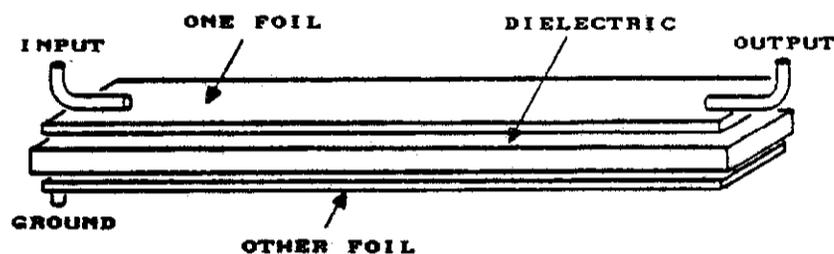


FIGURE 9. Distributed constant delay line cylindrical construction.

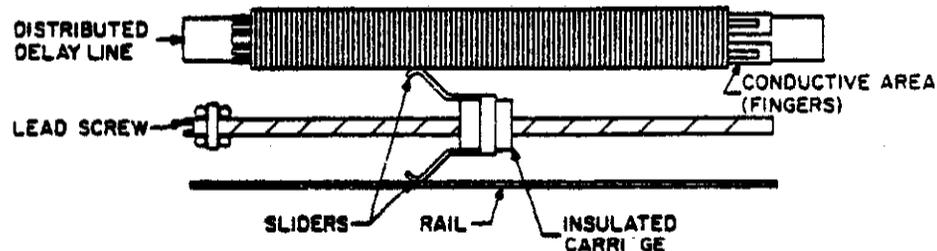


FIGURE 10. Variable distributed constant delay line.

11.3.4 Military designation. The applicable military specification for distributed constant delay lines is MIL-D-23859, Delay Line, Electromagnetic and Fixed. There are no delay lines included in MIL-STD-975.

11.3.5 Electrical characteristics. Distributed delay lines are generally limited to modest delays, lower T_d/T_r ratios, higher attenuations, and higher temperature coefficients than coaxial delay lines as well as low operating voltages (100 V or less) and a modest range of impedances. Low impedances require high capacitance-low inductances and high impedances require very low capacitance-high inductance. Both cases are difficult to obtain in a distributed delay line design.

11.3 DELAY LINES, DISTRIBUTED CONSTANT

Distributed delay lines are usually more likely to generate external fields than lumped constant delay lines. Distributed delay lines also tend to be more susceptible to external fields created by other components. Care is required in specifying, positioning, and shielding the distributed line. The cable type delay line (Figure 8) is not shielded against magnetic fields and although it appears to be like a coaxial cable, it has very few of the coaxial cable's immunities to the electromagnetic environment.

11.3.6 Environmental considerations. The environmental considerations associated with distributed delay lines are:

- a. Electromagnetic environment
- b. Reduced barometric pressure
- c. Humidity
- d. Operational temperature range
- e. Shock and vibration.

Due to inherent design configurations of the cable type distributed delay line, susceptibility to the electromagnetic environment is a possibility. This interaction can be inductive leakage to, or pickup from, other portions of adjacent circuitry or stray capacitance.

Altitude also affects delay line performance. Certain materials (electrical and packaging) have the tendency to outgas at reduced barometric pressure. This outgassing can eventually lead to the degradation of the part which can be physical or electrical in nature or a combination of both.

Properly packaged delay lines are relatively immune to the effects of humidity. However, potted or encapsulated devices can develop cracks and fissures if the materials used are incorrect for the application or if the correct materials are applied incorrectly by the vendor.

Because dc delay lines are assemblies of many components and their interconnections, the operational and storage temperature range is an important consideration. The inductance and capacitance values can change as a function of temperature. This change can range from 50 to 200 parts per million and may not be a linear function of temperature.

Packaging is also affected by temperature. Molded cases, which are formed with elevated temperature and pressure are immune to degradation over the temperature range of -55 to +125 °C. However, potted and encapsulated devices can experience degradation at elevated temperatures especially when coupled with reduced barometric pressure.

11.3 DELAY LINES, DISTRIBUTED CONSTANT

The mandril and capacitor designs of distributed delay lines are relatively immune to the effects of shock and vibration when properly packaged and mounted. Because vibration can introduce spurious signals into the line and cause the cable to undergo mechanical fatigue, cable applications require careful dressing and mounting.

11.3.7 Reliability considerations. After a dc delay line has been assembled, the number of reliability screening options is limited. Radiographic inspection is usually not an advisable screen. Unless the internal construction of the delay line consists of a few macroscopic component parts with simple interconnections, the resultant radiographs will only exhibit indiscernable clutter.

Burn-in under voltage is an inexpensive and relatively fast reliability screen. Thermal shock under voltage is another inexpensive and relatively fast reliability screen. Both tests should be preceded and followed by functional electrical testing.

11.4 DELAY LINES, LUMPED CONSTANT

11.4 Lumped constant.

11.4.1 Introduction. Distributed constant delay lines tend to be more space efficient, less environmentally susceptible, offer a wider range of impedances, and modest signal delay.

Lumped constant (LC) delay lines using discrete inducting and capacitors, were developed to provide the electrical advantages of both coaxial and distributed constant delay lines.

The accuracy of the approximation to coaxial or distributed constant delay lines depends on the length of the equivalent line increment. As the length of the line modeled becomes shorter, the approximation becomes closer if the resistance and parasitic reactance of the inductors and capacitors are low.

Where the length of the coaxial line is a small fraction of the wavelength, the increment can be characterized by two resistors, one inductor, and one capacitor. If the line has low loss, the resistors can be ignored entirely. The line increment can then be equated to a series inductor (L) and a shunt capacitor (C). The impedance (Z) of the line increment can be expressed as:

$$Z = \sqrt{L/C}$$

and time delay (T_d) can be expressed as:

$$T_d = \sqrt{LC}$$

Note that both equations are independent of frequency. This is the basic advantage for using a coaxial cable to achieve a delay. Therefore, the LC delay line which is an approximation of the coaxial design with the augmented inductances and capacitances of the distributed design exhibits longer signal delays, but is limited by frequency. However, the LC delay line is more space efficient, can have a variety of impedances, and exhibits large delays due to relatively larger inductive and capacitive elements.

11.4.2 Usual applications. LC delay lines can be used in essentially the same applications as distributed constant delay lines. One such application is pulse inversion. LC devices can be used to invert the polarity of a pulse. The output of a line can be shorted and the reflected (and inverted) pulse will emerge from the input terminals of the delay line after a period equal to twice the delay of the line.

11.4 DELAY LINES, LUMPED CONSTANT

It is also feasible to derive a train of pulses from a single pulse with a line having several taps. A single input pulse will appear, in turn, at each successive tap. The summed pulses form a binary word. It should be noted that each tap will reflect some energy back to the input (and likewise deduct some energy from the single traveling pulse) so there must be some compromise between the amount of tap loading and the characteristics of the output waveform.

11.4.3 Physical construction. LC delay lines consist of discrete multilayer capacitors and core inductors in a ladder configuration. Capacitors are chosen that exhibit negative temperature coefficients and inductors are selected that exhibit a positive temperature coefficient. The resulting cancellation yields a device which is relatively stable over the temperature range of -55 to +125 °C.

Digital integrated circuits that use transistor-to-transistor logic (TTL) or emitter coupled logic (ECL) can also be incorporated into LC delay line design. The delay line assembly compensates for the inherent propagation delay of the integrated circuit (IC) and enables the delay line to process digital information.

Delay line packaging consists of three types: molding, potting, and encapsulation. Consideration should be given to outgassing characteristics. Molding utilizes high pressure and elevated temperature resins to package the assembly. Potting usually consists of housing the assembly in a mold. A resin is then introduced which completely surrounds the assembly. When the resin has cured the mold is removed. Encapsulation involves the dipping of the assembly in a high viscosity thixotropic resin to give a conformal coating of 10 to 50 mils on the surface. Dual in-line packages (DIPs) and single in-line packages (SIPs) are most common. Their popularity arises from their ability to be machine inserted onto printed circuit boards (Figure 11).

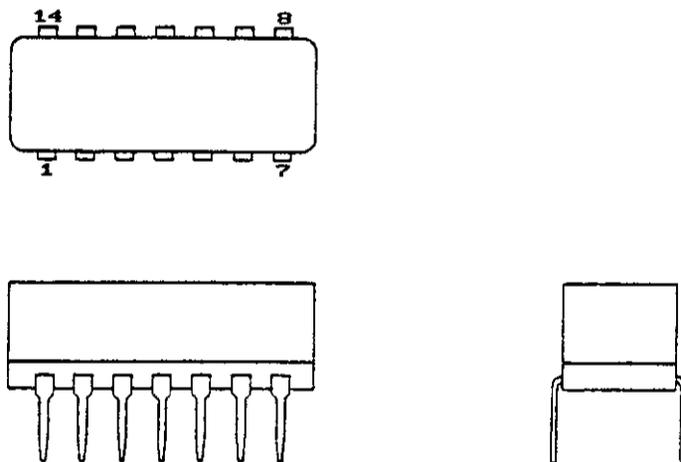


FIGURE 11. Lumped-constant delay line dual in-line package.

11.4 DELAY LINES, LUMPED CONSTANT

Variable LC delay lines are also available. In Figure 12A taps are at the position of shunt capacitors to achieve delay variation in incremental steps and because an external shunt capacitance does not degrade electrical performance as much as when the tap is placed in the coil.

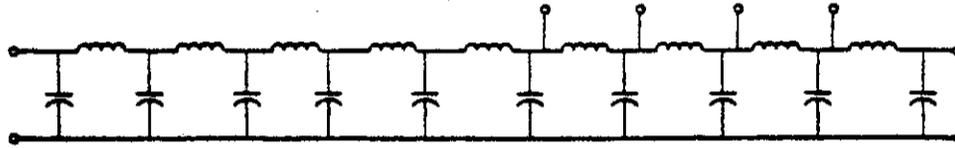
A signal impressed at the input terminals will proceed to the loaded tap where a part of it will leave the line, part of it will be transmitted, and part will be reflected back to the input. The reflected signal will modify the signal appearance at previous taps. The reflection occurs whether power is withdrawn or only stored and then discharged by a pure reactance.

Figure 12B, shows an example where only the last few sections of a line are tapped in order to trim the delay line length to a desired value. Figure 12C shows several lines (usually within a single enclosure) designed to be electrically connected in cascade. A number of the mismatch and reflection problems are avoided, although a delay line is not necessarily a good termination for another delay line.

Figure 12D shows a way to avoid the effects of tap loading by using the tap as the input point. The two disadvantages to this configuration are: First that the insertion loss is a minimum of 3 dB and the input impedance is 0.5 of the line impedance; the second is that the reflections will still occur if several intermediate taps are used simultaneously.

Electronically variable versions of LC delay lines are also available. Figure 13A shows the simplest form wherein a change in a direct current voltage bias causes a change in the capacitance of semiconductor varactor diodes. A disadvantage of this type of line over that made of passive components is that the signal can only be a small part of the bias level if the signal is not to be affected. Figure 13B shows an electronically variable LC delay line using both semiconductor varactor diodes (as capacitors) and saturable reactors. It is possible to devise a control system that will give variable delay while keeping the ratio L/C (impedance) constant. This design also requires a relatively small signal to avoid distortion.

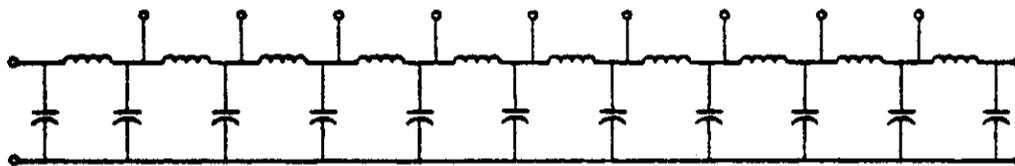
11.4 DELAY LINES, LUMPED CONSTANT.



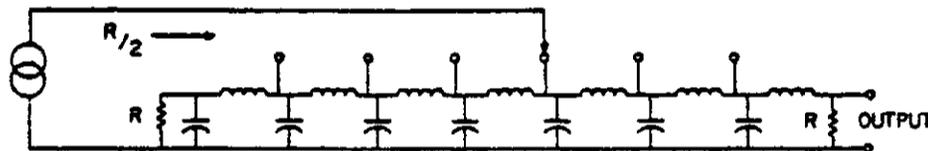
A. Lumped-constant line with trimming tap increments



B. Segmented lumped-constant line



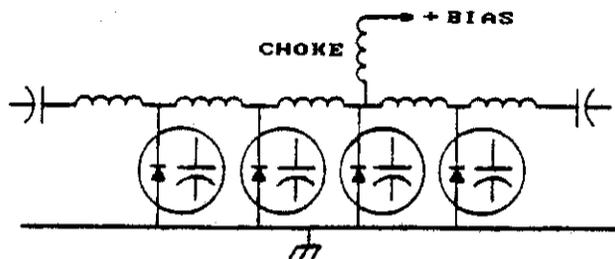
C. Lumped-constant line with equal tap increments



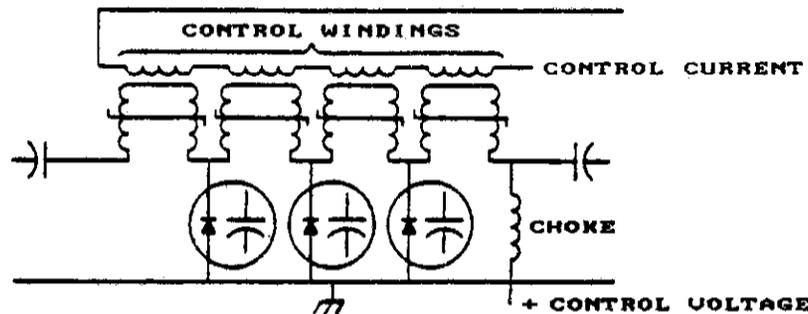
D. A tap used as input to avoid reflections

FIGURE 12. Variable lumped constant delay line configurations.

11.4 DELAY LINES, LUMPED CONSTANT



A. Nonconstant-impedance electronically-variable line



B. Constant-impedance electronically-variable line

FIGURE 13. Electronically variable lumped-constant delay line.

11.4.4 Military designation. The applicable military specification for lumped constant delay lines is MIL-D-23859. There are no delay lines included in MIL-STD-975.

11.4.5 Electrical characteristics. All delay lines will degrade the response of the input pulses because these devices function as low-pass filters that attenuate the higher frequencies. This is especially true with the relatively high inductances and capacitances encountered in lumped delay lines.

When observing the response of the delay line to an input signal, the input waveform rise time, the oscilloscope rise time, and the output rise time degradation (due to the delay line) must be considered. If the delay line rise time characteristics are to be referred to the ideal rectangular pulse, the following relation applies.

$$T_{r(ob)} = \left[T_{r(i)}^2 + T_{r(osc)}^2 + T_{r(o)}^2 \right]^{0.5}$$

11.4 DELAY LINES, LUMPED CONSTANT.

where

$T_r(ob)$ = observed output rise time

$T_r(i)$ = input pulse rise time

$T_r(osc)$ = rise time of the oscilloscope

$T_r(o)$ = output pulse rise time.

In video applications, the LC delay line is usually used in group delay. It is expected that the output pulse will be a faithful reproduction of the input, and usually a band of frequencies is involved. Pulse fidelity is obtained

when

$$\frac{\theta}{F} = K$$

where

θ = radians

F = radians/second.

A similar condition applies to passband delay for faithful reproduction of a waveform envelope within the passband,

$$\frac{d\theta}{dF} = K$$

In each of these cases (zero dispersion), not only is the relation of $\theta/F = K$ required, but the attenuation must be constant also. Attenuation arises from two factors: those that are internal to the delay line and those external to the delay line. Internal factors are primarily the dc resistance of the inductors, losses in the dielectric of the capacitors, and losses in the ground planes. External factors consist of termination mismatch and tap loading.

11.4.6 Environmental considerations. The environmental considerations associated with LC delay lines are:

- a. Electromagnetic environment
- b. Reduced barometric pressure

11.4 DELAY LINES, LUMPED CONSTANT

- c. Humidity
- d. Operational temperature range
- e. Radiation hardness.

Passive and digital LC delay lines are usually furnished in molded, potted, or encapsulated packages. For this reason, the device is susceptible to various types of interaction with the electromagnetic environment. This interaction can be inductive leakage to, or pickup from, other portions of adjacent circuitry, stray capacitance from a nearby chassis or shield, or waveform changes from the effect of a hand used to hold the delay line in a test jig.

Altitude also affects delay line performance. Certain materials (electrical and packaging) have the tendency to outgas at reduced barometric pressure. This outgassing can eventually lead to the degradation of the part. The degradation can be either physical or electrical.

Properly molded, potted, or encapsulated delay lines are relatively immune to the effects of humidity. However, molded, potted, or encapsulated devices can develop cracks and fissures if the materials used are incorrect for the application or if the correct materials are applied incorrectly by the vendor. Of particular concern is the lead egress area of the package. Incorrectly applied potting or encapsulation can separate from the lead material creating an entry point for moisture.

Because LC delay lines are assemblies of many components and their interconnections, the operational and storage temperature ranges are an important consideration. The value of inductors and capacitors will change as a function of temperature. This change can range from 50 to 200 parts per million and may not be a linear function of temperature.

Discrete passive LC delay lines are relatively insensitive to nuclear radiation effects. Semiconductors used in adjacent circuitry will usually fail before a discrete passive delay line. However, digital delay lines use ICs that reside in the same package as the passive components. If radiation hardness is called for, radiation-hardened versions of ICs are available.

11.4.7 Reliability considerations. After an LC delay line has been packaged, the number of reliability screening options decreases. For this reason pre-assembly screening of the lumped components (inductors and capacitors) is advisable. In digital LC designs, preassembly screening of the IC used is also recommended.

11.4 DELAY LINES, LUMPED CONSTANT

After a design has been assembled and packaged, only certain screening tests are advisable. For example, radiographic inspection is usually not a conclusive screen. Unless the internal construction of the delay line consists of a few macroscopic component parts with simple interconnections, the resultant radiographs would only exhibit indiscernable clutter.

Burn-in and thermal shock under voltage are inexpensive and relatively fast reliability screens. Both of these screen tests should be preceded and followed by functional electrical testing.

Where digital delay lines are utilized, it is advisable that radiation hardened ICs with high mean time between failures (MTBFs) be used.

11.5 DELAY LINES, QUARTZ-GLASS SONIC

11.5 Quartz-glass sonic.

11.5.1 Introduction. Delay in electrical circuits does not have to be confined to electromagnetic delay phenomena. One technology utilizes the transformation of an electrical signal to a sonic signal. The sound wave then passes through a mechanical medium and the resultant delayed signal is transformed back into an electrical signal.

It should be noted that the use of the term "sonic" is a misnomer. In actuality the signal is ultrasonic and can extend into the gigahertz region. Ultrasonic frequencies are utilized because of the greater information handling capabilities of that portion of the spectrum. This section is concerned with the utilization of fused quartz or glass as a bulk medium to achieve delay.

11.5.2 Usual applications. The applications of quartz-glass delay lines can be divided into analog information processing and digital information processing categories.

Analog information can be processed in the form of a modulated carrier centered in the passband of a nondispersive line. One such application is in moving target indicator radars where the stored result of one sweep is subtracted from the result of a subsequent sweep. Only those targets not appearing in the same position of the subsequent sweep are displayed by the radar.

Another application is passing a narrow signal envelope through a quartz-glass delay line so that it emerges as a longer (and lower in amplitude) frequency modulated (fm) pulse. In a radar system, this pulse can be amplified and transmitted at relatively high average power with the reduced possibility of insulation and parts overload. This wider fm pulse is received and inverted by passage through a delay line and the received pulse is narrowed once again. This permits the display of, and discrimination between, many small targets that might otherwise be undetectable or indistinguishable.

A common digital application of quartz-glass delay lines is as memory devices. The delay line is used as a storage device in which the input signal is injected into the medium, extracted at the output side of the delay line, and then externally rerouted to the input side multiple times. This signal recirculation can be achieved with or without the benefit of external pulse shaping circuitry to preserve its integrity.

Signal recirculation can also be accomplished internally to the delay line itself. A signal can be injected into the input side of the delay medium, reach the output side, and be reflected back to the input side for another pass through the medium. This is referred to as an echo line. The echo line is very often used to create a train of pulses differing in amplitude from the previous pulse by a few decibels. Usually the signal is inserted into the echo line by a single burst of a few cycles of the nominal operating frequency and the input transducer is used as a detector to ascertain the number of pass throughs.

11.5 DELAY LINES, QUARTZ-GLASS SONIC

11.5.3 Physical construction. Material for the delay medium of this class of device falls into two categories: amorphous and single crystal. The material called amorphous may not be truly amorphous but actually polycrystalline. However, because the crystals are usually very small and of random orientation they can be regarded as amorphous for practical purposes. The single crystal delay media are usually quartz and the amorphous media are fused quartz or "delay line glass."

Single crystal material has the advantage of low attenuation. However, single crystal media as a category have different propagation velocities and temperature coefficients depending on orientation. The orientation of the crystal (right handed or left handed) will also give rise to performance variations.

Amorphous material tends to be less critical than single crystal material of direction of propagation but does exhibit more attenuation. Moreover, various latent material faults and variations may make a quartz-glass blank undesirable in some applications. Adverse effects are noted in lines designed to furnish the greatest delay as these lines require the greatest and most complex paths.

Regardless of the medium, transducer positioning is critical to delay line function. The transducer is attached to the surface of the delay medium by means of a bonding agent (Figures 14 and 15). After the transducer is attached, the bonding agent is applied to the rear of the transducer and a backing is attached to it. The transducer, the bonding agents, the backing, and the delay medium form an acoustical system that provides the efficient transfer of power to and from the delay medium over a broad bandwidth.

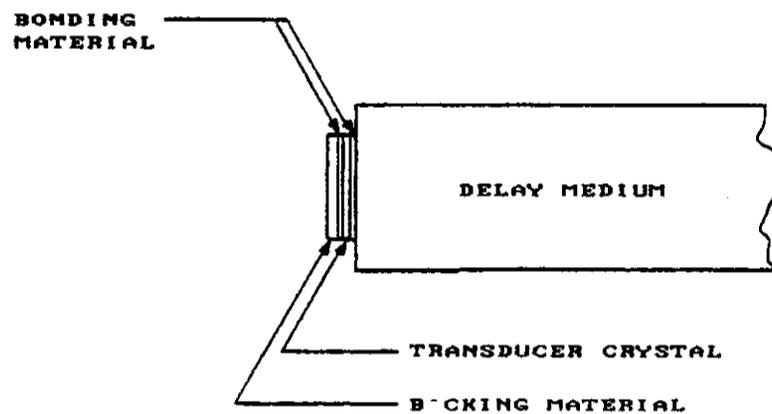


FIGURE 14. Transducer attachment.

11.5 DELAY LINES, QUARTZ-GLASS SONIC

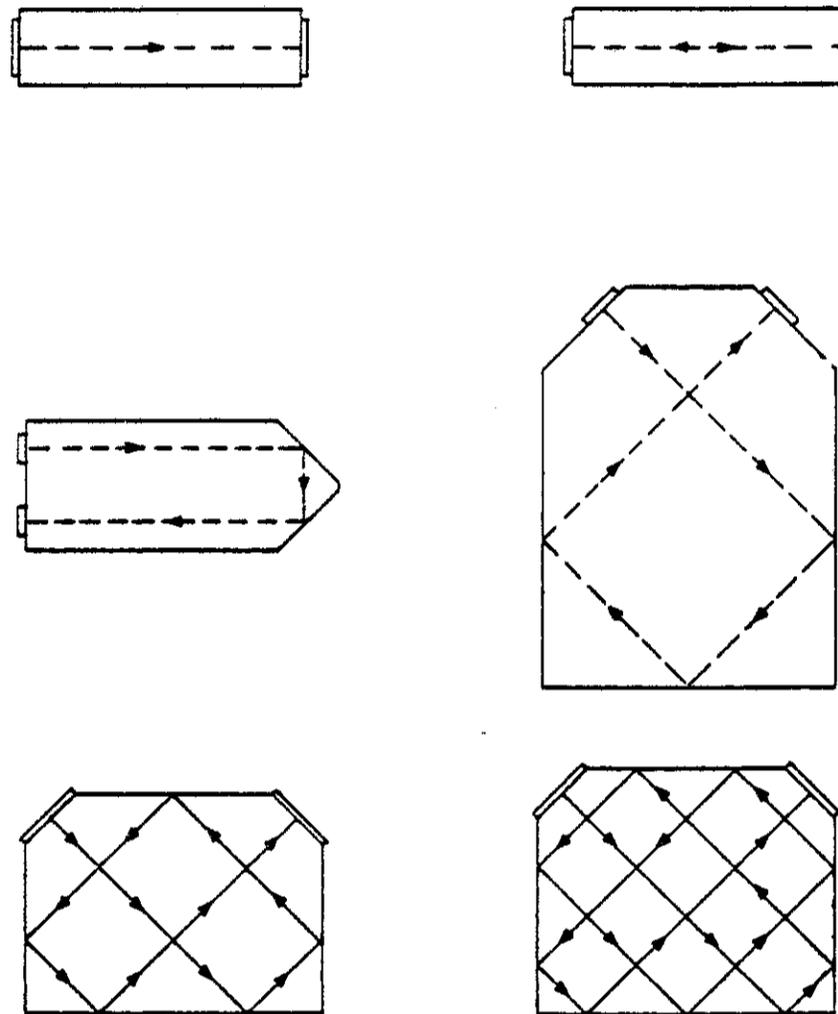


FIGURE 15. Basic quartz and glass delay line construction.

Variable quartz delay lines are also available (Figure 16). The moving delay medium can be adjusted either manually or by external control circuitry. This configuration permits variation of delay up into the microsecond region. To a lesser extent, a fixed delay line can also be used to obtain a variable delay. Variation of delay is achieved by the relative movement between an optical pickup system and the quartz/glass delay medium (Figure 17).

11.5 DELAY LINES, QUARTZ-GLASS SONIC

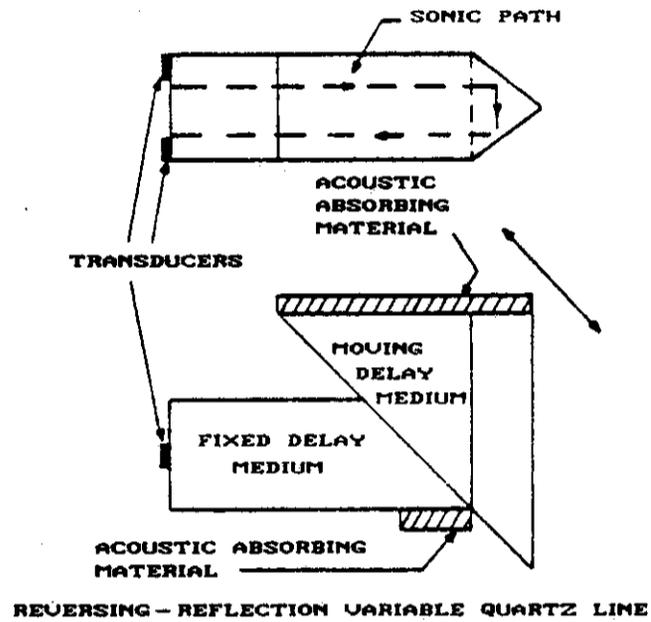


FIGURE 16. Variable quartz/glass delay line.

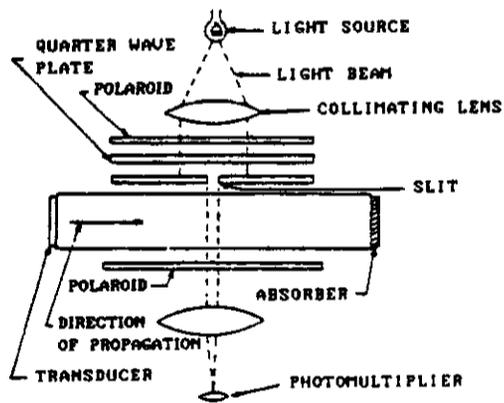


FIGURE 17. Variable quartz/glass delay line with optical pickup.

11.5 DELAY LINES, QUARTZ-GLASS SONIC

11.5.4 Military designation. There are no directly applicable military specifications or standards for quartz/glass delay lines. Some relevant military documents are:

MIL-C-3098	Crystal Units, Quartz, General Specification for
MIL-F-18327	Filters; High Pass, Low Pass, Band Pass, Band Suppression, and Dual Functioning, General Specification for.

There are no delay lines included on MIL-STD-975.

11.5.5 Electrical characteristics. Three techniques in making glass-quartz media are used to improve electrical performance by minimizing the effects of spurious signals. One uses wide apertures to minimize the generation of spurious lobes in the reflection pattern. Another uses large reflection angles which cause spurious lobes to strike subsequent facets at less than optimal angles. The third uses long propagation paths in high attenuation media. Such long paths expose spurious lobes to greater attenuation as they traverse the delay medium.

11.5.6 Environmental considerations. The environmental considerations associated with quartz-glass delay lines as a whole are:

- a. Electromagnetic environment
- b. Reduced barometric pressure
- c. Humidity
- d. Operational temperature range
- e. Shock and vibration
- f. Radiation hardness.

Virtually all of the environmental considerations associated with quartz-glass delay lines are concerned with the transducer components rather than the delay medium used in an application. The two exceptions to this are temperature and nuclear radiation.

Quartz-glass transducers can exhibit sensitivity to the electromagnetic environment, reduced barometric pressure, humidity, temperature excursions, shock-vibration and nuclear radiation. However, transducer sensitivity to the environment is no greater than that of associated electrical and electronic components.

Temperature coefficient problems can be avoided by using an oven to stabilize the delay medium. The high temperature inertia of a large blank and the general requirements for high stability temperature control mean that critical lines are usually within two or more nested ovens. The outer oven provides the rougher temperature control and the inner oven, the finer temperature control.

11.5 DELAY LINES, QUARTZ-GLASS SONIC

Amorphous delay media are relatively insensitive to nuclear bombardment. The transducer used in the application will most likely fail first. Single crystal delay media will usually exhibit a much greater susceptibility to nuclear effects. It should be noted that these are generalizations and that specific media must be evaluated for specific behavior in this kind of environment.

11.5.7 Reliability considerations. Once a quartz-glass delay line has been packaged, the number of reliability screening options is limited. Radiographic inspection is usually not an effective screen. Unless the internal construction of the delay lines consists of a few macroscopic component parts with simple interconnections, the resultant radiographs will only exhibit indiscernable clutter.

Burn-in and thermal shock under voltage are inexpensive and relatively fast reliability screens. Thermal shock is especially important as it provides a means of ensuring the integrity of the transducer attachment. Both of these screen tests should be preceded and followed by functional electrical testing.

Vibration is also a recommended reliability screen. By their very nature, these devices are physically massive and their fabrication is labor intensive. Vibration reveals the mechanical integrity of the basic design and the transducer attachment.

11.6 DELAY LINES, METAL SONIC

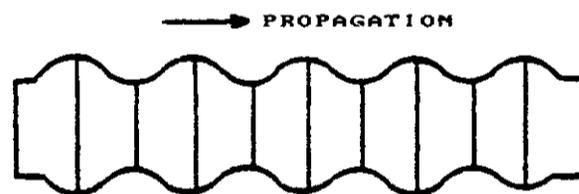
11.6 Metal sonic.

11.6.1 Introduction. The delay of electrical signals may be accomplished by propagation of an electromagnetic wave through a dielectric medium or by translating the electrical wave to some other form (i.e., sonic waves) where it can pass through an appropriate delaying medium and be retranslated back to electrical signals after the delay. This section deals with use of metals as sonic delay media.

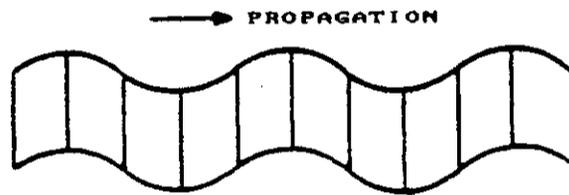
The most common propagation modes in metals are longitudinal and shear (Figure 18). These modes can be excited either with direct electrical to mechanical transducers or by magnetostriction. Nickel is a particularly good magnetostrictive material exhibiting longitudinal and shear distortion when subjected to a longitudinal magnetic field.

11.6.2 Usual applications. The most common application for a metal sonic delay line is for digital memories which can be used as buffer memories for teletype, process control, and numerically controlled machines. They can also be used as main memories for special purpose computers, main frame computers, differential analyzers, and shift registers.

A number of the applications of metal sonic delay lines have been superseded by surface acoustic wave delay lines operating in higher frequency ranges. While such lines are obviously attractive for equipment operating in these ranges (such as the intermediate or signal frequencies of radar), data processing equipment can likewise benefit from the wider bandwidths available in metal sonic delay lines.



A. Compression wave



B. Metal sonic delay line propagation modes

FIGURE 18. Metal sonic delay line propagation modes.

11.6 DELAY LINES, METAL SONIC

11.6.3 Physical construction. The simplest forms of metal sonic delay lines consist of the delay medium and a transducer at each end of the sonic path (Figure 19). Absorptive material is used at the ends of the line to prevent unwanted reflections of the sonic wave.

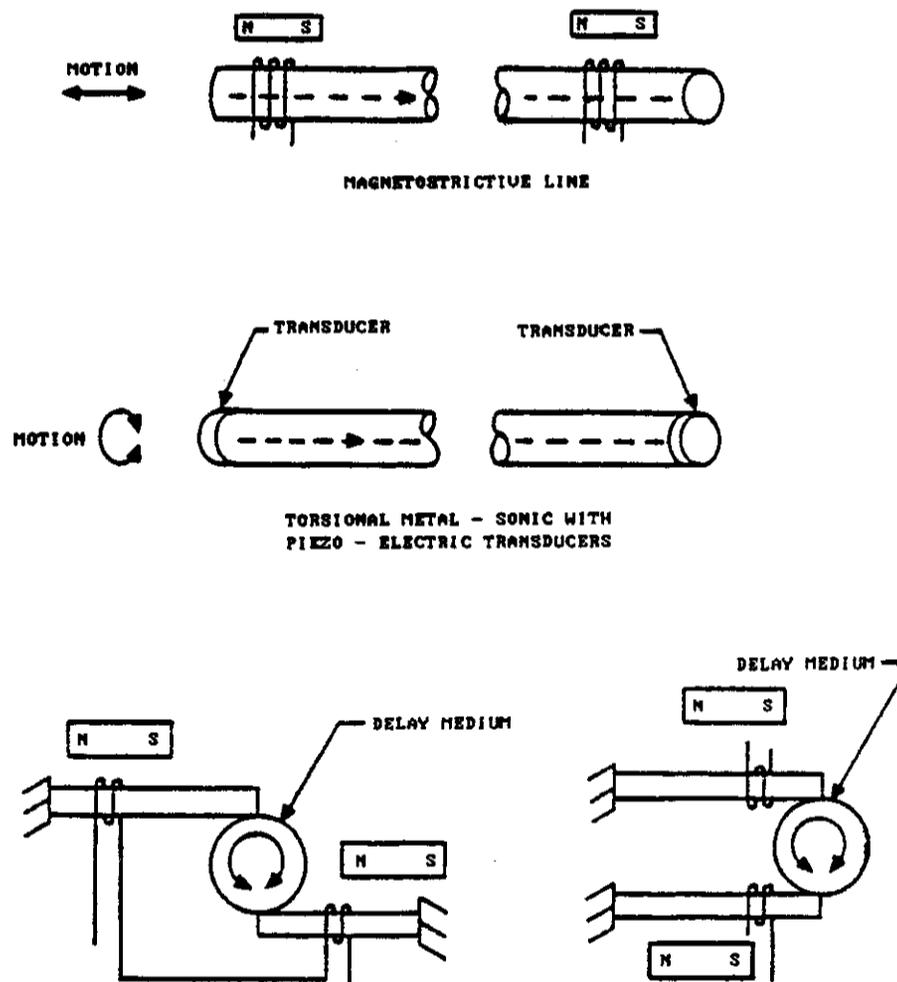
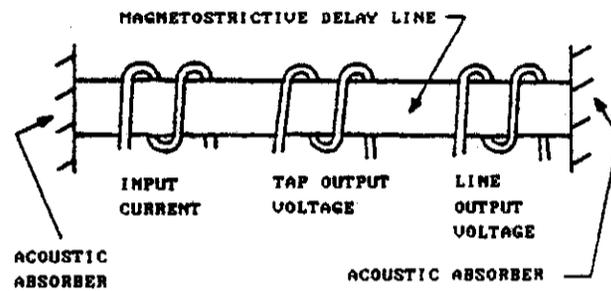


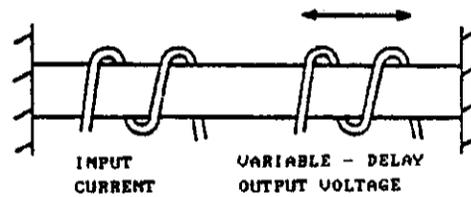
FIGURE 19. Metal sonic delay line magnetostrictive transducers.

An attractive feature of metal sonic delay lines with magnetostrictive transducers is that it is possible to change the position of one or both transducers without major difficulty and thus adjust the length of the line by as much as several microseconds. In order to achieve a variable delay, metal sonic devices can be tapped by the addition of a magnetostrictive transducer (Figure 20).

11.6 DELAY LINES, METAL SONIC



A. Tapped magnetostrictive delay line



B. Variable magnetostrictive delay line

FIGURE 20. Variable metal sonic delay line with magnetostrictive tap.

11.6.4 Military designation. There are no directly applicable military specifications or standards.

11.6.5 Electrical characteristics. The nature of the disturbance propagated through metal delay media and the nature of the transducers used cause considerable alteration in the form of the transmitted signal. For this reason metal sonic delay lines are more suitable to the processing of digital rather than analog information because digital signals lend themselves more to reshaping after the delay phenomenon has been accomplished.

There is also considerable attenuation in metal media devices, making pre- and postamplification desirable. Combining amplifiers and signal shaping circuits in the sealed delay line package make unity gain delay lines possible.

11.6 DELAY LINES, METAL SONIC

The longer lines, relying upon sonic propagation due to a torsional stress, pose special problems. A considerable length of wire is required to achieve useful delays. For reasons of space efficiency, the wire is coiled. However, this configuration, although more space efficient, leads to performance problems. In the coiled form, the delay medium must be mechanically supported and these support points can contribute both attenuation and reflection.

The temperature coefficient of the transducer material is also important because it controls the mode of operation of the delay line. There are three typical modes of operation, and a delay line is usually designed for just one of the three. These modes of operation are: the return to zero (RZ or RTZ), the non-return to zero (NRZ), and the bipolar mode. In typical applications, the NRZ can store twice the information (number of bits) of the RZ mode. The advantage of the RZ mode is that it permits the simplest device specification. The advantage of bipolar operation is greater immunity to noise.

11.6.6 Environmental considerations. The environmental considerations associated with all sonic delay lines are:

- a. Electromagnetic environment
- b. Reduced barometric pressure
- c. Humidity
- d. Operational temperature range
- e. Shock and vibration
- f. Radiation hardness.

Virtually all of the environmental considerations associated with metal media delay lines are concerned with the transducer components rather than the delay medium used in an application. The exceptions to this are temperature and nuclear radiation.

Metal media delay line transducers can cause, in the electromagnetic environment, reduced barometric pressure, humidity, temperature excursions, shock-vibration, and nuclear radiation. However, transducer sensitivity to the environment is no greater than that of associated electrical and electronic components.

By their very nature metal media exhibit different physical and electrical characteristics as a function of temperature. For this reason, it is advisable this type of delay line be operated only in a moderate thermal environment.

Metal delay media are relatively insensitive to nuclear bombardment. The transducer used in the application and adjacent electronic components will most likely fail before significant delay medium deterioration will occur.

11.6 DELAY LINES, METAL SONIC

11.6.7 Reliability considerations. After a metal media delay line has been packaged, the number of reliability screening options is limited. Radiographic inspection is typically not an effective screen. Unless the internal construction of the delay line consists of a few macroscopic component parts with simple interconnections, the resultant radiographs will only exhibit indiscernable clutter.

Burn-in and thermal shock under voltage are inexpensive and relatively fast reliability screens. Thermal shock is especially important as it provides a means of ensuring the integrity of the transducer. Both of these screen tests should be preceded and followed by functional electrical testing.

Vibration is also a recommended reliability screen. By their very nature, metal media devices are physically massive and labor intensive. Vibration reveals the mechanical integrity of the basic design and the physical integrity of the transducer.

11.7 DELAY LINES, SURFACE ACOUSTIC WAVE

11.7 Surface acoustic wave.

11.7.1 Introduction. The delay of an electrical signal need not occur only in a dielectric medium. Electrical signal delay can also be achieved by translating an electrical wave into a sonic wave, passing the sonic wave through a bulk sonic medium, and then translating the delayed wave back into an electrical signal. This procedure is attractive because the velocity of sound in most bulk media is much less than that available in electrical circuitry. This sonic signal transmission through a bulk medium is by a compressed longitudinal wave (Figure 21A) or a shear transverse wave (Figure 21B).

The surface acoustic wave (SAW) has both longitudinal and transverse components. However, the majority of the energy concentrated in SAW is within one acoustic wavelength along the surface of the bulk medium. The energy decreases exponentially normal to the surface of the medium (Figure 21C).

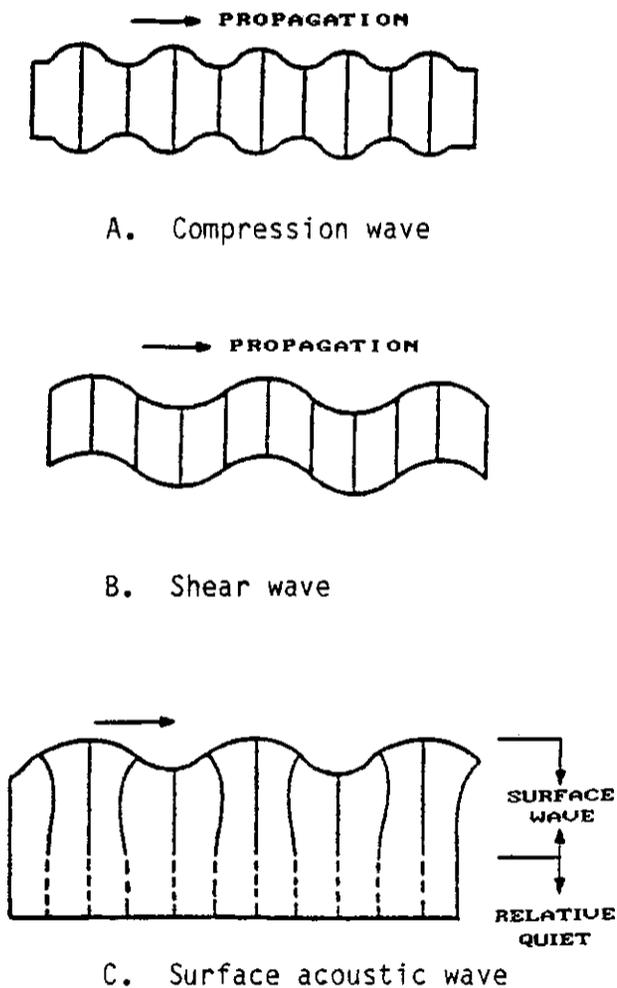


FIGURE 21. Sonic propagation modes in solids.

11.7 DELAY LINES, SURFACE ACOUSTIC WAVE

Many sonic bulk delay lines are made from amorphous materials such as quartz/glass (see subsection 11.4) and metal (see subsection 11.5). However, there exists the problem of exciting the sonic wave in such substances. The problem is greatly minimized if a piezoelectric crystal is used as the bulk material.

A pair of electrical conductors laid on the surface of a piezoelectric crystal will cause the crystal to distort mechanically when an electrical potential energizes the conductor pair. If several parallel pairs of conductors are placed on the surface and simultaneously energized, the combined distortion will create a surface acoustic wave at particular frequencies.

A similar pair (or pairs) of conductors placed on the surface in the path of this wave will be electrically excited when the wave passes under them. The efficiency of electrical coupling increases as a function of the contact between the conductors and the crystal because the crystals are essentially insulators; the conductors may be deposited directly onto the crystal surface. This is done in the same manner as deposition of conductive paths on integrated circuit substrates.

The velocity of an SAW is on the order of 10^{-5} that of an electrical wave in free space. As wavelength equals velocity-frequency, conductor arrays of the size of several wavelengths are practical at desirable frequencies. Proper design of such an array permits practically all of the electrical signal to excite the SAW mode and very little of the energy excites bulk modes.

11.7.2 Usual applications. Dispersive SAW devices are used in chirp radars. In this application, a narrow high intensity pulse may be stretched to a wide linear fm pulse suited to peak limited amplifiers. This results in an increase in effective radar power.

After reflection from the radar target and reception at the receiver, the wide pulse may be passed through a matching filter and be compressed into a narrow pulse. The resulting narrow pulse is used in the radar to discriminate between two or more targets at the same range.

Use has also been made of the fact that as the SAW travels along the delay medium that there is an accompanying electric field. With an amplifier on the surface to detect and amplify the field as it travels, it becomes possible for the signal to emerge as large or larger than it was at the input.

When the delay medium surface is coated with photosensitive resistive material, it will act as an electrical load. In response to an optical image focused on the surface of the resistive material, this load can be made variable. Choice of input waveform and electrode geometry will then allow the processing of the optical information. Direct optical information can be obtained from the traveling SAW by reflection, or when the crystal is transparent by variation in transmission caused by dimensional distortion accompanying the SAW.

11.7 DELAY LINES, SURFACE ACOUSTIC WAVE

11.7.3 Physical construction. The simplest form of the SAW line consists of two transducers and a piezoelectric delay medium. The practical way to achieve this is to deposit metallic fingers on the surface of the crystal to act as the electrical portion of the transducers.

The length, width, shape, and material of the transducer fingers all affect the production and pickup of the SAW and the creation of spurious reflections within the delay medium. Soft metals such as gold or aluminum are preferred for transducer materials.

Whereas the narrowband versions of the SAW delay line may have evenly spaced fingers, the wider band devices will vary both finger spacing and finger width (Figure 22). In addition, the structures may vary the finger overlap, weighting, or apodizing the response. The impulse response of the SAW line will approximate the appearance of an apodized line because the amplitude of a wave generated by a finger pair is controlled by the amount of overlap of the fingers.

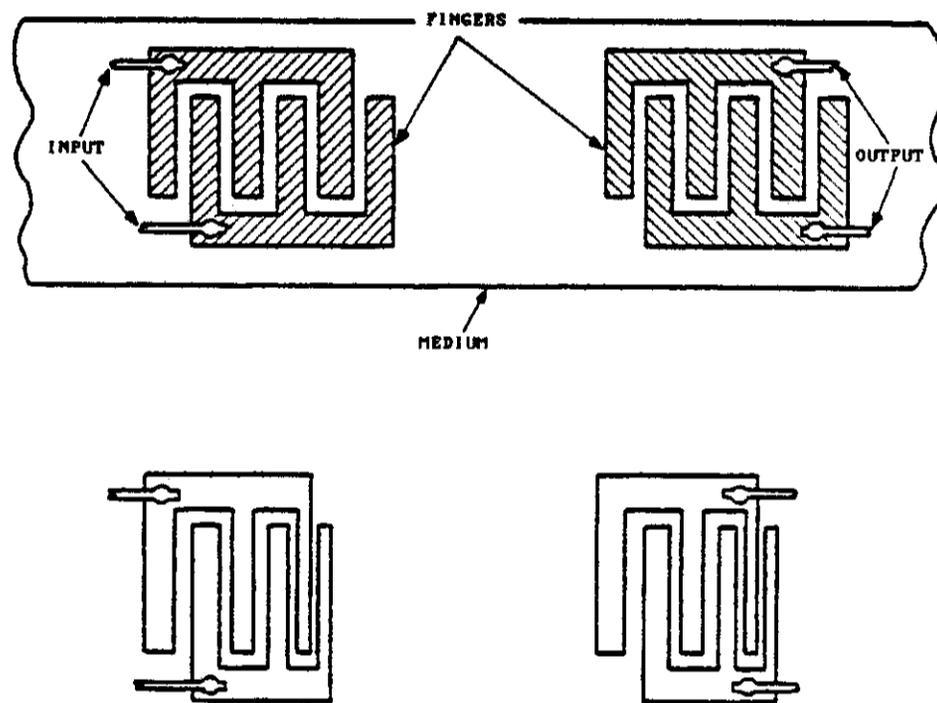


FIGURE 22. SAW delay line narrow and wide band transducer arrays.

11.7 DELAY LINES, SURFACE ACOUSTIC WAVE

Variation in the amount of overlap will cause a variation in the amount and distribution of acoustic reflection of a wave passing under the fingers. To minimize this variation, it is often useful to fill this space with dummy fingers of the same width as the missing section; these fingers are not connected to either electrical conductor.

The SAW generated by a finger array is bidirectional with approximately one-half of the energy moving in each direction. That portion of the energy moving in the undesired direction tends to reflect from the end of the medium and acts as an inverted signal. A common way to minimize this undesired reflection is to pad that end of the medium with an acoustic absorber.

Similarly, SAW energy not absorbed by the pickup transducer, will tend to reflect off the opposite end of the medium and reach the pick up transducer a second time if that end is not also padded. Note that it is possible to achieve a good match of the electrical signal into the medium by choice of physical configuration of the fingers and simple electrical circuitry such as a loading coil. The resulting goodness of the match is dependent on the bandwidth.

Choice of the delay medium depends on a number of factors. Bismuth germanium oxide has an attractively low acoustic velocity, lithium niobate has a high coupling constant, and ST quartz has a low temperature coefficient over temperature. Lithium niobate has low loss characteristics but stable operation requires its use in temperature controlled ovens. Lithium niobate also has some tendency to fracture from mechanical shock. Where higher insertion loss is permissible, ST quartz is a good choice for space and other critical environments.

Conductor pairs at intermediate positions along the delay line medium may be used to sample the SAW as it passes. This is effectively a tap and a large number of such taps may be used simultaneously. The amount of energy so withdrawn at a tap will lessen the signal's energy and may in extreme cases cause reflections back toward the origin of the signal. However, taps cause fewer problems on low coupling factor material (e.g., ST quartz) than high coupling factor material (e.g., lithium niobate).

11.7.4 Military designation. There are no applicable military specifications or standards for surface acoustic wave delay lines.

11.7.5 Electrical characteristics. SAW devices afford long delays with minimal distortion provided the signal is an alternating current wave whose frequencies are well within the passband. This includes a modulated pulse or pulse train if the sideband frequencies are also within the passband. However, compared with other types of sonic delay lines the SAW passband is narrow.

The requirements for this passband region are that the delay line phase response increases linearly with frequency and that the loss be constant as a function of frequency. The degree by which the line departs from these criteria determines the degree of distortion of the desired wave.

11.7 DELAY LINES, SURFACE ACOUSTIC WAVE

SAW devices have three undesirable electrical characteristics which are: direct feed through spurious, crosstalk, and other spurious. Direct feed through spurious occurs when electrical coupling takes place without a signal having traversed the delay medium. Crosstalk occurs when one or more delay medium is positioned in close proximity to another and their respective transducers pickup each other's signals. Other spurious pertains to any other types of undesirable characteristics such as reflections from the ends of the delay medium.

SAW devices usually have lower losses than other acoustic devices. However, insertion losses within the passband can be as great as 50 or 60 dB. As a result, the user needs to decide which frequency or frequencies for measurement are needed for the application. If a simple constant delay with no waveform change is needed, measurement of the center frequency insertion loss and group delay over the desired bandwidth is usually sufficient.

11.7.6 Environmental considerations. The environmental considerations associated with SAW delay lines as a whole are:

- a. Electromagnetic environment
- b. Reduced barometric pressure
- c. Humidity
- d. Operational temperature range
- e. Shock and vibration
- f. Radiation hardness.

Virtually all of the environmental considerations associated with SAW delay lines are concerned with the transducer components rather than the delay medium used in an application. The exceptions to this are temperature and nuclear radiation.

SAW delay line transducers can exhibit sensitivity to the electromagnetic environment, reduced barometric pressure, humidity, temperature excursions, shock-vibration and nuclear radiation. However, transducer sensitivity to the environment is no greater than that of associated electrical and electronic components.

SAW delay media are relatively insensitive to nuclear bombardment. The transducer utilized in the application and adjacent electronic components will most likely fail before significant delay medium deterioration will occur.

11.7.7 Reliability considerations. After a SAW delay line has been packaged, the number of reliability screening options is limited. Radiographic inspection is usually not an effective screen. Unless the internal construction of the delay line consists of a few macroscopic component parts with simple interconnections, the resultant radiographs will only exhibit indiscernable clutter.

Burn-in and thermal shock under voltage are inexpensive and relatively fast reliability screens. Thermal shock is especially important as it provides a means of ensuring the integrity of the transducer. Both of these screen tests should be preceded and followed by functional electrical testing.

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11.7 DELAY LINES, SURFACE ACOUSTIC WAVE

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12.1 MOTORS, GENERAL

12. MOTORS

12.1 General.

12.1.1 Introduction. This section discusses motors used for NASA applications. These devices are not included in MIL-STD-975.

Selection of the proper motor for a given application is extremely important. A motor that is too small can result in overheating and premature failure whereas a motor that is too large can waste power and space. In addition, selection of the wrong type of motor can cause excessive power consumption or poor response.

This section is restricted to the selection and application of conventional ac and dc motors.

12.1.1.1 Applicable military specifications. Motors are covered by the following military specifications:

<u>Mil Spec</u>	<u>Title</u>
MIL-M-7969	Motors, Alternating Current, 400 cycle, 115/200 Volt System, Aircraft, General Specification
MIL-M-8609	Motors, Direct Current, 28 Volt System, Aircraft, General Specification
MS 33543	Criteria - Temperature and Altitude Range Self Cooled Electric Equipment

12.1.2 General definitions.

Accelerating torque. The torque exerted by a motor during the entire acceleration time. A speed-torque curve is required to find this torque.

Breakaway torque. The torque that a motor is required to develop to break away its load from rest to rotation.

Full-load current. The current which the motor draws when the rated voltage and frequency are applied and when the motor is operated at rated power output (full load torque and speed).

Full-load torque. The torque developed by a motor during rated power output when the rated voltage and frequency are applied.

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12.1 MOTORS, GENERAL

Locked-rotor current. The current that the motor draws when rated voltage and frequency are applied and the rotor does not rotate. Fuses, circuit breakers, and wire must be selected to accommodate this initial surge.

Locked-rotor torque. The minimum torque that a motor develops for all angular positions of its rotor when the rated voltage and frequency are applied and the rotor does not rotate.

Pull-in-torque of a synchronous motor. The maximum constant torque under which the motor will pull its connected inertia load into synchronism at rated voltage and frequency.

Pull-out torque of a synchronous motor. The maximum sustained torque which the motor will develop at synchronous speed with rated voltage and frequency applied.

Pull-up torque of a synchronous motor. The minimum torque developed by the motor during the period of acceleration from rest to the speed at which break-down torque occurs with rated voltage applied at rated frequency.

Starting torque. The minimum torque developed at zero rotation with rated voltage and frequency applied. This is normally expressed as a percentage of full-load torque.

Synchronous torque of a synchronous motor. The torque developed after voltage is applied. This is the total steady-state torque available to drive the load.

12.1.3 General device characteristics. An electric motor has a stator (which has a laminated steel core and insulated copper wire coils), a rotor, or armature, and a frame to position the stator and rotor. The rotor is attached to the shaft, and the rotor and shaft assemblies rotate on bearings. In an ac motor the rotating element is the rotor, and in a dc motor the rotating element is the armature.

A further classification to ac and dc motors may be made according to their construction and method of operation.

A dc motor is classified by the type of motor field. This includes permanent magnet, series-wound, shunt-wound, and compound wound designs. This motor is characterized by large changes in speed and current with changing loads.

An ac motor is classified as either induction or synchronous. Induction motors operate at speeds below synchronous speed but are still relatively constant speed devices. The torque developed in induction motors is the result of relative motion between an induced rotating field in the rotor and a rotating electrical field in the stator. Synchronous motors differ in that they operate at a constant speed which is synchronized with the speed of the rotating field in the stator. Synchronous motors supply torque when their speed is locked into the synchronous speed of the windings.

12.1 MOTORS, GENERAL

A dc motor is used in applications which do not require speed regulation or the fine control that is typical of a servo system. Shunt-wound dc motors are reversible during rotation and have good starting torque. However, the dc motor creates radio frequency interference due to arcing at the commutator. The dust contributed by the brush in normal wearout can cause failure of the bearings and voltage leakage paths.

An ac motor generally operates at constant speed and has moderate to high starting torques, but the direction of rotation cannot be reversed without stopping. The ac motor is simpler and has a more reliable design than the dc motor because it has no rotating windings, brushes, or commutator.

12.1.3.1 Motor enclosures. The correct enclosure should be selected for the type of motor to be used. Most motors used in NASA environments are totally enclosed nonventilated or explosion-proof. Following are definitions of the major types of motor enclosures.

Open motors. An open motor is one which has ventilating openings to permit passage of external cooling air over and around the windings.

Drip proof motors. Drip-proof motors are open motors in which the ventilating openings are constructed so that successful operation is not affected if drops of liquid or solid particles strike or enter the enclosure at any angle up to 15 degrees from the vertical.

Splash-proof motors. Splash proof motors are open motors in which the ventilating openings are constructed so that successful operation is not affected if drops of liquid or solid particles strike or enter the enclosure at any angle up to 100 degrees from the vertical.

Totally enclosed motors. These motors are enclosed to prevent free exchange of air between the inside and outside of the case. However, the enclosure might not be airtight.

Totally enclosed nonventilated (TENV) motors. TENV motors are not equipped for cooling by means external to the enclosure.

Totally-enclosed fan-cooled (TEFC) motors. TEFC motors are cooled by a fan attached to the motor outside of the enclosure.

Explosion-proof motors. Explosion-proof motors are totally enclosed motors with enclosures designed to withstand an explosion of a gas within them and to prevent ignition of the external environment by sparks, flashes, or explosions within the motor housing.

12.1 MOTORS, GENERAL12.1.4 General parameter information.

12.1.4.1 Speed. The synchronous speeds of a motor can be calculated using the following formula:

$$\text{Synchronous speed} = \frac{120 \times \text{line frequency}}{\text{number of poles}}$$

The most commonly used speeds are given in Table I.

TABLE I. Commonly used synchronous speeds

No. of Poles	400 Hz	60 Hz
2	24,000 rpm	3,600 rpm
4	12,000 rpm	1,800 rpm
6	8,000 rpm	1,200 rpm
8	6,000 rpm	900 rpm
10	4,800 rpm	720 rpm
12	4,000 rpm	600 rpm

Induction motors operate at less than a synchronous speed, depending on the torque they are delivering and the shape of their speed-torque curve. The difference between synchronous speed and actual speed is the "slip" of the motor. This is usually expressed as a percentage of synchronous speed and is calculated using the following formula:

$$\text{Percent slip} = \frac{(\text{synchronous speed} - \text{full load speed}) \times 100\%}{\text{synchronous speed}}$$

As an example, if a 2-pole, 400-Hz motor with a synchronous speed of 24,000 rpm ran at 20,000 rpm, the slip would be 4,000 rpm or 16.7 percent.

The amount of slippage varies with the horsepower rating, number of phases, starting torque, and other factors. The induction motor speeds most commonly used are given in Table II.

TABLE II. Commonly used induction speeds.

No. of Poles	400 Hz	60 Hz
2	22,500 - 20,000 rpm	3,250 - 3,000 rpm
4	11,200 - 10,000 rpm	1,675 - 1,550 rpm
6	7,500 - 7,000 rpm	1,100 - 1,050 rpm
8	5,500 - 5,300 rpm	825 - 800 rpm
10	4,400 - 4,200 rpm	680 - 650 rpm
12	3,700 - 3,500 rpm	530 - 500 rpm

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12.1 MOTORS, GENERAL

12.1.4.2 Full-load torque. To properly select a motor, it is necessary to know the continuous operating torque required and the speed at which the load must be driven. If a gear head is applied to a motor, it is necessary to know the gear ratio desired in the gear head and the efficiency of the gear head, so that the speed and torque output for the motor can be calculated.

12.1.4.3 Locked-rotor torque. Locked-rotor torque is the minimum torque a motor develops at zero rotation when the rated voltage and frequency are applied. This torque is generally expressed in percent of full-load torque. Thus, a motor rated at 1/10 hp, 10,500 rpm, and 340 percent locked-rotor torque develops a 9.6 ounce-inch torque minimum at 10,500 rpm and 32.6 ounce-inch torque at zero rpm.

The locked-rotor torque must be known to select the correct motor. A high locked-rotor torque motor generally has a lower full-load speed than a motor of the same rating with a lower locked-rotor torque. Therefore, a low-slip motor (high full-load speed) is obtained by having relatively low locked-rotor torque.

Very high locked-rotor torque requirements require a larger motor than is ordinarily used.

A typical motor performance curve is shown in Figure 1.

12.1.4.4 Developed power. The developed power may be calculated from either of the two formulas given below:

$$\begin{aligned} \text{Horsepower (hp)} &= \frac{\text{torque (inch-ounce)} \times \text{rpm}}{1,008,000} \\ &= \text{torque (inch-ounce)} \times \text{rpm} \times 10^{-6} \end{aligned}$$

$$\begin{aligned} \text{Watts out} &= \frac{746 \times \text{torque (inch-ounce)} \times \text{rpm}}{1,008,000} \\ &= 746 \times \text{torque (inch-ounce)} \times \text{rpm} \times 10^{-6} \end{aligned}$$

12.1 MOTORS, GENERAL

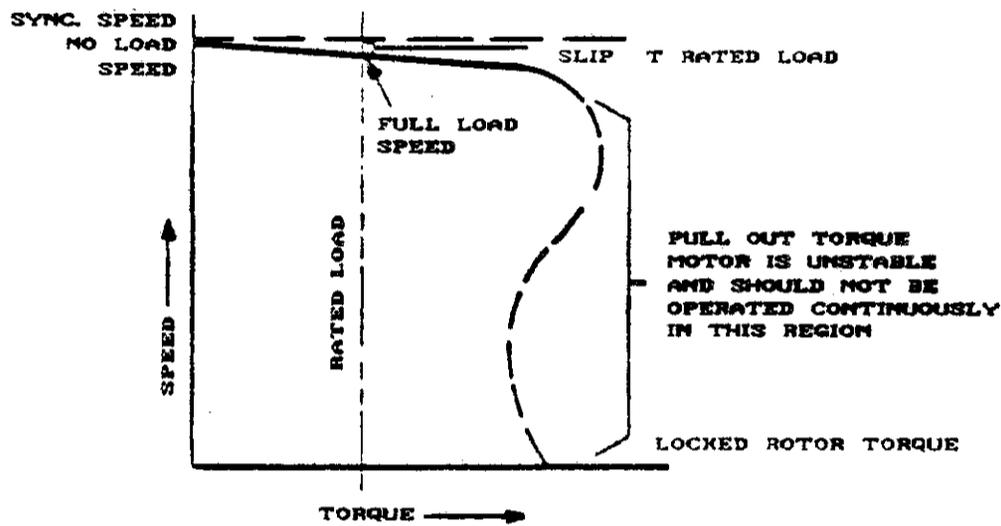


FIGURE 1. Typical motor performance curve.

12.1.4.5 Torque. The torque (T) may be calculated from the formula below.

$$T = \frac{hp \times 1,008,000}{rpm}$$

$$= \frac{hp}{rpm} \times 10^6$$

12.1.4.6 Volt ampere input. The volt-ampere input may be found as follows:

- for a dc motor = $E \times I$
- for a single-phase ac motor = $E \times I$
- for a three-phase ac motor = $\sqrt{3} \times E \times I$

where E and I are the line voltage and current, respectively

12.1.4.7 Power input. The power input may be found as follows:

- for a dc motor = $E \times I$
- for a single-phase ac motor = $E \times I \times \text{power factor}$
- for a three-phase ac motor = $\sqrt{3} \times E \times I \times \text{power factor}$

12.1 MOTORS, GENERAL

12.1.4.8 Efficiency. The efficiency (Eff) may be calculated from the following formula:

$$\text{Eff} = \frac{\text{Developed power}}{\text{Power in}} \times 100\%$$

12.1.4.9 Required input current. The required input current may be found as follows:

$$\text{for dc motors:} \quad I = \frac{\text{hp} \times 746}{E \times \text{Eff}}$$

$$\text{for single-phase ac motors:} \quad I = \frac{\text{hp} \times 746}{E \times \text{Eff} \times \text{power factor}}$$

$$\text{for three-phase ac motors:} \quad I = \frac{\text{hp} \times 746}{\sqrt{3} \times E \times \text{Eff} \times \text{power factor}}$$

where E is the line voltage.

12.1.4.10 Power factor. For ac motors, the power factor (PF) is:

$$\text{PF} = \frac{\text{watts in}}{\text{volt amperes in}}$$

12.1.4.11 Rise time. Another factor in some applications is the time required to accelerate a given load to a given speed. The time (T), in seconds, for a motor with an inertia (I_M) and supplying a torque (L) to accelerate an inertial load (I_L) to a speed (N) can be found from the expression:

$$T = 0.0162 (I_M + I_L) (N/L)$$

where

I_M is in inch-ounce² and may be obtained from published data or calculated from the formula:

$$I_M = 0.5 m r^2 \text{ where } m \text{ is the rotor's weight in ounces, } r \text{ is the radius in inches}$$

I_L is the inertia of the load in inch-ounce²

N is the speed in revolutions per second, and L is the torque supplied by the motor in inch-ounces

When a hysteresis synchronous motor is used, the starting torque may be substituted for L. For other motors, the value of the average torque must be used for L.

12.1 MOTORS, GENERAL

12.1.4.12 Duty cycle. Motors are designed to operate in various duty cycles, depending upon their application. These duty cycles include:

Continuous duty. Operating at approximately a constant load for an indefinite period. If load is on for more than 15 minutes, this is considered continuous.

Intermittent duty. Intermittent duty includes the following:

- a. Load and power-off periods
- b. Load and no-load periods
- c. Load, no-load, and power-off periods.

To select the correct motor for a given application, it is necessary to have a correct definition of this duty cycle including the time and ambient temperature for each phase of the duty cycle.

12.1.5 General guides and charts. An important consideration in selecting a motor is the configuration of its speed torque curve. The following are the typical speed torque curves of the different types of motors discussed later.

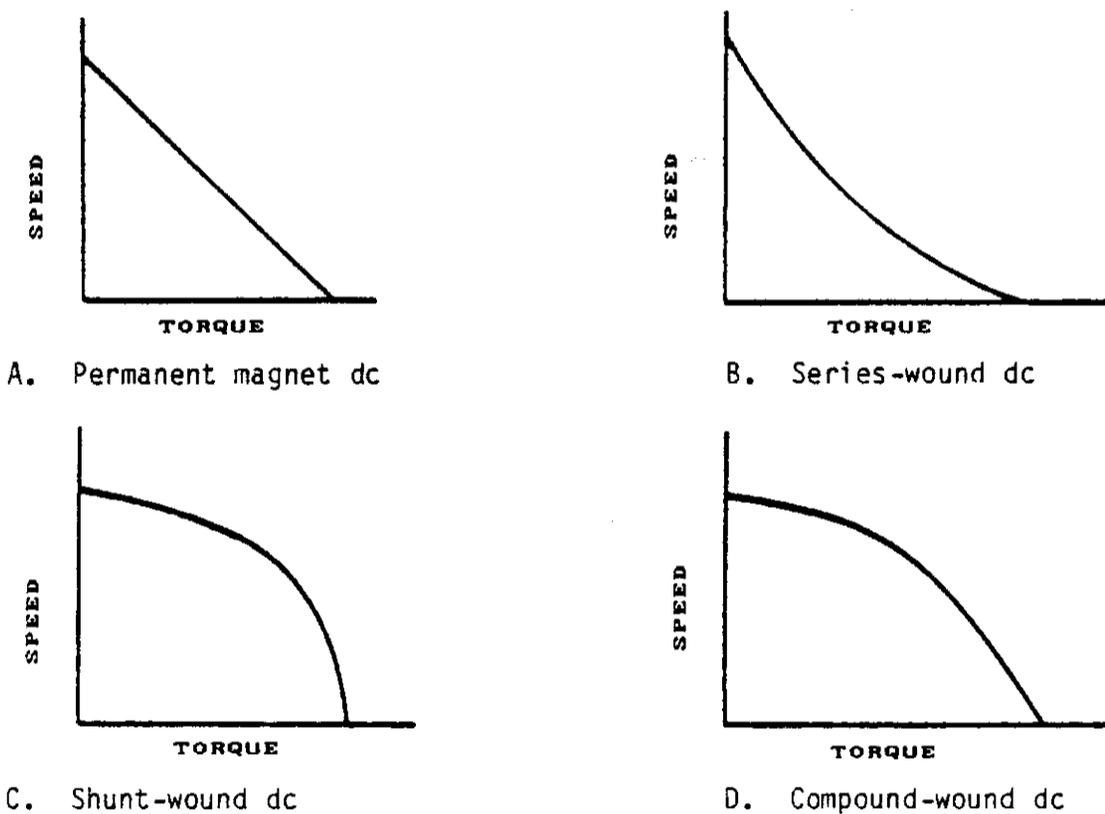
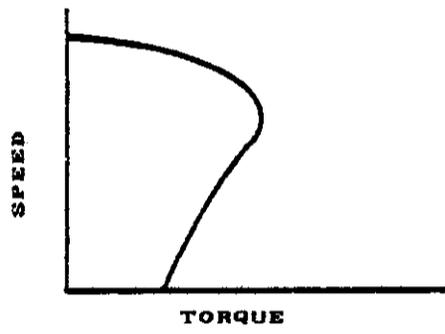
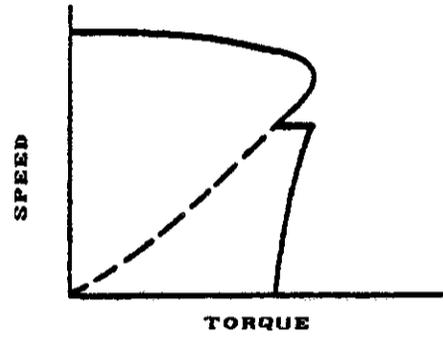


FIGURE 2. Typical speed torque curves.

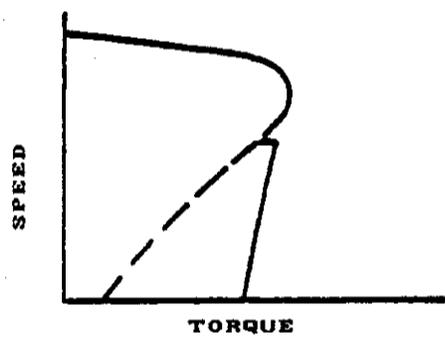
12.1 MOTORS, GENERAL



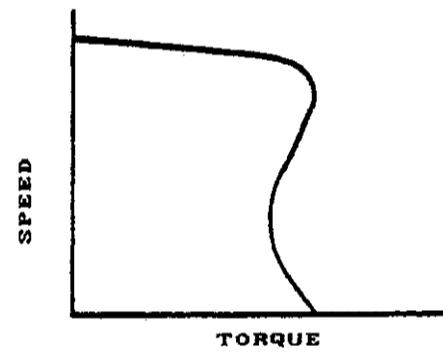
E. Permanent split capacitor, ac



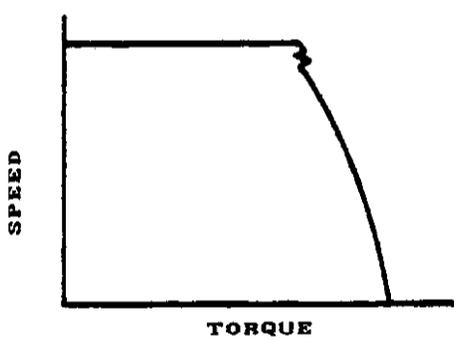
F. Capacitor-start, ac



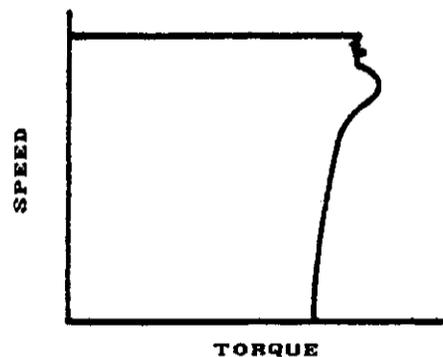
G. Capacitor start-capacitor run, ac



H. Three-phase induction, ac



I. Reluctance ac



J. Hysteresis permanent magnet induction ac

FIGURE 2. Typical speed torque curves (Continued).

12.1 MOTORS, GENERAL

12.1.6 General reliability considerations. The motor should be able to deliver the torque required for the application in the given environment. A motor that provides satisfactory performance at room temperature or at sea level may be inadequate at higher temperatures or altitudes. The overheating that results from selecting a unit that is not satisfactory for its duty cycle or environment leads to either catastrophic failure or shortened bearing and insulation life due to excessive winding temperatures.

When selecting motors to operate over wide temperature ranges, care should be taken to ensure that the bearing design is adequate for the application. This is particularly true if a unit capable of operating continuously at high temperatures must also operate at extreme cold temperatures.

In designing a motor in a system, care should be taken to ensure that the unit is mounted to a sufficiently large heat sink. Improper heat sinking can cause the same types of failure as overloading. The availability of cooling air will help minimize problems caused by overheating.

As with all electromechanical components, motors should be protected from extremely high voltage.

In operation, small decreases in voltage are actually more harmful than small increases. Decreased voltage reduces the torques available, reduces full-load speed, increases full-load currents, and causes increased winding temperatures. An increase in voltage will increase the magnetic noise from the motor.

In an ac motor, fluctuations in line frequency affect performance and reliability. The speed of an ac motor is proportional to the frequency. Starting torque and overload capacity are reduced by increased frequency, whereas winding temperatures are increased by reduced frequency.

For a three-phase motor, care should be taken to protect the motor from operating in single phase if one phase should be lost. Single-phase operation will cause the unit to operate at a continuous stall condition and overheat or, if the unit is rotating at the time, will cause it to be overloaded and develop excessive winding temperatures.

Since bearings or rotating parts are damaged by shock loads, care should be taken in handling units before they are installed in equipment. Bearings become brinelled by excessive shock loads resulting in noisy operation and premature bearing failures.

For a three-phase motor, the rated performance assumes a balanced power supply. The voltage and power factor, as determined for each lead or pair of leads, should be equal when the motor is delivering full load. The effects of unbalanced voltage are measurable in the motor current, speed, torque, and temperature rise. An unbalance of as little as 3.5 percent between highest phase voltage and the average of all three voltages in a three-phase motor will

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result in a temperature rise of about 25 percent above normal. Also, the current per phase resulting from a severe power unbalance makes it quite difficult to protect and control the motor.

Another factor in selecting a reliable motor is the trade-off between bearing life and motor size. Slower rotating motors have a longer bearing life and require less gear reduction. Higher speed motors develop more horsepower than a slower speed motor of the same size.

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12.2 MOTORS , AC

12.2 AC motors.

12.2.1 Introduction. The two major categories of ac motors are induction and synchronous. There are several types in each category.

12.2.1.1 Induction motors. The induction motor, which is the most widely used, is often referred to a "squirrel cage" motor because of the squirrel cage configuration of the electrical conductors when the steel core is removed from the rotor. There are no brushes or slip rings which need adjustment as a result of wear. The rotor consists of either a series of bars connected together by end rings or a single unit of cast aluminum. The only friction or wearing of parts occurs in the shaft support bearings. The induction motor runs at less than synchronous speed.

The induction motor is so named because current is induced in the rotor with no outside connection. This induced current reacts with the stator's field, causing rotor rotation. These motors are preferable to dc motors for constant speed work because the initial cost is less and the absence of slip rings or permanent magnets eliminates problems inherent in dc designs.

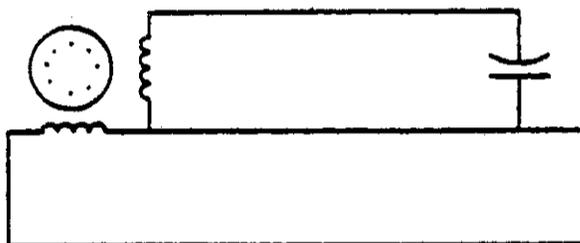
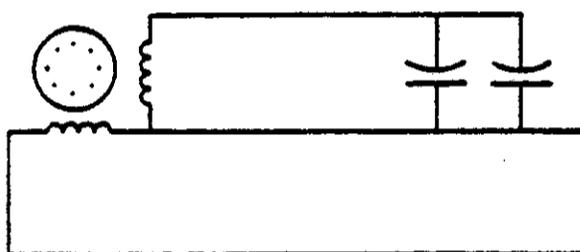
Single-phase and three-phase motors are the types most commonly used. Single-phase motors require some means of starting rotation. If single-phase power were supplied to a motor without an auxiliary starting winding, the motor would not run unless some external force were applied to start shaft rotation. Single-phase motors are generally referred to by designations which describe the starting method used. They exhibit more noise in operation and less efficiency than polyphase motors.

Induction motors may be subdivided into three types, they are discussed in the following paragraphs.

Permanent split-capacitor. Permanent split-capacitor units have a capacitor placed between one phase winding and the power supply. The other phase is connected directly across the power line as shown in Figure 3.

Motors of this type have a characteristically low locked-rotor torque, generally 30 percent of full-load torque (FLT). Permanent-split capacitor motor windings usually differ in capacitor phase and main phase to obtain the best possible performance and capacitor-size. This is the most common type of single-phase motor.

Capacitor start--capacitor run. To correct the starting torque limitation of the permanent-split capacitor motor, an external relay is used to put a second capacitor in parallel with the running capacitor during starting as shown in Figure 4. This second capacitor is then eliminated from the circuit as the motor attains a predetermined speed or current value. By this means, locked-rotor torques of 500 percent are obtainable. The major limitations of this system are the externally mounted relay and two capacitors, which also increase cost.

FIGURE 3. Permanent split-capacitor induction motor.FIGURE 4. Capacitor start-capacitor run induction motor.

Three-phase motors. Three-phase motors provide the most horsepower per weight and are highly stable units. Exceptionally high locked-rotor torques are obtained, with typical locked-rotor torques ranging from 250 to 300 percent. Higher locked rotor torques are readily obtained and extremely high locked-rotor torques are provided through proportionate increases in size and weight. Efficiency is higher than that for single-phase motors, but the power factor is less than that of a permanent-split capacitor type. The schematic for a typical three-phase motor is shown in Figure 5.

This section does not include a discussion of such types as wound rotor induction motors, repulsion-induction motors, repulsion-start induction motors, split-phase induction motors, or universal ac and dc motors, since these units are not often used in NASA applications.

12.2 MOTORS, AC

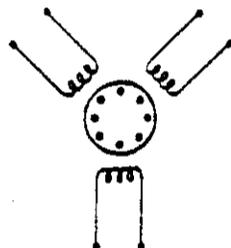


FIGURE 5. Three-phase induction motor.

12.2.1.2 Synchronous motors. In contrast to induction motors, synchronous motors rotate in exact step with line frequency. In large motors, this is accomplished by connecting the rotor to a dc supply. This provides a strong field in the rotor that locks in step with the rotating field of the stator.

Reluctance. Except for having flats on the periphery, reluctance synchronous (salient pole) motors have the same type of squirrel cage rotors as induction motors. The flats affect the reluctance (resistance to establishing a magnetic field) which causes synchronism. The poles tend to align with the flux field. Current alternating off and on continually pulls the poles around in exact step. Motors of this type approach synchronous speed as induction motors and, after effecting synchronism, run synchronously as reluctance motors.

In cases where the motor load has relatively high inertia, the reluctance type motor is limited. It may be incapable of attaining synchronism, since it runs as an induction motor until near-synchronous speeds. The torque exerted by the rotor pole must be sufficient to accelerate the rotor and the load into synchronism rapidly. This must happen in time for the rotor to rotate one-half pole pitch. Therefore, a reluctance motor may start a load that it cannot pull into synchronism and the motor will continue to run as an induction motor.

Considerable cogging effects in starting, together with greater than normal induction motor noise and vibration, are present because of salient pole construction.

The major advantage of the reluctance synchronous motor is that it will phase (synchronize at a fixed angular position). Therefore, a two-pole unit phases in two positions, 0 and 180 degrees apart, and a four-pole unit phases in four positions, 0, 90, 180, and 270 degrees apart. The output of a reluctance motor is only about 40 percent of the same size induction motor.

Hysteresis. If the discrete poles of a reluctance motor are eliminated, the rotor still becomes magnetized but the flux is not channeled as in the reluctance motor and, thus, it is not as efficient. The rotor's tendency to

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maintain a magnetized condition in one position is called hysteresis. The hysteresis rotor uses this principle to run as the induced rotor poles are pulled around by the alternating field.

The rotor of a hysteresis synchronous motor is constructed of a ring of magnetic materials such as 37 percent cobalt and alnico.

Because of its smooth rotor, the hysteresis motor is free of magnetic pulsation resulting from pole saliency or slots, and is much quieter and generates less vibration than other types of synchronous motors.

A constant synchronous speed is obtained, but a load angle is assumed that changes with variations of load or applied voltage. There are no phasing or locking points for this type motor.

The hysteresis synchronous motor, unlike the reluctance type, is capable of synchronizing high-inertia loads. Because there is no cogging, smooth and uniform starting conditions prevail.

Permanent magnet induction (synchronous). Permanent magnet induction motors are often referred to as polarized hysteresis motors and are used in applications where lock-in position is important. The rotors of these units may be either smooth or salient pole.

This type of motor provides the advantage of overcoming the limitations of the hysteresis type and improving the reluctance-type phasing properties. Permanent magnet synchronous motors lock-in at one-half the number of positions required by a reluctance motor. Lock-in of a two-pole unit occurs in one place, and a four-pole in two places. A disadvantage lies in the reduced horsepower output per frame size, as compared to hysteresis types.

12.2.2 Usual applications. Wherever constant speed motion is required, ac motors are used. Although this is normally angular motion, it can be linear by use of a special gearing.

These motors are used to drive rotating devices such as antennas, tape systems, fans, pumps, timers, and counters, and to operate sliding devices such as valves and doors.

12.2.3 Physical construction.

12.2.3.1 Typical size and ratings. The mechanical dimensions of a motor are dependent upon the ratings. A typical outline configuration is shown in Figure 6. The ratings for a given motor frame size depend upon rotational speed, ambient temperature, heat dissipation through the frame, and duty cycle and thus cannot be demonstrated here.

12.2 MOTORS, AC

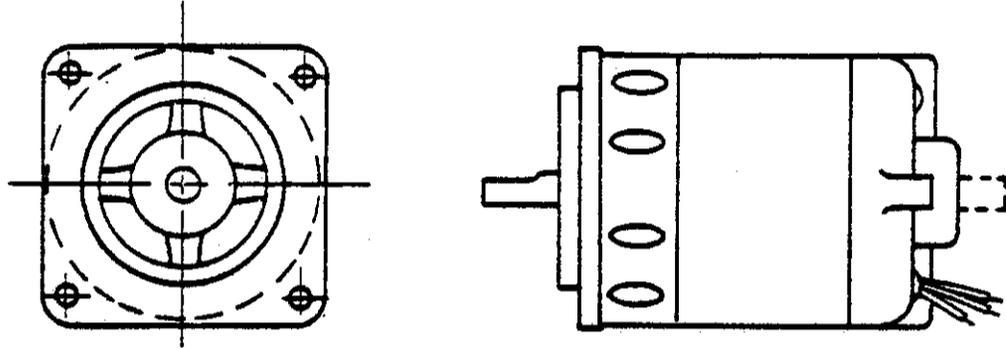


FIGURE 6. Typical outline of an ac motor.

12.2.4 Military designation. The ac motors used in aerospace electronics have not been standardized to the point where a military part designation is available.

However, motors are available that conform to the requirements of MIL-M-7969 for 400-Hz airborne applications.

12.2.5 Electrical characteristics. Most AC motors are designed to operate with 400-Hz, 115- or 200-V single- or three-phase power supply. However, because 400-Hz, 28-V power is also available in many applications, the same motors are available with stators wound for that voltage.

Electrical Current requirements generally vary according to the power requirements of the motor.

12.2.6 Environmental considerations. Motors conforming to MIL-M-7969 will withstand the environments of many aerospace applications. This specification describes two types of motors which differ only in temperature and altitude requirements. Class A motors are operable under the temperature-altitude conditions of Curve II of MS33543, while Class B motors are operable under the conditions of Curve I of MS33543, as shown in Figure 7.

These motors will not operate in high altitude or vacuum environments.

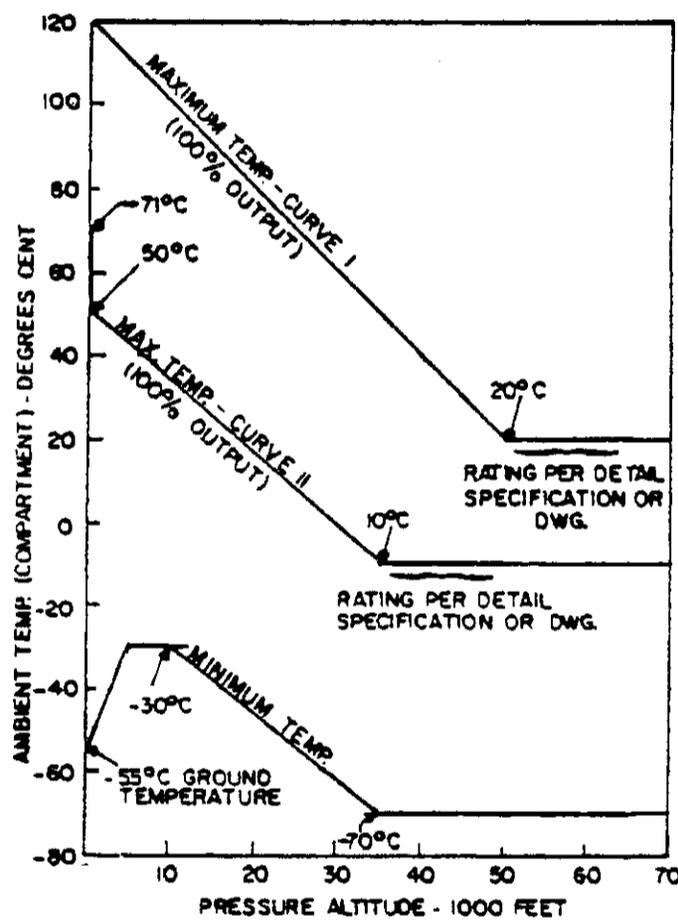
12.2.7 Reliability considerations. Because of their construction, ac motors have fewer failure modes than dc motors. As discussed in paragraph 15.1.7, these modes include:

- a. Brinelling or damaging of bearings from shock loads during handling
- b. Overloading of the motor by the application of too large a load, causing overheating
- c. Failure of one phase of a three-phase motor, causing it to overheat
- d. Operation at too low a voltage, causing overheating of the windings

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- e. Defective wire and/or insulation in the manufacturing process, causing shorts or open circuits during system application
- f. Bearing failures from misapplication of a bearing design.

Most motors are screened 100 percent to ensure that they operate at rated current and speed with rated input. In addition, all motors are given a high voltage potential test to determine the integrity of the insulation system.



Curve I indicates the maximum temperatures to which equipment must be designed to operate satisfactorily in unconditioned compartments of high performance aircraft powered by air consuming engines.

Curve II indicates the maximum temperature for which the equipment must be designed to operate satisfactorily in conditioned compartments of high performance or low performance aircraft.

FIGURE 7. Altitude vs ambient temperature operation.

12.3 MOTORS, DC

12.3 DC motors.

12.3.1 Introduction. DC motors may be classified into four categories.

12.3.1.1 Permanent magnet. Permanent magnet motors are used in applications requiring less than 0.1 hp or in power motors of less than 1 1/4-inch case diameter. Permanent magnet motors run cooler than wound field types because no power is expended to maintain a magnetic field. Placing the motor in an excessive external magnetic field can demagnetize the permanent magnet material. Furthermore, demagnetization can occur if the applied voltage is significantly higher than the design voltage of the armature. Excessive current flow created by instantaneous reversal of applied voltage will also cause demagnetization. A remedy is to use an armature winding with sufficient resistance to safely limit this current during reversal or to provide a current limiting component, usually a low-value resistor, in the circuit.

The performance of a permanent magnet motor is varied by changing the armature voltage or winding. Figure 8 shows the schematic for this type of motor.

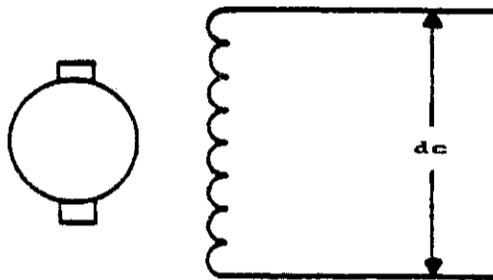


FIGURE 8. Permanent magnet dc motor.

12.3.1.2 Series-wound. The armature (rotor) and stator windings are connected in series with current passing through both.

Field strength in a series motor is a function of the armature current flowing through the field. Heavy motor loads cause more armature current to flow through the field and, thus, increase the field strength. Series motors are characterized by high stall torques and high no-load speeds. For this reason, a series motor should not be used in a system that will permit it to run without load.

Speed varies with load. These units are not usually used where power requirements are below 0.02 hp because permanent magnet motors perform the job more efficiently.

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Small-size units can be operated on either dc or ac power and are called universal motors. Operation on ac is possible only with laminated field poles (Figure 9).

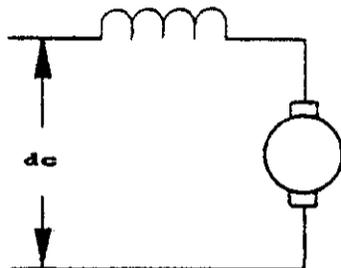


FIGURE 9. Series-wound dc motor.

12.3.1.3 Shunt-wound. The armature and stator are connected in parallel. Voltage across the armature and the stator is equal. This type gives a lower starting torque, and a constant speed with varying load. Under constant voltage, the motor operates like a permanent magnet motor because its magnetic field strength is held constant. As is the case with the series wound dc motor, this design is less efficient than a permanent magnet motor where power requirements are below 0.1 hp. This type is recommended for applications requiring constant speed (Figure 10).

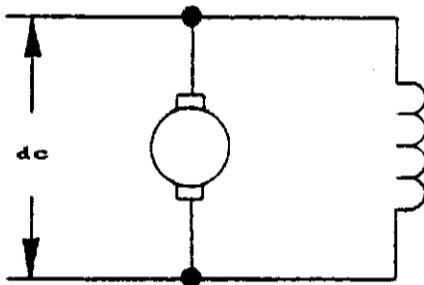


FIGURE 10. Shunt-wound dc motor.

12.3.1.4 Compound-wound. A compound-wound motor is a combination of series-wound and shunt-wound. Speed variation is much less than the series-wound motor but greater than the shunt-wound. This type is used for heavy starting loads or where loads change suddenly (Figure 11).

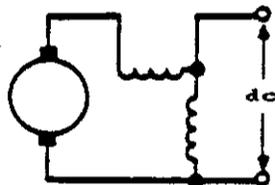


FIGURE 11. Compound-wound dc motor.

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12.3.2 Usual applications. As with ac motors, dc motors are used wherever angular motion or linear motion is desired. The dc motors are used where speed control is desired (by voltage adjustment) or where speed change with changing loads is desired.

12.3.3 Physical construction. As with ac motors (paragraph 12.2.3.1), there are many factors that determine the output of a given size dc motor. For comparison purposes, a typical outline is shown in Figure 12.

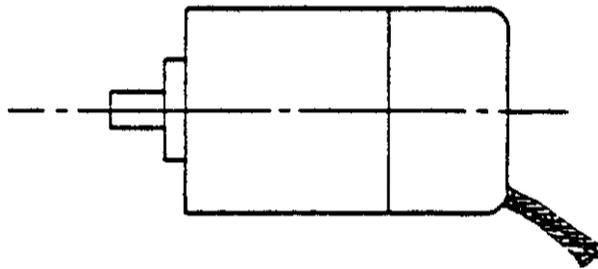


Figure 12. Typical outline of dc motors.

12.3.4 Military designation. Dc motors have not been standardized for NASA applications.

However, motors are available which conform to the requirements of MIL-M-8609 for 28 V dc airborne applications.

12.3.5 Electrical characteristics. Most dc motors are designed to operate at 28 V dc; however, where other dc voltages are required, units can be wound for these voltages.

The dc motors designed to meet the requirements of CC-M-645 are designed for either 115 V ac or 230 V dc. As with aerospace motors, these may also be designed for different voltages.

12.3.6 Environmental considerations. Motors conforming to MIL-M-8609 will withstand many aerospace applications. This specification describes two types of motors which differ only in temperature and altitude requirements. Class A motors are operable under the temperature-altitude conditions of Curve II of MS 35543 while Class B motors are operable under Curve I of the same specification. This military standard is included in the environmental section shown earlier for ac motors (refer to Figure 7). The dc motors have disadvantages associated with brush wear; brushes wear out faster at higher altitudes.

12.3 MOTORS, DC

12.3.7 Reliability considerations.

12.3.7.1 Failure modes. Because brushes are required, dc motors, have more failure modes and are more prone to failure than ac motors. Failure modes for these designs include:

- a. Fracture of brushes
- b. Rapid brush wear at high altitude
- c. Contamination of bearings resulting from wear of brushes
- d. Brinelling or damaging of bearings from heavy shock loads
- e. Insufficient voltage to drive the required load, causing overheated windings
- f. Bearing failure from misapplication of bearing designs
- g. Defective wire or insulation resulting in shorted windings or open circuits
- h. Demagnetizing of permanent magnets in permanent magnet motors caused by exposure to electromagnetic fields or rapid reversal of voltage polarity.

DC motors should be inspected for speed and current at a specified voltage. They are also subjected to high voltage potential inspection to determine the integrity of insulation systems.

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1 MARCH 1988

SUPERSEDING
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15 MARCH 1984

MILITARY HANDBOOK

NASA PARTS APPLICATION HANDBOOK

(VOLUME 5 OF 5)
CONNECTORS, PROTECTIVE DEVICES,
SWITCHES, RELAYS, WIRE, CABLE



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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

WASHINGTON, D.C. 20546

NASA Parts Application Handbook

1. This handbook is approved for use by all elements of the National Aeronautics and Space Administration and is available for use by all departments and agencies of the Department of Defense.
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FOREWORD

This handbook provides a technological baseline for parts used throughout NASA programs. The information included will improve the utilization of the NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List (MIL-STD-975) and provide technical information to improve the selection of parts and their application, and failure analysis on all NASA projects. This handbook consists of five volumes and includes information on all parts presently included in MIL-STD-975.

This handbook (Revision B) succeeds the initial release. Revision A was not released. The content in Revision B has been extensively changed from that in the initial release.

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13.1 CONNECTORS, GENERAL

13. CONNECTORS, POWER

13.1 General.

13.1.1 Introduction. This section contains general information for multipin power connectors and contacts identified in MIL-STD-975 NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts Lists for use in Grade 1 and Grade 2 applications. The connectors are classified as rack and panel, circular, and printed circuit types. Within each type are categories of contact size, operating temperature range, environment-resistance capability, contact density, and coupling mechanisms.

The information in this section is intended to help the equipment designers select electrical connectors and associated hardware from those devices contained in MIL-STD-975. The information pertaining to specific connector types is contained in its appropriate section. The most important decision a user must make is which of the numerous types of connectors and associated hardware will be the most suitable for use in the particular equipment he is designing. Proper selection, in its broadest sense, is the first step in building reliable equipment. To effectively and properly select the connectors and associated hardware to be used, the user must know as much as possible about the types from which he can choose. He should know the advantages and disadvantages, the behavior under various environmental conditions, the construction, the effect upon the circuits, the effect of the circuit upon the connector, as well as what causes connectors to fail.

13.1.1.1 Applicable military specifications.

MIL-C-5015	Connector, Electrical, Circular Threaded, General Specification for
MIL-C-22992	Connector, Plugs and Receptacles, Electrical, Waterproof, Quick Disconnect, Heavy Duty Type, General Specification for
MIL-C-24308	Connectors, Electrical, Rectangular, Miniature Polarized Shell, Rack and Panel
MIL-C-26482	Connectors, Electrical, (Circular, Miniature, Quick Disconnect, Environmental Resisting) Receptacles and Plugs, General Specification for
MIL-C-38999	Connector, Electrical, Circular, Miniature, High Density, Quick Disconnect (Bayonet Threaded, and Breech Coupling), Environmental Resistant, Removable Crimp and Hermetic Solder Contacts, General Specification for

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MIL-C-39029 Contact, Electrical Connector,
General Specification for

MIL-C-55302 Connector, Printed Circuit
Subassembly and Accessories

13.1.1.2 Applicable NASA specifications.

40M38277 Connectors, Electrical, Circular,
Miniature High Density Environment
Resisting, Specification for

40M38298 Connectors, Electrical Special,
Miniature Circular, Environment
Resisting

40M39569 Connectors, Electrical, Miniature
Circular, Environment Resisting
1200 °C, Specification for

GSFC S-311-P-4 Connectors (and Contacts), Electri-
cal, Rectangular, for Space Flight
Use, General Specification for

GSFC S-311-P-10 Connectors, Subminiature, Electri-
cal and Coaxial Contact, for Space
Flight Use

ASTM-E-595 Standard test method for total mass loss
and collected volatile condensable
materials from outgassing in a vacuum
environment.

13.1.2 General definitions. Terms which are commonly used in electrical
connector engineering practice and generally accepted by the electrical and
electronic industries are as follows:

Adapter. An intermediate device to provide for attaching special accessories
or to provide special mounting means.

Accessories. Mechanical devices, such as cable clamps, that are added to con-
nectors.

Ambient temperature. The temperature of the environment, usually air, surround-
ing a connector.

Arc resistance. The tendency of insulating material to resist breakdown or
passage of current on the surface between contacts and between contacts and
ground.

Back-mounted. A connector mounted from the inside of a panel or box with its
mounting flange inside the equipment.

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Barrel, crimp. The section of the terminal, splice, or contact that accommodates the stripped conductor.

Barrel chamfer. The bevel at the end of the conductor barrel which facilitates entry of the stripped conductor.

Basis metal. Metal from which the connector components are made and on which one or more metals or coatings may be deposited.

Bayonet coupling, rotary. A quick coupling device for mating connectors utilizing pins on one connector and ramps on the mating connector. Mating and unmating is accomplished by rotating the coupling ring.

Body, connector. The main portion of a connector to which contacts and other components are attached. This term is not used with connectors incorporating nonintegral shells in their construction.

Bonded assembly, electrical. An assembly whose supporting frame and metallic noncircuit elements are connected so as to be electrically shorted together.

Boot. A form placed around the wire terminations of a multiple contact connector as a protective housing or as a container for potting compound.

Braid. 1. A flexible conductor made of a woven or braided assembly of fine wires. 2. A fibrous or metallic group of filaments interwoven in cylindrical form to form a protective covering over one or more wires.

Bused. The joining of two or more circuits.

Cable clamp. A mechanical clamp attached to the cable side of the connector to support the cable or wire bundle, provide strain relief, and absorb vibration and shock, that would otherwise be transmitted by the cable to the contact/wire connection.

Cable clamp adapter. A mechanical adapter that attaches to the rear of a connector to allow the attachment of a cable clamp.

Cable sealing clamp. A device consisting of a gland nut and sealing member designed to seal around a single jacket cable.

Cable shielding clamp. A device consisting of a sealing member and cable support designed to terminate the screen (shield) of an electrical cable.

Circumferential crimp. The type of crimp in which the crimping die completely surrounds a barrel, resulting in symmetrical indentations in the barrel.

Closed entry. A contact or contact cavity design in the insert or body of the connector which limits the size or position of the mating contact or printed circuit board to a predetermined dimension.

13.1 CONNECTORS, GENERAL

Configuration control. The discipline providing for uniformity of materials, processes, geometry, and performance in manufactured items.

Connector, electrical. A device, either a plug or a receptacle, used to terminate or connect the conductors of individual wires or cables and which provides a means to continue the conductors to a mating connector or printed circuit board.

Connector set, electrical. Two or more separate connectors (plug connector and receptacle connector) which are designed to be mated together. The set may include mixed connectors mated together such as one connector plug and one dummy connector receptacle or one connector receptacle and one dummy electrical plug.

Connector classes. Categories based on the performance capabilities of the connectors. Classification categories include environment-resistant, firewall, and hermetically sealed connectors.

Contact. The conductive element in a connector which makes physical contact with another contact on the mating connector for the purpose of transferring electrical energy.

Contact area. The area in contact between two conductors, two contacts, or a conductor and a contact permitting the flow of electrical current.

Contact arrangement. The number, spacing, and arrangement of contacts in a connector.

Contact, crimp. A contact whose conductor barrel is a hollow cylinder that accepts the conductor. After a bared conductor is inserted, a crimping tool is applied to swage or form the contact metal firmly against the conductor. An excellent mechanical and electrical joint results. A crimp contact is often referred to as a solderless contact.

Contact engaging and separating force. The force needed to either engage or separate mating contacts.

Contact, female. A contact of such design that the mating contact is inserted therein; this is similar in function to a socket contact.

Contact, fixed. A contact which is permanently included in the insert material. It is mechanically locked, bonded, or embedded in the insert.

Contact float. The overall side play and angular displacement of contacts within the insert cavity.

Contact, insertable/removable. A contact that can be mechanically joined to, or removed from, an insert. Usually, special tools are required to lock the contact in place or remove it for repair or replacement.

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Contact, male. A contact of such design as to make contact by insertion into a mating contact. This is similar in function to a pin contact.

Contact, open entry. A socket contact whose engaging end is split and therefore vulnerable to distortion or damage from test probes or other wedging devices.

Contact, pin. A male type contact designed to slip inside the mating female contact member.

Contact resistance. Electrical resistance of a pair of engaged contacts. Resistance may be measured in ohms or millivolt drop at a specified current through the engaged contacts.

Contact retainer. A device either on the contact or in the insert that retains the contact in an insert or body.

Contact retention. The axial load in either direction that a contact can withstand without being dislodged from its normal position within an insert or connector body.

Contact shoulder. The flanged portion of the contact that limits its depth into the insert.

Contact size. An assigned number denoting the size of the contact engaging end.

Contact, socket. A female-type contact that is designed to accept the mating pin contact member.

Contact, solder. A contact having a cup, hollow cylinder, eyelet, or hook to accept a conductor and retain the applied solder.

Contact wipe. The distance of travel (physical engagement) made by one contact with another during its engagement or separation or during mating or unmating of the connector halves.

Corona. A luminous discharge of electricity due to ionization of the air appearing on the surface of a conductor when the potential gradient exceeds a certain value.

Coupling ring. That portion of a plug which aids in the mating or unmating of a plug and receptacle and holds the plug to the receptacle.

Cover, electrical connector. An item which is specially designed to cover the mating end of a connector for mechanical and environmental protection.

Creep distance. The shortest distance on the surface of an insulator separating two electrically conductive surfaces.

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Creepage path. The path across the surface of dielectric between two conductors. Lengthening the creepage path reduces the possibility of arc damage or tracking.

Crimp. The physical compression (deformation) of a contact barrel around a conductor to make an electrical and mechanical connection.

Crimping. A pressure method of mechanically securing a terminal, splice, or contact to a conductor.

Crimping die. The portion of the crimping tool that shapes the crimp.

Crimping tool. The mechanism used for crimping.

Current-carrying capacity. The current a conductor of given size and length is capable of carrying safely without exceeding its temperature limitations.

Cutout, connector. The hole, usually round or rectangular, cut in a metal panel for mounting a connector. The cutout may include holes for mounting screws or bolts.

Depth of crimp. The distance the indenter penetrates radially into the crimp barrel.

Dielectric. A material having electric insulating properties.

Dummy connector assembly, electrical. Two or more electrical dummy connectors having a common mounting or mounted on each other, each one capable of being independently replaced. Excludes items which are furnished as mated pairs or sets.

Dummy connector, receptacle. A connector receptacle which does not have provisions for attaching conductors. It is generally used for stowage of a cable assembly connector plug.

Environmentally sealed. A device that is provided with gaskets, seals, grommets, potting, or other means to keep out moisture, dirt, air, or dust which might reduce its performance. Does not include nonphysical environments such as rf and radiation.

Extraction tool. A device used for removing removable contacts from a connector.

Ferrule. A short tube used to make solderless connections to shielded or coaxial cables. Also used in connectors to reduce transmission of torque to the grommet.

Flange, connector. A projection extending from or around the periphery of a connector with provisions to permit mounting the connector to a panel.

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Flux. A material used to promote fusion or joining of metals in soldering, brazing, or welding.

Follower. A sleeve used to compress the grommet, thus tightening the seal around the conductors entering the termination assembly.

Front mounted. A connector mounted on the outside of a panel or box with its mounting flange outside the equipment.

Grid spaced. The arrangement of contacts in a multiple contact termination assembly by spacing in a geometric pattern.

Grommet, connector. An elastomeric seal used on the cable side of a connector to seal the connector against moisture, dirt, and air.

Grope free. A situation in which a connector coupling system can easily be mated and locked, usually with one hand. A coupling ring that is held in the proper position to start the mating cycle while uncoupled.

Ground, electrical. A point of common potential in an electric circuit used for common connections and reference voltage.

Guide pin. A pin or rod extending beyond the mating faces of a connector designed to guide the mating or unmating of the connector to ensure proper engagement or disengagement of the contacts.

Housing, connector, electrical. The connector less the insert, but with the insert-retaining and positioning hardware required by standard construction.

Impedance. The total opposition (resistance and reactance) a circuit offers to the flow of electric current. It is measured in ohms and its reciprocal is admittance, usually expressed in siemens.

Insert arrangement. The number, spacing, and arrangement of contacts in a termination assembly.

Insert, closed entry. An insert having openings that restrict the entry of devices larger than the specified contact.

Insert, electrical connector. An insulating element with or without contacts that is designed to position and support contacts in a connector.

Insertion tool. A device used to insert contacts into a connector and to insert taper pins into taper pin receptacles.

Inspection hole. A hole placed at one end of a crimp barrel to permit visual inspection to see that the conductor has been inserted to the proper depth in the barrel prior to crimping.

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Insulation support. The portion of a barrel that supports but does not compress the conductor insulation.

Interface. The two surfaces on the contact side of mating connectors that face each other when mated.

Interfacial gap. Any gap between the faces of mated inserts.

Interfacial seal. A sealing of mated connectors over the whole area of the interface to provide sealing around each contact.

Jacket. The outermost layer of insulating material of a cable or wire.

Jackscrew (screwlock). A screw attached to one half of a two-piece multiple contact connector used to draw and hold both halves together. A jackscrew may also be used to separate two connector halves.

Key. A short pin or other projection which slides in a mating slot, hole, groove, or keyway to guide parts being mated. Keys are generally used in circular connectors to obtain polarization.

Keyway. A slot or groove in which a key slides.

Mold, potting, electrical connector. A device, consisting of one or more pieces, designed to be used as a hollow form into which potting compound is injected and allowed to cure (or set) to seal the back of an electrical connector. The potting may eliminate the need for a backshell of the connector. The form may or may not be removable after the potting compound cures.

Nest. The portion of a crimping die which supports the barrel during crimping.

Operating temperature. The maximum temperature resistance capabilities of a connector in continuous service.

Panel. The side or front of a piece of equipment, usually metal, on which connectors are mounted.

Pin contact. A contact having an engagement end that enters the socket contact.

Plating. The deposition of a thin coating of metal on metallic components to improve conductivity, provide for easy soldering, or prevent corrosion.

Plug, connector. An electrical fitting with pin, socket, or pin and socket contacts, constructed to be affixed to the end of a cable, conduit, coaxial line, cord, or wire for convenience in joining with another electrical connector and not designed to be mounted on a bulkhead, chassis, or panel.

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Polarization. A coded arrangement of keys, keyways, and insert positions which prohibits the mating of mismatched plugs and receptacles. Polarization allows connectors of the same size to be lined up side by side with no danger of making the wrong connection. The polarization of rectangular connectors is usually accomplished by the design of the shell or the mating hardware so that mating is possible in only one orientation.

Polarizing pin, key, or keyway. A device incorporated in a connector to accomplish polarization.

Potting. The permanent sealing of the cable end of a connector with a compound or material to exclude moisture and to provide a strain relief.

Pretinned. Solder applied to either or both the contact and conductor prior to soldering.

Pull-out force. The force necessary to separate a conductor from a contact or terminal, or a contact from a connector, by exerting a tensile pull.

Quick disconnect. A type of connector or splice which permits relatively rapid locking and unlocking of mating parts.

Rack-and-panel. A type of connector that is attached to a panel or side of equipment so that the connector is engaged when these two members are brought together.

Radio frequency (rf). The frequency spectrum from 15 KHz to 10,000 MHz.

Radio frequency interference (rfi). Electromagnetic radiation in the radio frequency spectrum from 15 KHz to 10,000 MHz. The best shielding materials against rfi are copper and aluminum alloys. The term EMI should not be used in place of rfi because shielding materials for the entire electromagnetic frequency spectrum are not available.

Range, wire. The sizes of conductors accommodated by a particular crimp barrel. Also the diameters of wires accommodated by a sealing grommet.

Receptacle, connector. An electrical fitting with contacts constructed to be electrically connected to a cable, coaxial line, cord, or wire to join with another electrical connector, and designed to be mounted on a bulkhead, wall, chassis, or panel.

Scoop-proof. The feature that prevents connector pins from bending during mating or unmating.

Sealing plug. A plug which is inserted to fill an unoccupied contact aperture in a termination assembly. Its function is to seal all unoccupied apertures in the assembly, especially in environmental connectors or junctions.

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13.1 CONNECTORS, GENERAL

Sealing plug. A plug which is inserted to fill an unoccupied contact aperture in a termination assembly. Its function is to seal all unoccupied apertures in the assembly, especially in environmental connectors or junctions.

Service life. A period of time over which a device is expected to perform in accordance with its specified requirements.

Service rating. The maximum voltage or current which a connector is designed to carry continuously.

Shell, electrical connector. The outside case of a connector into which the dielectric material and contacts are assembled.

Shield, electrical connector. An item especially designed to be placed around that portion of a connector which contains the facilities for attaching wires or cables. It is used for shielding against electrical interference or mechanical injury and usually has provisions for passage of the wire or cable.

Socket contact. A contact having an engagement end that will accept entry of a pin contact.

Solder cup. The end of a terminal or contact in which the conductor is inserted prior to being soldered.

Solder eye. A solder type contact provided with a hole at its end through which a wire can be inserted prior to being soldered.

Solderless connection. The joining of two metals by pressure means without the use of solder, braze, or any method requiring heat.

Solderless wrap. A technique of connecting stripped wire to a terminal post containing a series of sharp edges by winding the wire around the terminal.

Strip. To remove insulation from a conductor.

Threaded coupling. A means of coupling mating connectors by engaging threads in a coupling ring with threads on a receptacle shell.

Tubular adapter. An accessory attached to the rear of a termination assembly, usually metallic, used to extend the shell far enough to support a sealing gland or to give mechanical support for a cable or conductor harness.

Umbilical connector. A connector used to connect cables to a rocket or missile prior to launching and which is unmated from the missile at the time of launching.

Wiping action (see contact wipe). The action of two electrical contacts which come in contact by sliding against each other.

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Work curve. A graph which plots the pull out force, indent force, and relative conductivity of a crimp joint as a function of various depths of crimping.

Working voltage. The maximum voltage at which a connector is rated to operate.

13.1.3 General device characteristics. A connector must be capable of mating and unmating in its intended application, provide electrical continuity (when mated) adequate for the functions of the circuitry involved, and maintain adequate insulation between the conductors. These requirements are fundamental; however, there are many other factors to be considered that influence the selection of connectors. Some of these factors are:

- a. Limitation on size, weight, and cost
- b. Current carrying capacity and voltage withstanding ability
- c. Ease of installation and repair
- d. Ease of operation (mating and unmating).

Connectors should be used within the design limitations stated in the manufacturer's ratings and in the class of service for which they are intended. However, the following special design considerations should always be considered:

- a. What are the current and voltage requirements?
- b. How many signal circuits?
- c. How many power circuits?
- d. What number and type of conductors?
- e. Shielded, rf, coaxial?
- f. How many of each size conductor?
- g. Type and physical characteristics of wire?

13.1.3.1 Human engineering. The human engineering aspects of the connector application must be considered by the equipment designs. For example, if a coupling ring must be unscrewed by hand, room must be allowed around the connector so that a hand can do this. Can it be reached and manipulated from where a person must stand? Can maintenance and repair operations be carried out with reasonable convenience?

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13.1.3.2 Safety. The connector application should be planned so that the recessed female contacts are used on the "hot" or power side of the circuit. This is done so that there is no danger of shock when handling or touching the connector. The mating connector having exposed male pins should be used on the "dead" side of the connection. The exposed pins should be enclosed in the connector housing to prevent damage.

When more than one connector of the same type is used on the same panel, connectors with alternate mating keyways should be used to prevent mismatching.

13.1.3.3 Alignment. Many connectors are designed with a certain amount of float in the contacts or in the housing. This permits the contacts or housing to align themselves in the proper position during mating. It is important that the mounting of the connector and the dress of the wire bundle behind the connector be arranged to allow wire slack so as not to restrict the float. If the rear surface of the connector is to be potted after wiring, a resilient potting material must be used. It is desirable to have the connector mated with its mating half or an equivalent fixture during potting so that the contacts are in proper alignment as the potting material sets up. Care in cable dressing is especially important with connectors having removable contacts. To maintain proper contact alignment and prevent possible damage to the contact, the cable should not make a sharp right angle bend at the end of the connector.

13.1.3.4 Number of circuits carried. When a large number of leads are to be accommodated, several separate connectors may prove more satisfactory than a single connector. Spare contacts in a quantity of 10 percent of the total contacts required should be provided. In addition to the difficulties in coupling a large connector, consideration should be given to the total current the connector must carry and the temperature rise in the connector due to contact resistance. Contact resistance can be expected to increase approximately 75 percent after exposure to a salt atmosphere.

13.1.3.5 Insulation material. The insulation material separating the connector contacts requires prime consideration. A wide variety of insulators have been developed. Each has a long list of electrical and mechanical properties, any one of which might be the dominant characteristic for the application. Some of the common materials used are listed in Table I.

13.1.4 General parameter information. Connector selection must be based on electrical, mechanical, and environment-resistant qualities as required by the application (see Table II).

The principal sources of information on performance characteristics are the military specifications, data from connector manufacturers, and the user's history of previous performance.

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13.1 CONNECTORS, GENERAL

TABLE I. Properties of insulating materials

Parameter	Neoprene	Silicone	Diallyl Phthalate	Epoxy	Phenolic
Dielectric strength (volts/mil)	300	500	350	350	350
Temperature resistance (°C) (1000 hours)	120	250	175	230	200
Tensile strength (psi)	1,200	4,000	6,000	11,000	10,000
Low temperature properties	Poor	Excellent	--	--	--
Oil resistance	Good	Excellent	--	--	--
Compression set	Fair	Excellent	--	--	--
Outgassing at 55 °C	High	Low	Low	Very low	Low
Insulation resistance	Fair-Good	Excellent	Excellent	Good	Good

TABLE II. Connector specification considerations

Performance Classification	Parameters to Consider	
Electrical	Voltage Current Insulation resistance Contact resistance	Grounding rf characteristics Number of contacts
Mechanical	Coupling system Connector mating alignment Contact retention	Wire terminations Seals Interfaces
Environmental	Temperature Pressure differential Vibration Shock Thermal cycling	Acceleration Moisture Sand and dust Fluids

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13.1 CONNECTORS, GENERAL

Two areas where the application of specification information vitally affects expected reliability are:

- a. Exceeding any one specified rating
- b. Applying more than one published performance characteristic at maximum rating.

When either of the two conditions noted above must be encountered, the selection of the connector used should only be made after manufacturers and users have been contacted to determine best performance trade-off.

13.1.5 General guides and charts. General characteristics of power connectors are given in Table III.

TABLE III. General characteristics of power connectors

Connector	Contact Size	Contact Termination	Coupling Type	Temperature Range	Moisture Resistant
MIL-C-5015 (Power)	#0 to #16	Solder and crimp	Threaded coupling	-55 to +125 °C	Yes
MIL-C-22992 (Heavy duty)	#0 to #16	Solder	Threaded coupling	-55 to +125 °C	Yes
MIL-C-24308 (Rack and Panel)	#20,22	Solder and crimp	Jackscrews	-55 to +125 °C	No
MIL-C-26482 (Circular)	#12,16,20	Solder and crimp	Bayonet	-55 to +200 °C	Yes
MIL-C-38999 (Miniature)	#12 to #22	Solder and crimp	Bayonet	-65 to +150 °C	Yes
MIL-C-55302 (Printed Circuit)	#22	Solder and crimp	Jackscrews	-65 to +125 °C	No

13.1.6 General reliability considerations. The reliability of typical connectors compares favorably with that of other components used in electrical and electronic equipment. Factors that tend to degrade connector reliability include the added complexity of new multicontact designs with their multiplicity of parts, the variety of contact termination methods, and the more extreme environmental requirements imposed by complex modern systems. Failure rate prediction data for electrical connectors is presented in MIL-HDBK-217.

13.1 CONNECTORS, GENERAL

13.1.6.1 Mechanical effects. Achieving good electrical contact in a connector is a function of surface films (oxides and sulphides), surface roughness, contact area, plastic deformation of the contacting materials, and load applied.

Because even the best machined, polished, and coated surfaces look rough and uneven when viewed microscopically, the common concept of a flat, smooth contact is grossly oversimplified. In reality, the connector interface is basically an insulating barrier with a few widely scattered points of microscopic contact. The performance of the connector is dependent upon the chemical, thermal, and mechanical behavior at these contact points.

13.1.6.2 Electrical effects. Current flow between mating metals is constricted at the interface to the small points on the surfaces which are in electrical contact. This flow pattern causes differences of potential to exist along the contact interface, and causes higher current flow at points of lower resistance. As a result, contact resistance and capacitance are introduced into the circuit, and certain chemical effects evolve (see following paragraph on chemical effects).

13.1.6.3 Thermal effects. The total contact resistance of a pin and socket contact pair is a parallel combination of many higher resistance-areas of point contact. The effect is that of a series of localized hot spots, all contributing to an average integrated temperature rise at the interface. When high currents are conducted through many contact pairs, the cumulative heat rise in the connector can be appreciable. This heat rise above the ambient or body temperature is a deciding factor in connector reliability.

Excessive temperature can cause failure of connectors by breakdown of insulation or by breakdown in the conductivity of the conductors. Either malfunction can be partial or complete.

A typical breakdown caused by excessive temperature occurs progressively as follows: as operating temperature increases, the insulation tends to become more conductive, and simultaneously, the resistance of the conductors increases. Higher resistance causes the temperature of the conductor and of its insulation to rise further. This pyramiding effect can raise the temperature of conductors and connector contacts beyond their maximum operating temperatures with resultant damage occurring to the contacts and conductive platings. Complete breakdown will occur if the operating temperature reaches the point where the conductor melts, breaking electrical conductivity; or where the insulation fails, causing a short.

Maximum operating temperatures are the sum of the ambient temperature and the conductor temperature rise caused by the passage of current. A maximum conductor operating temperature of 125 °C, for example, is based on an ambient temperature of 100 °C, plus a rise of 25 °C due to the conductor carrying current.

13.1 CONNECTORS, GENERAL

Extremely low temperatures introduce the same kind of problems to the design of the interconnection system as to any other portion of a system. Metals and non-metals tend to become brittle and shrink at different rates. The relative importance of each characteristic depends on the application. Most high-performance connectors will operate at temperatures as low as -55 °C. Operation at lower temperatures may require special materials.

Temperatures below normal ambient are not usually the cause of conductivity problems in interconnection systems (more current can be carried by a given conductor at lower temperatures). However, extremely low ambient temperatures do produce mechanical failures, which occur most often in the nonmetallic portions of connectors, wires, and cables. The coefficients of expansion of most polymers and elastomers are so different from those of the metals used in structural members that they will contract enough at extremely low temperatures to open seals. An open seal will not cause a malfunction unless contaminants can enter through the opening. If a seal opens after the temperature of a connector falls below the freezing point of the contaminants present, and then seals itself before the melting point of the contaminants is reached, foreign matter will never enter. If, however a connector seal will open at a temperature where all liquid or gaseous contaminants have not been frozen, a more suitable type of material or mechanical arrangement is required.

13.1.6.4 Chemical effects. Most connector failures resulting from chemical effects are induced by the growth of films at points of contact. These films can cause increased contact resistance or an open circuit. Contact resistance gives rise, as explained above, to interfaces at higher temperatures than the surroundings, thus increasing the chemical activity.

Normally, the purity of metals in contact is not considered a reliability problem. However, ions in impurities or contamination in the surface pores will migrate to the points of highest potential, which are frequently the localized hotspots. Ions interfacing with electrons and other constituents at the points of high chemical activity generate films, which are usually nonconducting. There is also a continuous supply of material for the growth of insulating films from environments where there are corrosive elements such as hydrogen sulfide, water vapor, oxygen, ozone, hydrocarbons, and various dusts.

13.1.6.5 Cycling effects. Repeated mating and unmating of connectors exposes the contacts to a fresh supply of local corrosive contaminants during each mating cycle. There is also the problem of physical wear on the connecting interfaces in this type of operation. The result is increased interface resistance, temperature, and degradation of the connection.

13.1.6.6 Materials. Materials capable of emitting vacuum condensable, noxious or toxic gases when tested in accordance with ASTM E 595 shall not be used. Materials with a total mass (TML) of less than 1.0% and collected volatile condensable materials (CVCM) less than 0.10% are generally considered acceptable for NASA use. The outgassing requirements are not controlled in MIL-STD-975. Consult the project parts engineer for recommendations.

**13.2 CONNECTORS, CIRCULAR,
MIL-C-26482**

13.2 Circular, MIL-C-26482.

13.2.1 Introduction. This specification covers environment-resisting, bayonet coupling, and circular connectors with solder or rear release removable crimp contacts.

Receptacles are available in the following mounting styles: square flange, jam-nut, and solder mount.

Four service classes are available: grommet seal, hermetic seal solder contacts, hermetic seal crimp contacts, and fluid resistant.

Some available accessories are: protective covers, stowage receptacles, strain relief clamps, and rfi backshells.

These connectors meet the requirements of MIL-C-26482, Series 2.

13.2.2 Usual applications. These connectors are intended for use in environment-resisting applications where an operating temperature range of -55 to +175 °C or -55 to +200 °C is experienced. Contacts are available with crimp or solder terminations. Crimp contacts are preferred because they can be replaced individually. Solder contacts are molded in the connector and damage to any one contact requires the replacement of the entire connector. These connectors are moisture sealed and can operate, when mated, in 95 percent relative humidity.

13.2.3 Physical construction. Figures 1 through 4 illustrate various types of circular connectors.

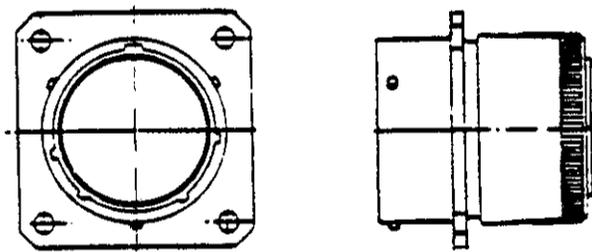


FIGURE 1. Square flange, MIL-C-26482.

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**13.2 CONNECTORS, CIRCULAR,
MIL-C-26482**

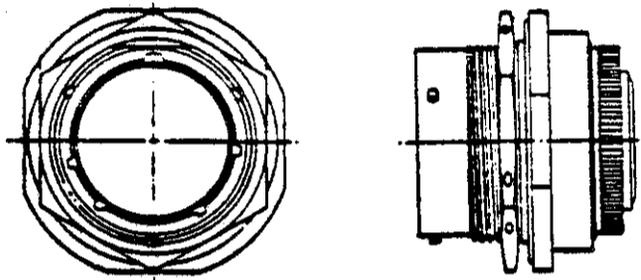


FIGURE 2. Jam-Nut, MIL-C-26482.

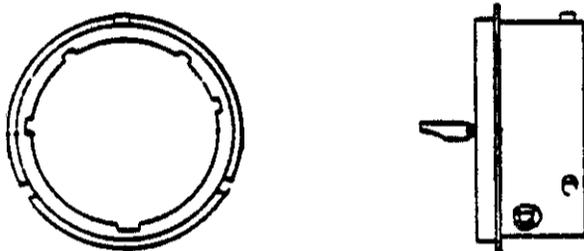


FIGURE 3. Solder mount, MIL-C-26482.

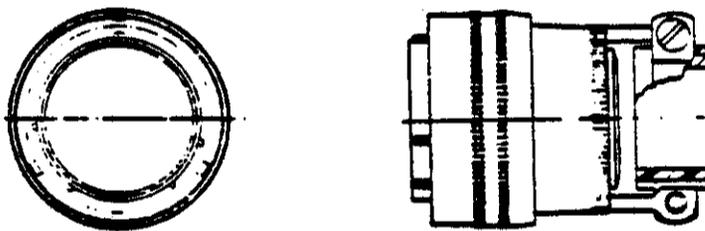


FIGURE 4. Plug, MIL-C-26482.

13.2.3.1 Insert design and construction. Inserts are of voidless construction and are secured to prevent rotation within the shell. The inserts are not removable from the shell and are installed in the position specified on the applicable military standard. A wire sealing grommet is an integral part of the insert assembly. The design will permit the removal and reinsertion of individual crimp contacts using the applicable tools.

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13.2 CONNECTORS, CIRCULAR,
MIL-C-26482

13.2.3.2 Contacts. Solder contacts (Class H) are not removable from the connector. Solder cups are designed so that the connector will not be damaged and no liquid solder will escape during soldering.

Crimp contacts are designed to meet the requirements of MIL-C-39029. Connectors are designed to permit individual insertion and extraction of contacts without removing the insert or sealing members. Insertion and extraction is done from the wire side of the connector and with the aid of tools listed in the applicable military standard.

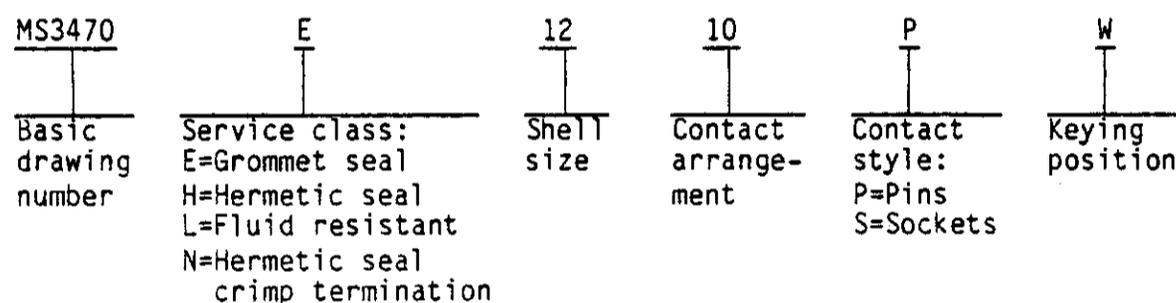
13.2.3.3 Contact spacing. Two service ratings are available. The service ratings differ in the contact center-to-center spacing and the minimum dielectric thickness as specified in MIL-C-26482.

13.2.3.4 Bayonet coupling. Coupling is done by clockwise rotation of the coupling ring; uncoupling is done by counterclockwise rotation. The coupling rings are knurled to provide a gripping surface.

13.2.3.5 Polarization. Polarization is done by matched integral keys and keyways of counter-part connectors. Only connectors with matching keys and keyways will mate. To assure proper contact alignment, keys and keyways are designed to engage before the pin contact enters the socket contact.

13.2.3.6 Interfacial seal. An interfacial seal is provided to minimize air voids between adjacent contacts and between contacts and the shell. This feature improves high-altitude voltage performance. This seal also prevents the entrance of moisture, salt, fog, and fuels when the connector is mated.

13.2.4. Military designation. Qualified connectors are procured by military standards drawings. The following is an example of a complete number for a MIL-C-26482 connector. Basic drawings provide detail information for connectors supplied to MIL-C-26482.



13.2.5 Electrical characteristics.

13.2.5.1 Contact resistance limits. Contact resistance for mated contacts is often expressed as millivolt drop measured across the mated contact pair. The maximum allowable voltage drop is specified in MIL-C-26482 for various combinations of contact size, wire gauge, test current, and service class.

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**13.2 CONNECTORS, CIRCULAR,
MIL-C-26482**

13.2.5.2 Working voltages. Maximum working voltages are specified in MIL-C-26482.

13.2.5.3 Insulation resistance. The insulation resistance of MIL-C-26482 connectors is 5,000 megohms for series 2 connectors when measured as specified in the military specification. Insulation resistance decreases with increasing temperature. Insulation resistance derating curves are presented in MIL-C-26482.

13.2.6 Environmental considerations. There are no unusual environmental considerations for these connectors. MIL-C-26482 connectors are considered to be environment resistant. Performance requirements, including the environmental qualification tests are listed in MIL-C-26482.

13.2.7 Reliability considerations. These connectors have been widely used in military systems for many years, and the design is considered to be mature. Some problems that occasionally arise are contact retention failures, difficulty in intermating connectors supplied from different vendors, and loss of electrical continuity in mated contact pairs.

Contact retention failures are most often due to improper assembly techniques or faulty contact retention systems in the connector. Operator training is usually sufficient to eliminate problems due to improper assembly techniques. Increased attention by the vendor to manufacturing processes and quality control will generally eliminate faults in the contact retention system.

Difficulty in mating connector halves which have been supplied from different vendors is often traced to dimensions which are out of tolerance. This problem can be eliminated with better control over the manufacturing process and increased process and increased emphasis on quality control.

Electrical failures of the contacts are most often caused by contamination, improper heat treatment of the socket springs, or dimensions which are out of tolerance.

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**13.3 CONNECTORS, CIRCULAR, MINIATURE,
HIGH DENSITY, MIL-C-38999**

13.3 Circular, miniature, high density, MIL-C-38999.

13.3.1 Introduction. These connectors are circular, high density, quick-disconnect, environment-resisting connectors. These connectors are classified as follows: Series I - scoop-proof and Series II - low-silhouette. Series I and II connectors are not interchangeable or intermateable.

This specification employs removable crimp contacts, except in the hermetic types, where nonremovable solder type contacts are used. A three-pin bayonet coupling mechanism is employed with these connectors. Receptacle mounting is accomplished three ways: flange mount, jam-nut mount and solder mount. Three types of seals are used: grommet seal, potted seal and hermetic seal. Available accessories include protective covers and caps, potting boots, strain relief clamps, adapters, and stowage receptacles.

13.3.2 Usual applications. These miniature circular connectors are designed for use in environment resisting applications where an operating temperature range of -65 to +200 °C is encountered. These connectors offer high contact density, light weight, and low profile characteristics which are desirable where weight and available space must be considered. Removable crimp contacts are preferred because it is possible to replace any one contact. Solder contacts are glass-sealed molded in the connector, and any damaged contact requires the replacement of the entire connector.

13.3.3 Physical construction.

13.3.3.1 Typical configurations. MIL-C-38999 connectors are available in five configurations: straight plug, box mounting receptacle, wall mounting receptacle, jam-nut mounting receptacle and solder mounting receptacle. Typical configurations are shown in Figures 5 through 9.

13.3.3.2 Polarization. Polarization is accomplished by means of five integral keys and matching keyways on the counterpart connector half. Polarization is accomplished before the initial engagement of the coupling ring is possible. During axial engagement the pins do not touch the sockets or the insert face until polarization has been achieved.

13.3.3.3 Insert design and construction. Inserts are of voidless construction and are secured to prevent rotation within the shell. The inserts are non-removable from the shell, and are positioned in the shell in accordance with the applicable military standard drawing or specification sheet. Individual crimp contacts may be removed or inserted using appropriate military standard contact insertion or withdrawal tools.

13.3.3.4 Contact arrangement. The contact positions are delineated in the applicable military standard drawing specification sheet.

MIL-HDBK-978-B (NASA)

**13.3 CONNECTORS, CIRCULAR, MINIATURE,
HIGH DENSITY, MIL-C-38999**



FIGURE 5. Straight plug, MIL-C-38999.

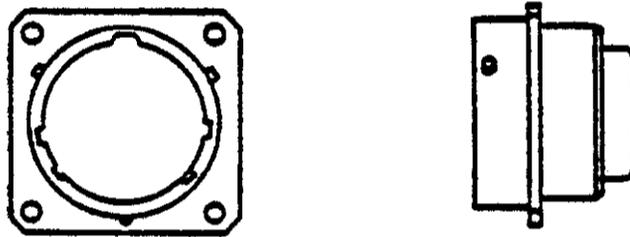


FIGURE 6. Box mounting receptacle, MIL-C-38999.

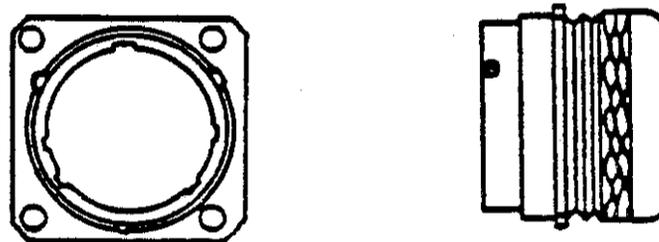


FIGURE 7. Wall mounting receptacle, MIL-C-38999.

MIL-HDBK-978-B (NASA)

**13.3 CONNECTORS, CIRCULAR, MINIATURE,
HIGH DENSITY, MIL-C-38999**

13.3 Circular, miniature, high density, MIL-C-38999.

13.3.1 Introduction. These connectors are circular, high density, quick-disconnect, environment-resisting connectors. These connectors are classified as follows: Series I - scoop-proof and Series II - low-silhouette. Series I and II connectors are not interchangeable or intermateable.

This specification employs removable crimp contacts, except in the hermetic types, where nonremovable solder type contacts are used. A three-pin bayonet coupling mechanism is employed with these connectors. Receptacle mounting is accomplished three ways: flange mount, jam-nut mount and solder mount. Three types of seals are used: grommet seal, potted seal and hermetic seal. Available accessories include protective covers and caps, potting boots, strain relief clamps, adapters, and stowage receptacles.

13.3.2 Usual applications. These miniature circular connectors are designed for use in environment resisting applications where an operating temperature range of -65 to +200 °C is encountered. These connectors offer high contact density, light weight, and low profile characteristics which are desirable where weight and available space must be considered. Removable crimp contacts are preferred because it is possible to replace any one contact. Solder contacts are glass-sealed molded in the connector, and any damaged contact requires the replacement of the entire connector.

13.3.3 Physical construction.

13.3.3.1 Typical configurations. MIL-C-38999 connectors are available in five configurations: straight plug, box mounting receptacle, wall mounting receptacle, jam-nut mounting receptacle and solder mounting receptacle. Typical configurations are shown in Figures 5 through 9.

13.3.3.2 Polarization. Polarization is accomplished by means of five integral keys and matching keyways on the counterpart connector half. Polarization is accomplished before the initial engagement of the coupling ring is possible. During axial engagement the pins do not touch the sockets or the insert face until polarization has been achieved.

13.3.3.3 Insert design and construction. Inserts are of voidless construction and are secured to prevent rotation within the shell. The inserts are non-removable from the shell, and are positioned in the shell in accordance with the applicable military standard drawing or specification sheet. Individual crimp contacts may be removed or inserted using appropriate military standard contact insertion or withdrawal tools.

13.3.3.4 Contact arrangement. The contact positions are delineated in the applicable military standard drawing specification sheet.

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**13.3 CONNECTORS, CIRCULAR, MINIATURE,
HIGH DENSITY, MIL-C-38999**



FIGURE 5. Straight plug, MIL-C-38999.

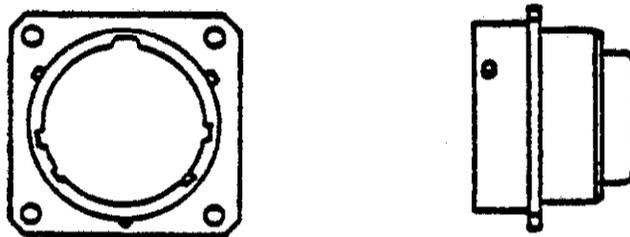


FIGURE 6. Box mounting receptacle, MIL-C-38999.

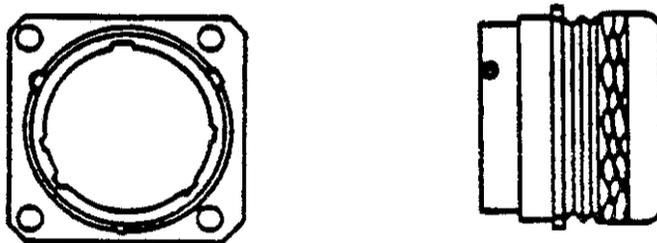


FIGURE 7. Wall mounting receptacle, MIL-C-38999.

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**13.3 CONNECTORS, CIRCULAR, MINIATURE,
HIGH DENSITY, MIL-C-38999**

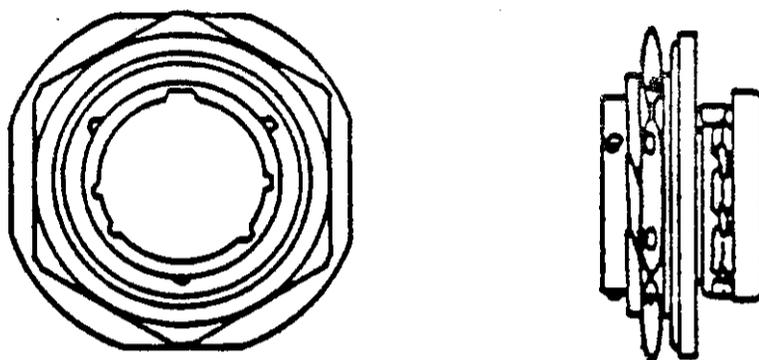


FIGURE 8. Jam-nut mounting receptacle, MIL-C-38999.



FIGURE 9. Solder mounting receptacle, MIL-C-38999.

13.3.3.5 Contacts. MIL-C-38999 removable contacts conform to MIL-C-39029. Dimensional data for solder contacts are given in MIL-C-38999. The solder cups for the nonremovable contacts are designed so that connector components will not be damaged and no liquid solder will escape during soldering. Care is required when removing crimp contacts so as not to damage the rear grommet.

13.3.3.6 Interfacial seal. Plugs and receptacles with pin inserts have a resilient face. This provides an interfacial seal with the hard face of the socket insert in the mated condition. This seal minimizes air voids between adjacent contacts and between contacts and the shell, thereby improving high-voltage performance at altitude. The seal also prevents entrance of moisture, salt fog, and fuels when the connector is mated.

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**13.3 CONNECTORS, CIRCULAR, MINIATURE,
HIGH DENSITY, MIL-C-38999**

13.3.4. Military designation. Qualified connectors are procured by military standard drawings. The following is an example of a complete part number for a MIL-C-38999 connector. Basic drawings provide detailed information for connectors supplied to MIL-C-38999.

MS 27467	E	13	F	8	P	A
Basic MS drawing no.	Class:	Shell size	Finish:	Insert arrange- ment	Contact style:	Keying positions:
	E=Environment resisting		C-Anodic		P-pins S-sockets	A,B,C,D
	P=Potting, includes pot- ting form		D-Fused tin plate			No letter required for normal position.
	T=Environment resisting without rear accessories		E-Passivated (corrosion resistant steel only)			
	Y=Hermetically sealed		F-Conductive electro- less nickel coating			

13.3.5 Electrical characteristics.

13.3.5.1 Contact resistance. Contact resistance for mated contacts is often expressed as millivolt drop measured across the mated contact pair. The maximum allowable voltage drop across the mated contact pair is specified in MIL-C-38999 for various combinations of contact size, wire gauge, test current, and service class.

13.3.5.2 Working voltage or test voltage. The voltage rating of a connector decreases with increasing altitude. This parameter may be expressed as a test voltage or working voltage. The maximum voltage ratings are given in MIL-C-38999.

13.3.5.3 Insulation resistance. The insulation resistance requirements for MIL-C-38999 connectors vary between the different series and classes. The insulation resistance values are listed in MIL-C-38999. The insulation resistance decreases with increasing temperature. Insulation resistance requirements at elevated temperatures are presented in MIL-C-38999.

13.3.6. Environmental considerations.

13.3.6.1 Temperature. All MIL-C-38999 Series I and II connectors are capable of continuous operation within a temperature range of -65 to +150 °C. The maximum temperature for selected classes is +200 °C.

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**13.3 CONNECTORS, CIRCULAR, MINIATURE,
HIGH DENSITY, MIL-C-38999**

13.3.6.2 Humidity. To assure engagement of the moisture seal, plugs and receptacles with pin contacts have a resilient face. The resilient interfacial seal provides individual contact seals in the mated condition to insure circuit isolation between contacts and between contacts and the shell. To assure an adequate seal at the wire end of the connector, only wires with the proper insulation dimensions should be used. The proper insulation sizes to achieve moisture seals are listed in MIL-C-38999.

13.3.6.3 Environmental qualification. MIL-C-38999 connectors are considered to be environment resistant. The performance requirements, including the environmental qualification tests are listed in MIL-C-38999.

13.3.7 Reliability considerations. These connectors have been widely used in military systems. Some problems that occasionally arise are contact retention failures and loss of electrical continuity in mated contact pairs.

Contact retention failures are most often due to improper assembly techniques or faulty contact retention systems in the connector. Operator training is usually sufficient to eliminate problems due to improper assembly techniques. Increased attention to manufacturing processes and quality control will generally eliminate faults in the contact retention system.

Electrical failures of the contacts are most often caused by contamination, improper heat treatment of the socket springs, or dimensions which are out of tolerance.

**13.4 CONNECTORS, RACK AND
PANEL, MIL-C-24308**

13.4 Rack and panel, MIL-C-24308.

13.4.1 Introduction. Connectors covered by this section are rectangular, nonenvironment-resisting and are available with solder or crimp type contacts. These connectors may be mated by being pushed together without the aid of a coupling mechanism or by use of screwlock hardware. Proper connector orientation during mating is accomplished by a polarized shell. Mounting holes provided in the connector shell, or screwlock hardware may be used to mount the connector to a rack or panel.

13.4.2 Usual applications. These connectors may be used with cables and are available with hoods and cable clamps to protect the assembled cable.

As their description suggests, these connectors are primarily used where a rack or drawer must plug into a panel. The mounting hardware and polarizing shells act as guides. These connectors are also available with float mounting hardware to aid in mating.

MIL-C-24308 connectors are available in six classes as follows:

G = General purpose

D = General purpose (spaceflight)

N = Nonmagnetic

M = Nonmagnetic (spaceflight)

H = Hermetic

K = Hermetic (spaceflight)

13.4.3 Physical configuration. A typical pair of polarized shell rack and panel connectors is shown in Figure 10.

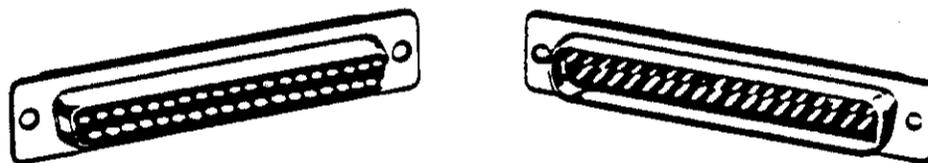
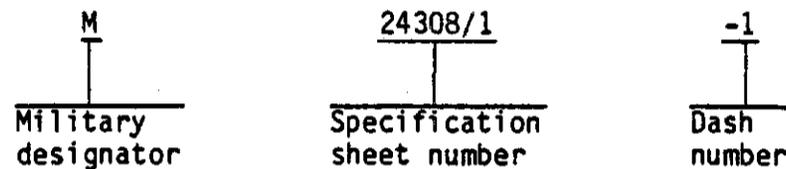


FIGURE 10. Polarized rack and panel connectors, MIL-C-24308.

13.4 CONNECTORS, RACK AND PANEL, MIL-C-24308

13.4.4 Military designation. Polarized rack and panel connectors meeting MIL-C-24308 are identified by military specification sheets. The military part number consists of the letter M, the basic number of the specification slash sheet, and a specific part number as follows:



13.4.5 Electrical characteristics.

13.4.5.1 Contact resistance limits. The electrical resistance of mated pairs of pin and socket contacts is controlled by the test methods and requirements of MIL-C-24308. The maximum allowable voltage drop across the mated pair of contacts at the required test current is specified in MIL-C-24308.

13.4.5.2 Dielectric withstanding voltage. The connectors are designed to show no evidence of flashover between contacts or contacts and shell at the voltages specified in MIL-C-24308.

13.4.5.3 Insulation resistance. The rack and panel connectors are designed to have an insulation resistance of not less than 5,000 megohms under normal conditions. After being subjected to 10 days humidity conditioning, minimum insulation resistance of 1 megohm must be met.

13.4.6 Environmental considerations.

13.4.6.1 Operating temperature range. These connectors are suitable for operation throughout a temperature range of -55 to +125 °C.

13.4.6.2 Vibration. These connectors are designed to perform, when properly mated, over the range of mechanical vibration as specified in MIL-C-24308 without cracking, breaking or loosening of parts and to exhibit no loss of electrical continuity for the specified time period.

13.4.6.3 Mechanical shock. When properly mated, these connectors will perform without cracking, breaking or loosening of parts when subjected to the mechanical shock as specified in MIL-C-24308.

13.4.6.4 Moisture resistance. These connectors are not designed to seal against moisture accumulation. Condensation can form on contacts. The materials used will maintain 1000 megohm insulation resistance after exposure to high humidity, but the connectors are not tested for high potential at a high humidity state. They are tested for high potential only when no surface moisture is present.

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**13.4 CONNECTORS, RACK AND
PANEL, MIL-C-24308**

13.4.7 Reliability considerations. This connector design is produced by many suppliers of commercial connectors. An emphasis on quality control by both the connector vendor and the user is desirable for these types of connectors which are to be used in aerospace systems. Connectors of this design have been used in commercial and military systems for many years, and the design is considered to be mature. Some problems that occasionally arise are a loss of electrical continuity in mated contact pairs and dimensions which are out of tolerance.

Electrical failures of the contacts are most often caused by contamination, improper heat treatment of the socket springs, or dimensions which are out of tolerance.

MIL-HDBK-978-B (NASA)

**13.5 CONNECTORS, CIRCULAR,
POWER, MIL-C-5015**

13.5 Circular, power, MIL-C-5015.

13.5.1 Introduction. This specification covers circular electrical connectors with solder or removable crimp contacts (both front and rear release).

The connector types chosen for standardization under this specification are those with rear release crimp removable contacts and those with non-removable solder type contacts.

The rear release crimp contact series connector is available in two classes as follows:

Class L - Fluid resistant

W - General purpose

The solder type contact series connector is available as follows:

Class R - Grommet seal

13.5.2 Usual applications. These connectors are generally used in high-power designs, such as power supplies. A contact range from size 0 to 16 is offered. Due to their high power capabilities, contact arrangements are less dense than in other connectors used in digital designs where space is critical and high power is not a design factor.

13.5.3 Physical construction. Coupling is accomplished with a threaded coupling nut on the plug which mates with threads on the receptacle. Insert alignment is assured through the use of key and keyway locators. The connectors are designed so that polarization of the mated pairs occurs prior to coupling.

Receptacles are available in cable-connecting, wall mount, box mount, and jam nut configurations.

Some accessories which are available for use with these connectors include cable clamps, shield terminations, 90 °C and 45 °C backshells, strain reliefs and stowage connectors.

13.5.3.1 Connector configurations. Typical connector configurations are as shown in Figures 11 through 15.

13.5.3.2 Contact arrangement. The contact positions are delineated in the applicable military standard drawing.

13.5.3.3 Contacts. Contacts conform to MIL-C-39029. Contacts are designed to prevent damage to the contact retention device and sealing device during insertion or removal of the contact and are designed to insure proper mating with their counterpart without damage.

MIL-HDBK-978-B (NASA)

**13.5 CONNECTORS, CIRCULAR,
POWER, MIL-C-5015**

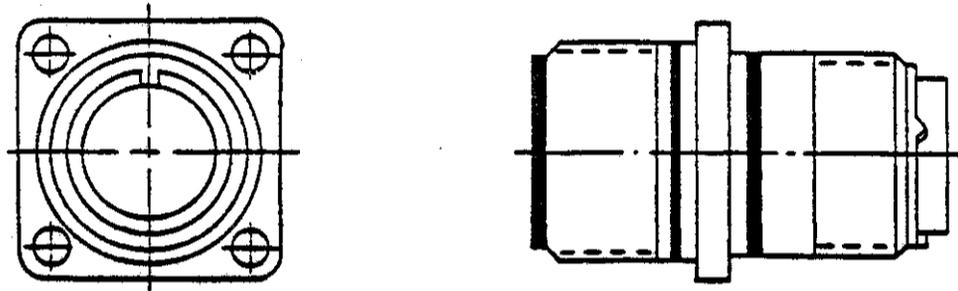


FIGURE 11. Wall mounting receptacle, MIL-C-5015.

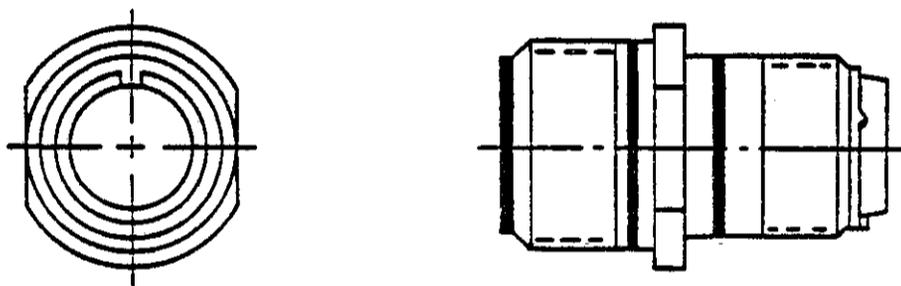


FIGURE 12. Cable-connecting receptacle, MIL-C-5015.

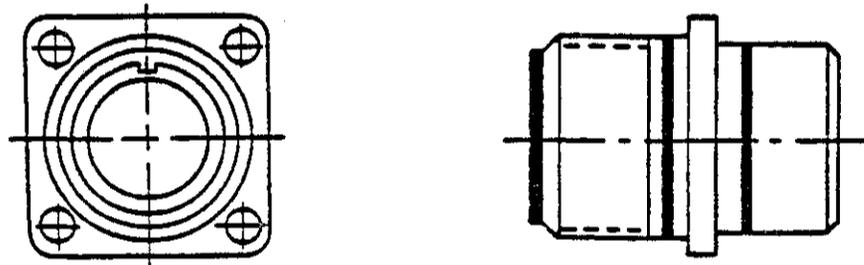


FIGURE 13. Box mounting receptacle, MIL-C-5015.

MIL-HDBK-978-B (NASA)

**13.5 CONNECTORS, CIRCULAR,
POWER, MIL-C-5015**

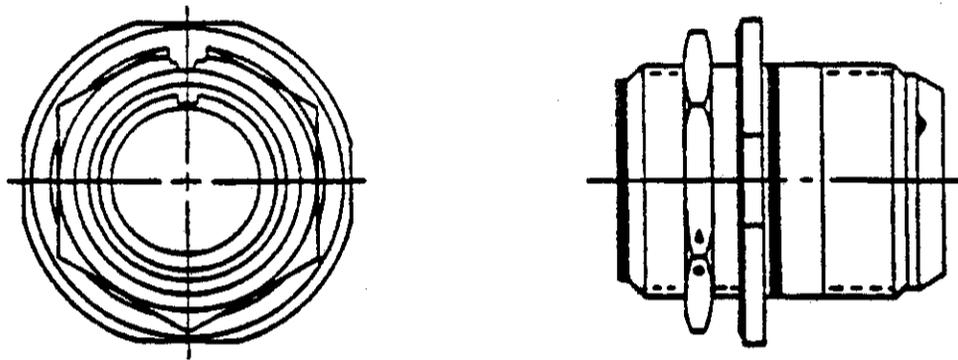


FIGURE 14. Jam nut receptacle, MIL-C-5015.

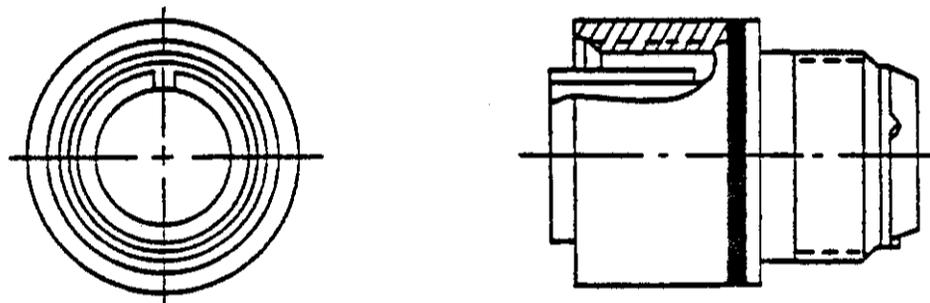


FIGURE 15. Straight plug, MIL-C-5015.

Contact insertion and removal tools are listed in MIL-C-5015. Crimp tools conform to M22520 series and applicable military standard drawings.

Crimp contacts are designed to permit individual insertion and removal.

13.5.3.4 Insert design and construction. Inserts are of voidless construction and are secured to prevent rotation within the shell. Removable inserts are keyed to prevent rotation. Slots and markings to allow for alternate clocking positions are as indicated in the applicable military standard. Inserts are installed in the position indicated by the part number.

13.5.3.5 Engagement seal. Connectors with pin inserts have an interfacial seal with raised sealing barriers around each pin contact. The barriers seal off moisture upon contact with the socket insulator mating face. Connector plug shells are provided with a static peripheral seal to insure shell to shell sealing.

MIL-HDBK-978-B (NASA)

**13.5 CONNECTORS, CIRCULAR,
POWER, MIL-C-5015**

13.5.3.6 Wire range accommodations. The proper insulated wire diameters must be selected to assure an adequate moisture seal around the wire as it penetrates the rear grommet. The proper insulated wire diameters are listed in MIL-C-5015.

13.5.4. Military designation. The part number for qualified connectors procured in accordance with MIL-C-5015 conforms to the following example. Basic drawings provide detail information for connectors supplied to MIL-C-5015.

Example MS3400W 18 - 10PW

MS3400	W		18	10	P	W
Basic part no.	Class	Material designator	Shell size	Insert arrangement	Contact style	Insert position
	L=High temp, Fluid resistant W=General purpose R=Grommet seal	S=stainless steel Blank=aluminum				

13.5.5 Electrical characteristics.

13.5.5.1 Contact resistance. Contact resistance for mated contacts is often expressed as a voltage drop measured across the mated contact pair. The maximum allowable voltage drop is specified in MIL-C-5015 for various combinations of contact size, wire gauge, and test current.

13.5.5.2 Working voltage. The maximum working voltage at sea level is specified in MIL-C-5015.

13.5.5.3 Test voltages. The maximum test voltages at sea level and altitude (70,000 ft.) are specified in MIL-C-5015.

13.5.5.4 Insulation resistance. Insulation resistance decreases with increasing temperature; the requirements at elevated temperature are specified in MIL-C-5015. Requirements for MIL-C-5015 connectors vary between classes and insulation resistance values are specified in MIL-C-5015.

13.5.6 Environmental considerations.

13.5.6.1 Temperature. These connectors are capable of continuous operation within a temperature range as follows:

Class L: -55 to +200 °C

R: -55 to +125 °C

W: -55 to +125 °C

**13.5 CONNECTORS, CIRCULAR,
POWER, MIL-C-5015**

The expected service life decreases with increasing temperature. Service life and hot spot temperature curves are presented in MIL-C-5015.

MIL-C-5015 connectors are considered to be environment resistant. The performance requirements, including the environmental qualification test requirements, are listed in MIL-C-5015.

13.5.7 Reliability considerations. These connectors have been widely used in military systems for many years, and the design is considered to be mature. Some problems that occasionally arise are contact retention failures, difficulty in intermating connectors supplied from different vendors, and loss of electrical continuity in mated contact pairs.

Contact retention failures are most often due to improper assembly techniques or faulty contact retention systems in the connector. Operator training is usually sufficient to eliminate problems due to improper assembly techniques. Increased attention to manufacturing processes and quality control will generally eliminate faults in the contact retention system.

Difficulty in mating connector halves which have been supplied from different vendors is often traced to dimensions which are out of tolerance.

Electrical failures of the contacts are most often caused by contamination, improper heat treatment of the socket springs, or dimensions which are out of tolerance.

MIL-HDBK-978-B (NASA)

**13.6 CONNECTORS, PRINTED
CIRCUIT, MIL-C-55302**

13.6 Printed circuit, MIL-C-55302.

13.6.1 Introduction. This section covers connectors and their accessories for printed circuit assembly.

13.6.2 Usual applications. These printed circuit connectors are for use with single-sided, double-sided and multilayer printed wiring boards conforming to MIL-STD-275, MIL-P-55110 and MIL-P-55424.

These connectors are designed for printed-wiring-board to printed-wiring-board or printed-wiring-board to cable interconnections of miniaturized equipment sub-assemblies with low power requirements. These connectors have an operating temperature range of -65 to +125 °C.

13.6.3 Physical construction. Figures 16 through 19 illustrate various types of printed circuit connectors.

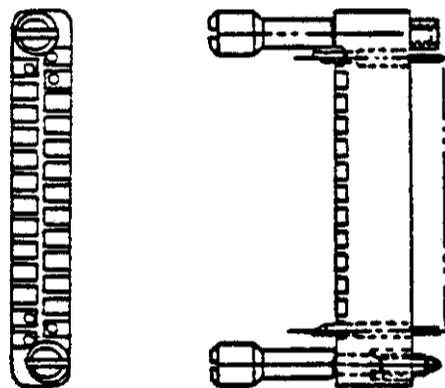


FIGURE 16. Plug with pin contacts and turning jackset shown (also available with socket contacts and alternate mounting hardware), MIL-C-55302.

MIL-HDBK-978-B (NASA)

13.6 CONNECTORS, PRINTED
CIRCUIT, MIL-C-55302

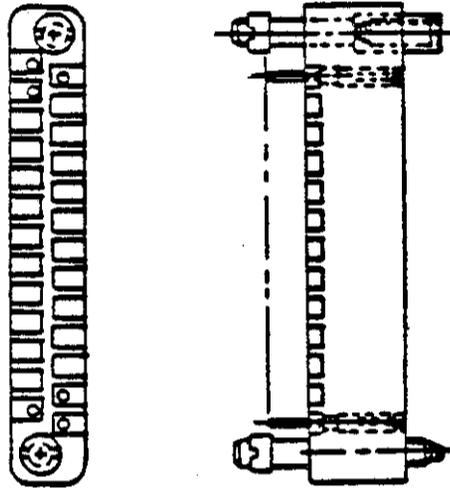


FIGURE 17. Receptacle with socket contacts and fixed jackset shown (also available with pin contacts and alternate mounting hardware), MIL-C-55302.

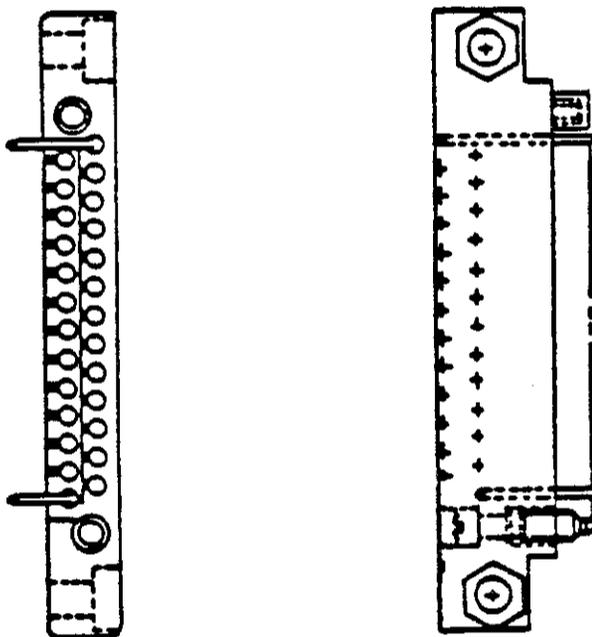


FIGURE 18. Right angle plug with pin contacts and fixed jackset shown (also available with alternate mounting hardware), MIL-C-55302.

MIL-HDBK-978-B (NASA)

**13.6 CONNECTORS, PRINTED
CIRCUIT, MIL-C-55302**

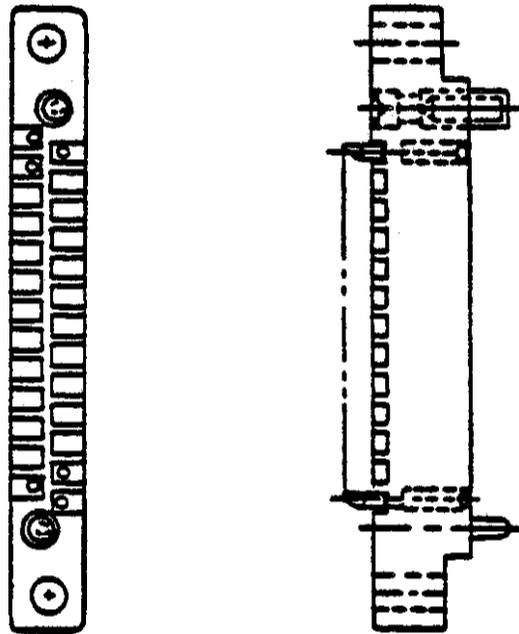


FIGURE 19. Receptacle with socket contacts, round guideset, and mounting ears (also available with alternate mounting hardware), MIL-C-55302.

13.6.3.1 Bodies. The printed circuit connector consists of two plastic bodies containing pin and socket contacts. They contain integral aligning features to assure proper mating of the connectors.

13.6.3.2 Contacts. Contacts may be either removable crimp type or solder type. The contact pairs have sufficient compliance to assure proper mating.

The contacts are designed to assure that the mating and unmating forces transmitted to the connection joining the contact to the board will not degrade the connection through usage.

The connectors covered herein have contacts spaced at 0.100 inch.

13.6 CONNECTORS, PRINTED
CIRCUIT, MIL-C-55302

13.6.3.3 Polarization. Polarization is accomplished through jackscrew or guide pin hardware, to ensure correct mating of the plug and receptacle.

13.6.3.4 Terminations. Figures 20 thru 23 illustrate various types of terminations available for these connectors.

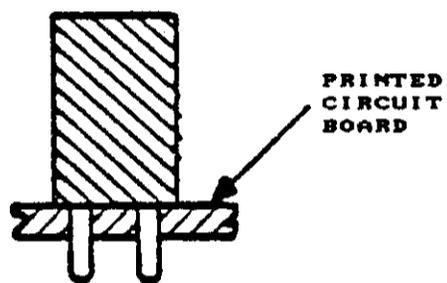


FIGURE 20. Dip solder or flex circuit termination.

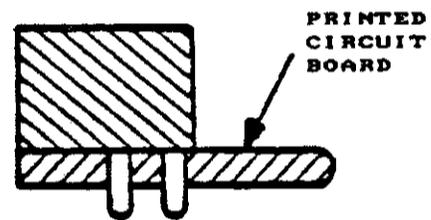


FIGURE 21. Dip solder termination (right angle).



FIGURE 22. Solder cup.

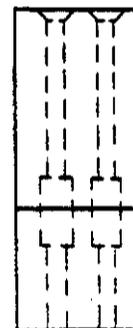


FIGURE 23. Removable crimp contacts.

MIL-HDBK-978-B (NASA)

**13.6 CONNECTORS, PRINTED
CIRCUIT, MIL-C-55302**

13.6.3.5 Connector assembly. Table IV identifies the connector by specification sheet number and associates it with its configuration type, contact quantities, and terminations available. Refer to MIL-STD-975 for connector selections.

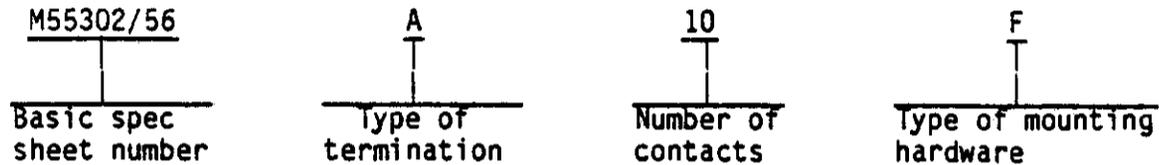
TABLE IV. Connector configurations

Spec Sheet Number	Configuration Figure No.	Contact Quantities	Termination Type-Figure No.			
			20	21	22	23
M55302/55	16	10 - 70	X		X	
M55302/56	17		X		X	
M55302/57	18			X		
M55302/58	19		X		X	
M55302/59	18	90 - 120		X		
M55302/60	19		X		X	
M55302/61	18	10 - 70		X		
M55302/62	17		X		X	
M55302/63	16		X		X	
M55302/64	17		X		X	
M55302/65	19					X
M55302/66	17					X

MIL-HDBK-978-B (NASA)

13.6 CONNECTORS, PRINTED
CIRCUIT, MIL-C-55302

13.6.4. Military designation. Qualified connectors are procured through military specification sheets. The following is an example of a complete number:



13.6.5 Electrical characteristics. The electrical characteristics are as specified in Table V.

TABLE V. Electrical characteristics

Spec Sheet Number	Dielectric Withstanding Voltage		Contact Resistance		Contact Current Rating
	Sea Level	Altitude	Individual	Mated Pair	
M55302/55 thru M55302/66	1000 V(RMS)	300 V(RMS)	0.010 Ohm	0.020 Ohm	5.0 Amps maximum

13.6.6 Environmental characteristics. These connectors are not designed to be environment resistant. The operating environment for these connectors is specified in MIL-C-55302.

13.6.7 Reliability considerations. These connectors are widely used in military and aerospace systems. Some problems that occasionally arise are contact retention failures, loss of electrical continuity in mated contact pairs, and dimensions which are out of tolerance.

Contact retention failures are most often due to improper assembly techniques or faulty contact retention systems in the connector. Operator training is usually sufficient to eliminate problems due to improper assembly techniques. Increased attention to manufacturing processes and quality control will generally eliminate faults in the contact retention system.

Electrical failures of the contacts are most often caused by contamination, improper heat treatment of the socket springs, or dimensions which are out of tolerance.

**13.7 CONNECTORS, CONTACTS, ELECTRICAL
CONNECTOR, MIL-C-39029**

13.7 Contacts, electrical connector, MIL-C-39029.

13.7.1 Introduction. These contacts are removable crimp type pin and socket contacts intended for use in electrical connectors.

13.7.2 Usual applications. These contacts are normally supplied with the connector with which they are intended to be used. Removable contacts are also available separately, and may be obtained under the military part number.

13.7.3 Physical construction.

13.7.3.1 Typical configurations. MIL-C-39029 contacts are available in pin and socket configurations as shown in Figures 24 and 25. Wires are terminated to the contacts by means of a crimp joint.

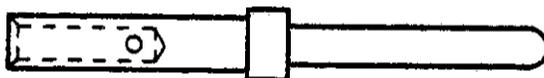


FIGURE 24. Pin contact with crimp termination, MIL-C-39029.

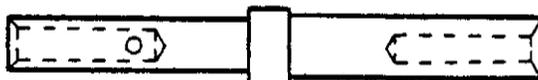
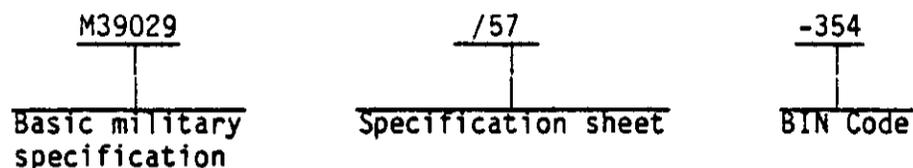


FIGURE 25. Socket contact with crimp termination, MIL-C-39029.

MIL-HDBK-978-B (NASA)

**13.7 CONNECTORS, CONTACTS, ELECTRICAL
CONNECTOR, MIL-C-39029**

13.7.4 Military designation. The military part number for qualified contacts in accordance with MIL-C-39029 conforms to the following example:



13.7.5 Electrical characteristics. Electrical characteristics are in accordance with the provisions of MIL-C-39029 and the applicable military connector specifications.

13.7.6 Environmental considerations. The environmental considerations for the contacts follow the environmental considerations for the electrical connectors in which the contacts are intended to be installed. The environment resistant characteristics of the contacts are covered by the environment resisting requirements of the applicable connector specifications and MIL-C-39029.

13.7.7 Reliability considerations. These contacts are widely used in many connectors. The most common contact failures are due to loss of electrical continuity in mated contact pairs. These failures are most often caused by contamination, improper heat treatment of the socket springs, or dimensions which are out of tolerance. Increased awareness of contamination control during assembly and use of the connector can alleviate these problems. Other common contact failures are caused by plating defects and improper hardness of the crimp barrel.

MIL-HDBK-978-B (NASA)

**13.8 CONNECTORS, CIRCULAR,
HEAVY DUTY, MIL-C-22992**

13.8 Circular, heavy duty, MIL-C-22992.

13.8.1 Introduction. This specification covers heavy duty waterproof circular connectors with threaded coupling mechanisms and solder contacts.

Available service classes include Class C (pressurized) and Class R (environment resisting).

Plugs are available in two configurations. Straight plugs are normally used to make cable to box connections. Cable connecting plugs without coupling rings (that perform the same function as in-line, cabled receptacles) are used for making cable to cable interconnections.

Receptacles are available with wall flange mounts, box flange mounts, wall jam-nut mounts, and box jam-nut mounts.

Available accessories include protective caps, stowage receptacles, and cable sealing adapters.

13.8.2 Usual applications. The connectors are intended for use on ground support equipment in environment resisting applications. The connectors are waterproof and are suited for use in high humidity environments. The connectors are designed to withstand abuse due to rough handling.

These connectors are designed to have an operating temperature range of -55 to +125 °C.

13.8.3 Physical construction. Figures 26 through 31 illustrate various MIL-C-22992 connectors.

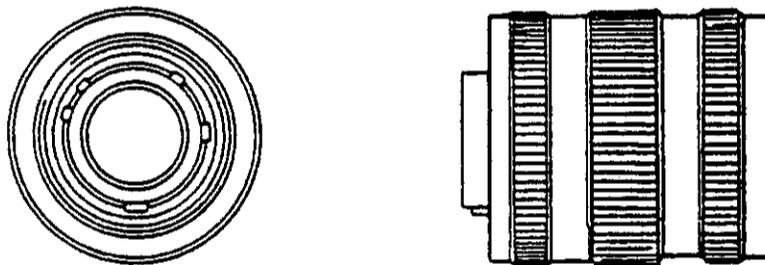


FIGURE 26. Straight plug, MIL-C-22992.

MIL-HDBK-978-B (NASA)

**13.8 CONNECTORS, CIRCULAR,
HEAVY DUTY, MIL-C-22992**

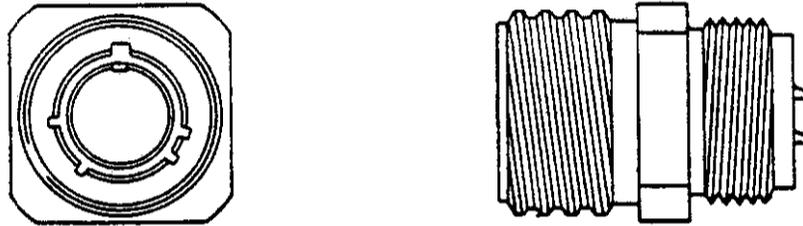


FIGURE 27. Cable connecting plug without coupling ring
(cabled receptacle), MIL-C-22992.

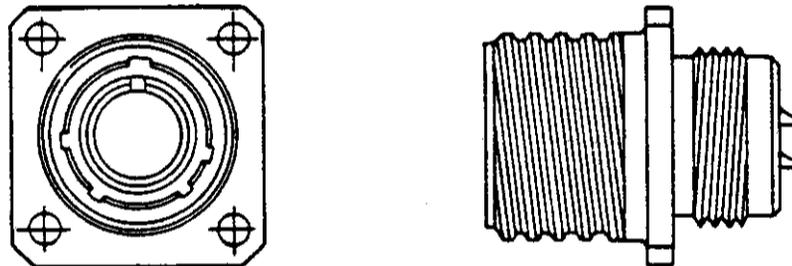


FIGURE 28. Wall mount receptacle, MIL-C-22992.

MIL-HDBK-978-B (NASA)

**13.8 CONNECTORS, CIRCULAR,
HEAVY DUTY, MIL-C-22992**

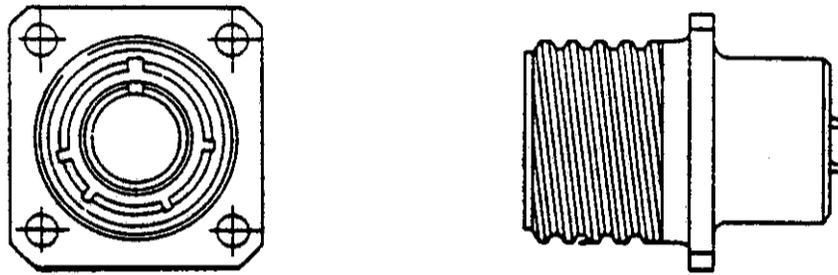


FIGURE 29. Box mount receptacle, MIL-C-22992.

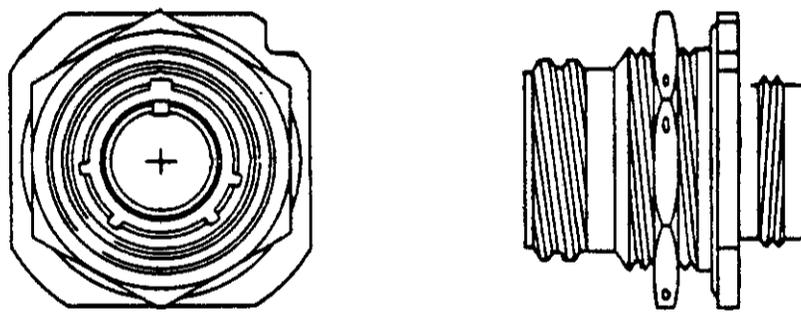


FIGURE 30. Wall jam-nut receptacle, MIL-C-22992.

MIL-HDBK-978-B (NASA)

**13.8 CONNECTORS, CIRCULAR,
HEAVY DUTY, MIL-C-22992**

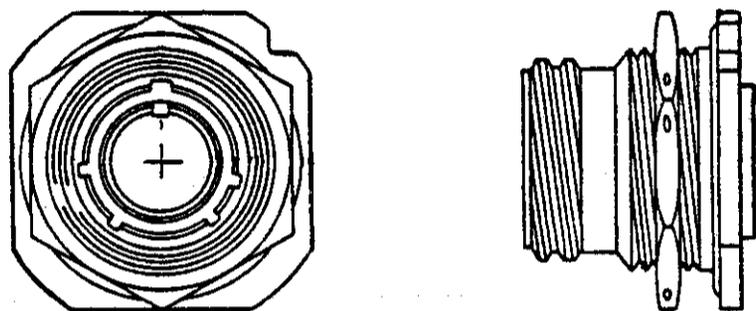


FIGURE 31. Box jam-nut receptacle, MIL-C-22992.

13.8.3.1 Insert design and construction. Inserts are of voidless construction and are secured to prevent rotation within the shell. The inserts are not removable from the shell for Classes C and R. The inserts are installed in the position specified on the applicable Military Standard.

13.8.3.2 Contacts. Solder contacts are utilized for Class C and Class R connectors. Contact sizes 0, 4, and 8 may be designed so that the contacts may be removed from the connector to facilitate soldering. Contact sizes 12 and 16 are not removable from the connector. Contact design is in accordance with MIL-C-22992 and MIL-C-39029.

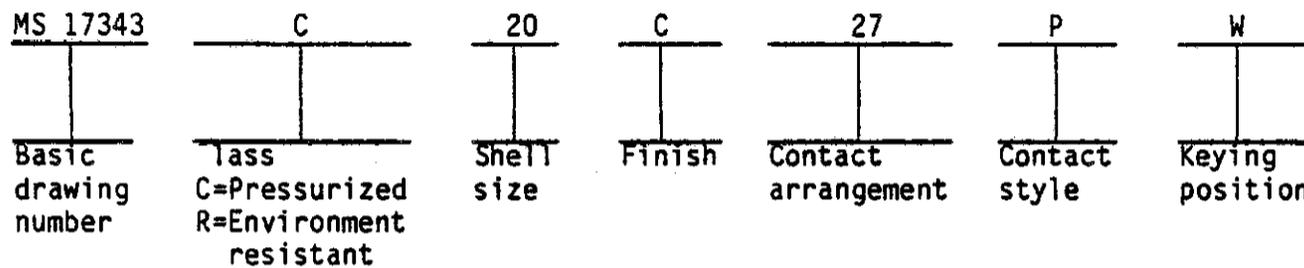
13.8.3.3 Coupling. Coupling is accomplished by clockwise rotation of the coupling ring. Uncoupling is accomplished by counterclockwise rotation of the coupling ring. The coupling rings are knurled or fluted to provide a gripping surface.

13.8.3.4 Polarization. Polarization is accomplished by matched integral keys and keyways of counterpart connectors. Only connectors with matching keys and keyways will mate. To assure proper contact alignment during connector mating, polarization and engagement of the shells occurs prior to engagement of the contacts.

MIL-HDBK-978-B (NASA)

**13.8 CONNECTORS, CIRCULAR,
HEAVY DUTY, MIL-C-22992**

13.8.4 Military designation. The military standard part numbers for qualified connectors procured in accordance with MIL-C-22992 conform to the following example. Basic drawings provide detailed information for connectors supplied to MIL-C-22992.



13.8.5 Electrical characteristics.

13.8.5.1 Contact resistance. The electrical resistance of mated pairs of pin and socket contacts is controlled by the test methods and requirements of MIL-C-22992. The maximum allowable voltage drop across the mated contact pair is specified in MIL-C-22992 for the applicable contact size.

13.8.5.2 Insulation resistance. The insulation resistance of MIL-C-22992 connectors is 5,000 megohms when measured in accordance with MIL-C-22992.

13.8.6 Environment considerations.

13.8.6.1 Temperature. These connectors are rated for operation within a temperature range of -55 to +125 °C. The upper temperature limit is the maximum hot spot temperature resulting from any combination of ambient temperature and the temperature rise induced by electrical loads.

13.8.6.2 Fluid immersion. To ensure resistance to immersion in fluids, the connector shells are equipped with sealing gaskets. These connectors are designed to withstand the effects of immersion in water, petroleum based hydraulic fluid, and aircraft lubricating oil.

13.8.7 Reliability considerations. These connectors have been widely used in heavy duty ground support equipment. Some problems that occasionally arise are improper keying or clocking of the connectors, difficulty in intermating connectors supplied from different vendors, and loss of electrical continuity in mated contact pairs.

Keying, clocking, and intermateability problems can be eliminated with better control over the manufacturing process and increased emphasis on quality control.

Electrical failures of the contacts are most often caused by contamination, improper heat treatment of the socket springs, or dimensions that are out of tolerance.

MIL-HDBK-978-B (NASA)

**13.9 CONNECTORS, CIRCULAR,
NASA 40M SERIES**

13.9 Circular, NASA 40M series.

13.9.1 Introduction. These specifications cover environment resistant and hermetic bayonet coupling circular connectors with removable crimp or solder type contacts.

Receptacles are available with flange mounts, solder mounts, jam-nut mounts, and thru-bulkhead jam-nut mounts.

These connectors are available in accordance with the following specifications:

40M 38277	Connectors, Electrical, Miniature, High Density, Environment Resisting
40M 38298	Connectors, Electrical, Special, Miniature, Circular, Environment Resisting, 200 °C
40M 39569	Connectors, Electrical, Miniature Circular, Environment Resisting, 200 °C

13.9.2 Usual applications. These connectors are intended for use on space vehicles while exposed to earth atmosphere, space vacuum, or crew compartment environments. These connectors are similar to the connectors specified by MIL-C-26482 and MIL-C-38999, except for the additional requirements which make these devices suitable for operation in a space environment.

13.9.3 Physical construction. Figures 32 through 36 illustrate various 40M series circular connectors.



FIGURE 32. Straight plug.

MIL-HDBK-978-B (NASA)

**13.9 CONNECTORS, CIRCULAR,
NASA 40M SERIES**

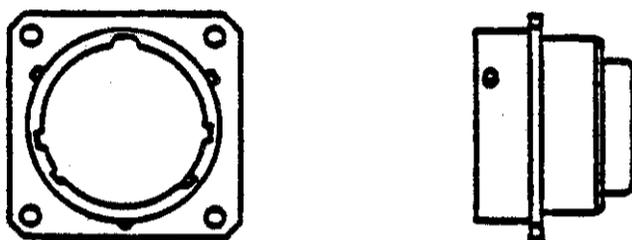


FIGURE 33. Receptacle with flange mount.

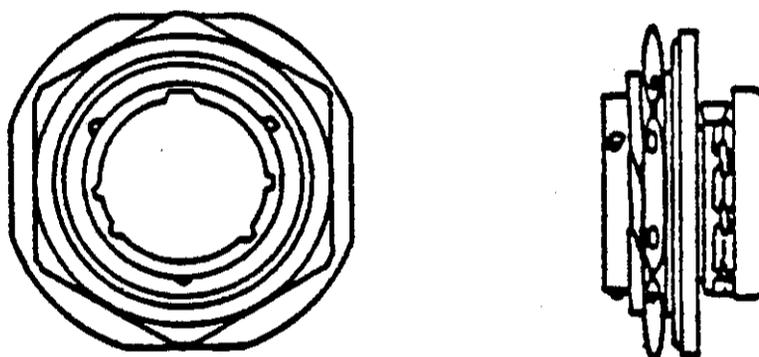


FIGURE 34. Jam-nut mount receptacle.

MIL-HDBK-978-B (NASA)

**13.9 CONNECTORS, CIRCULAR,
NASA 40M SERIES**

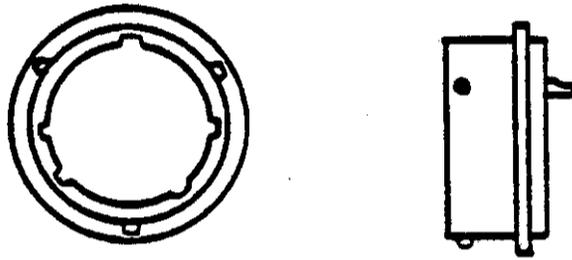


FIGURE 35. Solder mount receptacle.

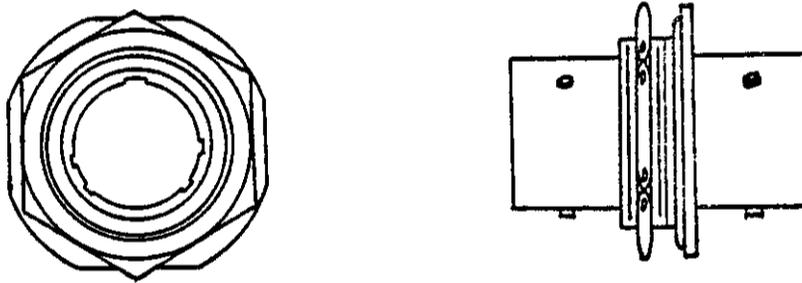


FIGURE 36. Bulkhead receptacle.

MIL-HDBK-978-B (NASA)

**13.9 CONNECTORS, CIRCULAR,
NASA 40M SERIES**

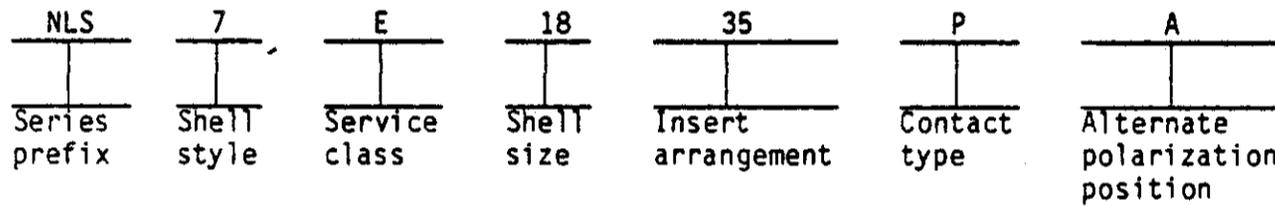
13.9.3.1 Flammability, odor, and outgassing. The connectors are capable of meeting the NASA requirements for flammability, odor, and outgassing for spaceflight applications.

13.9.3.2 Explosive atmosphere. The connectors are capable of operating in an explosive atmosphere without burning or igniting the atmosphere. The connectors are capable of meeting the explosive atmosphere requirements as specified in the NASA 40M series specifications.

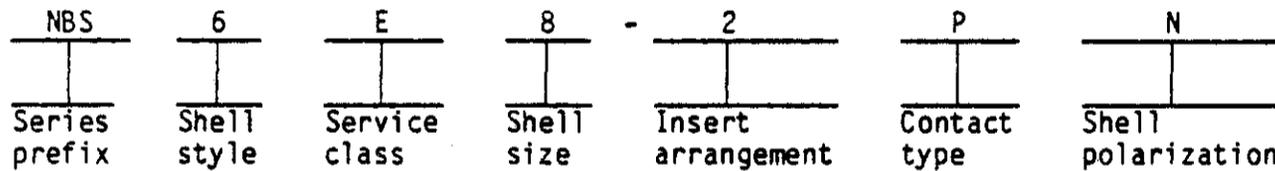
13.9.3.3 Stress corrosion prevention and control. The materials and processes used in the construction of the connectors are controlled in order to prevent and control the occurrence of stress corrosion.

13.9.4 MSFC designations. The Marshall Space Flight Center part numbers for connectors procured in accordance with the following specifications are as indicated in the following examples:

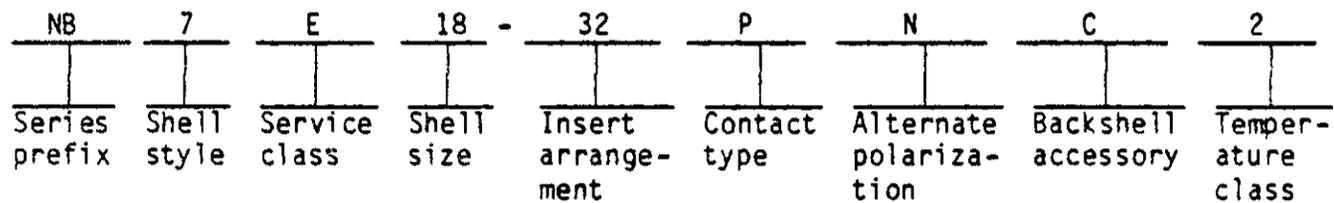
40M 38277



40M 38298



40M 39569



MIL-HDBK-978-B (NASA)

**13.9 CONNECTORS, CIRCULAR,
NASA 40M SERIES**

13.9.5 Electrical characteristics. The electrical characteristics are similar to those for MIL-C-26482 and MIL-C-38999 connectors, and are as specified in the 40M specifications.

13.9.6 Environmental considerations. The environment resisting characteristics are similar to those for MIL-C-26482 and MIL-C-38999 connectors. Additional requirements as outlined in 13.9.3 herein and as specified in the 40M specifications ensure that these connectors are suitable for space flight applications.

13.9.7 Reliability considerations. These connectors are similar to those as specified in MIL-C-26482 and MIL-C-38999, and are subject to the same reliability considerations see 13.2.7 and 13.3.7.

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**13.10 RACK AND PANEL,
NASA GSFC-S-311 SERIES**

13.10 Rack and panel, NASA GSFC-S-311 series.

13.10.1 Introduction. Connectors covered in this section are rectangular and nonenvironment-resisting. These connectors may be mated by being pushed together without the aid of a coupling mechanism or by use of screwlock hardware. Proper connector orientation during mating is accomplished by a polarized shell. Mounting holes provided in the connector shell, or screwlock hardware may be used to mount the connector to a rack or panel.

13.10.2 Usual applications. These connectors may be used with cables and are available with hoods and cable clamps to protect the assembled cable.

As their description suggests, these connectors are primarily used where a rack or drawer must plug into a panel. The mounting hardware and polarizing shells act as guides. These connectors are also available with float mounting hardware to aid in mating.

GSFC S-311-P-4/07 connectors have high density contact arrangements, and are intended for spaceflight use. These connectors utilize removable crimp contacts and are available with two levels of residual magnetism.

A = residual magnetism not specified
B = 200 gamma

GSFC S-311-P-4/09 connectors have standard density contact arrangements, and are intended for spaceflight use. These connectors utilize removable crimp contacts and are available with two levels of residual magnetism.

A = residual magnetism not specified
B = 200 gamma

GSFC S-311-P10 connectors are intended for spaceflight use and are available with nonremovable coaxial, high voltage and solder type power contacts. These connectors are available with three levels of residual magnetism.

B = 200 gamma
C = 20 gamma
D = 2 gamma

**13.10 RACK AND PANEL,
NASA GSFC-S-311 SERIES**

13.10.3 Physical construction. A typical pair of polarized shell rack and panel connectors is shown in Figure 37.



FIGURE 37. Polarized rack and panel connectors, GSFC-S-311.

13.10.4 NASA designation. High density polarized shell rack and panel connectors meeting GSFC-S-311-P-4/07 are identified by the GSFC part number as shown below:

311P407	-	1	-	P	-	B	-	12
GSFC		Contact		Contact		Residual		Mounting
Prefix		arrangement		type		magnetism		hole diameter

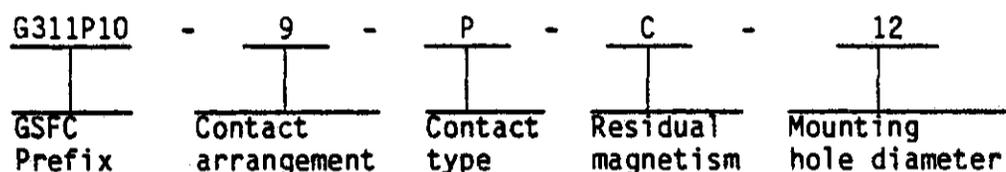
Standard density polarized shell rack and panel connectors meeting GSFC-S-311-P-4/09 are identified by the GSFC part number as shown below:

311P409	-	1	-	P	-	B	-	12
GSFC		Contact		Contact		Residual		Mounting
Prefix		arrangement		type		magnetism		hole diameter

MIL-HDBK-978-B (NASA)

**13.10 RACK AND PANEL,
NASA GSFC-S-311 SERIES**

Polarized shell rack and panel connectors with coaxial, high voltage and power contacts meeting GSFC-311-P-10 are identified by the GSFC part number as shown below:



13.10.5 Electrical characteristics. The electrical characteristics are similar to those for MIL-C-24308 connectors, and are specified in the GSFC-S-311 specifications.

13.10.6 Environmental considerations. The environmental resistor characteristics are similar to those for MIL-C-24308 connectors. These connectors are not designed to seal against the environment.

13.10.7 Reliability considerations. This connector design has been produced in a similar military version (MIL-C-24308) by many suppliers for many years, and the design is considered to be mature. Some problems that occasionally arise are a loss of electrical continuity in mated contact pairs and dimensions which are out of tolerance.

Electrical failures of the contacts are most often caused by contamination, improper heat treatment of the socket springs, or dimensions which are out of tolerance.

14.1 CONNECTORS, GENERAL

14. CONNECTORS, COAXIAL, RADIO FREQUENCY

14.1 General.

14.1.1 Introduction. This section contains information covering coaxial, radio frequency (rf) connectors included in MIL-STD-975 NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List for use in Grade 2 applications.

Early military specifications lacked definition of the connectors in terms of their rf characteristics. Though dimensions were spelled out, manufacturing controls did not guarantee rf performance because variations in dielectric materials and tolerance fits of metal parts in the assembly can completely alter rf behavior. As a result, many users issued specifications reflecting their particular application needs; thus, making standardization impossible, creating confusion, and making correct field replacement extremely difficult.

MIL-C-39012 is a rf connector specification which defines the electrical characteristics and delineates the mating face dimensions. Recent additions to this specification are categories C and D which standardize cable strip dimensions and in some types, parts connecting to the cable are defined.

The rf connectors can be classified into specific series using several criteria such as size, coupling method, electrical characteristics, and applications such as high voltage or close impedance. The connectors in this section are classified by size.

14.1.1.1 Applicable military specifications. MIL-C-39012, Series, N, TNC, and SMA.

14.1.2 General definitions. The following is a list of terms which are commonly used with rf connectors. Refer to Section 13 Panel Connectors, for additional definitions applicable to connectors in general.

Jack. A jack is the mating unit for a plug and all mating features fit inside the plug when mated. It is also essential to have means for securing it to a cable. There are three types of jacks. A cable jack is secured to the end of a cable. A panel jack is secured to a cable but mounted on a panel by a square flange. A bulkhead jack is mounted on a panel by means of a shoulder and a hex nut.

Plug. The term plug defines the mating characteristics and can be broadly stated as that unit which, when mated, encompasses or fits over its mate. It is attached to the end of the cable and usually has no other means of mounting. It is called a cable plug although it is sometimes referred to as the male connector because it generally contains the male contact.

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14.1 CONNECTORS, GENERAL

Receptacle. A receptacle is similar to a jack without the cable clamping parts. The receptacle is open wired with the center conductor soldered onto the unit. The receptacle is mounted to a panel by means of a mounting flange or to a bulkhead by means of a shoulder and hex nut. Receptacles usually have female center contacts.

Semirigid cable connectors. This term refers to a connector that accepts a coaxial cable with a metallic outer jacket. Copper is commonly used for the outer jacket. However, it is also available in aluminum and stainless steel. The dielectric is polytetrafluoroethylene (Teflon) and the inner conductor is silver-plated solid copper wire.

Shorting plugs. A shorting plug is a unit with the mating features of a plug or receptacle which will short out the center contact to the body of the connector. A nonshorting plug is similar to the shorting plug without the center contact and has an added insulating backing disk.

14.1.3 General device characteristics. Series TNC, N, and SMA connectors are covered separately in subsection 14.2.

14.1.4 General parameter information. This subsection contains information on four electrical parameters generally associated with rf connectors as opposed to multipin connectors. The parameters are voltage standing wave ratio, insertion loss, rf leakage, and corona level.

14.1.4.1 Voltage standing wave ratio (VSWR). Any impedance mismatch in an rf transmission line will reflect energy that, in turn, will set up standing waves along the line with maximums and minimums occurring at one-quarter wavelength intervals. The VSWR is the ratio of the maximum to the minimum of these standing wave voltages.

The rf connectors are usually a compromise design in which one desirable characteristic is sacrificed to some extent in favor of another. This is especially true with voltage rating and VSWR. For example, a 5000 V connector does not have the broadband VSWR performance of a 500 V connector. VSWR characteristics are usually not measured for high-voltage connectors. On the other hand, a broadband connector with a VSWR of 1.2 at 10 GHz would not have high voltage (5000 V) characteristics. Long leakage paths in high-voltage connectors usually upset broadband VSWR performance due to characteristic impedance change.

14.1.4.2 Rf insertion loss. Insertion loss of a mated connector pair is defined as the increase of a loss due to the insertion of a mated connector pair in a cable. This includes the reflection losses to the cable and the dissipating losses in the pair.

Losses in an rf transmission line can be a design consideration but in practice, losses in connecting cables are large enough to make connector loss insignificant. For example, an N connector has an insertion loss of only 0.05 dB at

14.1 CONNECTORS, GENERAL

10 GHz, but RG-8/U cable has a loss of 1 dB per foot. Similarly, transmission loss of cables usually masks that of a connector. A connector VSWR of 1.35 would result in transmission loss of less than 0.1 dB, as compared with 1 dB per foot cable loss. The insertion loss varies with the square root of the frequency.

14.1.4.3 Rf leakage. Rf leakage describes both the loss of internal energy from a mated connector pair and the entrance of energy into the mated connector pair. Coaxial connectors have rf leakage values of 60 to 90 dB down from the signal level depending on the type of connector. But generally the shielding, even on bayonet-coupled units, is such that connector leakage is less than that introduced into the transmission system by a typical coaxial cable. For example, the leakage of a standard RG-59/U is about 30 dB down. In applications where rf leakage is critical, it is best to use triaxial connectors and cables or precision connectors with solid sheathed cables. The leakage is about 60 dB down for the triaxial equivalent of RG-59/U and at least 120 dB down for the precision connectors.

14.1.4.4 Corona level. Corona is the result of ionization of air. Corona tests are usually specified at 70,000 feet and are used to detect the presence of air pockets in the connector's dielectric and in the insulation in the cable. A high corona level can cause a sustained electrical discharge which can damage the connector and cable assembly.

14.1.5 General guides and charts. Figures 1 and 2, and Table I relate frequency range, voltage rating and RG cables used, respectively, for various connector series.

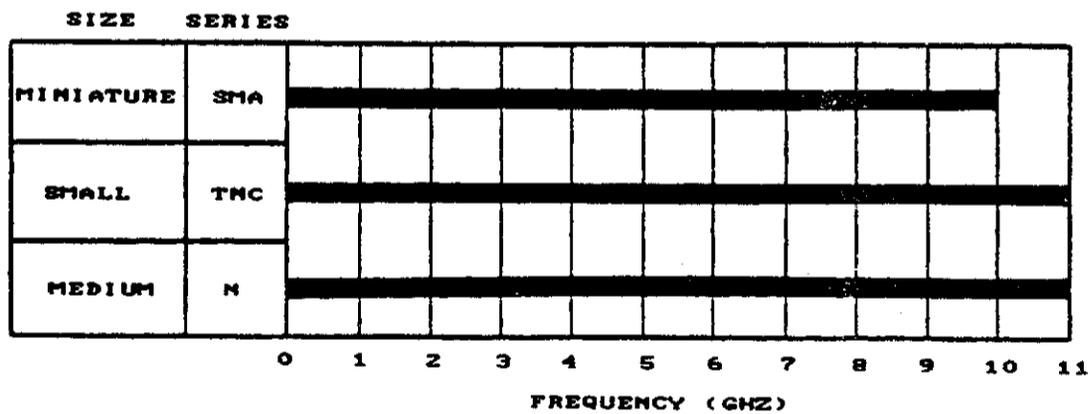


FIGURE 1. Frequency range vs series and size (typical).

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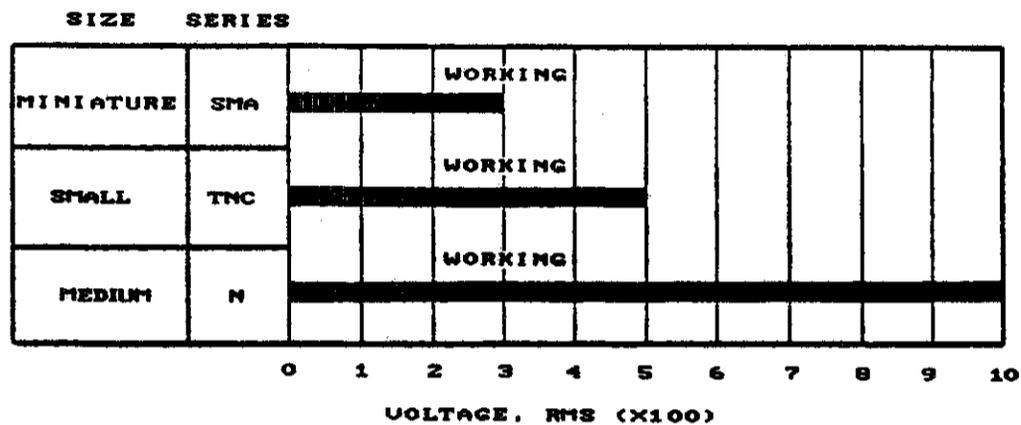


FIGURE 2. Voltage vs series and size (typical) at sea level.

TABLE I. Typical RG cables used with connectors

Size	Series	Cable OD	Typical RG cables used	Coupling	VSWR (typical)	Max Temp ^{1/} (connector)	Weather proof
Mini-ature	SMA	Up to 0.150	RG 178 RG 316	Threaded	1.2	200 °C	Yes
Small	TNC	0.150 to 0.350	RG 58, 141, 142	Threaded	1.3	100 °C	Yes
Medium	N	0.350	RG 213, 214, 215	Threaded	1.35	200 °C	Yes

^{1/} Maximum temperature will depend on the maximum temperature of the cable that is used with the connectors.

14.1.6 General reliability considerations.

14.1.6.1 Design considerations. Generally, one of the following objectives takes precedence of a particular coaxial line connector.

- a. The connector must be designed to have the same characteristic impedance as the mating cable. The objective is to make the connector an electrically homogeneous extension of the cable. This is almost invariably done in the frequency range above 1000 MHz.

14.1 CONNECTORS, GENERAL

- b. The connector must be designed to allow the cable to take its maximum rated power. Expansion, caused by temperature, may cause a discontinuity by separating the cable from its clamping device. Designs using large contact areas should be used. It is also recommended that the center conductor of a connector be mechanically held in a fixed position (captive contact construction).
- c. The connector must be designed to allow the cable to take its full peak voltage rating (e.g., pulse cables). To accomplish this objective, physical discontinuities (steps in conductor diameters, etc.) where high voltage gradients may occur, must be kept to a minimum in number and size. Also, because of their higher electrical strength, dielectrics other than air are preferable throughout the connector. Provisions should be made to avoid the development of air pockets at the mating boundaries of connector pairs.
- d. The coupling mechanism must be highly functional to do the service intended. Where long, massive cables are to be joined, the coupling nut and associated retaining rings must be correspondingly strong. Where frequent movement or vibration is anticipated, the joining must be by strong, positive, vibration-free means.

14.1.6.2 Assembly considerations.

Dielectric joints. Where assembly instructions show butting of the cable and connector dielectrics, every precaution should be taken that the assembly method insures a positive butt. Where the connector is to be used at high frequency, where a low VSWR is a design-assembly aim, the development of air pockets, loose butt joints, or rounded corners of the dielectric will give rise to mismatch which is proportional to frequency.

If the assembly is of a high voltage cable, for pulse applications, air pockets or loose joints materially reduce the peak voltage capability of the entire connector. Loose butt joints, usually develop unless the dielectric trimming process is one of the last assembly operations. The clamping mechanism on the outer conductor should stress the butt joint.

Rounded corners usually develop either due to excess heating during soldering or through a mistaken notion that all "sharp edges should be avoided."

It is extremely important that dielectrics should be cut at a perfect right angle to the inner conductor. No notches should be permitted.

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14.1 CONNECTORS, GENERAL

Air pockets between the inner conductor and the dielectric of the cable usually develop due to excess heat when soldering the center contact of the connector onto the inner conductor of the cable. Some of the dielectric is softened, and through movement of the inner conductor, a larger hole is formed.

Center pin alignment. Precautions should be taken that the center contact of the connector rests at its proper lateral position.

In many connectors, the exact axial distance between a point on the connector shell and the tip of the pin is an electrical matching circuit. This is the case, for example, in Type N connectors where the male pin steps down before entering the female pin, of the mating connector, leaving a deliberate radial notch which is compensated for by the overhung iris in the ID of the outer conductor.

Many times, a misalignment results from assembling connectors to both ends of a relatively long cable while it is coiled. When it is uncoiled, the ends of the center conductor may assume a different position with respect to the ends of the outer braid.

For similar reasons, connectors should not be assembled to a cable, or replaced on a cable, under temperature extremes.

14.1.6.2.1 Soldering operations. In general, joining the center contact of the connector to the inner conductor of the cable constitutes the only soldering operation.

There are two major precautions to observe.

- a. A good solder bond should be made between the pin and the cable inner conductor over the entire length of the cable conductor that extends into the pin. Otherwise, a significant inductive reactance can be created. (The hole in the pin and the cable center conductor can form the conductors of a small, short-circuited coaxial line having significant electrical length at elevated frequencies.)
- b. Excess solder should be removed so that the step contour between the pin and cable conductor corresponds to the original dimensions. Design has generally taken this step into account and compensated for it in the overall design. A change in dimension by excess solder is in effect a circuit change. Excess solder appears as a shunt capacitance.

14.1.6.2.2 Chemical considerations. Contact of dissimilar metals in a connector can result in electrolytic couples which may promote corrosion through galvanic action. For this reason, the finish of the component parts of a connector must be compatible.

The standard base metal on coaxial connectors is brass. The standard plating is gold.

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14.2 CONNECTORS, MIL-C-39012

14.2 MIL-C-39012.

14.2.1 Introduction. This subsection contains information covering rf coaxial connectors which meet the requirements of military specification MIL-C-39012. There are three series, or sizes, of connectors included in MIL-C-39012, denoted by one or more alphabetical characters: N, TNC, and SMA. Miscellaneous hardware, CW & MX, associated with these connectors is also covered in this specification. MX denotes such things as caps, hoods, and armor clamps. CW designates a cover and is used with caps only. These connectors are used with flexible rf cables and certain other types of coaxial transmission lines.

14.2.1.1 Classification. There are two classes of connectors. A Class I connector is intended to provide superior rf performance at specified frequencies and all rf characteristics are defined. A Class II connector is intended to provide a mechanical connection within an rf circuit and it provides specified rf performance.

14.2.1.2 Categories. MIL-C-39012 connectors have been divided into the following categories.

- Category A - These are connectors which are field replaceable and do not require the use of special tools (standard wrenches, pliers, soldering equipment are not considered special tools).
- Category B - These are not field replaceable and require special crimping tools to assemble. These connectors may be used for original installations, and field replacement is intended to be made using category A or C connectors.
- Category C - These are field replaceable connectors which require a specific crimp tool and cable strip dimensions as designated by the military specification.
- Category D - This is a category of field replaceable connectors in which the military specification defines the following: crimp tool, cable strip dimensions, center contact and crimp ferrule configurations.
- Category E - Semirigid cable connectors - These are field replaceable using a standard cable stripping tool.

14.2.1.3 Characteristics. When classifying connectors into their respective series, there are three main defining characteristics. The first is the size of cable for which they are designed and is classified as small, medium, or large.

The second criterion of classification is the method of coupling or mating. The best method of coupling is by the use of threads. The jacks and receptacles have external body threads and the plugs have internal threads on the coupling nut.

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14.2 CONNECTORS, MIL-C-39012

The third criterion used when classifying connectors is the electrical application, such as high voltage, impedance matched circuits, and dc pulse circuits.

14.2.2 Usual applications. Currently, there are over 30 slash specification sheets covering the series N, TNC, and SMA. Some application considerations for each series are given below.

14.2.2.1 The N series. The N is a threaded coupling connector and is by far the most popular of the medium size connectors. The average cable diameter for the N connector is 0.400 in., but due to its popularity, the diameter ranges from around 0.200 to 0.900 in. for special applications. There are N connectors designed to match 50 or 70 Ω cables. These connectors have a maximum peak voltage rating of 1500 V and a practical frequency limit of 10,000 MHz. They are designed for use with medium size cables and are covered in specification sheets M39012/1 through M39012/5.

14.2.2.2 The TNC series. A TNC connector is a threaded BNC type. This type was developed because the two-ear bayonet locking device of the BNC connector tends to rock during vibration, setting up rf noise in the circuit. The electrical characteristics of the TNC connector are similar to those of the BNC. Specification sheets M39012/26 through M39012/34 cover these connectors.

14.2.2.3 The SMA Series. The SMA is a miniature connector which was originally designed to fill the need for a low VSWR general purpose connector at microwave frequencies. The frequency range specified in MIL-C-39012 for connectors using flexible cable is 0 to 2.4 GHz. Due to its size and inherent ruggedness it is suitable for lower frequency as well. These connectors are covered by M39012/55 through M39012/62.

A new addition of SMA series connectors can be found on specification sheets M39012/79 through M39012/83. This series uses semirigid cable RG402/U and RG405/U. These connectors and cables, when used together, provide better electrical characteristics than the standard flexible cable. They also provide lower attenuation, and no radiation, and permit usage at higher frequencies with good VSWR characteristics. MIL-C-39012 specifies a frequency range of 0 to 18 GHz. However, these connectors have been used at frequencies to 26 GHz in certain applications.

The disadvantage of semirigid cable assemblies is the preparation necessary for using them. Complex drawings defining bends and compound angles are usually necessary to properly route cable and orient the connectors. Assembly is critical due to the solid copper jacket and inner Teflon dielectric problems in termination. A typical problem is cold solder joints between connector and cable which causes electrical intermittencies and weakened terminations. The dimensional stability of the Teflon in the copper jacket is highly susceptible to temperature. The expansion and contraction of Teflon during heating causes many problems in the areas of VSWR and impedance.

There are many experienced cabling houses which have developed processes to minimize these types of problems and provide excellent operating assemblies.

14.2 CONNECTORS, MIL-C-39012

14.2.2.4 Miscellaneous hardware. Miscellaneous hardware such as covers, shields, clamps, and shorting plugs used with MIL-C-39012 connectors is covered in specification sheet M39012/25.

14.2.3 Physical construction and mechanical characteristics.

14.2.3.1 Physical construction. Figure 3 shows the individual parts of a TNC connector, which is a typical rf coaxial connector covered in MIL-C-39012. A brief discussion of the individual parts is given below.

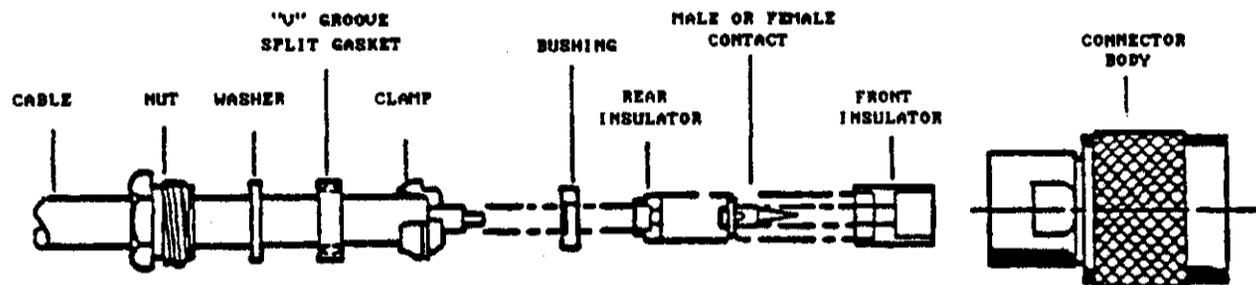


FIGURE 3. TNC connector.

Center contacts. There are two types of center contacts, male and female, which are terminated to the center conductor of the cable. The male contact, sometimes referred to as the male pin, almost always has a solder pocket in one end and is tapered at the other. The female or socket contacts may have a variety of terminations such as a solder pocket, flattened and pierced, or turret but the front end is hollow and generally slotted. The male contact is made from 1/2 hard brass for ease in machining and is plated, usually with gold. The majority of female contacts are beryllium copper. This provides good spring action even after many insertions and withdrawals. Again, the plating is gold. Both male and female contacts may be captivated. The captive contact is made by adding a shoulder of some type on the contact and then physically holding it into the connector. It is usually held in place by putting it between two insulators which are in turn held stationary by a clamp nut or a staking operation.

The crimp type contact is deformed on assembly, causing a discontinuity. However, by designing the inner contact to compensate for these crimping changes, excellent performance is possible and many new designs feature this.

Insulator. The dielectric of the connector is referred to as the insulator. The insulator varies in configuration depending upon the connector style and type. The materials also vary with the applications. The major insulation materials are polystyrene and Teflon.

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14.2 CONNECTORS, MIL-C-39012

Outer contact. This part of the connector is electrically connected with the outer shield of the cable and serves to carry a signal, to act as a shield, or as a grounding member of the circuit. In the case of jacks and receptacles, the body of the connector is the outer contact. The plug may have an outer contact and a coupling nut or just a coupling nut which acts as the outer contact. The term "outer contact" is used only when referring to the tinned portion of the body and is generally made out of nickel-plated beryllium copper because of its good spring action and electrical characteristics.

Coupling nut. The coupling nut is that portion of the connector that mechanically joins two connectors. The coupling nut material is usually 1/2 hard brass with nickel plating, or corrosion resisting steel that has been passivated.

Connector body. The bodies of coaxial connectors may utilize the same material and plating, or finish as the coupling nut. Body configurations depend on the type of connector and will be covered under the type description.

Cable retention methods. There are three basic methods of attaching the cable to the connector: soldering, clamping, and crimping. Soldering of center contacts was discussed previously. In some of the earlier UHF series connectors, the braid was soldered to the connector. The clamping method of attachment requires the use of additional piece parts, the braid clamp and "V" groove gasket. The clamp is usually made of brass or nickel-plated phosphor bronze. The gasket is rubber. The washer is usually brass or phosphor bronze and nickel plated. The last part is the clamp nut which, when tightened into the body of the connector, supplies the force which holds the cable into the connector. As the clamp nut is tightened down, the "V" groove gasket splits, allowing metal to metal contact in the retention mechanism, and forms a seal between the cable jacket and the connector. In the crimping method, the braid clamp, gasket, washer, and clamp nut are replaced by a ferrule clamp nut assembly or ferrule which is an extension of the body. The cable dielectric and center conductor are put inside the ferrule and the cable braid and compressed by means of a crimping tool. The inner female assembly is of nickel-plated brass but the outer ferrule must be of a softer copper alloy because it must be deformed. The biggest advantage of the crimping method of assembly is that it is a much easier method than clamping. Also, the cable stripping dimensions are not as critical. There is no combining of the braid nor is there a problem with the tightening torque necessary to obtain satisfactory cable retention. Its main disadvantage of crimping is the need for special tools.

14.2.3.2 Typical mating configuration. Figures 4 through 8 show typical SMA and N series connector configurations.

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14.2 CONNECTORS, MIL-C-39012

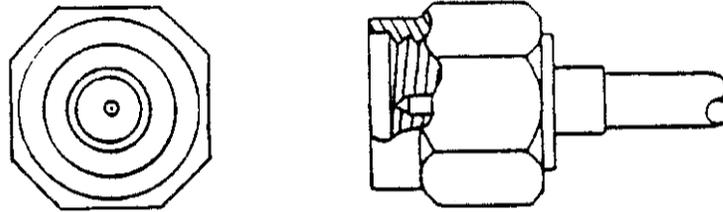


FIGURE 4. Typical SMA plug.

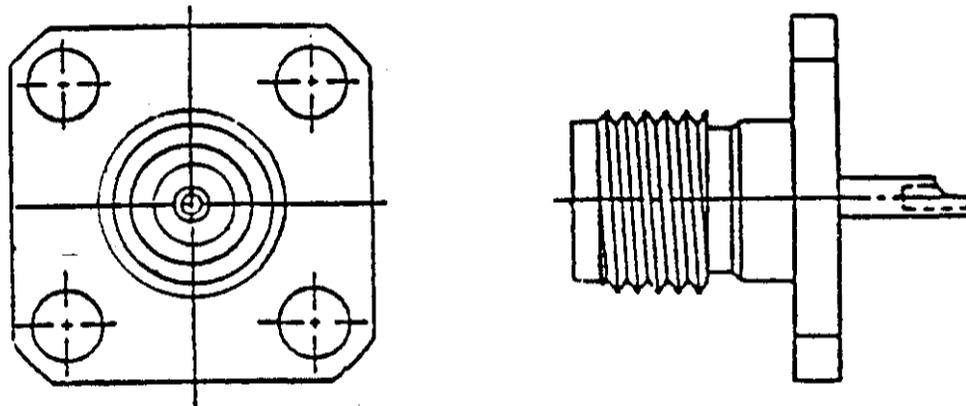


FIGURE 5. Typical SMA flange-mount receptacle.

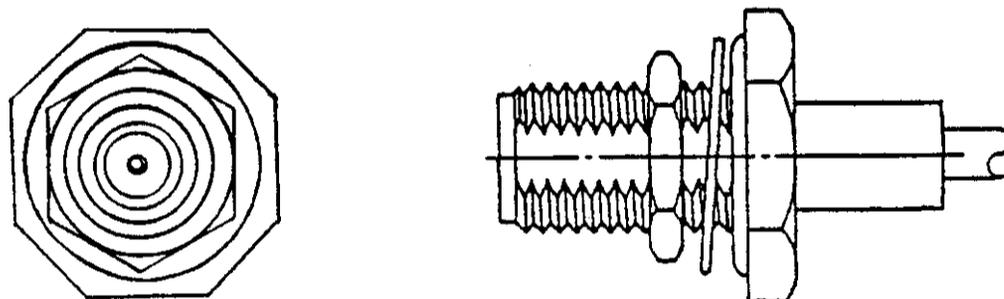


FIGURE 6. Typical SMA type bulkhead-mount jack.

14.2 CONNECTORS, MIL-C-39012

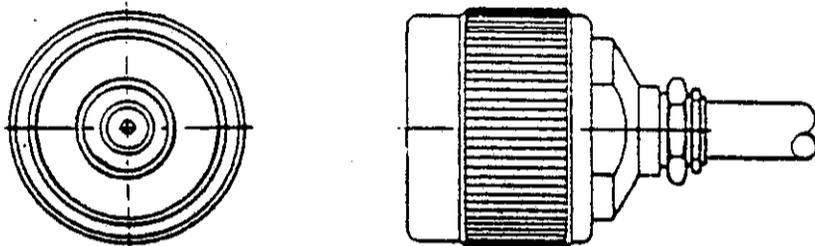


FIGURE 7. Typical N-type plug.

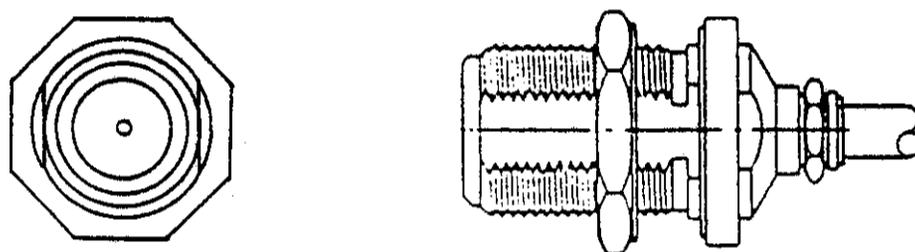
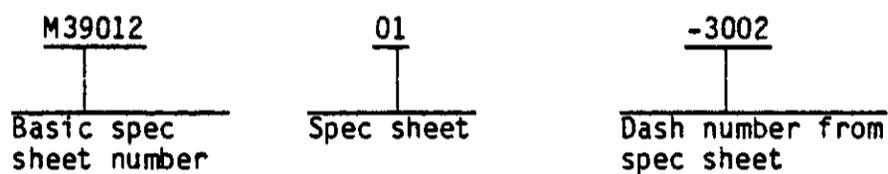


FIGURE 8. Typical N-type bulkhead-mount jack.

14.2.4 Military designation. The part number consists of the letter M followed by the basic specification sheet number and a sequentially assigned dash number.



The first digit of the dash number denotes material

- 0 - Brass
- 1 - Phosphor bronze
- 2 - Aluminum
- 3 - Corrosion resistant steel
- 4 - Beryllium copper

The example above represents a series N plug made of corrosion-resistant steel.

14.2 CONNECTORS, MIL-C-39012

14.2.5 Electrical characteristics.

14.2.5.1 Impedance. Coaxial connectors are designed to keep a constant impedance throughout the connector. For maximum energy transfer, the connector's nominal impedance should be the same as the characteristic impedance of the cable to be used with the connector. Connectors covered in MIL-C-39012 have a nominal impedance of 50 Ω , although N connectors may be obtained to match 70 Ω cables.

14.2.5.2 Voltage rating. Rf connector design involves compromises where one desired characteristic is sacrificed to some extent to obtain another characteristic. This is true regarding impedance matching and high voltage characteristics. These two parameters are not compatible. To obtain a high voltage rating, especially at the junction of the cable core and connector insulator, requires a long overlap of cable core, which in turn presents an inductive discontinuity. This is partially compensated for by using an adjacent section of low impedance line, generally in the form of an oversize center contact.

14.2.5.3 Frequency range. At the lowest frequencies (dc, for example) cable connections consisting of simple solder joints to both conductors are sufficient, if mechanically firm.

As application frequency progresses into the low megahertz range, such connections allow rf leakage and it becomes necessary to provide contact to the center conductor (as before) and 360 degree contact to the outer conductor to completely contain the conducted electromagnetic fields between the two conductors. The characteristic impedance of the section of line represented by the inner and outer diameters of the connector is generally not important. The familiar "UHF" series of connectors is an example of this frequency range.

When application frequency reaches the hundreds of megahertz, it is important that the characteristic impedance of the connector be the same as that of the cable. Also, any physical discontinuities (e.g., connector pin diameter differing from the cable inner conductor diameter) must be held to a minimum. These discontinuities behave as shunt capacitors or series inductances.

The adverse effect of these reactive discontinuities increases with application frequency. Therefore, to maintain a given level of performance, the physical size of the discontinuities underlying them must be kept smaller and smaller as frequency is raised. It is not always possible to avoid all discontinuities and, at the same time, maintain a sound mechanical joint. The frequency range of MIL-C-39012 connectors is given in Figure 1.

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14.2.5.4 Insulation resistance. This is the resistance offered by the insulating members of a component part to an impressed dc voltage; it tends to produce a leakage current through or on the surface of these members. Insulation resistance is an important parameter in the design of high impedance circuits. Low values of insulation resistance are indicative of large leakage currents which can disturb the operation of circuits intended to be isolated, for example, feedback loops. Excessive leakage currents can contribute to insulation deterioration as a result of heating and dc electrolysis. Some factors affecting insulation resistance are temperature, humidity, and residual charges.

14.2.5.5 Contact resistance. The millivolt drop between the braid and outer conductor at the point of contact, between outer contacts and between center contacts, is measured with a current of 1 A passing through the contacts under test. At this current, the contact resistance in milliohms is numerically equal to the measured millivolt drop.

14.2.5.6 Dielectric withstanding voltage. (Also called hipot, voltage breakdown, and dielectric strength). To obtain this parameter, a voltage higher than the rated voltage is applied for a specified time between insulated portions of a component part or between insulated portions and ground. This tests the ability of the component part to operate safely at its rated voltage and withstand momentary over-voltages. This test is not intended to cause insulation breakdown or for corona detection; rather it serves to determine whether or not insulating materials and spacings in the component part are adequate. Some factors affecting the dielectric strength of a material are: temperature, pressure, humidity, rate of amplification of test voltage, time duration of test voltage, shape of test specimen, and form of electrodes.

14.2.5.7 Rf high potential withstanding voltage. This test is similar to the dielectric withstanding voltage test. However, the frequency of the applied voltage is on the order of 5 MHz versus 60 Hz used in the dielectric withstanding voltage test. The test is performed to determine the ability of a cabled, mated pair of connectors to withstand high rf voltages without excessive leakages or electrical breakdown.

14.2.5.8 Voltage standing wave ratio (VSWR). Impedance mismatch (or input impedance versus output impedance) in a transmission line causes reflected power which will interfere with the transmitted power in such a way as to set up standing waves on the line. These standing waves will have maximums and minimums at one-quarter wavelength intervals. The ratio of maximum to minimum of the standing wave voltages along the transmission line is called the voltage standing wave ratio (VSWR). This ratio is a figure of merit for the matching of input to output impedance of a connector, and a VSWR of 1 indicates no reflected power. In rf coaxial connectors, VSWR varies with the frequency and over a broad frequency range. The VSWR will be greater at the higher frequencies.

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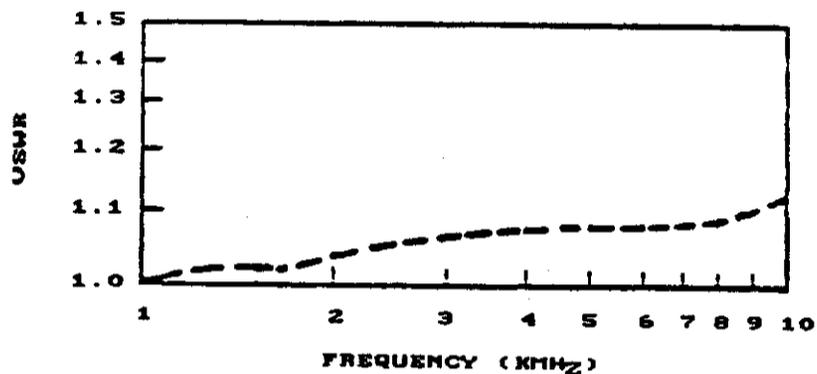


FIGURE 9. Typical VSWR curve for series N connector.

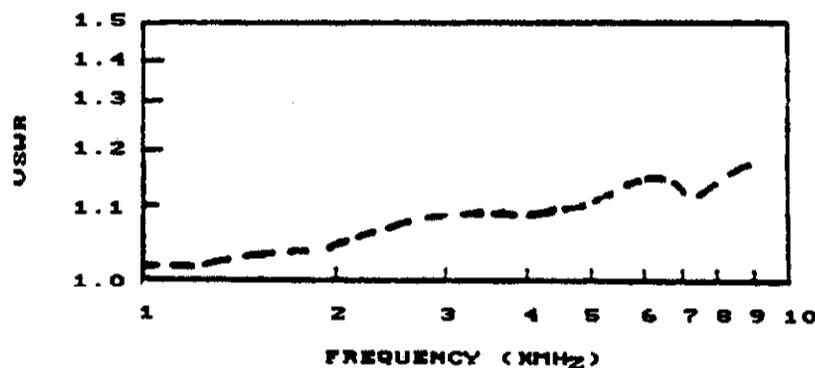


FIGURE 10. Typical VSWR curve for serial TNC connector.

14.2.5.9 Corona level. Corona is the result of ionization of air. A high corona level can cause a sustained electrical discharge which can damage the connector and cable assembly. Corona tests can be used for the detection of air pockets in the insulation or dielectric of connector and cable assemblies. The presence of corona is detected by testing at a specified barometric pressure, usually 70,000 ft, and a specified voltage. The connectors are cabled and mated for this test.

Typical VSWR curves for series N and TNC connectors from a selected vendor's catalog are shown in Figures 9 and 10 respectively, for mated plugs and jacks using medium size 50 Ω cable.

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14.2.5.10 Rf leakage. The word leakage is used to describe both the loss of internal energy (rf) from a microwave circuit and the entrance of energy into a microwave circuit. Errors due to leakage can sometimes be 1 dB or greater. Usually leakage is due to poor mechanical conditions involving damaged or dirty parts or poor assembly. It can also result from poor design.

14.2.5.11 Rf insertion loss. Insertion loss is the ratio of power delivered to a matched load by a matched generator before and after the insertion of a component (connector) into the line. It is a combination of two losses: mismatch loss (reflective) and attenuation (dissipative). Mismatch loss is the ratio of ratio of power that would be absorbed by the device if it were perfectly matched to that of the actual power absorbed by the device with its mismatch in impedance. Attenuation is the ratio of power into a component to the power out under matched conditions and represents the actual power dissipated within the component. Where a component is matched perfectly into the line and load, this mismatch loss is zero and the insertion loss is the same as the attenuation.

Refer to Table II for typical values of the previously discussed parameters for the various series of MIL-C-39012 connectors.

14.2.6 Mechanical requirements. Connectors have their mechanical requirements itemized on the appropriate sheet of MIL-C-39012. Among the mechanical aspects to be considered are: material, finishes, dissimilar metal restrictions, center contacts, allowed displacement, engage and disengage force, coupling tightening torque, mating characteristics, and permeability of nonmagnetic materials.

14.2.7 Environmental considerations. There are no extraordinary environmental considerations for these connectors.

14.2.8 Reliability considerations. Coaxial connector failures are usually mechanical in nature. Generally, they can be attributed to poor workmanship during assembly of the connector to the cable. Poor soldering techniques of the solder connections are common failure modes. Cable stripping dimensions should be made with care. Poor workmanship can cause degradation of electrical parameters, such as VSWR, rf leakage, and insertion loss. It can also result in a low cable retention force and a possible separation of cable and connector. In high voltage connectors, corona at reduced barometric pressure can result from air pockets in the connector dielectric, due to construction or poor assembly techniques. Pressurized connectors have the additional problem of defective seals. Rough handling, rather than construction, would probably be the most common cause of leaking seals.

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TABLE II. Typical electrical characteristic values

Series	Minimum insulation resistance (megohm)	Millivolt drop-contact resistance (millivolts, max)		Dielectric withstanding voltage (Vrms @ sea level, min)	Rf High potential withstanding voltage (Vrms @ 5 MHz, min)
		Center contact	Outer contact		
N	5,000	1.0	0.20	2,500	1,500
TNC	5,000	2.1	0.20	1,500	1,000
SMA	200	4.0	2.0	500 <u>2/</u>	335 <u>2/</u>

TABLE II. Typical electrical characteristic values (Continued)

Series	VSWR (500 Hz to 12.4 GHz, max)	Corona level (@ 70 K Ft, min)	RF leakage (2-3 GHz, dB, min)	Insertion loss (@ 10 GHz, dB, min)	Voltage rating (Vrms max. work. volt.)	Frequency range (GHz)
N	1.30	500	-90	0.15	1,000	0-11
TNC	1.30	250	-60	0.20 <u>1/</u>	500	0-11
SMA	1.20 <u>3/</u>	125 <u>2/</u>	-60	0.60 <u>4/</u>	170	0-12.4

- 1/ 500 Hz to 5 GHz
2/ Using RG178 cable
3/ 500 Hz to 12.4 GHz
4/ At 6 GHz

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15.1 PROTECTIVE DEVICES, GENERAL

15. PROTECTIVE DEVICES

15.1 General.

15.1.1 Introduction. In designing circuit protection, the whole electrical system--power source, transmission path and individual components--must be considered. In addition to minimizing damage to a malfunctioning unit, the fuse or circuit breaker must also protect the rest of the system from any fault that may occur. When one element within the system fails, the protective device should isolate it so that other portions of the system continue to function normally.

This section will discuss protective devices which are included in MIL-STD-975 NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List as well as thermal circuit breakers that are not included.

15.1.1.1 Applicable military specifications.

MIL-F-23419	Fuses, Instrument Type, General Specification
MIL-C-5809	General Specification for Trip-Free Aircraft Circuit Breakers
MIL-C-39019	General Specification for Magnetic, Low-Power, Sealed Trip-Free, Circuit Breakers.

Only MIL-F-23419 and MIL-C-39019 devices are included in MIL-STD-975.

15.1.2 General definitions. The following is a list of terms commonly used with circuit protective devices.

Adjusted circuit or bolted-fault level. The current in a circuit measured under short circuit conditions with the leads normally connected to the circuit breaker bolted together.

Ambient temperature compensation. A method used in protective devices to compensate for the effects of ambient temperatures so as to prevent a change in calibration.

Arc extinction. Use of special methods to control contact arcing in circuit breakers.

Circuit breaker. An automatic device which, under abnormal conditions, will open a current-carrying circuit without damaging itself.

Circuit breaker cascade system. A system wherein the protective devices are arranged in order of ratings such that those in series will coordinate and provide the required protection.

Circuit breaker, nontrip-free. A circuit breaker that can be kept closed by manual override action while a tripping condition persists.

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15.1 PROTECTIVE DEVICES, GENERAL

Circuit protection. Automatic protection to minimize the danger of fire or smoke as well as the disturbance to the rest of the system that may result from electrical faults or prolonged electrical overloads.

Coordination. Arrangement of a lower-rated breaker in series to trip before the higher-rated one trips.

Crowbar. An action that effectively creates a high overload on the actuating member of the protective device. This crowbar action may be triggered by a slight increase in current or voltage.

Current-limiting fuse. A protective device that prevents a dangerous short circuit current by opening the circuit before the development of the peak available current.

Fuse. A protective device with a circuit-opening fusible member that is destroyed by the passage of overcurrent.

Instantaneous. A circuit breaker that trips above a predetermined value of current without any purposely delayed action.

Joined actuator. A multipole breaker that trips all poles when one pole trips. When the faulted pole is trip-free (will not close the contact), the other poles may be kept in position by the restraining actuator.

Let-thru current. The current that actually passes through the breaker under short-circuit conditions.

Manual switching. A circuit breaker feature permitting manual opening and closing of the circuit.

Rupture or interrupting capacity. The maximum current that a protective device will interrupt. It is specified as the number of interruptions in amperes without change in calibration or failure of dielectric strength.

Self-wiping contacts. A switch or relay contact designed to move laterally with a wiping motion when engaging with or disengaging from a mating contact. This feature assures low contact resistance, which is particularly important on low-current circuit breakers.

Short time limits. Values of minimum and maximum trip time measured at various percentages of overload.

Time constant. A measure of the response of a system expressed as the time required to reach 63.2 percent of the total change in value.

Trip-free. A safety device construction in which the contact arm is independent of the operating handle or mechanism, making it impossible to manually hold the contacts closed during an overcurrent or short circuit fault.

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15.1 PROTECTIVE DEVICES, GENERAL

Trip-free breaker. A breaker so designed that it will not maintain the circuit closed when carrying overload currents regardless of the restraint placed on the actuator. All poles of a multiple breaker should trip free on a single-pole fault or overload.

Trip indication. The means of indicating that the circuit breaker has tripped or is in open position.

Ultimate trip current. The smallest value of current that will cause tripping of the circuit breaker under a given set of ambient conditions.

Ultimate trip limits. The specified limits for ultimate trip currents: maximum ultimate trip current and minimum ultimate trip current. At the maximum specified ultimate trip current the breaker will open within the specified time, and at the minimum specified ultimate trip current the breaker will not open.

15.1.3 General device characteristics. The general characteristics of fuses and circuit breakers are discussed below. Both fuses and circuit breakers are available for ac or dc applications. Some circuit breakers must be selected for either ac or dc application (see the discussion below).

15.1.3.1 Fuses. A fuse is a protective device that melts and breaks the circuit when the current exceeds the rated value. Its purpose is to protect the other elements of the circuit from a possibly damaging overload.

Fuses are generally a glass-bodied cartridge with nickel-plated brass caps. High-amperage fuses are made of a punched zinc alloy element. Low-amperage fuses are made with filament wires of various materials such as copper, platinum, or alloy.

15.1.3.2 Circuit breakers. There are, essentially, two types of circuit breakers: thermal breakers and magnetic breakers. (Only magnetic breakers are included in MIL-STD-975). Both provide overcurrent protection; the difference lies in how they perform this function. Following is a brief description of the two types.

Thermal breakers. Thermal breakers convert current into heat, which operates the tripping mechanism. Unless compensated, thermal breakers tend to be sensitive to ambient temperature. Thermal breakers have an inherently slow response to current variations and therefore are not generally frequency-sensitive or prone to nuisance tripping from transients. However, they are sensitive to shock and vibration. The most common form of thermal breaker utilizes a bimetallic heat sensing element.

One type of thermal breaker is a hot wire breaker in which the expansion and contraction of a hot wire element operates the tripping mechanism. Because the hot wire has less thermal lag than the bimetallic type, it offers faster response time.

15.1 PROTECTIVE DEVICES, GENERAL

Magnetic breakers. Magnetic breakers are strictly current devices and can be quite fast. Many magnetic breakers incorporate a hydraulic dashpot that provides an inverse time delay on values of overload below the instantaneous trip point. Magnetic breakers supplied without the dashpot are referred to as "instantaneous." The effect of mechanical vibration is a function of the particular breaker design. Different breakers are needed for dc, 60 Hz, or 400 Hz. Tripping varies with frequency but can be compensated by proper breaker calibration. The tripping coil is frequently actuated by a circuit other than the one that is protected. This arrangement allows use of a magnetic breaker in combination "control--protection" circuits.

15.1.4 General parameter information.

15.1.4.1 Fuses. Fuses are the safety valves of electrical circuits; it is important that they operate or "blow" before damage occurs in the equipment or the wiring being protected. Conversely, fuses must not blow too easily and cause open circuits when the equipment is operating normally.

15.1.4.2 Circuit breakers. Thermal, compensated thermal, and thermal-magnetic circuit breakers may be used on either ac or dc because their tripping properties are thermally dependent. However, hydraulic-magnetic circuit breakers are dependent on magnetic pulses and must therefore be used on the electrical frequency for which they were designed.

15.1.5 General guides and charts. The specific military specifications must be used in the selection of specific fuses and circuit breakers.

15.1.6 General reliability considerations.

15.1.6.1 Fuses. When the fuse is used in a circuit other than the test circuit, its rating can change. It is the circuit designer's responsibility to determine that the fuse is the proper one for the application.

Most small fuses are inherently able to withstand most environmental conditions, the most severe of which are vibration and shock. The smaller the fuse, the more resistant it is to mechanical shock and vibration because the fuse link is short and does not easily resonate.

15.1.6.2 Circuit breakers. Circuit breakers are subject to both electrical and mechanical failure. The most significant factor in circuit breaker reliability is proper selection of a suitable type for the application.

Two basic types of circuit breakers are commonly used: magnetic and thermal. The magnetic type is used in airborne equipment and similar environments where wide excursions of temperature may occur. The thermal type should be limited to environments that provide relatively stable and low temperatures (below 100 °C), as the current-carrying capabilities of the thermal sensing element are highly sensitive to temperature variations.

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Hermetically sealed devices should be used where it is practical to eliminate failures caused by external contamination.

Snap action switching of contacts should be designed to minimize arcing time and should have a wiping action.

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**15.2 PROTECTIVE DEVICES,
CIRCUIT BREAKERS**

15.2 Circuit breakers.

15.2.1 Introduction. A circuit breaker is a device for closing or interrupting a circuit between separable contacts under normal and abnormal conditions. Normal refers to interruption of currents within the continuous current rating of the circuit breakers. Abnormal refers to interruption of currents exceeding the continuous current rating such as short circuits. Ordinarily, circuit breakers are required to operate relatively infrequently although some classes of breakers are suitable for frequent operation. Only magnetic circuit breakers are included in MIL-STD-975.

15.2.1.1 Applicable military specifications.

MIL-C-5809	General Specification for Trip-Free Aircraft Circuit Breakers
MIL-C-39019	General Specification for Magnetic Low-Power, Trip-Free Sealed Circuit Breakers.

15.2.1.2 Circuit breaker types. Four basic kinds of circuit breakers are discussed in this section:

- a. Thermal
- b. Magnetic (included in MIL-STD-975)
- c. Thermal-magnetic
- d. Compensated thermal and thermal-magnetic.

Mechanical breakers, both thermal and magnetic, require an appreciable time to operate. Magnetic types are the fastest. For a total short circuit, the mechanical mechanism of a fast magnetic breaker will act within three or four milliseconds. This is not fast enough for certain kinds of diodes and silicon controlled rectifiers with heat sinks that do conduct heat away with sufficient speed. However, magnetic mechanical breakers do operate rapidly enough for most protection, at least when some limit to current exists in the circuit. Figure 1 illustrates the time-to-trip of a typical magnetic breaker, up to 10 times rated current. The band on the curve indicates that the breakers will not trip below the lower line of the band, may trip anywhere inside the band, and will trip above the band.

15.2.1.3 Thermal breakers. Thermal breakers use heating elements that activate bimetallic devices to trip out a contact. In effect, the bimetallic element pulls the trigger of a spring loaded switch, which then flies open.

Thermal circuit breakers are best suited to protect wire, because the thermal element within the breaker is a reasonable analog of the protected wire.

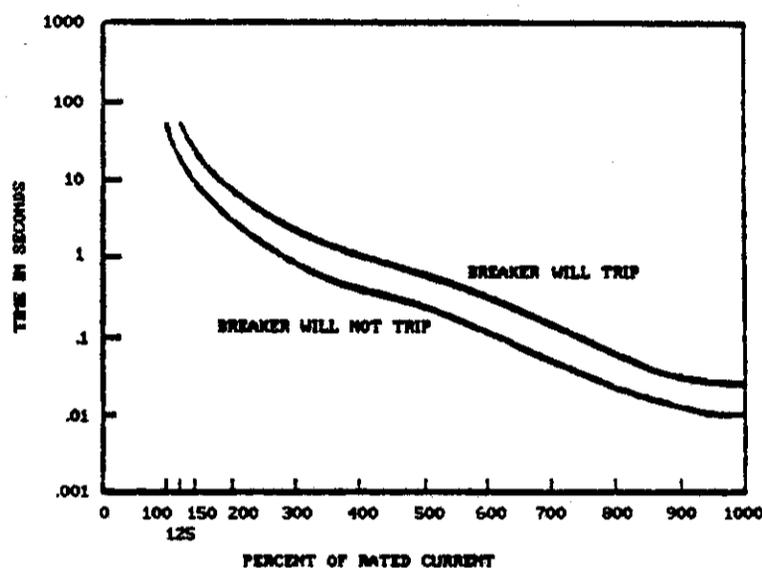
15.2 PROTECTIVE DEVICES,
CIRCUIT BREAKERS

FIGURE 1. Trip-time characteristics of a typical magnetic breaker.

Selecting the correct thermal breaker is more complex than simply matching the breaker rating with the wire rating. One must also consider the ambient operating temperature, the allowable voltage drop, and the heat sink provided. Although thermal breakers are necessarily temperature sensitive, proper design permits some compensation for ambient temperature change.

Trip action is obtained with bimetals and trimetals. The low-expansion side may be Invar, the center may be copper for low resistivity or nickel for high resistivity. Metals used in the high expansion side vary considerably.

The speed of a thermal breaker varies directly with temperature and with the square of the current.

15.2.1.4 Hydraulic-magnetic breakers. Most circuits are subject to momentary currents that exceed the normal continuous current that can be carried safely by the wiring or the equipment load. A motor, for example, draws high current when power is first applied.

As the motor approaches running speed, the current requirement quickly drops to normal. Because of the comparatively short duration of the high current, most circuitry and equipment can safely tolerate it. Therefore, it is not desirable to have circuit interruptions for these transient overloads, but it is necessary to interrupt prolonged or heavy overcurrents.

The hydraulic-magnet circuit breaker provides for both situations; it will interrupt a circuit instantaneously at 1000 percent rated current, and it will provide an inverse time delay to allow for normal inrush surges and tolerable overloads.

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**15.2 PROTECTIVE DEVICES,
CIRCUIT BREAKERS**

Time delay is provided by the action of the fluid-damped movable core in the magnetic element. Overall delay characteristics are determined by several design variables, the most important being the viscosity of the damping fluid. Through adjustment of these variables, virtually any time-delay curve (trip time versus percent load) can be obtained.

Properly specified and applied, time delay never results in loosening of the protection standards. It permits closer tolerances because it can be tailored to the operating characteristics, both starting and running, for the circuit to be protected. With time delay there is no need for overrating (with consequent loss of protection against small, sustained over-currents) to avoid nuisance power interruptions.

There are some applications where only extremely small and very brief current changes can be tolerated, and where time delay is not desirable. For such applications, breakers are available that are designed to trip instantaneously (i.e., without any deliberately imposed delay) above a predetermined current level.

In the time-delay range between the minimum and instantaneous trip-points, there is a definite and desirable ambient temperature effect. With a higher ambient temperature, it is still necessary to carry the rated load, but overloads cannot be tolerated for as long a period. With lower ambient temperatures, overloads can be more prolonged without damage to the equipment. Because fluid viscosity is an inverse function of temperature, the time delay is self-adjusting. As temperatures drop, the fluid becomes more viscous, lengthening the delay period and providing extra time for starting up cold equipment with safety. As the temperature rises, viscosity (and the time delay) decreases, giving a safe margin of protection at higher ambients.

This self-adjusting time delay does not affect the current rating of the breaker.

With the instantaneous trip point establishing the maximum current point on the time delay curve, it is necessary to establish the minimum point at which the circuit breaker will trip.

With the hydraulic-magnetic breaker the minimum trip point is not affected by ambient temperature, and derating is not required.

15.2.1.5 Thermal-magnetic breakers. These are similar to thermal breakers except that a magnetic feature has been added to give an instantaneous tripping point when current reaches a certain level. This provides the normal time delay of a thermal breaker and the high instantaneous overload protection of a magnetic breaker.

15.2.1.6 Compensated thermal and thermal-magnetic breakers. These are similar to thermal and thermal-magnetic breakers, except that an additional bimetallic element is used with the overload element. As the ambient temperature increases, the additional bimetallic element compensates for the temperature change.

15.2 PROTECTIVE DEVICES, CIRCUIT BREAKERS

15.2.2 Usual applications. A circuit breaker, regardless of its application, must accomplish four basic tasks:

- a. When closed, it should be an ideal conductor.
- b. When open, it should be an ideal insulator.
- c. When closed, it must be able to interrupt the assigned current promptly without causing dangerous overvoltages.
- d. When open it must be able to close promptly, possibly under short circuit conditions, without being damaged by contact welding, etc.

Often automatic reclosing of circuit breakers is required. In about 80 percent of the cases, by the time a circuit interruption takes place the cause no longer exists and the breaker closes again with no problem. About 20 percent of the causes persist and the circuit breakers may have to interrupt another short circuit immediately after reclosing.

A circuit breaker must have sufficient short circuit rupture capacity under all conditions to isolate the fault, even if directly shorted at the load terminal of the breaker.

Because circuit breakers have trip-point tolerance, this must be considered in their selection; the individual specification sheet must be examined because breakers do not trip at just above rated value.

15.2.2.1 Trip-free versus nontrip-free construction. In most applications, trip-free construction is required. With this construction, the contact arm is independent of the manual operating handle while a fault exists. It is impossible to hold the contacts manually closed against an overload or short circuit.

There are applications, primarily in aircraft operation, where a nontrip-free construction is desirable. These are cases where the function performed by the equipment is more important than the equipment itself; in these cases, non-trip-free construction permits the contacts to be held closed manually against the fault.

The most common internal circuit arrangements and representative applications are presented here. Many other forms are not illustrated here including several coils in a single breaker, additive or bucking magnetic fields, and one design that even protects against both overloads and reverse current.

15.2.2.2 Series-trip. Series trip is the standard circuit breaker construction. Coil and contacts are in series with the line and load terminals. Overcurrent sensing and circuit interruption are therefore in the protected circuit itself. This is the conventional circuit breaker arrangement used as a main switch and short-circuit protection for supply voltage wiring, or for overload protection of equipment or electronic components such as filament and plate supply transformers, motors and operating coils of solenoids.

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**15.2 PROTECTIVE DEVICES,
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15.2.2.3 Shunt-trip. Shunt trip permits tripping from circuit-closing contacts in remote control, or safety devices responsive to changes in temperature, pressure, air flow, or any other measurable function. Shunt-trip breakers utilize line voltage; relay trip breakers (15.2.2.4) can also be tripped from remote contacts, but they are usually employed when the coil and contacts must be in separate circuits using different voltages or types of current.

A shunt-trip coil does not provide a time delay unless specified. Designed for noncontinuous duty, it must be de-energized when the breaker contacts open. Within limits, continuous-duty coils can be furnished. The combined current through the load and shunt-trip terminals must not exceed contact rating of breaker contact.

15.2.2.4 Relay trip. Relay trip provides a separate control circuit that may be actuated from a remotely located control or safety device. Electrical isolation of the coil from the contacts permits the use of a different current from that of the lead circuit.

Voltage-sensing relay trip coils are normally supplied without time delay and are for noncontinuous duty; i.e., the coil must be de-energized when the contacts open. Time-delay construction is also available and, within limitations, continuous-duty coils may be furnished.

Current-sensing coils are normally for continuous duty, and are furnished with a time-delay characteristic. Nontime-delay coils are also available.

When selecting relay trip construction, always specify the current and voltage values for the coil and contacts separately.

15.2.2.5 Auxiliary switches. Auxiliary contacts are furnished as miniature, snap-action switches mounted on the back of a circuit breaker. They are mechanically coupled to the breaker's switching mechanism but electrically isolated from it. When it is desirable to relate breaker operation to the control of indicators, alarms, lights, fans, and similar circuits, auxiliary switches offer a convenient and dependable solution.

An auxiliary switch can be supplied on each pole of a multipole breaker. The only limitation is that the pole must be of series-trip or switch-only construction to accommodate the switch. Single-pole, double throw (SPDT) contacts furnish a choice of open or closed operation, regardless of circuit breaker contact position. As shown, the normal condition of the auxiliary switch (N.O.: contacts open, N.C.: contacts closed) is with the breaker in the "off" position.

15.2.2.6 Dual rating. Dual rating permits the use of one circuit breaker with two ratings to protect equipment designed to operate at two different current levels. Breakers have two load terminals, each with a different rating. The ratings are usually provided in a 2-to-1 ratio; however, ratios up to 4-to-1 can be provided. Minimum dual ratings are 1 and 2 amperes; maximum 25 and 50 amperes.

15.2 PROTECTIVE DEVICES, CIRCUIT BREAKERS

Breakers with dual-rating construction can also be used with equipment designed to operate on two different voltages, such as 6/12 V dc or 110/220 V ac. For example, a dual rating of 10 A at 110 V and 5 A at 220 V can be provided.

15.2.2.7 Calibrating-tap. A calibrating tap permits two loads to be controlled by a single breaker, but trips in response to overloads in the main circuit only. Though it resembles a shunt-trip schematically, calibrating-tap construction uses current-sensing coils and, like series-trip, can be furnished with time-delay characteristics. Auxiliary loads connected to the calibrating-tap terminal are switched on and off with the main load, but they cannot trip the breaker. The combined main and auxiliary loads must not exceed the breaker's contact rating.

This circuit is called calibrating-tap because the breaker's rating and trip points can be raised by shunting the coil with a resistor. This function is limited to breakers with ratings of one ampere, and the maximum practical increase is four to five times the breaker's rating.

15.2.2.8 No-voltage-trip and remote-trip. Breakers are available with a special back-mounted electromechanical relay that is adjusted to provide either a no-voltage-trip or a remote trip function. Breakers furnished for no-voltage trip will trip instantly in response to a loss of voltage in a normally energized circuit. The lost voltage can be the breaker's own line voltage, or an ac voltage from 120 V to 480 V. Once the relay is de-energized, the tripped breaker cannot physically be reset until the relay voltage is restored.

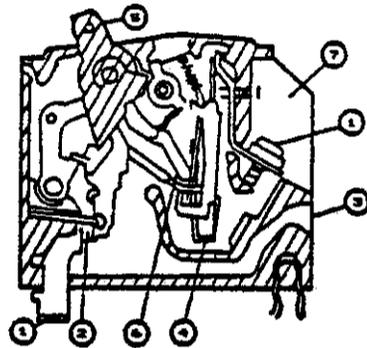
Breakers equipped for remote tripping will trip instantly in response to the closing of remotely located control or alarm contacts. The breaker cannot be reset until the contacts open.

15.2.3 Physical construction and mechanical characteristics. As noted in the previous discussion, two basic means are used to obtain circuit interruption with a circuit breaker: thermal and magnetic. Our discussion here will be limited to a description of these two basic devices.

Thermal breakers use a bimetallic element which deforms because of the heat generated by current going through it or an auxiliary heating element; this triggers a spring-loaded switch which opens. An example of a thermal device is shown in Figure 2.

Magnetic circuit breakers depend on changes in magnetic flux caused by load current changes in a solenoid coil. The sensing and tripping functions are performed by separate elements--the solenoid coil and a mechanical latch. Since this is an almost instantaneous tripping action, time delays are usually built in by the use of hydraulic damping. An example of a typical magnetic circuit breaker is shown in Figure 3.

**15.2 PROTECTIVE DEVICES,
CIRCUIT BREAKERS**



- 1. TERMINALS
- 2. CONTACTS
- 3. BLOWOUTS UENT
- 4. BIMETAL
- 5. HANDLE
- 6. LATCH
- 7. CASE

FIGURE 2. Outline of a typical thermal breaker.

As long as the current flowing through the unit remains below 100 percent of the rated current of the unit, the mechanism will not trip and the contacts will remain closed as shown in Figure 3. Under these conditions the electrical circuit can be opened and closed by moving the toggle handle on and off.

If the current is increased to between 100 percent and 125 percent of the rated current of the unit, the magnetic flux generated in coil 5 is sufficient to move the delay core 3 against spring 11 to a position where it comes to rest against pole piece 12 as shown in Figure 3.

- 2 Armature
- 3 Delay core
- 4 Terminal
- 5 Coil
- 6 Contact bar
- 7 Contact
- 8 Contact
- 9 Terminal
- 10 Toggle
- 11 Spring
- 12 Pole piece

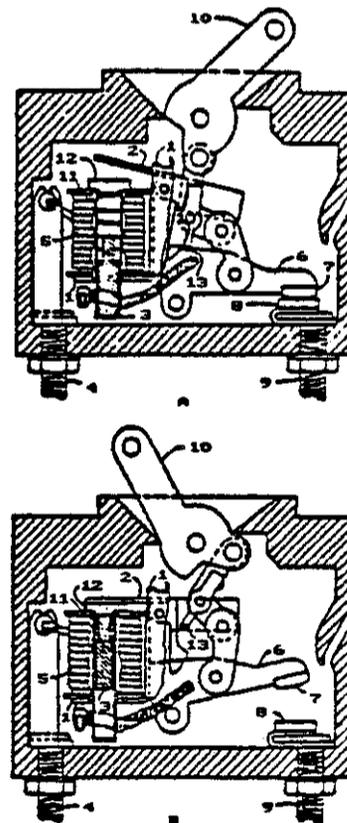


FIGURE 3. Operating mechanism of a typical magnetic circuit breaker.

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CIRCUIT BREAKERS

15.2.4 Military designation.

MIL-C-5809. The military part number is on MS drawing which refers to MIL-C-5809. The designation is:

M25304 - XX
| |
Basic Dash
drawing number

MIL-C-39019. This device is included in MIL-STD-975. The military designation is as follows:

M39019 - /XX - XXX
| | |
Military Slash Dash
specification sheet number

15.2.5 Electrical characteristics. Refer to the applicable military specification and slash sheet for details of performance. In general, magnetic type circuit breakers are rated from 0.05 ampere current capacity to 20 amperes with a trip time either "fast" or "slow." Fast trip is less than 1 second at 200% of rated current at 25 °C. Slow trip is less than 9 seconds for the same conditions.

15.2.6 Environmental considerations. Circuit breakers conforming to either MIL-C-5809 or MIL-C-39019 meet the environmental requirements of most military applications. The environmental requirements of both specifications are quite similar except for the following:

- a. MIL-C-5809 circuit breakers are required to withstand a 50-G, 11 ms half-sine shock, whereas MIL-C-39019 circuit breakers are required to withstand a 100-G, 6 ms sawtooth shock.
- b. MIL-C-39019 circuit breakers are sealed devices and have a thermal shock requirement over a temperature range of -65 to +125 °C whereas MIL-C-5809 has no thermal shock requirements.

15.2.7 Reliability considerations. Consult paragraph 15.1.7, for general reliability considerations of circuit breakers.

15.3 PROTECTIVE DEVICES, FUSES

15.3 Fuses.

15.3.1 Introduction. The fuses discussed in this section include only those normally used in electronic equipment. The current ratings of these fuses range up to 15 A. The voltage ratings range up to 125 V.

A fuse is generally considered the weakest link in the circuit. This is not always true and can be understood by considering the normal functions of the fuse. These functions take two forms: equipment protection and fault isolation.

When protecting equipment, the fuse is selected to open before any other component is damaged or becomes a menace to safety.

When used for isolation, the fuse must clear the circuit after some component or wire has had an electrical fault that has been damaging to itself. The fuse must clear the circuit and isolate the faulted portion from the rest of the system so that the latter can continue to operate. The fuse rating is selected so that it is not the weakest link. It cannot prevent a component from failing; it protects the remainder of the system.

15.3.1.1 Applicable military specifications.

MIL-F-23419 Fuses, Instrument Type, General Specification.

15.3.2 Usual applications. Fuses are used to protect circuits and equipment from current overload.

The following questions should be answered when deciding on the proper fuse:

- a. What is the normal operating current?
- b. What is the abnormal current at which the fuse is required to open the circuit?
- c. What are the minimum and maximum times during which abnormal current is permitted?
- d. What voltage is applied to the fuse?
- e. What is the normal ambient temperature at the fuse locations?
- f. Are there pulse characteristics involved in the circuit?
- g. Are there other requirements special to this application (mechanical or electrical characteristics) beyond the normal requirements of a fuse?
- h. What is the physical size which can be tolerated?

15.3 PROTECTIVE DEVICES, FUSES

The proper size fuse is selected after considering the type of load current it will pass. Where equipment is being protected, the fuse should be selected so that the load current meets the MIL-STD-975 derating. Where transients of short duration and relatively high amplitude are encountered as a normal current, the fuse rating may have to be two, three, or more times the value of the equivalent rms current, particularly if the transient reoccurs at a regular interval.

Some transient spike currents exist for just a few microseconds, others are measured in milliseconds. There might be a dwell in between these spikes when the current subsides to zero or to a value of only a small fraction of the transient peak value. The fuse selection is then based on two factors; (1) the fuse link must not melt while the transient current is building up to its peak; and (2) the equivalent rms value of the current over a long period of time must be below the rating of the fuse. The ampere rating of a fuse is always an rms value that is a measure of the heating effect of the current. This rating must be considered particularly in rectifier circuits in which half-cycle currents flow. The direct current values of half-cycle duration are average currents and are of a magnitude less than the rms value.

The fuse link must not melt while transient current is building up to its peak. This is a particular problem for the application engineer using sub-miniature fuses because of the relatively low ampere ratings available in these fuses. The sub-miniature fuse is available in ampere ratings from 1/100 ampere to 5 amperes. The fuse link in this ampere range is always a wire of very small diameter. Because of its small size, the wire has very little mass and therefore will reach its melting temperature quickly under transient conditions. For example, a one ampere sub-miniature fuse is current-limiting when a 5 ampere load is impressed upon it. This means that if the circuit is set to deliver five amperes, and is closed with a one ampere fuse installed, the current will never reach five amperes because the fuse link will melt while the current is still building up.

This is not a disadvantage. The thermal capacity of such a fuse is low, but so is the thermal capacity of the equipment it is protecting, and this equipment cannot afford to be heated up by such transients. A good example is a milliammeter or millivoltmeter having a full scale deflection of 5 or 10 milliamperes. A 10 or 20 mA fuse becomes current limiting around 200 percent of rated value. This characteristic allows the fuse to clear quickly before either thermal or mechanical forces can injure the meter movement.

15.3.3 Physical construction. Fuses are designed to carry 100 percent of current rating and blow rapidly at very slight percentages of overload. For example, they will open the circuit at a 200 percent overload within a maximum of 5 seconds. The greater the overload, the more rapidly the circuit opens. There is very little mass in the filament used, therefore, the reaction can take place very quickly under shortcircuit conditions. The filament is generally made of silver, platinum, or other precious metal alloys.

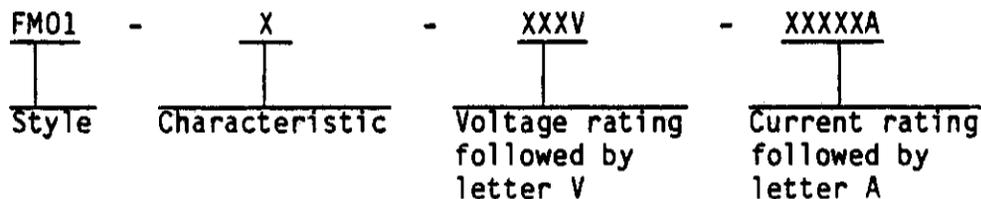
15.3 PROTECTIVE DEVICES, FUSES

These fuses are manufactured with wire diameters as low as 0.000020". Obviously this filament will open the circuit under adverse conditions before any damage can be done in other parts of the circuit.

There are three basic constructions for fast-acting fuses. The first, a bead-type construction, uses a small onyx bead holding two heavy wires, across which is placed a fine filament. The second is a filament-type construction employing a diagonal lineup to insure the constant blowing characteristics of the fuse; and the third, an element construction for heavy amperage fuses in this design.

By observing the characteristic curves found in MIL-F-23419, it will be noted that as the amperage rating increases there is an inherent lag, automatically built-in, due to the increase in mass of the fusible link. In the design of fast-acting fuses, a minimum amount of mass is desired in order to cause the fuse to open the circuit as quickly as possible.

15.3.4 Military designation. The military designation for fuses is:



15.3.5 Electrical characteristics.

15.3.5.1 Current rating. A fuse operated at its rated current consumes some electrical power which must dissipate in the form of heat. When operated above its rated current, a fuse must either operate or blow, which means that the fuse element has melted because of the additional heat.

Generally fuses are rated so that at 110 percent of rating, the temperature rise (from room temperature) at the hottest point on the fuse will not exceed 70 °C.

Under other mounting conditions or ambient temperatures the rating of a fuse must be adjusted up (uprated) or down (derated) for the increased or decreased subtracted heat provided by the mounting or environment.

15.3.5.2 Voltage rating. A fuse can be operated at any circuit voltage as long as the fuse is able to blow without suffering arc damage, and as long as it is sufficiently insulated.

When a fuse blows there is always a sharp break in the circuit that causes full circuit voltage to appear across the blown fuse; and in circuits where inductance is present, the voltage across the fuse may be substantially increased by inductive "kick." Under these conditions, a destructive electric arc may be formed within the fuse that continues to grow in size until the intense heat and pressure within the fuse literally causes it to explode.

15.3 PROTECTIVE DEVICES, FUSES

15.3.6 Environmental considerations. There are no unusual environmental instructions for fuses.

15.3.7 Reliability considerations. Consult paragraph 15.1.6 for reliability considerations associated with fuses. These considerations apply to most fuses used in space applications.

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15.3 PROTECTIVE DEVICES, FUSES

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16.1 SWITCHES, GENERAL

16. SWITCHES

16.1 General.

16.1.1 Introduction. This section contains information on the various types of switches used in electronic equipment. Switches are not included in MIL-STD-975 NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List. The term "switch" applies to a wide spectrum of electromechanical devices used to make or break an electrical circuit. Relays, circuit breakers, and choppers are sometimes thought of as switches. However, those devices are dealt with in other sections of this manual.

Because switches are electromechanical, many considerations must be made prior to inclusion in equipment. These considerations are discussed in this section.

16.1.1.1 Applicable military specifications.

MIL-S-8805	Switches and Switch Assemblies, Sensitive and Push (Snap Action), General Specification for
MIL-S-22885	Switch, Push Button, Illuminated, General Specification for
MIL-S-24317	Switches, Multistation, Pushbutton (Illuminated and Non-Illuminated), General Specification for
MIL-S-3950	Switch, Toggle, General Specification for
MIL-S-8834	Switches, Toggle, Positive Break, Aircraft, General Specification for
MIL-S-3786	Switches, Rotary (Circuit Selector, Low-Current Capacity), General Specification for
MIL-S-6807	Switch, Rotary, Selector, General Specification for
MIL-S-15291	Switches, Rotary, Snap Action
MIL-S-24236	Switches, Thermostatic, (Metallic and Bimetallic), General Specification for
MIL-S-9395	Switches, Pressure, (Absolute, Gauge and Differential), General Specification for

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16.1 SWITCHES, GENERAL

16.1.1.2 Switch types. The most widely used switches in military equipment can be grouped into five basic types.

- a. Push and sensitive
- b. Toggle
- c. Rotary
- d. Thermostatic
- e. Pressure.

16.1.2 General definitions. The following is a list of terms which are commonly used.

Contact resistance. The resistance of a pair of closed contacts as measured from terminal to terminal. This includes the circuit resistance of the individual contact members.

Contact voltage drop. The voltage drop across a pair of closed contacts as measured from terminal to terminal.

Pretravel. The distance or angle through which the actuator moves from free position to operating position.

Releasing force or torque. The releasing force or torque is the value to which the force or torque on the actuator must be reduced to permit the contacts to return to the unoperated position.

Sensitive switch. A switch having a snap action such that the speed of the moving contacts is independent of the speed of the actuator.

Switch life. The number of cycles of operation during which the electrical and mechanical performance of the switch will meet predetermined and stated life-limiting criteria.

Switch rating. The load-carrying and breaking ability of a switch.

16.1.3 General device characteristics. This section discusses the general characteristics of each type of switch. All types are available for ac or dc applications.

16.1.3.1 Push and sensitive switches. These switches are actuated by a reciprocating plunger and are available as very sensitive devices, or as devices carrying large current loads.

16.1 SWITCHES, GENERAL

16.1.3.2 Toggle switches. These switches are actuated by a bat handle and are available as very sensitive devices, or as switches carrying large current loads. Many types of toggle switches are available:

- a. Single-, double-, triple-, and four-pole
- b. Single or double throw
- c. Snap-action or momentary contact
- d. Locking or non-locking.

16.1.3.3 Rotary switches. These switches are actuated by the rotary motion of a shaft for the selection of any one or more of a number of circuits. Many types of switching arrangements are available for use in varying one or more functions such as voltage, frequency and/or resistance.

16.1.3.4 Thermostatic switches.

These switches are actuated by temperature change. They are available in two types of construction: Bimetallic element and nonbimetallic element actuated. Thermostatic switches are used for temperature protection, accurate temperature control, and as detection devices.

16.1.3.5 Pressure switches. These switches are actuated by pressure changes in liquid or gas applications. The switch consists of an electrical snap switch actuated through the displacement of a pressure-sensing device having an inherent spring rate where its displacement is proportional to the applied pressure. These pressure-sensing devices are of three general types:

- a. Capsule sensor
- b. Bourdon tube
- c. Bellows elements.

16.1.4 General parameter information. Both physical and performance factors must be considered when selecting manual switches. Human factors are equally important for efficient and accurate system performance.

Design considerations applicable to switch selection include the following:

- a. Type of action
- b. Contact rating
- c. Switching speed
- d. Environmental considerations

16.1 SWITCHES, GENERAL

- e. Capacitive effect on associated circuitry
- f. Electrical considerations
- g. Life requirements (in number of cycles).

16.1.4.1 Type of action. The type of action required is determined by the switch application. Obviously toggle switches cannot provide the multiswitching capability of rotary switches.

16.1.4.2 Contact rating. Contacts are usually given multiple ratings dependent on the type of load being acted upon. These ratings are resistive, lamp, motor, and inductive loads. Most switches are rated with the resistive load capability and, in most instances, at least one additional rating of the four listed herein. Table I is an example of a typical switch rating.

Current ratings are established at 25 °C. The values shown for multiple switches are amperes per pole, except for motor load.

For some loads, the inrush current at the instant that the switch makes contact is considerably higher than the current that flows during normal operation. Lamp, motor, and capacitive loads are examples of this. Table I shows the lamp load rating as about one-fourth the rating for a normal resistive load. Inrush currents for motor loads may be as much as 12 times the normal running load because of the lack of back emf. Inductive load ratings for both ac and dc are lower than the resistive load rating because of the longer duration of the arc on current break. Inrush current in a capacitive load circuit is high because the capacitor acts as a virtual short circuit until it has acquired some charge.

Table I gives ratings for not only maximum inrush current but also emergency current-breaking capacity. In many instances, the maximum current-making capacity may be the limiting factor rather than the running current capacity of the switch. The current, voltage and the characteristics of the current during make, break, and continuous duty must be carefully considered.

Ratings of contacts are usually given for room ambient temperature and include some safety factor to take care of the temperature rise of the switch. Temperature has a marked effect on switch current ratings, as is shown in Figure 1.

16.1.4.3 Switching speed. Switching speed may be an important parameter to consider when choosing a switch. The term "switching speed" means the duration of contact travel during a make or break function. This parameter is important from the standpoint of reducing any interruptive arc which may be formed, or decreasing the flashover time during the making of a switch. During the actual making of contact, a spring action is sometimes used to increase the speed of closing. This decreases the length of time that an arc, caused by voltage breakdown between the contacts, can cause pitting or burning. If the arcing during the make function is very severe, the contacts may weld together.

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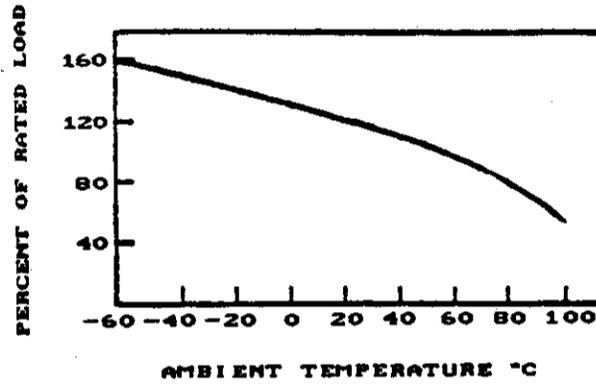


FIGURE 1. Switch current rating vs temperature for typical switch.

TABLE I. Typical switch ratings

Rating <u>1/</u>	Voltage (volts)	Current (amperes)
Lamp load	15 dc	10
	30 dc	7
	125 ac	4
Resistance load	15 dc	50
	30 dc	25
	125 ac	15
Inductive load	15 dc	25
	30 dc	15
	125 ac	11
Motor load	15 dc	40
	30 dc	20
	125 ac	6
Continuous current-carrying capacity (any voltage):		
	15 min of each hour	60
	5 min of each 1/2 hour	90
	1 min of each 1/4 hour	110
	5 sec of each 1/2 hour	250

16.1 SWITCHES, GENERAL

TABLE I. Typical switch ratings (Continued)

Rating <u>1/</u>	Voltage (volts)	Current (amperes)
Current-making capacity (Maximum in-rush on any type of load):	15 dc	100
	30 dc	70
	125 ac	40
Emergency current-breaking capacity (Opening circuit under emergency conditions for 50 operations only):	15 dc	225
	30 dc	100
	125 ac	20

- 1/ These figures are maximum current ratings for 10,000 operations. Longer life (more operations) can be expected when operated at less than maximum rating of switch.

During the break function, an attempt is made to decrease the speed to give the stored energy time to dissipate slowly, because an instantaneous opening will produce heavy transient currents across the contacts. This is especially true of dc circuits and inductive ac circuits. Pitting of contacts is usually more severe during the opening of the contacts. With switches used to interrupt heavy currents, it is sometimes necessary to employ arc suppressors or arc extinguishers. These may take the form of either a capacitor across the contacts to act as an energy sink, or a permanent magnet near the contacts to deflect the arc.

Contact Chatter. Whenever two objects collide, a force develops which causes them to rebound. The extent of rebound depends on the forces tending to restore them to contact, the relative masses, and the natural frequency of the supporting means.

With lamp loads, where the initial current inrush might be 10 or 15 times normal current, a switch exhibiting contact bounce might actually be making the circuit at 15 times the normal current for five or six bounces. This means that a switch being tested for 5,000 cycles might actually be closing this high inrush current at six times 5,000 or 30,000 cycles.

16.1.4.4 Environment. There are four environmental conditions that affect switch performance: temperature, altitude, mechanical vibration and shock.

- a. Ambient temperature. High ambient temperature may affect the lubricants used in some switches, reducing the viscosity and allowing the lubricants to flow out of the bearing areas and, in some cases, onto the contacts or insulating surfaces. If this occurs, improper lubrication of the shaft and/or plunger may hasten mechanical wear of

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these parts, and cause early mechanical failure. A film of lubricant on insulating surfaces has a tendency to trap dust and wear products which may reduce dielectric strength and insulation resistance. In switches with low contact pressure, lubricants present on the contact surface may increase contact resistance and, if dust particles are trapped, may even cause intermittent open circuits. While most switch manufacturers, through careful design and selection of lubricants, attempt to minimize these effects, it is still generally true that

continued operation in high ambient temperatures will shorten the life of a switch.

Extremely low ambient temperatures may also create problems. In switches that use lubrication on the contact areas, low temperatures may cause an increase in the viscosity of the lubricant to such an extent that it delays or prevents the closing of contacts, thus creating a high resistance contact.

- b. Altitude. The reduced barometric pressure encountered in high altitude operation reduces the dielectric strength of the air. This permits the arc to be struck at a lower voltage and to maintain itself longer, thus causing increased contact erosion. Use of switches in high altitude applications will therefore require derating in terms of loads and/or life.

In manned space flight, applications altitudes are such that the effect of gravity is lost. This permits particulate contamination to become weightless and float around in the switch housing. This can cause shorts between normally open contacts if the contaminant is conductive and can cause opens on normally closed contacts if the contaminant is nonconductive. This emphasizes the importance of hermetic sealing, clean room assembly, visual inspection and screening in these applications. Use of cadmium- or zinc-plated parts should be prohibited and nonmetallic parts used to a minimum because of outgassing and resulting contact contamination.

- c. Vibration. In any snap action switch, the contact pressure depends on the position of the actuating button. As the button is depressed, the operating point is approached, and the contact pressure decreases until the overcentering action takes place.

Most switches have good vibration resistance in either the free or the full overtravel position. Some switches do not withstand vibration if the operating plunger has been depressed almost to the operating point. Contact forces here have been reduced to a low value that depends on the remaining distance to be traveled until snap action occurs.

16.1 SWITCHES, GENERAL

If the switch is subjected to mechanical vibration, it is possible that the contact spring will reach a frequency which nearly corresponds to its fundamental frequency, or a harmonic. The vibration that results will cause the moving contact to chatter on the stationary contact.

Chatter caused by vibration can be of particular importance in digital circuits where making and breaking contacts can result in transmittal of erroneous information. The effect of vibration chatter can be minimized by using two sets of contacts in parallel. Each set should be selected to take the full current rating, since they will not make and break at the exact same time.

- d. Shock. Mechanical shock applied to switches may cause damage such as cracking of the switch body, insulating stack, or wafer; loosening of contacts from the wafers; or breaking of welds.

An electrical effect which is directly attributable to mechanical shock is contact bounce, which is a separation of contacts of the closed circuit. This is similar to contact chatter, but is usually of greater magnitude. Contact bounce is dependent on the contact mass and the spring constant of the contact arm, and may be severe enough to open a closed circuit or to close an open circuit.

Even if circuit malfunction does not result from contact bounce, the arcing it causes will materially shorten contact life and may generate electrical noise. Therefore, the susceptibility of switches to contact bounce must be studied with instrumentation of rapid response. Pulses produced by contact bounce must be considered in light of the circuit application. Voltage or current intermittencies which in one circuit would be completely inconsequential might badly disturb another, more sensitive circuit.

16.1.4.5 Electrical considerations. These are six basic factors that are important in the selection of a switch.

- a. Current Rating. This is based on the temperature developed within the switch under service conditions. Because of arcing and welding at the contacts, most heating (and consequent wear) occurs on making and breaking the circuit. Hence, the rate of operation also affects switch life. For ac loads, 60 operations per minute is generally considered the maximum rate at which full current capacity of the switch is available. The switch can, however, be operated at higher rates if the current is reduced, or if a decrease in switch life is acceptable. Ratings are normally based on continuous, steady-state current.
- b. Voltage. Maximum voltage rating depends on the air gap, or contact separation. A gap of 0.005 to 0.008 in. will permit a 250 Vac rating. A gap of 0.010 to 0.015 in. permits a 480 Vac rating, and a gap of 0.020 to 0.070 in. permits a 600 Vac rating.

16.1 SWITCHES, GENERAL

- c. Insulation resistance. This is the resistance between two normally insulated metal parts, such as a pair of terminals, measured at a specific high dc potential (usually 100 or 500 Vdc). Typical values for new switches are in the range of thousands of megohms. These values usually decrease during life as a result of surface contaminant build-up.
- d. Dielectric strength. This is a measure of the ability of the insulation to withstand high voltage without breaking down. During life, the deposition of contaminants and wear products on the surface of the insulation tends to reduce the dielectric withstanding voltage. In testing for this condition, a voltage considerably above rated voltage is applied, and the leakage current is measured. Test voltages are typically 1000 Vac plus twice the rated voltage.
- e. Effect of loads. On any switch, electrical erosion of the contacts occurs when an arc is drawn while breaking a circuit. This erosion normally tends to increase contact resistance, generate wear products by contamination of insulating surfaces, and reduce dielectric strength and insulation resistance. The amount of this erosion is a function of current, voltage, power, frequency, and speed of operation. The higher the current, the hotter the arc and the greater the erosion. The higher the voltage, the longer the arc will be maintained, resulting in greater erosion. In an inductive circuit the inductance acts as an energy storage device, which returns its energy to the circuit when the circuit is broken. The amount of erosion on an inductive circuit is in proportion to the amount of inductance.
- f. Contact resistance. This is the resistance of a pair of closed contacts which effectively appears in series with the load. Typical end-of-life criteria for this parameter is 20 m Ω .

16.1.4.6 Life requirements. In attempting to forecast life expectancy for a switch, both mechanical and electrical factors must be considered. In applications that do not require the switch to make or break large electrical loads, mechanical life is generally limited only by the fatigue characteristics of the spring. These characteristics, in turn, are influenced by the total travel required of the switch and the method of applying the actuating force. When overtravel is held within the prescribed normal limits, mechanical life can be expected to reach several million operations, regardless of the cycling rate. In most applications, however, electrical considerations are of major importance in determining switch life. Life is an inverse function of current made or broken. When determining life requirements, care should be taken to include the operations that take place during testing or screening of both the switch and the equipment that the part is used in. See Table II and Figure 2.

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16.1 SWITCHES, GENERAL

TABLE II. Typical electrical ratings comparing various manual switches

Switch type	Rated current (amperes)		Rated voltage (volts)
	Resistive	Inductive	
Pushbutton	5	3	30 dc
	5	3	125/250 ac
	15	15	125/250 ac
	1.5 1.5		110 dc 250 ac
Toggle	20	12	30 dc
	1.5		110 dc
	15	15	230 ac
Rotary	15	15	125/250 ac
	0.5		125 dc
	0.25		250 dc

16.1.4.7 General load and life discussion. The problem of switch rating arises from the wide variety of requirements placed on the switch in various applications, and the sensitivity of the switch to changes in requirements. If an attempt were made to establish life ratings for all possible applications, the result would be an almost infinite variety of ratings. In an effort to simplify the problem, switch manufacturers, in cooperation with switch users and the military, have established certain reference loads, life requirements, environments, duty cycles, and failure criteria. These are arbitrarily established, and give a relative basis for comparison between different switch designs. They do not, however, match the actual requirements for most applications.

16.1.5 Reliability considerations. Switches are electromechanical devices subject to both electrical and mechanical failure. Contact failure can result from high in-rush or sustained high currents, or from the inductive kick when an inductive circuit is opened. These currents may cause intense heat and the possible welding of contacts. Careful consideration to derating should be given in these cases.

In selecting a switch for high-reliability or space programs, hermetically welded sealed units should be used where available. Terminals should have a glass seal and be of a hook design to provide proper stress relief for the seal. Parts should be of corrosion-resistant material with no cadmium or zinc plating. Use of nonmetallic materials should be minimized.

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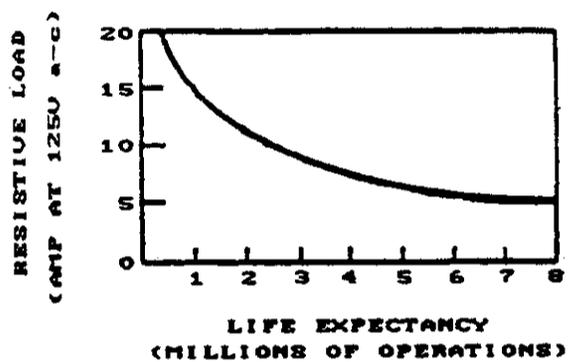


FIGURE 2. Life expectancy vs current for a typical switch.

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16.2 SWITCHES, PUSH AND SENSITIVE

16.2 Push and sensitive.

16.2.1 Introduction. Pushbutton and sensitive switches are actuated by a direct thrust in line with the button travel. They are available with numerous contact configurations and modes of operation. Common modes of operation are momentary contact, maintained contact, and sequential contact.

16.2.1.1 Applicable military specifications.

MIL-S-8805	Switch, Sensitive and Push
MIL-S-22885	Switch, Illuminated Pushbutton
MIL-S-24317	Switch, Multistation Pushbutton

16.2.2 Usual applications. Pushbutton and sensitive switches are a form of precision switch with integral pushbutton actuators. They are available with numerous contact configurations and modes of operation. Banks of pushbutton or sensitive switches can be individually operated or interlocked. The exact type or configuration is determined by the application.

16.2.3 Electrical ratings. Typical electrical ratings for push and sensitive switches are similar to electrical ratings for most other switches. Four types of loads are considered.

- a. Resistive
- b. Inductive
- c. Motor
- d. Lamp.

See Table I for typical electrical ratings.

16.2.4 Environmental considerations. Typical environmental requirements are found in MIL-S-8805, MIL-S-22885 and MIL-S-24317.

16.2.5 Reliability considerations. Refer to subsection 16.1 General, paragraph 16.1.5 Reliability considerations.

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16.3 SWITCHES, TOGGLE

16.3 Toggle.

16.3.1 Introduction. Toggle-action switches are available with two types of action, momentary contact and maintained contact. Additionally, they are available as two- or three-position switches with single, double, triple, or four poles.

A two-position toggle typically has these two positions at equal angles on each side of the vertical centerline of the mounting bushing. A three-position toggle has two extreme positions plus a center position. The center position is generally for the "OFF" condition with the two extreme positions representing "ON" conditions of the circuitry.

16.3.1.1 Applicable military specifications.

MIL-S-3950	Switch, Toggle
MIL-S-8834	Switch, Toggle
MIL-S-81551	Switch, Toggle, Hermetically Sealed

16.3.2 Usual applications. Toggle switches are widely used where simple make-and-break action is required, and are suitable for use on ac or dc circuits. Toggle switches are available with various actuating handles and subsequent switching actions in some 50 styles.

16.3.3 Electrical ratings. Typical electrical ratings for toggle switches are similar to electrical ratings for most other switches. Four types of loads are considered.

- a. Resistive
- b. Inductive
- c. Motor
- d. Lamp.

See Table I for typical electrical ratings.

16.3.4 Environmental considerations. Typical environmental requirements are found in MIL-S-3950, MIL-S-8834 and MIL-S-81551.

16.3.5 Reliability considerations. Refer to subsection 16.1 General, paragraph 16.1.5 Reliability considerations.

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16.4 SWITCHES, PRESSURE

16.4 Pressure.

16.4.1 Introduction. Pressure switches are switches whose response is a function of the gas or liquid pressure which they are designed to regulate and control.

The switching action of a pressure unit usually falls into two categories: the creep type and the snap-acting variety. The latter is preferable for applications where vibration capability and inductive load carrying capability are required. The creep type, however, can offer very close tolerances with regard to its pressure differential rating--as low as ± 1 percent.

16.4.1.1 Applicable military specification.

MIL-S-9395

Switch, Pressure

16.4.2 Usual applications. Applications include response to altitude, liquid pressure liquid level, and air flow. Switches may respond to absolute, gauge, or differential pressures. By means of adapting elements, these switches may be applied to a great variety of applications.

By using more than one switch, control of separate circuits at different pressure or temperature steps can be provided.

16.4.3 Electrical ratings. Typical electrical ratings for pressure switches are low, below 1 amp, and supply input to instruments. Usually a relay will be added to the circuit when the pressure switch must control a large load.

16.4.4 Environmental considerations. Typical environmental requirements are found in MIL-S-9395.

16.4.5 Reliability considerations. Refer to subsection 16.1 General, paragraph 16.1.5 Reliability considerations.

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16.5 SWITCHES, ROTARY

16.5 Rotary.

16.5.1 Introduction. Rotary switches are actuated with a twisting action and are generally available with either momentary contact or maintained contact. Rotation can be either unlimited (turned through more than one complete circle) or limited to 360 degrees or less, after which the direction of rotation must be reversed.

A rotary switch may have many points of actuation with several decks of contacts and combinations of actuating cans.

16.5.1.1 Applicable military specification.

MIL-S-3786	Switch, Rotary (Circuit Selector, Low-Current Capacity) General Specification
MIL-S-6807	Switch, Rotary, Selector, General Specification for
MIL-S-15291	Switch, Rotary, Snap Action
MIL-S-22710	Switch, Rotary (Printed Circuit)

16.5.2 Applications. Rotary switches are used for the selection of any one of a number of circuits or combinations of circuits with a rotary motion of the shaft. Rotary switches can handle a great many connections. The basic contact arrangements may be expressed by the number of poles and the switch positions or decks per pole.

16.5.3 Electrical ratings. Typical electrical ratings for rotary switches are low, below 2 amps. Usually a relay will be added to the circuit when the rotary switch must control a large load.

16.5.4 Environmental considerations. Typical environmental requirements are found in MIL-S-3786, MIL-S-6807, MIL-S-15291, and MIL-S-22710.

16.5.5 Reliability considerations. The contacts of a rotary switch are usually of the sliding type, providing the self-cleaning necessary for low contact resistance and long life. The rotary switch will usually exceed life and reliability referenced in subsection 16.1 General, paragraph 16.1.5 Reliability considerations.

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16.6 SWITCHES, THERMOSTATIC

16.6 Thermostatic.

16.6.1 Introduction. Thermal switches (thermostats) are used either for the indication or control of temperature, or in a protective function. The majority of thermal switches are simple thermostatic devices whose performance depends upon the action of a bimetallic element. The principal design factors relate to accuracy, repeatability, temperature range, and thermal lag.

16.6.1.1 Applicable military specification.

MIL-S-24236

Switches, Thermostatic, Bimetallic Actuated

16.6.2 Usual applications. Temperature-actuated switches provide switching action in response to temperature changes. Usually the objective is to maintain a specified temperature within the system. However, the devices are also used for over- or under-temperature protection.

16.6.3 Electrical ratings. Typical electrical ratings for thermal switches are low, below 1 amp, and supply input to instruments; usually, a relay will be added to the circuit when the thermal switch must control a large load.

16.6.4 Environmental considerations. Typical environmental requirements are found in MIL-S-24236.

16.6.5 Reliability considerations. Refer to subsection 16.1 General, paragraph 16.1.5 Reliability considerations.

17.1 RELAYS, GENERAL

17. RELAYS

17.1 General.

17.1.1 Introduction. The American Standards Association defines a relay as "an electrically controlled device that opens and closes electrical contacts to effect the operation of other devices in the same or another electrical circuit."

Although this definition states control in the "same" or "another" electrical circuit, an implicit characteristic of the electromechanical relay is the opening and closing of contacts to control a load, or loads, in circuits isolated from the controlling input and other contacts. Controlling an electromechanical relay is most commonly done with the application of a specified voltage or current to two input terminals. A coil within the relay translates control-signal electrical energy into mechanical energy, causing mechanical action to open or close the isolated contacts, controlling signals through the contacts.

This section contains information on the various types of relays used in electronic equipment. Many types of electrically controlled switching devices are described by the term "relay." However, the following discussion is confined to the selection and application of electromechanical, current-operated switching devices which find general usage in space applications. MIL-STD-975 NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List includes only MIL-R-39016 relays. The other relays are included for general information.

17.1.1.1 Applicable military specifications.

MIL-R-5757	Relay, Electromagnetic, General Specification for
MIL-R-6106	Relay, Electromagnetic (Including Established Reliability (ER) Types), General Specification for
MIL-R-28776	Relays, Hybrid, Established Reliability, General Specification for
MIL-R-28894	Relay, Hybrid or Solid State, Sensor, Established, Reliability, General Specification for
MIL-R-39016	Relay, Electromagnetic, Established Reliability, General Specification for
MIL-R-83726	Relay, Hybrid and Solid State, Time Delay, General Specification for
MIL-R-83407	Relay, Reed, Mercury Wetting, General Specification for
MIL-R-83516	Relay, Reed, Dry, General Specification for

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17.1 RELAYS, GENERAL

17.1.1.2 Applicable federal standard.

FED-STD-209

Clean Room and Work Station Requirements,
Controlled Environment.

17.1.2 General definitions. A list of terms which are commonly used in relay engineering and generally accepted by the electrical and electronic industries are as follows:

Ambient temperature. Ambient temperature is the temperature of the surrounding medium.

Armature. The armature is the hinged or pivoted moving part of the magnetic circuit of an electromagnetic relay. It is sometimes used in a general sense to mean any moving part which actuates contacts in response to a change in coil current.

Bounce time. Bounce time is the interval from the initial closure of the contacts until the uncontrolled making and breaking of contact ceases.

Chatter. Chatter is sustained rapid opening and closing of contacts. It is caused by uncompensated ac operation, mechanical vibration, and shock or other causes.

Coil. A relay coil is a magnetic winding to which energy is supplied to activate the relay.

Coil resistance. Unless otherwise specified, coil resistance is the dc ohmic resistance of the coil, measured at the coil terminals at 25 °C.

Contact bounce. Contact bounce is the uncontrolled making and breaking of the contacts when the relay contacts are moved to the closed position.

Contact nomenclature. Each of the moving contacts of a relay constitutes a pole of the relay.

- a. A combination of a stationary contact and a movable contact which are closed when the relay coil is deenergized is referred to as form B or normally closed contacts and is abbreviated NC.
- b. A combination of a stationary contact and a movable contact which are closed when the relay coil is energized is referred to as form A or normally open contact and is abbreviated NO.
- c. A combination of two stationary contacts and a common or movable contact which engages one of them when the coil is energized and engages the other when the coil is deenergized is called transfer, form C or double-throw contacts and is abbreviated DT.

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17.1 RELAYS, GENERAL

Contact resistance. Contact resistance is defined as the total electrical resistance of the contact system as measured at the terminals.

Contactor. A term sometimes used for a relay with heavy-duty contacts.

Control voltage. The voltage applied to an input which results in a change of state in the output of a relay.

Dropout current or voltage. Dropout current or voltage is the current or voltage at which all contacts will revert to the de-energized position. The relay will be considered released when all contacts have returned to their normal or nonoperated position.

Dropout time. Dropout time is the time interval from the removal of power to the coil until the NC contacts close; some specifications include contact bounce time in dropout time.

Dry circuits. Dry circuits are those circuits in which the relay does not perform a switching function on an energized circuit path.

Header. A header is the part of a hermetically sealed relay through which the electrical terminals pass.

Hold voltage. Hold voltage is the voltage at or above which the armature is maintained in its operated position.

Low-level circuits. Low-level circuits are those circuits which function in the millivolt or microampere range.

Miss. A miss is a failure of the contact, for any reason whatsoever, to establish a circuit as required within the limits as specified.

Operate time. Operate time is the time interval from the application of nominal power to the coil until the NO contacts close; some specifications include contact bounce time in operate time.

Pickup current or voltage. Pickup current or voltage is the maximum current or voltage required to operate the relay. The relay will be considered to have operated when all contacts have functioned.

Rated (nominal) coil voltage. Rated coil voltage is the coil terminal voltage at which the relay is designed to meet all specified electrical, mechanical, and environmental requirements.

17.1.3 General device characteristics. The following classifications define the general characteristics for various types of relays.

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17.1 RELAYS, GENERAL

17.1.3.1 Classification by configuration.

- a. Armature. The armature relay operation depends upon energizing an electromagnet which attracts a hinged or pivoted lever of magnetic material to a fixed pole piece. The hinged or pivoted lever is called the armature. The actuating coils may be operated with ac or dc voltage. Relays operated on direct current have greater life expectancy than ac relays.
- b. Hybrid. A relay with an isolated input and output. The input is a solid state device which controls an electromechanical output. The switching characteristics are dictated by the electromechanical output.
- c. Reed. A reed relay is operated by an electromagnetic coil or solenoid which, when energized, causes two flat magnetic strips to move laterally to each other. The magnetic reeds serve both as magnetic circuit parts and as contacts. Because of the critical spacing and the frailty of the arrangement, the reeds are usually sealed in a glass tube.
- d. Sensor relay. A sensor relay detects specified functions (for example, frequency, phase sequence, voltage level) and changes the output when the functions are within specified limits. The relay may incorporate time delay with the switching operation.
- e. Solid State. These are relays incorporating only semiconductor or passive circuit devices. There are no moving parts, so therefore there is no bounce or chatter, and they have fast response and long life; however, the number of designs available is still quite limited and at present only single pole devices are available.
- f. Time-delay.
 1. A delay in the operate time or the dropout time, or both, of the armature type relay may be obtained by placing a conducting slug or sleeve on the core in the proper position. This produces a counter magnetomotive force which produces a desired time delay. When the slug is placed on the core nearest the armature gap, a delay in operate time is obtained. Placing the slug farthest from the armature gap results in a delay in dropout time.
 2. The most common method of producing a time delay is the use of a separate circuit, usually in the same package, to produce either a fixed time delay or in some cases an externally adjustable delay in the time before the relay coil itself is energized.

17.1 RELAYS, GENERAL

17.1.3.2 Classification by application. Since classifications are of an arbitrary nature, any particular relay design may fall into one or more categories. For example, a low level relay may also be a latching relay or a sensitive relay.

- a. General purpose. Relays having an ac or dc voltage rated coil whose contacts are rated resistive up to and including 10 A. The term general purpose may be used when discussing nonlatching relays.
- b. Intermediate level. Relays used in a load application where there is insufficient contact arcing to effectively remove surface residue from the organic vapor deposits on the contact surface, although there is sufficient energy to cause melting of the contact material.
- c. Latching. A bistable polarized relay having contacts that latch in either position. A signal of the correct polarity and magnitude will reset or transfer the contacts from one given position to the other.
- d. Low-level. Relays intended specifically for the switching of low-level or dry circuits. In these circuits only the mechanical forces between the contacts affects the physical condition of the contact interface, that is, there are no thermal or electrical effects; e.g., arcing. The current and open circuit voltage are generally defined as being in the microampere, millivolt range.
- e. Power. Relays intended for switching loads in excess of 10 A.
- f. Sensitive. Relays which are defined in terms of coil resistance and maximum operate current. The reduced coil power required to operate the relay is characteristic of a sensitive relay. It is accomplished by increasing the ampere-turns, and thereby the resistance, of the coil.

17.1.4 General parameter information.

17.1.4.1 Design considerations. Physical and electrical characteristics of the relay can be varied almost infinitely to handle different requirements; however, there is a very strong interdependence among the relays' mechanical parts and electrical system. For these reasons, comparable performance curves for similar relays are almost meaningless.

Nevertheless, selecting the best relay in terms of required performance at minimum cost is relatively straightforward. It involves two primary steps:

- a. Determining all factors pertinent to the application.
- b. Translating the application requirements into a technically sound relay specification.

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Design considerations applicable to relay selection include the following:

- a. Coil
 1. Supply voltage, current, and type of power (dc or ac including frequency)
 2. Resistance and tolerance
 3. Duty cycle
 4. Operate pickup and drop-out values
- b. Contacts
 1. Load characteristics (dc or ac resistive, inductive, capacitive, motor, lamp. If ac, the frequency)
 2. Normal load current
 3. Worst case load level
 4. Normal watts load
 5. Open circuit voltage
 6. On-off cycle and frequency of operation
 7. Operate and release time
 8. Life
- c. Environmental considerations, including temperature effect on coil current.
- d. Physical considerations
 1. Size
 2. Mounting
 3. Terminations
- e. Definition of switching requirement (open, close, transfer, double contact arrangement, etc.)
- f. Assurance that circuit is designed to allow minimum stress on contacts.

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Table I lists relay requirements that may need consideration.

TABLE I. Relay requirement check list

Characteristic	Description
Control signal Type	ac, dc or combination
Sensitivity	Input power to control circuit at maximum rated input voltage
Maximum actuating signal	Level of input required to effect switching
Minimum release signal	Input level at which contacts return to normal condition
Maximum overdrive	Absolute maximum prolonged input voltage
Input transient protection	Maximum transient input signal expected
Thermal stability	Allowable variation in maximum actuating signal and minimum release signal over specified temperature range
Switching Arrangement	Number of switching circuits and configuration
Current rating	Magnitude resistive, inductive, lamp, or motor
Isolation	Required isolation from input circuit and between switching circuits
Leakage	Maximum permissible open-circuit leakage
Saturation	Maximum forward voltage drops at rated load
Response	Response time from application of input signal to completed switching
Transfer time	Transfer time of SPDT common lead, make-before-break or break-before-make and dwell-time snap action required
Load power	Frequency for ac and polarity for dc

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TABLE I. Relay requirement check list (Continued)

Characteristic	Description
Switching (Cont)	
Open-circuit with-standing	Maximum continuous load voltage a nonconducting contact must withstand
Transient protection	Maximum transient voltage anticipated in load circuit
Step voltage	Maximum anticipated rate of change of load voltage rise
Cross talk	Signal interference level acceptable in "multicontact" relays
Overload	Maximum nonrepetitive overload expected
Radio noise	Acceptability level of internally generated radio noise
Auxiliary power	
Voltage available	Type of auxiliary voltage available
Isolation	Normally common with load supply
Power level	Maximum power required from auxiliary supply
Voltage range	Maximum voltage variation of auxiliary power from nominal
Transient level	Maximum voltage transients anticipated on auxiliary voltage
Environment	
Temperature	Operating and nonoperating temperature extremes
Shock-vibration	Maximum levels anticipated in use
Package requirements	
Dielectric	Insulation properties between isolated terminals
Size	Maximum envelope dimensions
Termination	Required terminal wiring
Mounting	Mounting methods and dimensions

17.1 RELAYS, GENERAL

17.1.4.2 Coil requirements. The coil characteristics determine the pull on the armature. They may be expressed as ampere turns, amperes, watts, volts or resistance. Any one of these factors may dictate the use of a particular coil. The current available for the relay coil may be only a few milliamperes, in which case a large number of turns will be required to supply the required ampere-turns to make the relay operate properly. To assure positive operation under severe conditions, more than normal power input to the relay coil may be necessary.

- a. Coil resistance. Because of variations in wire, spool body dimensions, winding techniques, etc., it is impractical to attempt to wind a coil to a precise resistance. Generally, the coil resistance will vary from the nominal by ± 10 percent. For high resistance coils such as those used in sensitive relays the size of wire requires that this tolerance be increased to ± 15 percent or ± 20 percent. For low resistance coils a lower tolerance can be held although it is seldom required.
- b. Coil suppression. When a relay coil is turned off, the inductive energy stored in the coils magnetic field can create surge voltages of up to 1500 V on a dc power line. With the wide use of solid state circuitry, relay coils must be suppressed to limit the surge voltages. The suppression device absorbs and dissipates the energy in the coil. Suppression methods should be selected to minimize the affect on dropout characteristics which directly affect contact life. There are several different methods of suppression and each method has disadvantages.

The bifilar coil relay has two windings; a power winding and a shorting winding which absorbs the inductive energy from the power winding. The use of bifilar windings increases the dropout time and can increase bounce and arcing of the contacts. It is not polarity sensitive.

The use of a resistor in parallel with the coil is the simplest and oldest suppression technique. The resistor increases the power requirement for operation and like the bifilar coil, increases dropout time, bounce, and arcing.

The use of a diode is an effective method for suppression and is widely used. It is polarity sensitive and affects dropout, bounce, and arcing.

The use of back-to-back zener and diode or a zener-zener combination is the most effective method of suppression. These combinations have the least effect on dropout time, bounce, and arcing. The zener diode is polarity sensitive.

- c. Operate (pickup) and dropout voltage and current. Disregarding the increase in temperature resulting from the heating effect of the current flowing through the coil, coil resistance will be approximately

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140 percent higher at 125 °C than it is at the 25 °C ambient temperature. Under the same conditions when the voltage across the coil is held constant, the corresponding coil current and hence operating ampere-turns at 125 °C will be approximately 72 percent of the operating current at 25 °C ambient temperature. The increase in coil resistance and resulting decrease in power may cause marginal relay operation.

Coil resistances are normally specified at 25 °C. For the temperature of interest, the actual resistance can be approximated from the following formula:

$$R = R_{25} [1 + 0.0038 (T - 25)]$$

Where:

R = Actual coil resistance

R₂₅ = Coil resistance at 25 °C

T = Ambient temperature in °C

For example, at an ambient temperature of 125 °C, a 600-Ω coil will have a resistance of:

$$\begin{aligned} R &= 600 [1 + 0.0038(125 - 25)] \\ &= 600 (1 + 0.38) = 828 \Omega \end{aligned}$$

at -65 °C

$$\begin{aligned} R &= 600 [1 + 0.0038 (-65 - 25)] \\ &= 600 (1 - 0.342) = 394 \Omega \end{aligned}$$

The above examples considered only coil resistance change due to ambient temperature changes. Application of power to the coil causes heating and hence changes the coil resistance. To obtain the worst condition for high temperatures, this change must be added to the ambient temperature. Figure 1 shows the comparative values of coil resistance versus ambient temperature with and without coil voltage applied.

Care must be taken to insure that the relay will operate correctly at the maximum temperature. For example, a relay with a 600-Ω coil has a specified maximum pickup voltage of 13.5 V at 25 °C. The pickup voltage at any temperature can be calculated from the following formula:

$$E_{25} [1 + 0.0038 (T - 25)]$$

The formula is based upon a voltage shift of 0.38 percent per degree Celsius.

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From the above example, the maximum pickup voltage at 125 °C would be:

$$E_{125} = E_{25} [1 + 0.0038 (125 - 25)]$$

$$E_{125} = 13.5 (1 + 0.38)$$

$$E_{125} = 18.6 \text{ V}$$

Note that the pickup voltages listed in the various specifications are for relays where the case temperature is equal to ambient temperature. The designer cannot disregard the effects of temperature increase due to coil heating.

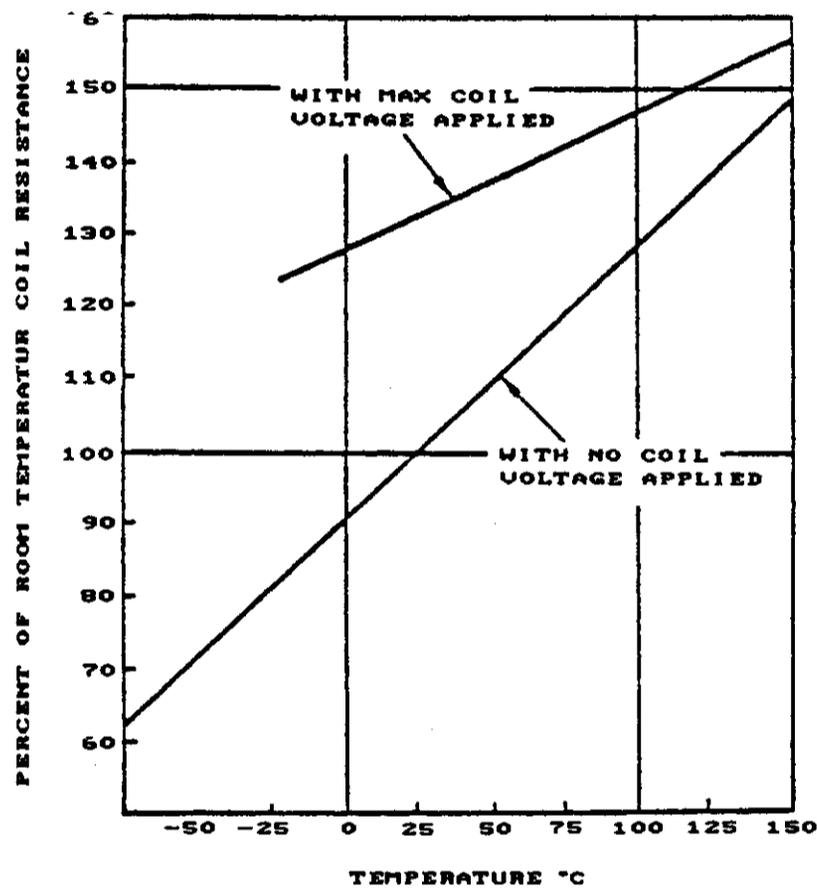


FIGURE 1. Coil resistance variation with temperature.

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The 18-V pickup at 125 °C is important since some military specifications require this for relays designed to operate on 26.5 V systems.

At low temperatures the operate voltage will decrease at a rate proportional to the decrease in coil resistance. Thus a relay which operates at 13.5 V at room temperature will operate at $13.5 (1 - 0.342) = 8.9$ V at -65 °C.

Maximum and minimum release voltages are also affected by ambient temperature in the same manner as the pickup voltage.

$$E \text{ (at } +125 \text{ °C)} = 1.38 \times \text{maximum release voltage at } 25 \text{ °C}$$

$$E \text{ (at } -65 \text{ °C)} = 0.658 \times \text{minimum release voltage at } 25 \text{ °C}$$

Ideally, operate and release currents should not be affected by changes in ambient temperature. Because of the changes in the spring modulus of the movable contacts, and because of changes in fits and clearances due to expansion or contraction of parts, considerable variation does occur. Over the range from -65 to +125 °C, operate and release currents may increase by as much as 10 percent over the 25 °C value and decrease by as much as 20 percent. Normally the value will increase at low temperature and decrease at high temperature, but this is not an infallible rule.

The relay will not operate below the rated coil voltage.

- d. Response speed. Mass and inertia of the armature and movable contact element, and electrical losses of the circuit, affect the operating speed. Some small relays operate in 2 to 3 ms, while larger units have pickup times in the 15 to 25 ms range. The greater the current or power over the minimum required to operate the relay, the faster the action.

Dropout time is usually faster than the operate time, and is almost entirely dependent upon parameters of the relay, such as spring rates and residual gaps.

17.1.4.3 Contact considerations.

17.1.4.3.1 Relay contact arrangements. An important consideration for relay applications is the contact arrangement. The contact arrangement is the various combinations of different basic contact forms that make up the entire relay switching structure. Relay contact notations are given in the following order.

- a. Poles. Each movable contact of a relay and its associated normally open and normally closed contacts constitutes a pole.

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- b. Throws. Double throw transfer contacts abbreviated DT, are a combination of two stationary contacts and a movable contact which engages one of the stationary contacts when the coil is energized and engages the other when the coil is unenergized. Contrasted with double-throw contacts, normally open or normally closed single contact are called single-throw contacts, abbreviated ST.
- c. Normal position. A combination of a stationary contact and a movable contact which are engaged when the coil is unenergized, is referred to as back, break, form B or normally closed contacts and is abbreviated NC. A combination of stationary contact and a movable contact, which is engaged when the coil is energized, is referred to as front make, form A or normally open contacts, and is abbreviated NO.
- d. Double break. A combination, abbreviated DB, in which a movable contact simultaneously makes and simultaneously breaks connection between two stationary contacts. For normally open contacts, this combination may be called doublemake contacts. All contacts are single break except when noted as double break; example: DPST NCDB designates double pole, single throw, normally closed, double-break contacts.

The common contact configurations are shown in Table II. Other configurations are available and are shown in the referenced relay specifications.

17.1.4.3.2 Contact configurations.

- a. Number of circuits. Basic to the operation are the number of circuits to be controlled and the type of switching required for each. Many possible combinations of contact arrangements can be used. Multiple contact sets can be combined within one relay package.
- b. Contact loads. After the number of contacts has been determined, the electrical load on each set of contacts must be analyzed. Open-circuit voltage and whether it is ac or dc must be specified. The ac frequency should be noted.

If the load is inductive, an effort should be made to determine the power factor and the amount of inductive reactance, or at least to describe the load, such as solenoid or contractor. If the load is a motor, complete information should be given about size, horsepower rating, and starting current.

A tungsten filament lamp load can be very difficult to switch. Contacts used to close the circuit to a cold filament must handle a current of ten times the steady state current. A 5-A contact, which might generally be considered adequate to operate 500 W from a 115 Vac source, is completely inadequate for a 500 W lamp, since the initial starting current would be on the order of 50 A. Such high surge-currents may result in sticking or welding of the contacts.

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TABLE II. Contact configurations

Form	Description	USASI symbol
A	Make or SPSTNO	
B	Break or SPSTNC	
C	Break, make, or SPDT (B-M),	
X	Double make or SP ST NO DM	
Y	Double break or SP ST NC DB	
Z	Double break, Double make SP DT NC-NO (DB-DM)	

Duty cycle and frequency of operation should also be specified in connection with contact performance. Some contact systems fail after relatively few operations when operated at a high rate of frequency, whereas the same contacts could handle the same loads many times when switched less frequently. Shortened life at high operational frequency is caused by heat developed at the contacts.

Refer to paragraph 17.1.4.6 Electrical considerations for an indepth discussion of contact ratings.

- c. Contact life. In many applications, relays may be required to operate many thousands or perhaps millions of times in order to provide an acceptable minimum life. A realistic appraisal must be made of each situation.

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As the contacts start to close, the material at the point of initial interface deforms until the contacting area supports the contact force. On a microscopic level, many point-contacts take place to form the electrical connection and carry the load current. If the current density is too high, localized melting takes place, and additional point contacts are engaged until there is sufficient surface area to carry the load.

The selection of contact material is based on the intended applications. Silver and silver alloys are normally used for power loads. Silver has excellent electrical and thermal characteristics. Arcing can precipitate carbonaceous products on the surface, normally in a ring around the primary contact points. The deposits can contribute to higher contact resistance, although the carbon is desirable to the extent that it minimizes sticking or welding in high current applications.

Gold plated silver contacts are commonly used in dry circuit or low level applications where stable contact resistance is essential. Gold has a very high resistance to surface film formation. It does exhibit poor resistance to sticking or welding, and it has poor mechanical deformation and wear characteristics.

High-current density at the point contact locations, during the first instant of contact closure, can cause contact surface melting, even at fractional ampere loads. This phenomenon is of practical concern in that it can result in contact sticking and metal transfer which potentially reduces long contact life. The effect is amplified by slow contact operation due to reduced coil voltage, slow rise time of the power supply, and circuit or line capacitance. Other factors that affect current density are high inrush current for lamp loads, switched motor loads, capacitive loads without current limiting resistors, and transformers. Figure 2 details the effect on contact life and thereby reliability by derating of the contact load.

With adequate arc suppression, contact load life can be extended to nearly the mechanical life expectancy. Initial values of resistance and capacitance for an RC arc suppression network may be calculated from:

$$R = (E/10) I(1+50/E), \text{ and } C = I^2/10,$$

Where:

R = resistance (ohms)

E = voltage before closing (volts)

I = current before opening (amperes)

C = capacitance before closing (microfarads)

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Peak rather than rms values of voltage and current must be used to calculate arc suppression of ac loads.

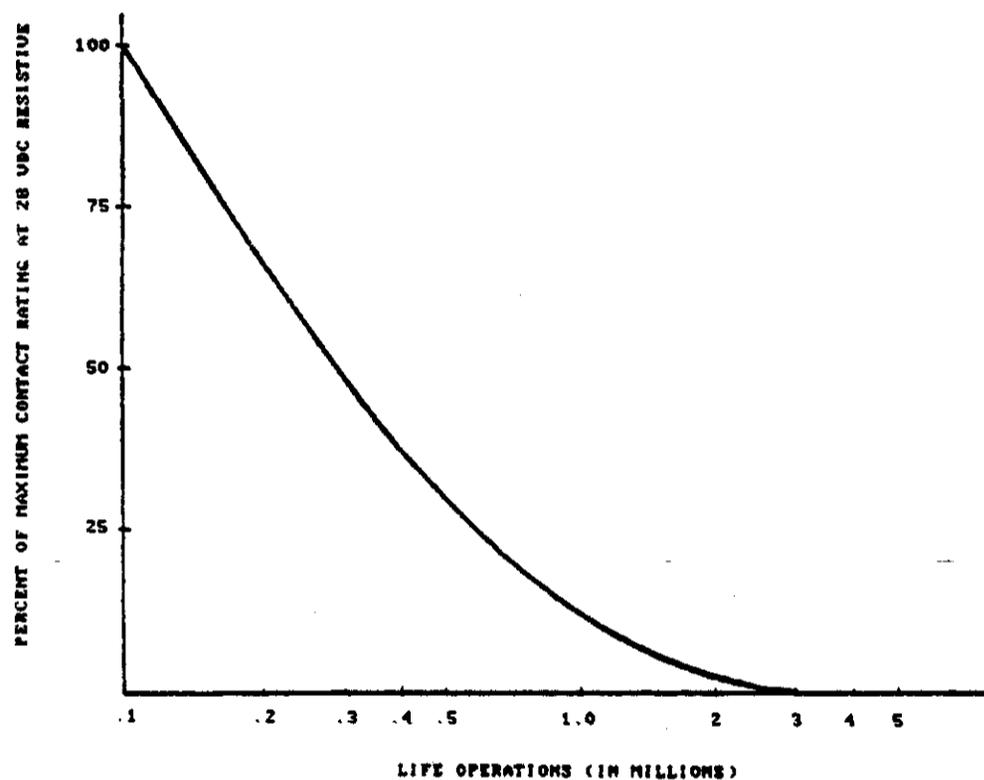


FIGURE 2. Effects of derating on contact life.

To ensure adequate suppression the RC network should be tested. If necessary, resistance and capacitance values can be adjusted to eliminate arcing at the contacts.

Several other systems are commonly used for arc suppression, such as resistors alone, capacitors alone, diodes, diodes and resistors, back-to-back diodes, neon glow lamps, magnetic blowouts, noninductive windings, and varistors.

For further details, consult paragraph 17.1.4.6 of this section.

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17.1.4.4 Environmental considerations. Ambient temperature affects operation of the relay and the ability of the contacts to switch and carry the circuit load. Because the coil of an electromagnetic relay is usually wound with copper magnet wire, compensations for the change in resistance due to change in temperature must be made. A MIL-R-5757, MIL-R-6106 or MIL-R-39016 relay requires operation at a maximum ambient temperature of 125 °C. If a 24 Vdc relay is considered for such an application, it must be capable of picking up at 18 V or less at the maximum ambient temperature. The relay must pull in at approximately 15 V at room temperature in order to pickup at 18 V at 125 °C. If this relay is required to dropout at 5 V at a temperature of -65 °C, it may be necessary to establish the dropout point when measured at room temperature as high as 10 or 11 V. Thus a close-differential relay may be required, which is another reason for keeping the pickup and dropout points as far apart as possible.

Other factors which may be grouped under general environmental conditions are high humidity, corrosive or explosive atmosphere, and dusty or dirty conditions. Explosive atmospheres are especially critical because of possible arcing of the contacts. These factors affect the protection provided for the coil, the plating or other finish required on metal parts, the insulating materials, and finish or treatments required to ensure serviceability. Many military applications dictate that hermetically sealed relays be used.

In most military applications, effects of shock, vibration, and linear acceleration have to be considered. It is possible to make certain types of relays that will resist vibration effects up to 3000 Hz at a level of 30 G and shock at a level of 100 G.

Environmental considerations are discussed in greater detail in the sections devoted to specific relay types.

17.1.4.5 Physical considerations.

- a. Enclosures. Many types of enclosures are available to protect the relay from external conditions, particularly high humidity and dirt. Relays may be loosely classified according to the degree of protection offered by the enclosure. Such classifications include the following: unenclosed, partially enclosed, enclosed, sealed, gasket sealed and hermetically sealed. It can be seen that, in common with other relay features, the relatively simple matter of enclosures is subject to considerable variation.

With the unenclosed relay, no effort is made to protect the relay or its parts from atmospheric conditions. Between the totally unprotected relay and the hermetically sealed type come the several variations noted previously in which the contacts or coils (but not both) may be enclosed, or in which the entire relay is enclosed but is not airtight, or the partially sealed type in which the coil or contacts (but not both) may be enclosed and contacts in an airtight container. The sealed relay has both coil and contacts in an airtight container, the seal of which may be broken for adjustment.

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The hermetically sealed relay is made airtight by a sealing process which involves fusing, or welding, and it does not use a gasket. Usually the air is removed and replaced by an inert gas under pressure. It cannot be unsealed and resealed without special equipment. This is usually performed under clean room conditions. In high-reliability applications, it is mandatory that the relays be hermetically sealed and that they be given an internal atmosphere of dry nitrogen. This activity should all be accomplished in a room meeting Class 100 cleanroom conditions as specified in FED-STD-209. A design approach that is sometimes used to reduce the chance of contact contamination is to have the contacts and coil in two separate adjacent hermetically sealed chambers in the same housing. This design increases the size and cost of the unit.

The different types of enclosures mentioned above offer different degrees of protection. The hermetically sealed enclosure offers the greatest protection, because it insulates against such elements as moisture, harmful gases, and foreign particles, which are the most common causes of relay failure. It also eliminates the increased arcing induced by low atmospheric pressures at high altitudes.

- b. Terminals. Terminals are available in a considerable variety of styles. They generally are made of round wire with material and size dependent upon type of enclosure, header design, and current and voltage requirements.

The number of terminals in the header will be determined by the number of poles and throws, and the necessary connections for the coil. An individual terminal with a glass or ceramic insulating bead for each may be used. The other scheme is to have a glass insulator with multiple terminals. The distance between terminals will determine the allowable voltage.

The type of terminals to use on a sealed relay depends upon the relay application. If, for example, it is to be used in a computer where convenient replacement is desired, then the plug-in type would be best. Plug-in type relays in the referenced military specifications have gold plated terminals to minimize terminal to socket resistance for plug-in applications. If the relay is mounted inside a piece of equipment subject to shock and vibration, then the solder type terminal is preferred. These terminals should be of a solder hook variety to reduce the stress on the header.

- c. Miniaturization. Miniaturization of relays has been going on for some time. A relay that was considered miniature several years ago may be big and awkward compared to its counterpart today. Relay manufacturers are devoting a lot of time and effort to the problem of miniaturization. The problem is complicated by the fact that the conditions requiring miniaturization also require the relay to withstand higher temperatures

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and more shock and vibration. It is not only a matter of reducing all dimensions, but also of utilizing new materials in new designs to obtain longer operating life, greater environmental resistance, and greater reliability. Relays are available in several designs packed in TO-5 transistor style cases with pin terminals or wire leads. This relay includes electromechanical, hybrid, and solid state devices.

17.1.4.6 Electrical considerations. The prime electrical considerations in selecting a relay are contact rating and coil voltage. The contact ratings and coil voltage of the relays described in this section are based upon MIL-R-5757, MIL-R-6106, and MIL-R-39016, the most commonly used military specifications.

The resistive load rating is that load which the relay can be expected to switch for 100,000 operations minimum at the maximum ambient temperature, (usually 125 °C) and at a cycling rate of 20 cycles per minute with a 50 percent duty cycle.

A change in any of these parameters will affect the expected life of the relay. For example, a reduction of the temperature by 50 °C can double the life expectancy of the contacts. Similarly, a reduction of the cycling rate or the ratio of on-time to off-time will have some effect on increasing the life expectancy.

The distinctive types of loads are:

- a. Resistive loads. The basic resistive load given for each relay on the individual spec sheets forms the base line rating for the relay. If the relay is tested under the conditions of MIL-R-6106 or MIL-R-39016 at the specified rating, it will perform under the rated load conditions for a minimum of 100,000 operations. Thus, the ratings and life expectancy figures do not represent the point at which the relay may be expected to fail, but represent the minimum number of operations at rated load which the relay can be expected to perform with high confidence levels.
- b. Inductive and motor loads. When the current to a dc inductive load is broken by the relay contact, the collapse of the magnetic field of the inductor or will induce a short-lived transient voltage which may be severe enough to cause considerable arcing.

The severity of this effect will depend upon the L over R ratio of the load, the normal operating current, and the speed with which the contact opens. Generally, such loads as circuit breaker coils, solenoids, or stalled dc motors provide high load currents and high L over R ratios, and will generate an arc sufficient to permanently damage or destroy a pair of contacts. The arc generated by this type of inductive load also serves to generate severe radio-frequency interference.

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The use of relay contacts, to break the inductive load of a second relay coil, will induce a high transient voltage even when the second relay coil is properly suppressed. For example, assume the contacts are switching a 28-Vdc line to a second relay coil which is suppressed to 42 V. The contacts will see the supply line of 28 V plus the 42-V transient of the coil during switching. The effective voltage of 70 V appears across the relay contacts during turn off of the second relay. Because most of the referenced military specifications rate the contacts at 28 Vdc, the load shown in this example exceeds the capability of the contacts, resulting in excessive arcing, shortened life, and potential failure. The effect of switching on unsuppressed coil could be catastrophic with surge voltage levels up to 1500 V.

Since each inductive load has its own time constant characteristics (its own L over R ratio) and its own load currents of a steady state nature, it is virtually impossible to establish a fixed inductive rating for each type of relay, which would be meaningful to the user. In general, however, the inductive load rating for any inductor of significance would be less than 50 percent of the normal resistive rating for the relay. There are two practical approaches to the solution of this problem, as follows:

1. Transient suppression of the load is recommended wherever possible. If proper suppression is utilized, the normal resistive ratings of the relay can be used, simplifying the acceptance testing procedure and the qualification-testing procedures necessary for the part. In addition, proper suppression provides protection against generation of radio-frequency noise, since the contact arc has been reduced considerably. For several typical suppression circuits, refer to the circuits shown in Figures 3 through 5 of this section.
 2. Consultation with the components engineer with knowledge of the specific load requirements will help in the selection of the proper relay for switching the load. When a relay has been selected, it should be tested under actual load conditions to insure that it will perform the function properly.
- c. Current loads. Between rated load conditions, as discussed above, and dry circuit contact loads, discussed later in this section, lies a vast region in which most contact applications fall. That portion of the contact load area falling between approximately 25 to 500 mA can generally be considered to be in the intermediate current area. It would normally be expected that contacts rated at 2 A, 30 Vdc would operate perfectly well in the intermediate current area. Such is not always the case. The relay may very well switch the load without any catastrophic failure of the contact system, but the failure criteria of contact resistance, if used, may very quickly fail the relay under intermediate current conditions.

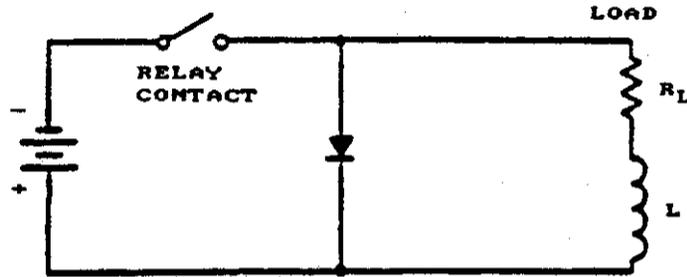


FIGURE 3. Diode arc suppression.

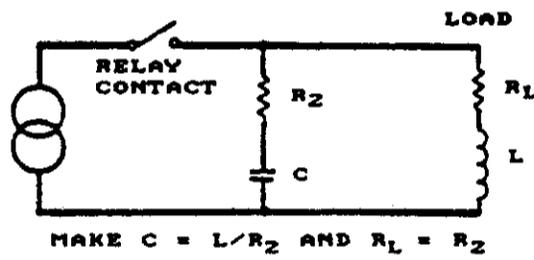


FIGURE 4. RC arc suppression.

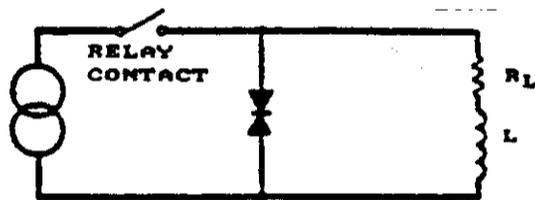


FIGURE 5. Diode arc suppression, ac.

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Under rated load conditions, the current density and arcing that occur at the contact surface are sufficient to provide a constant cleaning or burnishing action, keeping the contacts free of insulating contaminants which may be formed. In the case of low level circuit applications, the voltage and current levels are extremely low, and no arcing occurs.

In the case of the intermediate current application, the current density in the contact area will be sufficient to aggravate normal galling actions and to carbonize any hydrocarbons that may exist in a gaseous condition within the relay with any residues, developing deposits in and about the contact area. The result is that the relay will exhibit contact resistance which is out of specification under intermediate current load conditions, although failure of the relay to switch the prescribed load may not necessarily be impaired, and, in fact, is a condition which rarely occurs.

In the event that higher than normal contact resistance is a problem in the application of the relay in the intermediate current area, care should be exercised by the specifying engineer to inform the manufacturer that such is the case. This condition can be minimized or eliminated by proper controls of the manufacturing techniques, usually involving more critical handling and manufacturing methods, and additional precautions in the selection of materials utilized in the relay construction. Thus, for a relay designed specifically for low contact resistance in the intermediate current area, the basic cost is usually higher than for a standard production line unit.

- d. Low-level loads. Low-level loads are generally those which are in the millivolt and microampere region. The common military specifications (MIL-R-6106 and MIL-R-39016) have established 6 Vdc and 10 mA as maximum values for testing purposes. With voltages and currents of this value or less, there is insufficient voltage to break through a barrier film and insufficient current to produce an arc that might cleanse the contact area. For this reason, the contacts of the relay are of a noble metal (usually gold) to avoid corrosion, and are manufactured to maintain optimum cleanliness. The softness and inertness of gold reduces the susceptibility to and maintenance of film contamination.

For relays to be utilized in low-level applications, it has become common practice for the design engineer to specify the usage, and for the manufacturers of the relays to provide for a 5- or 10-thousand operation run-in miss test, with all contacts monitored. A test of this nature provides data on the ability of the relay to perform its low-level function, eliminates the infant mortality failures, and provides reliability assurance on the particular production lots in question. Relays which are to be used in low-level applications should not be used or tested at higher currents prior to usage in low-level applications.

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- e. Dry-circuit loads. Dry-circuit loads are applications in which the relay contact does not make or break the current. The contacts are closed before the power is applied through the contacts. Power is always removed before the contacts are opened. The contact may be any value within the specified rating, as the contact does not make or break the load current. There is no arcing to cause erosion of the contact surface. The life of the contacts used in dry circuit applications approaches the mechanical life of the relay. Contacts are normally plated with noble metal to avoid corrosion or oxide formation.
- f. Lamp loads. A tungsten filament lamp has the characteristic that when cold, it has a much lower resistance than when incandescent. Therefore, contacts which are used to switch lamp loads will see a surge current (inrush current) which may be 3 to 10 times the steady state current of the lamp. For most applications this is not too severe, since most of the lamp loads encountered are of the small miniature types. However, a lamp with a steady state current near the rated current of relay should be used with caution, since there is a possibility that the inrush current will weld the contacts. In the event that this may be a problem, a small series resistor may be used with the lamp to limit the inrush current. The use of series resistor has an additional benefit as it reduces the lamp voltage, thereby extending lamp life.
- g. Capacitive load. A capacitive load has the characteristic of a short circuit through the contacts for a short duration, depending on the size of the capacitor and the series resistance. Although the time interval may be as low as a few microseconds, sufficient current passes through the contacts to permanently damage or weld them. The practical solution to this type of load is to add sufficient series resistance to limit the inrush current to an acceptable level.
- h. Contact load and life curves. The life versus load curve shown in Figure 2 is the result of many hundreds of relays being tested under a variety of load conditions. Because the curve is intended to cover a broad spectrum of relay types, it has been very conservatively drawn, and can be referred to in establishing the resistive contact ratings for any of the MIL-R-5757, MIL-R-6106, or MIL-R-39016 relays. This curve does not specifically refer to low-level contact life, which must be considered under the special conditions of extremely clean contacts for the life of the relay.
- i. Contact arc suppression. To avoid damage to contacts when the load has an inductive component, one of several methods of arc suppression should be used. For dc loads, a simple diode may be employed as shown in Figure 3. The diode must be able to pass a current equal to the normal load current, and should have an inverse voltage rating higher

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than the source voltage of the circuit. For ac circuits, two methods are available. One is to make the load impedance appear resistive by the use of a series RC circuit in parallel with the load. (Refer to paragraph 17.1.4.3 of this section for equations calculating the values of R and C). The second method is by the use of special ac diode suppression available commercially. Figure 4 illustrates the resistive load principle, good for any frequency of operation, and Figure 5 shows the application of diodes, which must withstand load current and voltage. These devices should be separate components and not built into the relay. The use of additional components inside the package reduces reliability though particle contamination, outgassing characteristics, and the additional failure rate of the added parts.

17.1.5 General guides and charts. The effects of ambient temperature on relay performance is often overlooked. Because of the change in winding resistance with temperature, the force available to operate the armature decreases with increasing temperature. Figures 6 and 7 show this effect.

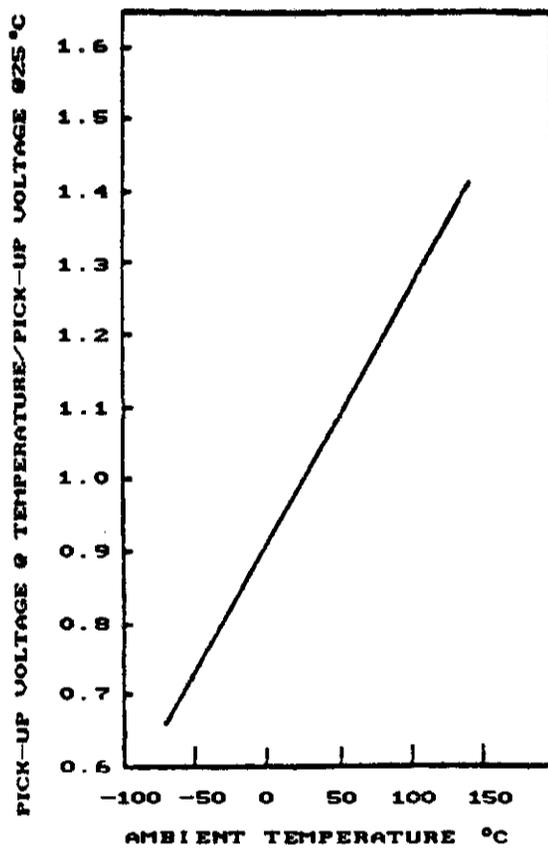


FIGURE 6. Pickup voltage vs temperature.

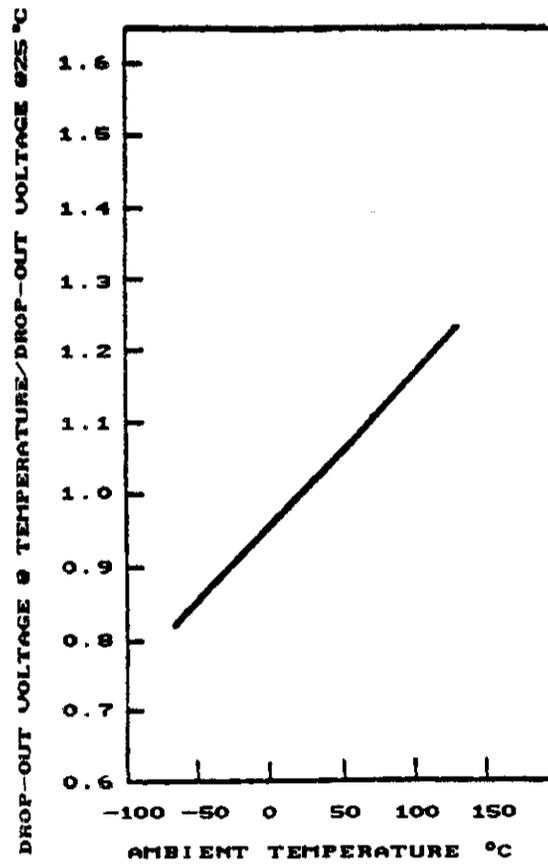


FIGURE 7. Dropout voltage vs temperature.

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17.1.6 General reliability considerations. Relays are electromechanical devices and are subject to both electrical and mechanical failure. Some causes of failures are poor contact alignment, loss of resiliency in springs, and open coils, as well as open, contaminated, or pitted contacts. Contact failure can result from high inrush or sustained high current, or from high voltage spikes when an inductive circuit is opened. High inrush currents occur in loads composed of motors, lamps, heaters, capacitive input filters, or other devices that have low starting resistance compared to operating resistance. These currents may cause intense heat with associated welding of the contacts.

In addition to overstressed contacts, contamination is the most common cause of relay failure. It is a particularly annoying cause of failure because it is often intermittent and difficult to verify. This may be nonmetallic or gaseous contamination, which periodically deposits itself on the contacts, causing an open condition; or metallic particles which cause shorted conditions or block movement of the mechanical parts. This cause of failure can be significantly reduced by process controls, use of welded hermetic sealed enclosures, small particle cleaning, assembly and back filling in Class 100 clean room facilities, as defined in FED-STD-209 precap visual inspection, and added screening after assembly.

There have been a number of failures attributed to vendor workmanship. Timely corrective action by the vendor has significantly reduced this type of failure.

Engineering selection of the proper relay for an application is the most significant factor of relay reliability.

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17.2 RELAYS, ARMATURES

17.2 Armatures.

17.2.1 Introduction. Armature relay operation depends upon energizing an electromagnet which attracts a hinged or pivoted lever of magnetic material to a fixed pole piece. The hinged or pivoted lever is called the armature. Another style, sometimes erroneously called a rotary relay, is pivoted between two pole pieces and the magnetic circuit contains two airgaps. Although the motion of the armature is circular, the relay is still basically an armature type. If necessary, a delay may be introduced into either the operating time, the release time, or both. This is discussed in subsection 17.4 Time delay. MIL-STD-975 includes armature relays in accordance with MIL-R-39016 for Grade 2 application.

17.2.1.1 Applicable military specifications.

MIL-R-5757	Relay, Electromagnetic
MIL-R-6106	Relay, Electromagnetic [Including Established Reliability (ER)]
MIL-R-28776	Relay, Hybrid, Established Reliability
MIL-R-39016	Relay, Electromagnetic, Established Reliability

Only MIL-R-39016 is included in MIL-STD-975.

17.2.2 Usual applications. Relays may be used in any of the following applications:

- a. To obtain isolation between the input circuit and the output circuit
- b. To invert the signal sense (from open to closed and vice versa)
- c. To increase the number of output circuits (so as to switch more than one load or to switch loads from different sources)
- d. To repeat signals
- e. To switch loads of different voltage or current ratings
- f. To retain an input signal
- g. To interlock circuits
- h. To provide remote control.

Proper application of relays requires consideration of characteristics such as input power or current available, the contact arrangements and load requirements which must be met, as well as size limitations and all other electrical, physical, and environmental characteristics which the relay must meet. If general understanding of the varieties of individual relay characteristics

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is possessed, it is possible to optimize a given characteristic which may be important in a given circuit. The military requirements are stated at a specific set or sets of conditions. For applications which vary from the specified conditions, the expected performance capability of the relays must be controlled by specification.

17.2.2.1 Misapplications. The following, extracted from MIL-STD-1346, discusses potential misapplications of relays.

- a. Improperly using existing military specifications by erroneous interpretation or even using the incorrect specification altogether. A given set or sets of conditions are provided in the specifications. Variations from these conditions will affect performance of the relay accordingly.
- b. Paralleling contacts to increase capacity. Contacts will not make or break simultaneously, and one contact carries all the load under the worst conditions. Contacts can be paralleled for added reliability in the low level or minimum current (contamination test current) areas.
- c. Circuit transient surges. Circuit designers must be careful not to expect relays to handle circuit transient surges in excess of their ratings.
- d. Using relays under load conditions for which ratings have not been established. Contact ratings should be established for each type of load. Many relays will work from low level to rated load, however, the designer should not expect such performance.
- e. Using relays on higher voltages than those for which they were designed; for example, switching 300 V power supplies with relays only rated for 115 V maximum.
- f. Contact ratings with grounded case. Relays switching 115 Vac with a grounded case must have the contact ratings significantly lower than in the ungrounded case mode of operation. The maximum ac rating of a nominally rated 28 Vdc, 2 A resistive relay is of the order of 0.300 A. Switching the ac loads with the relay case ungrounded results in a potential hazard to personnel.
- g. Transferring the load between unsynchronized power supplies with inadequately rated contacts. When the load is switched, the voltage amplitude can range from in phase to 180 degrees out of phase; therefore, the relay contact voltage may vary from zero volts to two times peak voltage and maximum current.
- h. Switching polyphase circuits with relays tested and rated for single phase only. A typical misapplication is the use of small multipole relays (whose individual contacts are rated for 115 V single-phase ac

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17.2 RELAYS, ARMATURES

in 115/220 V three phase ac applications. Phase-to-phase shorting at rated loads is a strong possibility in these instances, with potentially catastrophic results.

- i. Using relays with no established motor ratings to switch motor loads. In addition, caution should be used in applying relays to provide braking by reversing a motor, particularly where the motor can be reversed while running, commonly called plugging. This results in a condition where both voltage and current greatly exceed normal.

Many power relays should not be utilized in potential plugging situations unless so rated by the manufacturer.

- j. Using relays with no established minimum current capabilities. It must not be assumed that because a relay is used in an application considerably below its rated contact load that the consideration of a minimum current capability can be ignored; this is especially true if there is no established level of minimum current for the relay.
- k. Using relays rated for 115 Vac only on 28 Vdc or higher voltage dc applications. If contacts in these devices are of the single break form-A type, it may be necessary to derate severely for use on dc applications using 28 V or higher.
- l. Using relays at coil voltages below rated coil voltage. The operational parameters, life characteristics, and ability to withstand dynamic environments are based on operation at rated coil voltages. Reducing the coil voltage may extend time to operate, reduce contact closure forces, reduce holding force, increase arc duration, and reduce ability to operate properly during vibration or shock, or may preclude the relay from operating at all.
- m. Using relays when the coil is driven by a slowly rising power supply. The relay operates during some point during the increase in driving current from the power supply. Back electromotive forces (EMF) are produced when the armature closes to the pole face. The back EMF is the opposite polarity to the driving voltage causing the relay to release and then reoperate when the drive current increases to a sufficient level to overcome the back EMF.

17.2.3 Physical construction and mechanical considerations. Mechanical considerations briefly discussed in this section are size and shape, weight, terminals, method of mounting and contact arrangement. Physical construction is briefly discussed in paragraph 17.1.4 General parameter information.

Relays are constantly being made smaller, as are other component devices. Very complex relay devices are now available in a TO-5 transistor package. Ratings in this size vary up to 1 A. Other miniature sizes available with higher ratings include the crystal can series. They range from one-sixth crystal can size to full-size crystal can. The above devices are available as military standards with qualified suppliers.

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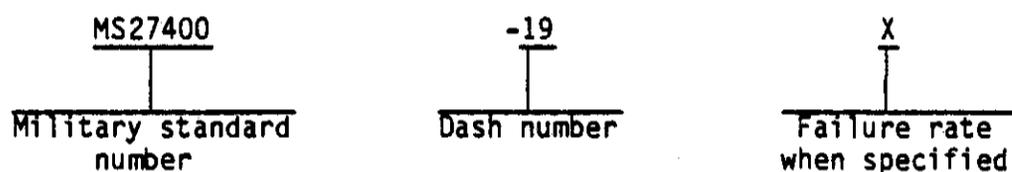
The most common terminals supplied with relays are screw, solder, plug-in, printed-circuit and quick-connect. Military applications usually require the solder, plug-in, or printed circuit type.

Hermetic sealing of relays used for military applications is generally required. MIL-R-5757, MIL-R-6106, MIL-R-28776, and MIL-R-39016 now require that cases be of welded construction. Previously soldered units were somewhat susceptible to interior contamination. Welding will increase the reliability of the hermetically sealed units.

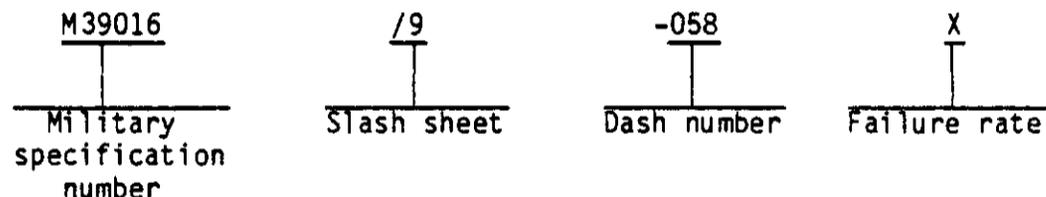
Figure 8 shows the internal construction of a typical armature relay with an associated parts list.

17.2.4 Military designation. Two methods of specifying relays are as follows:

a. Military standard



b. Military specification (only MIL-R-39016 is included in MIL-STD-975)



MIL-R-5757 is the general military specification for electrical relays with contact ratings up to and including 10 A. However, this specification is now being replaced by MIL-R-39016. As rapidly as sources become qualified for MIL-R-39016, they will replace equivalent MIL-R-5757 parts. Many styles, shapes, and sizes are available.

MIL-R-6106 is the general military specification for electrical relays for use in aerospace systems. They are capable of being mounted directly to the aircraft, missile, or spacecraft, and the relays are available in a wide range of contact ratings up to 400 A. The method of specifying these relays is by military standard reference or military specification number as appropriate.

MIL-R-28776 is the combination of a solid state input element functioning with an established reliability electromechanical relay that is hermetically sealed. The method of specifying these relays is by military specification number.

MIL-R-39016 is the general military specification for established reliability electromagnetic, hermetically sealed relays. These relays are designed to operate in low and medium power switching circuits with contact ratings up to 10 A.

17.2 RELAYS, ARMATURES

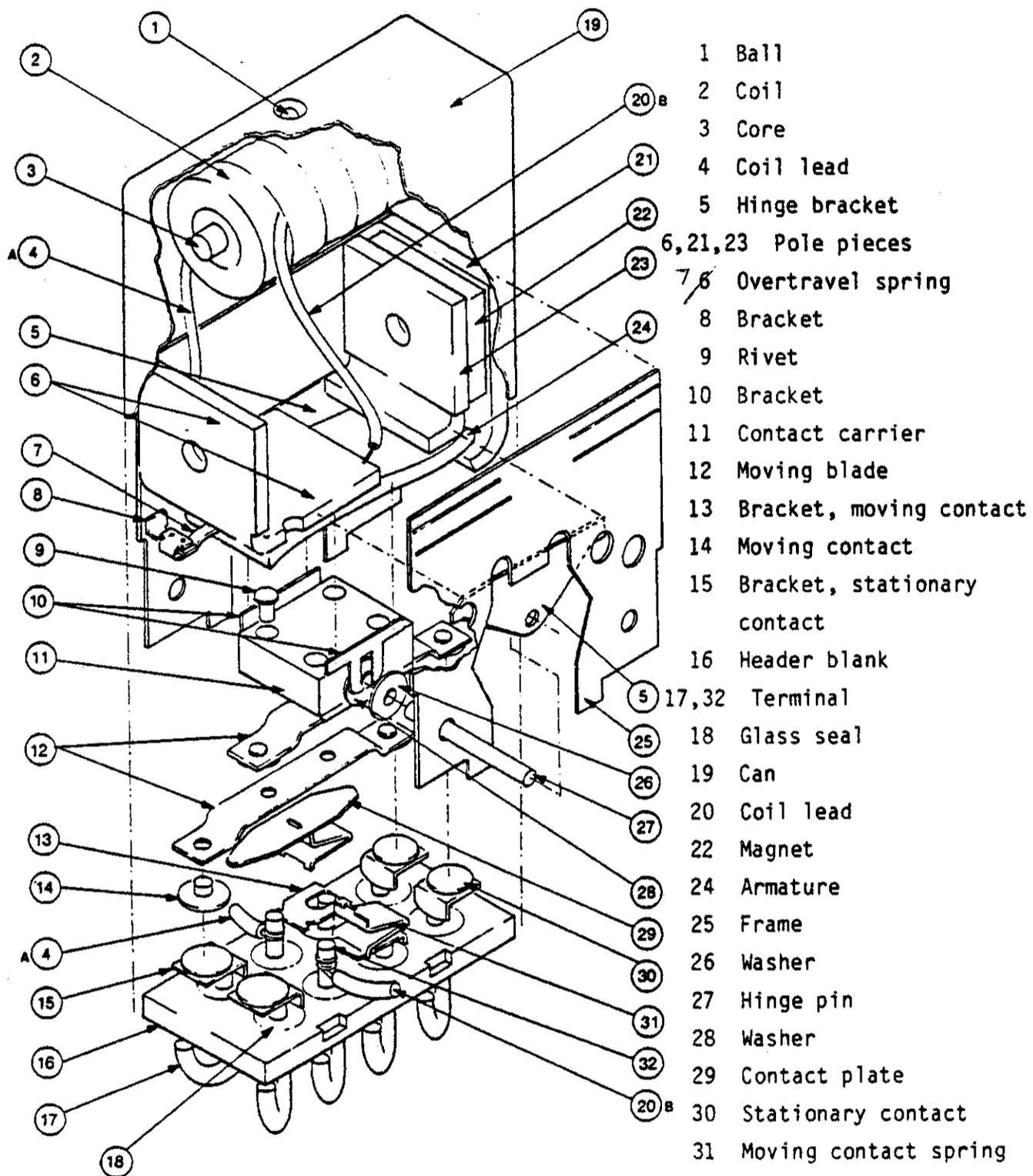


FIGURE 8. Typical armature relay.

17.2 RELAYS, ARMATURES

17.2.5 Electrical characteristics. The primary electrical considerations in selecting a relay are contact rating and coil voltage. Secondary considerations should be type of voltage (ac or dc), frequency (ac coils or load), thermal effects, types of loads (high power, intermediate, low level, inductive, motor, lamp, etc.), cycle rate, duty cycle, and transients. Each of these topics are covered in paragraph 17.1.4.

17.2.6 Environmental considerations. The environmental effects on relay operation are frequently overlooked during initial relay selection and application. While the basic temperature range of the relay is always considered, the effect of self heating from the coil and the IR heating across the contacts is often neglected. This increase in temperature is discussed in paragraph 17.1.4, as it affects operational parameters.

The dynamic environments of vibration and shock are the parameters where relays are frequently misapplied. Although the input levels of these environments are known and taken into account during selection, the amplification by a non-rigid mounting surface is often forgotten. When relays are mounted on an unsupported portion of a circuit board or structural surface it is not unusual to measure vibration levels 5 to 10 times higher than the input vibration level.

17.2.7 Reliability considerations. Paragraph 17.1.7 which was concerned with reliability failure data in general should be consulted for reliability and failure mode data.

Relays contained in the military specifications MIL-R-28776 and MIL-R-39016 for established reliability relays demonstrate a failure rate level of 0.01 to 3.0 percent in 10,000 relay operations at confidence levels of 90 percent for qualification and 60 percent for maintenance of qualification. Relays in military specification MIL-R-6106 that are identified as established reliability relays demonstrate a failure rate level of 0.1 to 1.0 percent in 10,000 relay operations at confidence level of 90 percent for qualification and 60 percent for maintenance of qualification.

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17.3 RELAYS, REED

17.3 Reed.

17.3.1 Introduction. The high speed, reduced potential for contamination, and limited number of moving parts of a reed relay provide industry with a bridge between electromechanical relays and solid-state switching devices.

The major component of a reed relay is the reed capsule commonly known as a reed switch. The switch is part of the magnetic circuit and the contact path of the relay, and it embodies the operating and switching parameters. Successful operation of a reed relay in a circuit will almost entirely depend on proper selection of the reed switch and close adherence to the switch specifications.

These devices are not included in MIL-STD-975.

17.3.1.1 Applicable military specification.

MIL-R-5757	Relay, Electromagnetic
MIL-R-83407	Relay Reed, Mercury Wetting
MIL-R-83516	Relay, Reed, Dry

17.3.2 Usual applications. Long electrical life of precious-metal contacts sealed in an inert atmosphere, the absence of wearing mechanical parts, immunity from circuit noise, high circuit isolation, and fast switching speed are the most significant features of reed relays. The small size is adaptable to high-density, printed-circuit-card mounting applications. Low price and a wide variety of contact forms and package configurations encourage use of reed relays in almost all types of circuit applications. Operating speed, under 2 ms, is acceptable for most control systems where the end function is operation of electromechanical devices. Commercial devices, test equipment, and numerous telephone circuits use reed relays.

The various types of reed relays described herein indicate the great versatility of both contact switching capabilities and types of packaging. The unique features of reed relays have caused it to be widely accepted for many types of circuit switching; therefore, the idea of selecting a reed relay to fill all types of switching applications may be very enticing. However, the specific application should still determine the style and type of switching method best suited for a particular designer's requirement.

For example, if complete sealing of the contact switching members and switching speeds in the millisecond range is not needed, a general purpose electromagnetic type relay may well be the most economical choice for multiple pole switching. If closed circuit resistance in the fractional ohm range and open circuit resistance in the multiple megohm range is not needed, but extremely fast switch-speed is needed, a solid state type of switch may be the best choice. The circuit designer should, therefore, consider the various parameters needed for a particular application and attempt to choose the component which is most suitable.

17.3 RELAYS, REED

17.3.3 Physical construction and mechanical characteristics.

17.3.3.1 Construction. The basic reed switch consists of two overlapping, flat, ferromagnetic reeds separated by an air gap, and sealed in a glass tube as shown in Figure 9. The reeds are sealed, one at each end of the tube, so that their free ends overlap in the center of the tube. During the sealing operation, dry nitrogen gas is forced into the tube, creating an inert atmosphere for the contacts. When the switch is introduced into a magnetic field, the reeds become flux carriers, and the overlapping ends become opposite magnetic poles which attract each other. If the magnetic attraction between the reed ends becomes strong enough to overcome the stiffness of the reeds, they move together and touch, completing the magnetic circuit and making electrical contact.



FIGURE 9. Reed switch (NO) construction.

To obtain a low and consistent contact resistance, the overlapping ends of the reeds are precious-metal plated, typically with gold, rhodium, ruthenium, or a combination of these. This plating not only affects the switch's electrical characteristics but also acts as a residual gap and largely determines the release point. The plating must be thin and uniform to ensure that the magnetic properties of the switch are not adversely affected. Plating which is too thick causes the switch to have a high or inconsistent contact resistance and a high release point. Contact resistance depends on the plated contact surface and the contact pressure. Contact pressure is dependent upon the attractive force between the reed ends. The force is a function of the flux density in the gap and is controlled by the size of the reeds, the gap, and the amount of overlap. These factors thus establish the operate characteristics of the switch. All parameters must be rigidly controlled to ensure the consistent operate and release points necessary for using reed switches in relays.

Mercury reed switches are available for applications where contact bounce cannot be tolerated. In this switch, one reed is wetted with mercury which is fed to the contact area by capillary action from a pool at the bottom of the switch. The mercury forms a fluid contact which eliminates bounce and provides a low and consistent contact resistance throughout the life of the switch. Mercury reed switches are suitable for ground based equipment only, since mounting must be maintained in a specified position and under normal gravitational forces.

17.3 RELAYS, REED

The basic reed switch has a normally open, Form A, contact. A normally closed, Form B, contact is made by biasing the normally open contact with a permanent magnet or by mechanically biasing one reed. A Form C contact is made by combining a Form A and a Form B in the same capsule. These contacts have the same ratings as the normally open switches.

Single-capsule Form C reed switches are available in many configurations which can be grouped into two main categories:

- a. Reed switches that use a magnetic bias to close one set of contacts
- b. Reed switches that use a mechanical bias to close one set of contacts

17.3.3.2 Mercury-wetted, 1 form D mechanically biased. Utilizing the characteristics of the mercury-wetted, 1 form C contacts and reducing the air gap, a make-before-break action can be achieved. Figure 10 shows the action of mercury wetted 1 form D contacts changing from the normally closed position, Figure 10A through intermediate positions, Figures 10B and 10C, to rupture of the mercury surface, Figure 10D, and the final state of normally open contacts closed, Figure 10E.

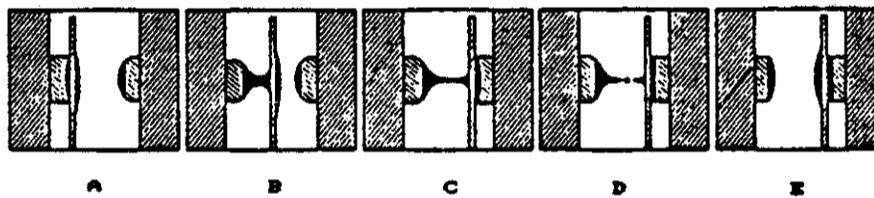


FIGURE 10. Mercury-wetted contact action 1 form D.

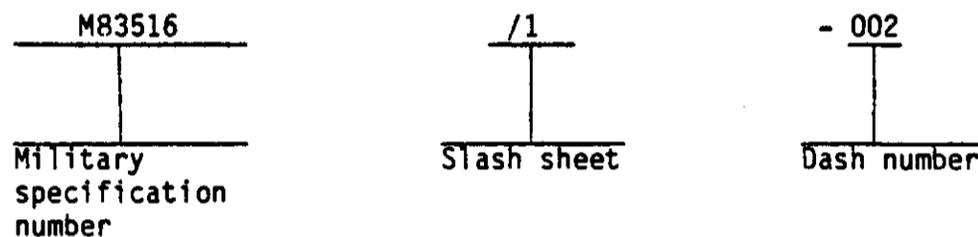
17.3.3.3 Latching relay. Multiple-wound coils are available for a variety of special applications such as logic gates, flip-flops, code-checking relays, and latching or bistable relays. The latching relay, one of the most common special assemblies, consists of a bifilar coil (two windings simultaneously wound on the bobbin), a magnet, and a reed switch. The magnet is strong enough to hold the reed switch closed once it operates, but cannot close the switch without the aid of a coil. A pulse to one winding, aiding the magnet, closes the reed switch, which is then held closed by the magnet. A pulse to the other winding, opposing the magnet, causes the switch to open. The major advantages of the latching relay are its speed, low operate power, and memory without power.

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17.3 RELAYS, REED

17.3.3.4 Relay packages. Reed relays are available in many different configurations to meet standard and special packaging and environmental requirements. Often space dictates which style assembly to use. The low-profile reed-relay package for printed-circuit-card mounting offers the greatest density and convenience.

17.3.4 Military designation. MIL-R-5757, MIL-R-83407, and MIL-R-83516 are the general military specifications for reed relays. Shown below is the proper method of specifying a military designation.



17.3.5 Circuit considerations. To obtain the best performance from reed relays in a circuit, the following points should be considered:

- a. The load to be switched should not exceed the switch rating. Inductive loads must be suppressed, and high inrush loads, such as lamps, should be current-limited.
- b. Although close-tolerance power supplies are not required for reed relays, there must be enough safety factor in the supply to ensure proper operation in spite of adverse temperature and voltage conditions.
- c. Control-pulse waveforms from limit switches and other similar devices must be free of bounce. The speed of the reed relay enables it to follow the discontinuities of some of the most commonly used input switches. Inputs to reed counters or registers should be buffered by devices such as mercury-wetted contact relays, that do not generate bounce.
- d. Whenever possible, the reed switch should not be used to switch loads. Using the reeds to establish a path and then switching the load from a single heavy-duty contact can add millions of operations to the life of the reeds.
- e. Because the reeds will resonate at their natural frequency of approximately 1000 Hz, a sufficient off time should be allowed before re-applying a holding voltage to prevent false reclosure. Dimpled reed switches, which mechanically damp the oscillation, may be used when the application does not permit sufficient off time.

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17.3 RELAYS, REED

17.3.5.1 Basic contact forms. The basic contact forms used with reed and mercury wetted contacts are Form A, Form B, and Form C. The Form C contact symbol usually refers to a single-pole, double-throw contact with a break-before-make contact switching action. In certain types of mercury-wetted reed relays, Form D make-before-break contacts are available. It has become conventional to show these with the same symbol as the Form C contact, while in more typical relay nomenclature, a different symbol is indicated.

Reed switches are normally available with a single contact form enclosed in an individual glass envelope. The various types of basic switches have been described in paragraph 17.3.3 Physical construction. Multiple contacts are not available within a single glass switch.

17.3.5.2 Contact rating. The most commonly used contact materials for reed switches are gold, rhodium, and various proprietary alloys. These switches are also available with silver, tungsten, of a variety of precious metal contacts. Careful choice of the particular contact material can definitely give optimum life for a given load.

The contact ratings for reed and mercury-wetted contact relays are to be considered as definitely maximum values. Not only must the maximum current and voltage be considered, but the volt-ampere rating, the product of the maximum current and voltage at the time of switching, must also be considered. If a given relay is operated at lower levels than maximum contact rating, extended life can normally be expected. For example, the contact rating of a standard single-pole, double-throw switch used on relays is rated at 10 V-A at 0.5 A maximum, or 250 V maximum, resistive load. At this maximum rating, a typical life expectancy would be 25 million operations. At approximately half this rating, the life expectancy would be 50 million operations. At a relatively low contact level, the life expectancy could well exceed 100 million operations.

Contact life cannot be derated by the use of heavier contact switching. With non-reed relays, it is very common to have derated life with overload switching conditions. For example, a conventional relay with 5-A contacts may have a life expectancy of approximately 300,000 operations at 5 A. At a level of 7 or 8 A, the typical life expectancy might well be 100,000 operations. In the case of reed relays, the maximum contact ratings indicated should definitely be considered a maximum and should not be exceeded in either the steady state or transient switching condition.

In order to obtain maximum contact life, it is necessary that any transient voltages or currents be kept within the maximum specified. In any case, appropriate arc suppression will always yield longer contact switching life.

17.3.5.3 Contact resistance. The apparent contact resistance which appears at the terminals when measured at very low levels of current and voltage can deviate considerably from given values. This is an inherent characteristic of any metallic contact and must be considered when using low level circuits.

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The contact ratings of all types of reed relays should be taken to indicate the maximum value, at the time of switching of the current, voltage, and the volt-amperes. These values should not be exceeded in either the steady state or transient conditions. Therefore, in circuits where transient conditions can exist, the use of arc suppression or quenching is necessary. To obtain maximum life under any switching conditions, arc suppression is recommended.

Various types of arc suppression are commonly used, the more typical types being diodes, resistor-capacitor networks, zener diodes, nonlinear voltage sensing resistors which exhibit high impedance to low voltages and low impedance to high voltages, and certain special devices designed specifically for arc suppression. The choice of type of arc suppression, and the particular values to use, depend upon the specific application involved. These transient conditions should be considered in all cases, and appropriate arc suppression used when necessary. The manufacturer of a reed relay can commonly give rules of thumb for a specific product, but it is very difficult to generalize on the proper type of arc suppression which would cover most cases. Additional information is contained in paragraph 17.1.4.

17.3.6 Environmental considerations. Reed and mercury-wetted relays are normally ideal types of devices for operation in adverse environmental conditions, as the contact switching elements themselves are hermetically sealed. However, depending upon the environmental conditions, consideration must be given to the packaging to obtain the greatest reliability. For extremely adverse conditions, completely hermetically sealed relays, or similar constructions are recommended.

Special consideration must be given to mounting limitations on mercury-wetted relays as they are normally position and gravity sensitive.

There is a low operating temperature limitation on mercury wetted relays. The limitation is due to the freezing point of mercury (-38.8 °C), and it must be considered for all applications.

The life of a relay is ordinarily limited by mechanical wear and electrical erosion of the contacts. As the contacts are in a closely controlled atmosphere, any deterioration due to atmospheric conditions is not possible with reed switches. Therefore, the contacts themselves are able to withstand extremely adverse environmental conditions such as humidity, low pressure due to high altitudes, salt spray, and so on. The only limitations are due to the mechanical construction of the external parts of the reed leads, or the external relay structure. The external magnetic reed members are either plated, electro-tinned, or hot tin dipped to allow ease of solderability to relay terminals.

In the mercury-wetted contact switch, suitable for ground applications only, the film of mercury provides an almost ideal contact surface. Contact resistance is constant, regardless of low-level, nearly zero current and voltage,

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or full-capacity loads. The mercury film cannot be squeezed off the contacts, and any mercury lost from the contacts is quickly replaced by the built-in wick action. In the normal production process, a mercury-wetted contact relay is operated millions of times to insure uniformity and stability. Erosion of the metal contact surfaces by the mercury is negligible. There is little change in the character of the armature and contacts because of the limited motion. Life is rated in billions of error-free operations.

Two factors which must be considered in all relay packages are reed mounting and magnetic shielding. Reed switches should be mounted so that stress is not transmitted to the seal area. Stress can cause the seals to fracture in handling, resulting in early contact failure. Although shielding is not required to make the relay operate, it does improve power sensitivity and reduce the effects of magnetic interaction with adjacent relays.

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17.3.5.4 Operate and release time. In general, the smaller the relay the faster the operating time. For fastest possible operating speeds, the micro-miniature types are recommended. The standard-size switches are typically slower than the miniature or microminiature types.

Faster operating speeds can normally be obtained by the selection of a minimum ampere-turn pull-in reed and can be available by a sorting process. For the fastest possible operating speeds, the use of series resistance and higher applied voltages will yield the best conditions. For example, a standard miniature Form A relay will have an average operate time of approximately 1 millisecond. If a resistor, having resistance equal to the coil resistance, is put in series with a relay, and a voltage of twice the coil nominal voltage is applied to the complete circuit, a reduction in operate time can be obtained. The greater the series resistance and the higher the overdrive voltage, the faster the operate time. However, an absolute minimum operating time occurs on all types of relays regardless of driving conditions.

Dropout times are based on the relay being operated at nominal voltage at room temperature, and the circuit opened through use of a high impedance contact switch of some type. This time also is based on no other series or parallel elements being connected to the coil circuit. Because diodes and similar devices are commonly used across relay coils for arc suppression of preceding contacts, it must be remembered that the use of these elements can appreciably delay the dropout time of the relay.

17.3.5.5 Contact bounce. All types of solid metallic contacts normally exhibit some contact bounce. This is true of reed relays as well as conventional general purpose relays. This contact bounce is normally caused by impingement of the mating contacts. In general, the contact bounce of a normally open contact on operation of the relay is considerably less than the contact bounce of a normally closed contact on the release of the relay.

In the circuits where absolutely no contact bounce is permitted, the use of a mercury-wetted reed relay or mercury-wetted contact relay is required. These types of relays will switch contacts with absolutely no bounce if the unit is properly constructed mechanically and has the proper magnetic biasing. The contact interfaces, which are covered with a film of liquid mercury, absorb the shock of operation, and the contacts will not open electrically, but remain closed after initial closing.

Note that mercury wetted relays must be mounted as specified and are thus suitable for ground based equipment only, as they are sensitive to the effects of gravity.

17.3.5.6 Insulation resistance. In the manufacture of the basic reed switch, the parts are carefully cleaned and assembled under closely controlled conditions, and the unit is filled with an inert gas. Because of this, the insulation resistance of the basic reed switch is quite high. The insulation resis-

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tance across open contacts is typically $10^{11} \Omega$ or greater. During normal manufacture and handling, and when the reed switch is built into an assembly, various parallel insulation paths are established. Therefore, the insulation resistance of a reed relay, measured at the terminals, can be expected to be lower than that which is obtained on the basic reed switch. For many types of relays a typical insulation resistance of $10^9 \Omega$, or greater, can be expected. However, as this parameter is not normally tested on a 100 percent basis, it is recommended that when leakage can be a problem in a given circuit, the required insulation resistance should be specified. By the use of proper insulating materials and closely controlled manufacturing techniques, many of the standard types of relays can be obtained with insulation resistance as high as $10^{12} \Omega$. This figure would represent measurements made on a clean relay at normal room temperature and humidity. If, subsequently, the relay is subjected to dust or dirt, or to high humidity conditions, insulation resistance measurements lower than this would be expected.

17.3.5.7 Contact capacitance. As the reed leads present parallel metallic surfaces at the contacting end in their open position, a measurable capacitance across the contacts is obtained on any type of reed relay. In addition, measurable capacitance is also obtained from the coil to the reed leads. Depending upon the type of mechanical construction and size of the unit, capacitances from approximately 0.5 pF to 5.0 pF are obtained across open contacts. With a given relay construction, the capacitance between open contacts and the coil, with the relay not energized, is usually about twice the value of the open contact capacitance. With the contacts closed and the coil energized, the capacitance from contacts to coil is about four times the value of the open contact capacitance. These are general "rules of thumb" and if inter-capacitance coupling is important in a given application, it is recommended that the maximum value, and points of measurements, be specified.

17.3.5.8 Electrostatic shielding. In certain applications, it is necessary to reduce stray pickup of rf noise and similar undesirable signals from the contacts of a reed relay. This can be accomplished by a nonmagnetic metallic shield which surrounds the switch capsule and is fastened to an appropriate grounding pin or terminal. Normally this shield is placed between the reed switch and the coil. On multiple contact relays, it is also common to have each of the individual switch elements shielded and in turn have all of these shields shielded from the coil and have this shield grounded. The shield will prevent the noise generated by the arcing of the reed switch contacts from affecting other elements closely adjacent to the relay in the circuit.

17.3.5.9 Contact protection. The opening of relay contacts in an inductive load normally produces a high transient voltage. For example, it is not unusual for a 24-Vdc circuit to produce transients in excess of 1,000 V. Certain other loads, such as motor loads and lamp loads, will have inrush currents upon contact closure considerably higher than the steady state value. Inrush currents of 10 times steady state currents are common in lamp loads.

**17.4 RELAYS, TIME-DELAY
AND SENSOR**

17.4 Time-delay and sensor.

17.4.1 Introduction. There are several basic methods of controlling delay time with respect to a switching function. The common time-delay relays or timing devices are thermal, pneumatic, hydraulic, motor-driven, and electronic. Each of the devices mentioned has certain advantages and should be studied for suitability to each application.

The type of action performed by a time delay-relay can be generalized by two basic actions. In the time-delay on "make," the time-delay occurs starting at the instant the power is turned on and actuates contacts some specific time later. In this case, power is available for use in the electronic circuitry during the time-delay period.

For the time-delay on "break," turning the power on activates a set of contacts. The removal of power initiates the time delay, after which the contacts return to their normal position. In this case, the external power may not be available during the timing period. Thus, the energy required for activating the time-delay circuit must either be stored within the device during the power "on" period, or must be obtained from an auxiliary power source to which the time-delay is connected at all times.

Combinations of more than one time-delay circuit of either or both of the types may be made to produce special timing devices, either with sequential action or with self-regenerating timing cycles.

The most commonly used time delay is the time delay on "make," either used singly or in combination circuits to provide sequential timing systems.

Sensor relays are available which sense voltage levels, under or over voltage condition, frequency of an ac voltage, phase sequence, or a combination of these conditions. It may incorporate time delay on "make" or "break."

These devices are not included in MIL-STD-975.

17.4.1.1 Applicable military specifications.

MIL-R-28894	Relay, Hybrid or Solid State, Sensor, Established Reliability
MIL-R-83726	Relay, Hybrid and Solid State, Time Delay

17.4.2 Usual applications. The unique features of the sensor and time-delay relays have caused them to be widely accepted for many types of circuit switching. The specific application should determine the style and type of the time delay relay best suited for the particular design requirement. The following operating characteristics and design information can be helpful in determining the various parameters to meet specific design applications.

17.4 RELAYS, TIME-DELAY AND SENSOR

17.4.2.1 Description of time delay relay functions. The following figures illustrate several time-delay relay functions. Figure 11 shows time delay on operate condition. Figure 12 is time delay on release for units with separate control and power inputs, and Figure 13 is time delay on release for common control and power input. Figure 14 shows an interval timer. Figure 15 shows a repeat cycle timer.

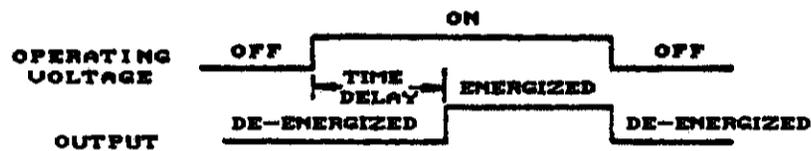


FIGURE 11. Time delay on operate.

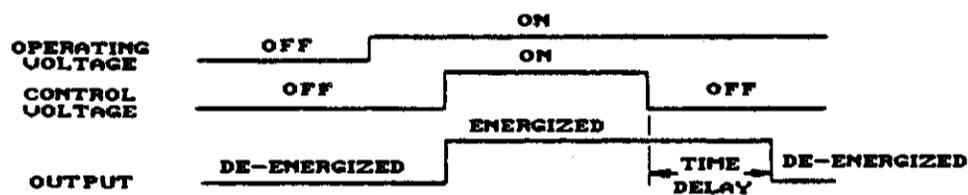


FIGURE 12. Separate control and power inputs.

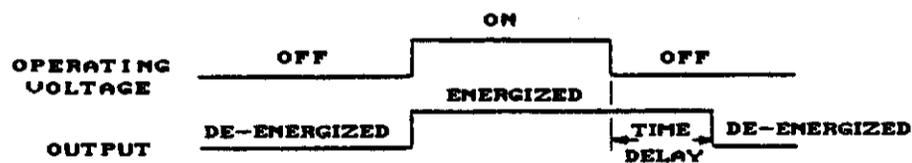
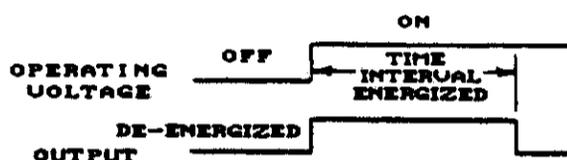
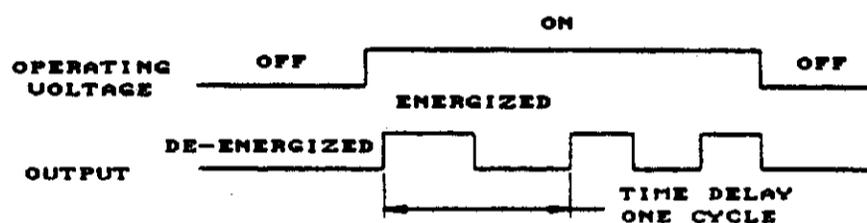


FIGURE 13. Common control and power input.

FIGURE 14. Interval time.FIGURE 15. Repeat cycle timer.

17.4.3 Physical construction and mechanical characteristics. The information pertaining to electromechanical output for armature relays in paragraph 17.2.3 is applicable to time-delay and sensor relays. Additional characteristics will be addressed in the following paragraphs.

17.4.3.1 Solid state (hybrid) time-delay relay. The most popular circuit used in solid state time-delay relay utilizes the RC, resistance and capacitance, charge principle and a programmable unijunction transistor. When a voltage is applied to the input the capacitor charges at a rate controlled by the RC network. At a voltage level controlled by the programmable unijunction transistor, the capacitor discharges through the relay coil and causes the relay to operate. A special input circuit is used which provides transient protection well over 1,000 V. The input is also protected against reverse polarity.

17.4.3.2 Copper slug time delay. The delay time is produced by placing a copper slug around the relay core. This produces a counter magnetomotive force (mmf) which produces the desired time delay on operate or release.

17.4 RELAYS, TIME-DELAY AND SENSOR

A time-delay can be produced on a dc relay by placing one or more shorted turns around the magnetic circuit, usually the core, to produce a counter mmf. This retards the buildup of the operating flux, and upon de-energization, provides mmf to retard the collapse of the flux. This shorted turn, or turns, is called a slug. Usually it consists of a copper collar on the core of the relay. In order to produce an effective delay, it is necessary to use a slug of considerable cross section, often equal to or larger than that of the coil itself. Although this method of time delay is applicable on any dc relay with sufficient space to accommodate this slug, it is most commonly used on telephone type relays and relays which have comparatively long coils and can easily accommodate this slug. A copper slug positioned at the armature end of the coil core retards magnetic build-up and provides long operate-delay with short-delay in release time. A copper slug positioned at the heel end of the coil core provides a long release time-delay with short delay in operate time.

The principle of operation of the slug is as follows: When the relay coil is energized, the flux build-up passes through the slug and by self-inductance produces an mmf that opposes the coil mmf. This opposing mmf delays the build-up of the magnetic field in the air gap to a strength that will cause the armature to close. The time delay on drop-out occurs in the opposite manner. When the coil is de-energized, the field starts to collapse, inducing a current in the slug. This in turn provides an mmf orientated to sustain the magnetic field and delay the drop-out. There are many factors that affect the time delay produced by slug relays. For this reason, each application must take into consideration coil power, armature gap, residual gap, amount of residual force required to move and close contacts, etc. to obtain required delay time.

17.4.3.3 Solid state-static output time delay relay. In this type of time-delay relay a solid state circuit performs the timing function, and an SCR or triac semiconductor performs the output switching. When voltage is applied to the input terminals, the timing delay is initiated. At the end of the electronically controlled period, a completely isolated solid state switch is energized. The switch leads are connected in series with the load and the supply voltage and are capable of handling ac current or dc current of either polarity. The device may switch the same voltage as applied to the control input or it may switch current to an isolated load from a secondary ac or dc current to a load from a dc power supply. Polarity need not be observed through the load switch; however, both the ac input voltage and the dc load current must be momentarily interrupted to reset the timer.

The solid state design of this time delay relay enables it to be used in specialized control applications where explosion-proof switching or operation in a corrosive atmosphere with arcless and bounceless time control is required.

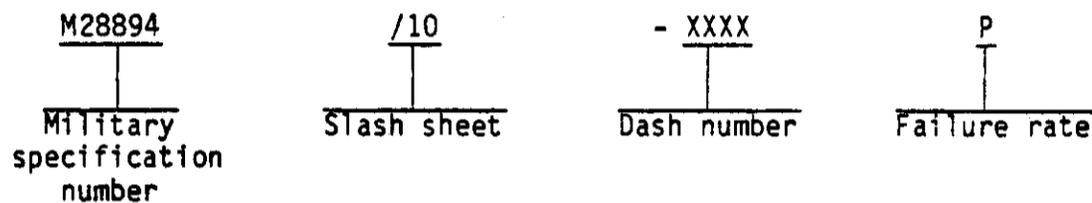
17.4.3.4 Sensor relays. Hybrid and solid state sensors are available with an under and over voltage, phase sensing, or frequency detection capability. In addition to these combinations, some sensors also incorporate a time delay on operate or release. The sensors are available with contact ratings ranging from low level to 10 A.

17.4 RELAYS, TIME-DELAY AND SENSOR

17.4.3.4.1 AC sensors. The sensors are capable of detecting an under voltage or over voltage condition, monitoring frequency, proper phase sequence, open neutral lines, or a combination of these features. Sensing accuracy is based on a true sinusoidal input waveform. Output status is provided by an electro-mechanic relay or solid-state logic.

17.4.3.4.2 DC sensors. Sensors are capable of detecting a specified dc voltage level and providing a status output when the prescribed requirements are met.

17.4.4 Military designation. MIL-R-28894 and MIL-R-83726 are the general military specifications for sensor and time-delay relays. The proper method of specifying the part number is:



17.4.5 Electrical characteristics. The basic considerations are covered in paragraph 17.1.4. The output characteristics of the time-delay and sensor relays are common to all electromechanical relays, and the ratings, configuration, and requirements are given in the referenced paragraph.

17.4.5.1 Reset. The reset time indicated for the various standard time delay relays is based on a typical time delay relay tested at the nominal voltage at room temperature. This number should not be construed as a maximum or minimum condition, but a typical value of reset time in milliseconds of a typical time delay relay.

17.4.5.2 Time-delay mode. The primary function of a time delay relay is to perform a given timing function, while satisfying the electrical and mechanical requirements of the application. The most popular timing modes are slow operate or slow release. However, time-delay relays are available in other timing modes, such as interval, totalizer, nontotalizer and momentary actuation. Functions are described in paragraph 17.4.2.

17.4.5.3 Time-delay tolerance. When choosing a time-delay relay, a major application consideration may be the repeatability tolerance of the relay. The standard repeatability tolerances are ± 1 percent, ± 5 percent and ± 10 percent. Other tolerances can be made available on special order. Recognition should be given to the fact that, in general, the broader the tolerance the lower the cost.

17.4.6 Environmental considerations. The typical environmental conditions are included in the referenced military specifications.

17.4.7 Reliability considerations. Paragraph 17.1.6 is concerned with general reliability data for relays and should be consulted for reliability considerations and failure mode data pertinent to time-delay relays.

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**17.4 RELAYS, TIME-DELAY
AND SENSORS**

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18.1 WIRE AND CABLE, GENERAL

18. WIRE AND CABLE

18.1 General.

18.1.1 Introduction. This section contains general information on electrical wire and cable listed in MIL-STD-975 NASA Standard Electrical, Electronic, and Electromechanical (EEE) Parts List. It is intended to help the designer select a wire or cable for a specific application. The referenced military specifications and MIL-STD-975 should be consulted.

In the following sections, specification requirements are discussed to assist the designer in narrowing the selection to a few choices. The most important decision is which of the numerous wire insulations, conductor configurations, and terminating hardware will be the most suitable for use in specific applications. Proper selection is the first step in building reliable equipment. To make the best wire and cable selection, the user must know as much as possible about the different wire types available. The designer should be aware of the advantages and disadvantages, the behavior of wire under various environmental conditions, the construction, the problems of termination, and the effects of chemical compounds that come into contact with the wire and its insulation. The designer should also be aware of the system requirements for temperature extremes, voltage/current requirements, and design constraints that could adversely affect wire and cable performance.

18.1.1.1 Applicable specifications.

MIL-W-22759	Wire, Electric, Fluoropolymer-Insulated, Copper or Copper Alloy
MIL-W-81381	Wire, Electric, Polyimide-Insulated, Copper or Copper Alloy
MIL-W-5086	Wire, Electric, Polyvinyl Chloride Insulated, Copper or Copper Alloy
MIL-W-16878	Wire, Electrical, Insulated, General Specification for
MIL-C-17	Cables, Radio Frequency, Flexible and Semirigid, General Specification for
MIL-C-27500	Cable, Electrical, Shielded and Unshielded, Aerospace

18.1.2 Definitions. The following terms in this section are to define words that might not be familiar to the designer. However, the intent is not to serve as a complete reference book of terms. If further information is needed, refer to the applicable military specification or consult an electrical engineering handbook.

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18.1 WIRE AND CABLE, GENERAL

Abrasion resistance. Ability to resist surface wear.

Aging. The change in properties of a material with time under specific conditions.

American wire gauge (AWG). The standard system used for designating wire diameter. Formerly referred to as the Brown and Sharpe (B&S) wire gauge. For solid wire, this gauge is characterized by a doubling in wire diameter for a six-step reduction in AWG number. Stranded wire, because of geometrical considerations of the strandings, is usually not equal in cross-section to a solid wire, but is designated by the closest AWG number.

Anneal. To heat and then gradually cool in order to relieve mechanical stresses.

Blocking. The property (usually considered undesirable) of the insulation of wire wound in closely spaced turns adhering to each other at elevated temperatures. All military specifications for wire require that a blocking test be performed to show that this does not occur at the specified temperature.

Bond strength. The amount of adhesion between bonded surfaces.

Bonding. The adhesion of insulation material to a conductor.

Breakdown (puncture). A disruptive electrical discharge through insulation.

Breakdown voltage. The voltage at which the insulation between two conductors will break down.

Cable. Two or more insulated conductors, solid or stranded contained in a common covering (sheath, shield, or jacket); two or more insulated conductors twisted or molded together without common covering; one or more insulated conductors with a metallic covering shield or outer conductor (insulated or uninsulated).

Characteristic impedance. A characteristic of coaxial cable which is a function of the two diameters of the conductors and the dielectric constant of the insulation material. Although called an impedance, it is actually a resistance that is independent of frequency and represents the ac resistance at rf frequencies of an infinitely long cable. Values of nominal characteristic impedance used in space programs are standardized at 50 ohms, 75 ohms, and 95 ohms.

Coaxial cable. A cable configuration consisting of a conductor in a cylindrical geometry with the center conductor insulated from the concentric outer conductor, with these dimensions accurately controlled. In general, these cables are designed for low loss, stable operation from relatively low frequencies to the high frequencies encountered in the microwave region of the frequency spectrum. Cables may also be used as circuit elements such as delay lines and for impedance matching devices. Coaxial cables are normally classified by their characteristic impedance.

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18.1 WIRE AND CABLE, GENERAL

Cold flow. The property (usually undesirable) of a solid insulation material (such as TFE), to flow or slowly displace away from a highly localized stress region caused by a point, corner, or even tight lacing. In severe cases, cold flow of the insulation will expose the wire conductor causing contact with the stress-producing object.

Conductivity. Reciprocal of volume resistivity. Conductance of a unit cube of a material.

Conductor. An electrical path which offers comparatively little resistance. A wire or combination of wires not insulated from one another, suitable for carrying a single electric current.

Conformal coating. A process of coating which generally follows the contours of the assembly, providing resistance to mechanical shock and environmental conditions.

Continuous lengths. Wire is normally manufactured in a continuous process. If, during the spark test, for example, an insulation failure is detected, that section of wire is cut out, leaving a gap in the reel. Military specifications require documentation listing the lengths of wire in the order in which they will be unspooled, so that installation personnel will be certain that the wire they are working with will be long enough for the immediate application.

Copper alloy, high strength. In applications where weight saving is critical and increased conductor resistance is not a factor, high strength copper alloy conductors are recommended. AWG 26 high strength copper alloy has approximately the same breaking strength as AWG 24 copper with two thirds the weight. The increase in resistance is approximately 4.5 times.

Copper conductor. Most electrical wire consists of a copper conductor of high purity and in an annealed condition which permits easy bending, and offers low electrical resistance. In wire sizes 24 AWG and smaller, the softness and small size cause the copper conductors to become increasingly more fragile with normal handling.

Corona. A luminous discharge due to ionization of the gas surrounding a conductor around which exists a voltage gradient exceeding a certain critical value.

Corona resistance. The time that insulation will withstand a specified level field-intensified ionization that does not result in immediate complete breakdown of the insulation.

Critical pressure region. High altitude region where the dielectric strength of air becomes less than 20 percent of the sea level dielectric strength of 70 volts/mil. This is approximately 65,000 ft (50 torr) to 310,000 ft (5×10^{-4} torr). Above this altitude, into "hard" vacuum, the dielectric strength rapidly increases. The minimum, at which the dielectric strength of air becomes essentially zero, is approximately 110,000 ft. This is also called the "Paschen" minimum. Because of the many uncontrolled factors involved, these limits are emphasized as being "approximate."

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18.1 WIRE AND CABLE, GENERAL

Cut-through. Resistance of solid material to penetration by an object under conditions of pressure, temperature, etc.

Delamination. The separation of layers in a laminate through failure of the adhesive.

Dielectric. Any insulation material between two conductors. This material is characterized by a high ohmic resistance in comparison to the conductor resistance.

Dielectric constant. The ratio of the capacitance of a capacitor with the given dielectric to the capacitance of a capacitor having air for its dielectric but otherwise identical. Also called permittivity, or specific inductive capacity.

Dielectric strength. The maximum voltage which an insulating material can withstand before breakdown occurs. See Voltage Gradient.

Dissipation factor. The tangent of the loss angle of the insulating material.

Elongation. Stretching characteristic of ductile flexible or elastomeric materials while under tension. If the yield point is exceeded (which is very low for annealed copper wire), the wire will not return to its original length when the load is removed. All military specifications define elongation rather than breaking stress as a requirement for annealed copper wire.

Embedment. A process by which circuit components are encased in a dielectric material, usually poured in as a liquid, then hardened. This includes both potting and encapsulation.

Encapsulation. An embedment process in which the resin is cast in removable molds.

ETFE. Common trade name "Tefzel".

FEP. Common trade name "Teflon".

Flammability test. A measure of the ability of insulation material to support combustion under specified test conditions.

Fluoropolymer. Plastic materials having fluorine and other elements, usually carbon. Some examples are polytetrafluoroethylene (PTFE), fluorinated ethylene propylene (FEP), and modified ethylene tetrafluoroethylene (ETFE) used as wire insulation materials in MIL-W-22759. Although the designation for the first is properly called out as "PTFE," to minimize confusion throughout this document, the shorter acronym "TFE" will be used to correspond with existing military specifications. These materials are characterized by toughness and flexibility at low temperatures (even down to absolute zero), lower dielectric constants than any other solid materials, and operation for years near their melting temperature of 327 °C without significant change in properties. TFE in the presence

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of oxygen has a rather low radiation resistance under high energy radiation ($10^7 - 10^8$ rads dose) becoming embrittled, and disintegrates, emitting gaseous products. Trade name "Teflon" for TFE.

Ground support equipment (GSE). Equipment not intended for flight, and designed for ground installation to support flight projects.

Harness. One or more insulated wires or cables, with or without helical twist; with or without common covering, jacket or braid; with or without breakouts; assembled with two or more electrical termination devices and so arranged that as a unit, can be assembled and handled as one assembly.

Impedance. The total opposition that a circuit offers to the flow of alternating current or any other varying current at a particular frequency. It is a combination of resistance R and reactance X measured in ohms and designated by Z.

Impulse. A voltage or current surge of unidirectional polarity.

Impulse dielectric test. A method of checking insulation integrity of wire or shielding by passing the wire through a bead chain on which high negative impulses are continuously applied. The wire or shielding is grounded during the test, and a circuit is used to detect voltage breakdown.

Insulation. Material having a high resistance to the flow of electric current, to prevent leakage of current from a conductor.

Insulation resistance. The ratio of the applied voltage to the total current between two electrodes in contact with a specific insulator. A measure of the insulation integrity or quality of the material in contact with conductors.

Insulation system. All of the insulation materials used to insulate a particular electrical or electronic product.

Jacket. An insulating material applied over the primary insulation of wire (MIL-W-5086) or over shielding (MIL-C-27500) for protective purposes.

Lay. The axial distance required for one conductor strand to be wound helically around a central strand.

Lay, concentric. A central strand surrounded by one or more layers of helically wound strands. It is optional for the direction of lay of the successive layers to be alternately reversed (true concentric lay) or in the same direction (unidirectional lay).

Lay, length of. Defined in MIL-W-81381 as not less than 8 nor more than 16 times the maximum conductor diameter as specified.

PFA. Perfluoroalkoxytetrafluoroethylene, material used for jacketing of cables.

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Polyimide. Polyimides are a family of high-temperature resistant thermoset and thermoplastic resins. Polyimide insulations are applied to wire listed in MIL-STD-975 tape form, 0.1 mil FEP, 1.0 mil polyimide, and 0.1 FEP, wound with 50 percent minimum overlap, then sintered or heated to fuse the layer together. Trade name "Kapton."

Potting. An embedment process in which the container (can) used in the embedment remains as part of the completed assembly.

Primary insulation. Insulation applied to conductors which determines the wire voltage rating. Other insulation materials, such as jackets, are not included.

Red plague. A deterioration problem encountered early in the space program through the use of silver-coated copper wire; however, controlled wire manufacturing processes will reduce the incidence of this problem. Red plague results from a combination of several factors. In wire manufacture, the insulation material is usually applied by a hot extrusion process over the wire in a continuous operation. Originally, this extruded wire was cooled by passing it through a water bath. In the cooling process, an occasional pin hole in the insulation sucked in the water during the cooling of the internal gases. Even though this pin hole would be later detected by suitable high voltage sparking techniques and the subsequent section of wire cut out, the water which had entered through the hole would travel a great distance up and down the wire strand. In the case of a silver coating on the wire, the coating is not continuous but is porous, exposing the copper underneath. The presence of moisture forms a battery by galvanic action producing cuprous oxide corrosion red in color and thus designated "red plague." Tin-coated and nickel-coated copper wire do not exhibit this reaction because they are much closer to copper than is silver in the galvanic series. Replacing the water quench with an air blast or other dry cooling during the wire insulating manufacturing process normally eliminates this problem. Also, silver-coated copper wire must be manufactured in a low-humidity environment.

Resistance. Property of a conductor that determines the current produced by a given difference of potential. The ohm is the practical unit of resistance.

Resistivity. The ability of a material to resist passage of electrical current either through its bulk or on a surface. The unit of volume resistivity is the ohm-cm; of surface resistivity, the ohm.

Surface resistivity. The resistance of a material between two opposite sides of a unit square of its surface. Surface resistivity may vary widely with the conditions of measurement and presence of surface contamination.

Surge. A transient variation in the current or voltage at a point in the circuit.

Tensile strength. The breaking stress of a material under tension, usually expressed in psi. For purposes of comparison with high strength copper alloy, which is 52,000 psi, annealed copper is assumed to be 36,000 psi.

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TFE. Common trade name "Teflon". See Fluoropolymer.

Thermoplastic. A classification of resin that can be readily softened and resoftened by repeated heating.

Thermosetting. A classification of resin which cures by chemical reaction when heated and, when cured, cannot be resoftened by heating.

Time delay. 1. The time required for a signal to travel between two points in a circuit. 2. The time required for a wave to travel between two points in space.

Velocity of propagation. The ratio of the speed of the flow of an electric wave in a coaxial cable to the speed of light, expressed as a percentage. The speed of the electric wave is governed solely by the properties of the dielectric medium and the permeability of the conductor through which it is transmitted. In free space, electromagnetic energy travels at the speed of light, 3×10^8 meters per second which is by definition 100 percent. In a coaxial cable with a uniform dielectric of a relative permeability greater than 1, the VP is less than 100 percent, in the same ratio.

The equation is $VP\% = \frac{100}{\sqrt{\epsilon}}$, where ϵ dielectric constant of the insulation.

For solid, extruded PTFE, $\epsilon = 2.07$

An equivalent equation is $VP\% = \frac{101670}{Z_0 C}$, where Z_0 = characteristic impedance
 C = capacitance in pf/ft.

Voltage gradient. A measure of the maximum stress placed on an insulating material between two conductors, usually expressed in volts/mil. For parallel plates the voltage gradient in volts/mil is the voltage applied to the plates divided by the separation in mils. For other geometries such as parallel or crossed cylinders (wires) or cylinder to plane, etc, the curved surfaces magnify the gradients over the parallel plate calculation, and different equations are required to determine the gradients.

Voltage standing wave ratio (VSWR). The ratio of the voltage maximum to the voltage minimum of the standing waves occurring in an r-f or coaxial cable not terminated in its characteristic impedance when connected to an r-f source.

Volume resistivity (specific insulation resistance). The electrical resistance between opposite faces of a 1-cm cube of insulating material, commonly expressed in ohm-centimeters. The recommended test is ASTM D 257-61.

Wicking. 1. Upward flow of solder underneath the insulation during soldering of stranded wire, stiffening it. 2. Flow of moisture or other contaminant underneath the insulation of wire.

Wire. A single metallic conductor of solid, stranded, or tinsel construction, designed to carry current in an electrical circuit. It may be bare or insulated, but does not have a metallic covering, sheath, or shield.

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18.1.3 General wire characteristics. The wire and cable section of MIL-STD-975 lists control specifications for standard wire and standard cable. These control specifications, with their specified slash sheets, give a wide range of characteristics and parameters for selection for a given application.

In addition to providing electrical continuity adequate for the functioning of the circuitry and interconnection of subsystems and maintaining adequate insulation between the various conductors to prevent shorting, many other factors must be considered which influence the selection of wire. Some of these factors are:

- a. Limitations on weight, size, and cost
- b. Rated voltage of the wire
- c. Ease of stripping, termination, and replacement
- d. Flexibility and bondability
- e. Environmental considerations
- f. Space versus ground support applications
- g. Resistance to solder and to damage from soldering irons
- h. Resistance to solvents
- i. Resistance to cut through, cold flow, and abrasion.

A first consideration in wire application is that the wire should be used within its design limitations and in the class of service for which it is intended. Because multiconductor cables, other than coaxial, are made up of individual wires, the limitations of the wire type should be considered. Special design considerations for electrical requirements are as follows:

- a. The current, voltage, and frequency requirements
- b. The number and type of conductors
- c. Shielded, rf, coaxial
- d. Type and physical characteristics
- e. Noise generation and emi considerations.

Depending on whether the application is for space or ground support, the following special environmental parameters are considered:

- a. The outgassing characteristics of the insulation material
- b. The expected temperature rise characteristics

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- c. Vacuum exposure effects
- d. Expected launch pad environments.

All of the considerations and constraints are discussed in separate paragraphs for guidance of the designer in comparing the various parameters of available wires.

18.1.4 Materials.

18.1.4.1 Conductor materials and plating. With the exception of the solid center conductors in the coaxial cables listed in MIL-STD-975 all of the wire conductors are stranded. Stranded is preferred over solid wire because it is easier to bend and is considered more reliable in vibration environments. The number of strands used in the conductor is limited to those which will give a perfect lay with all the strands the same diameter.

The conventional conductor material for electrical wire has been soft, annealed copper, as pure as possible, to achieve the lowest resistance. In applications where the weight is an overriding factor with resistance secondary, a high strength copper alloy is available. As an example, AWG 26 copper alloy has approximately the same breaking strength as pure AWG 24 copper but only two thirds the weight due to the smaller diameter. Also, the number of flexures that pure copper wire can withstand without work hardening or breaking is severely limited, compared to copper alloy, especially in the smaller sizes (26 to 32).

Because of the ductility of annealed copper, the military specifications have no tensile strength requirements. Rather, the minimum elongation is defined as that distance the wire is stretched before breakage of the first strand occurs. Elongation is also specified for copper alloy wire, and in addition, a tensile breaking stress is given. To enable comparison by designers, tensile strengths of the various AWG wire sizes are calculated by the product of the copper cross-sectional area and 36,000 psi, the accepted breaking strength of pure copper.

All copper and copper alloy wire conductors used in MIL-STD-975 are required to be plated or coated to inhibit corrosion of the conductor. Three plating materials are called out, each having a specific application which is a function of temperature and method of termination: tin, silver and nickel. Tin, the most common and least expensive coating used in electronic wiring, has an upper temperature limitation of 150 °C. Although intended for solder termination, tinned copper wire can also be used for crimping.

Silver- and nickel-coated wires have upper temperature limits of 200 °C and 260 °C, respectively. Silver-coated wire permits terminating both by soldering and crimping. However, silver coated wire is susceptible to cuprous oxide corrosion (red plague) when produced, stored, or used in a moist or high-humidity environment. Therefore, the environment must be controlled. The higher temperature ratings for silver- and nickel-coated wire ordinarily approach that of the insulation material itself, which is approximately 260 °C.

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Of the three materials, nickel is able to withstand corrosion much better than the other two and still maintain a good electrical connection. Because of the difficulty of soldering nickel, this wire is intended mainly for crimp-type terminations. Nickel wire can be soldered, provided that high temperature soldering equipment and active fluxes are used. Highly corrosive fluxes should be avoided.

Table I presents typical values for weight tensile strength, and resistance at common wire sizes.

TABLE I. Parameters for standard wires

Wire Size	Stranding (Number of Strands x AWG size of Strand)	Weight, Max ^{2/} lbs/1000 feet		Tensile Strength lbs ^{2/}			Resistance, Nominal Ohms/1000 feet ^{2/}		
		Bare Solid Copper Wire	Finished Wire	Copper		Copper Alloy	Solid Copper Wire	Nickel Plated Copper ^{1/}	Nickel Plated Alloy ^{1/}
				Solid	Stranded	Stranded			
30	7 x 38	0.303	1.00	3.1	3.2	5.17	104	114	129
28	7 x 36	0.481	1.36	5.0	4.9	8.16	65.3	68.6	79.0
26	19 x 38	0.765	1.90	7.9	8.6	14.2	41.0	41.3	49.4
24	19 x 36	1.22	2.58	12.7	13.4	22.4	25.7	26.2	30.1
22	19 x 34	1.94	3.68	19	21	36	16.2	16.0	18.6
20	19 x 32	3.10	5.36	32	34	58	10.1	9.77	11.4
18	19 x 30	4.92	7.89	49	54		6.39	6.10	
16	19 x 29	7.81	9.95	78	69		4.02	4.76	
14	19 x 27	12.4	14.9	124	109		2.52	3.00	
12	37 x 28	19.8	22.6	197	172		1.59	1.98	
10	37 x 26	31.4	35.1	314	265		1.00	1.24	
8	133 x 29	50.0	63.5	480	480		0.628	0.694	
6	133 x 27	79.4	99.9	763	759		0.395	0.436	
4	133 x 25	126	157	1213	1206		0.249	0.275	
2	665 x 30	201	245	1929	1882		0.156	0.177	
1	817 x 30	253	324	2423	2310		0.124	0.144	
0	1045 x 30	319	391	2984	2960		0.098	0.133	
00	1330 x 30	402	504	3763	3760		0.078	0.089	

^{1/} Silver and tin plated wires are not shown in this table.

^{2/} These parameters are typical. Consult the appropriate specifications for specific values.

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18.1.4.2 Insulation materials.

18.1.4.2.1 Types. Various types of insulation are used in the wire and coaxial cable callouts for MIL-STD-975. These include:

- a. Extruded tetrafluoroethylene. This insulation is applied over the conductor by means of a continuous extrusion process.
- b. Polyimide. Three layers of insulation fabricated over the conductor. The first layer is a fluorocarbon/polyimide composite tape consisting of 0.1 mil of fluorinated ethylene propylene (FEP) fluorocarbon resin, then 1 mil of polyimide film (trade name "Kapton") followed by another layer of 0.1 mil FEP. This is wound with a 50 percent (min) overlap over the conductor. The second layer is an identical film wound in a "cross lay" or opposite direction over the first layer. Finally, a modified aromatic polyimide resin coating 0.5 mil (min) is applied over the previously wound tapes. They are fused by a sintering process to become an integral insulation.
- c. An insulation system using polyvinyl chloride (PVC) as a primary insulation. In MIL-W-5086 wire, an additional nylon jacket is extruded over the PVC. This jacket serves to protect the PVC insulation from mechanical damage and adds a "slipperiness" to allow this wire to be pulled through conduits and other applications without the insulation binding. MIL-W-16878 wire has the PVC insulation without the jacket. Both of these PVC insulated wires are for GSE use only.
- d. Coaxial cable insulations are Type F-1 extruded polytetrafluoroethylene (PTFE, or by earlier designation TFE), which are preferred for the higher frequencies and configurations indicated in the specific slash sheets.

18.1.4.2.2 Color. The color designation codes for individual wire permitted by MIL-STD-975 are listed in Table II, which is taken from MIL-STD-681. Additional information can also be found in the referenced military specification.

18.1.4.2.3 Flammability. Wires with polyvinyl chloride insulation shall not be used in aerospace applications. Application in space transportation system (STS) payloads may require that the specific STS flammability hazards be addressed. Users are advised to consult the appropriate project systems safety officer.

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TABLE II. Color sequence for single wires

Base Color	Stripe or Band	Color Designation No.
Black		0
Brown		1
Red		2
Orange		3
Yellow		4
Green		5
Blue		6
Violet		7
Gray		8
White		9
White	Black	90
White	Brown	91
White	Red	92
White	Orange	93
White	Yellow	94
White	Green	95
White	Blue	96
White	Violet	97
White	Gray	98

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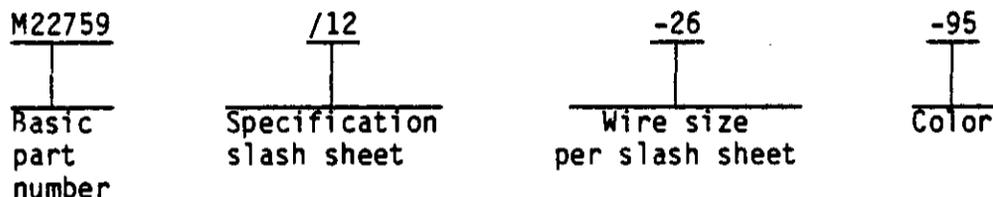
18.2 WIRE AND CABLE, FLUOROPOLYMER-INSULATED, MIL-W-22759

18.2 Wire, fluoropolymer-insulated, MIL-W-22759.

18.2.1 Introduction. This specification covers fluoropolymer-insulated single conductor electric wires made with tin-coated, silver-coated, or nickel-coated conductors of copper or copper alloy ranging from AWG 28 to AWG 00, depending on the particular slash sheet. The fluoropolymer insulation of these wires may be polytetrafluoroethylene (TFE) or ethylene-tetrafluoroethylene copolymer (ETFE).

18.2.2 Usual applications. Wires fabricated to this basic specification and applicable slash sheets may be used in ground support, aircraft, and space applications in accordance with their performance characteristics. Specifically, this includes cable harnesses and interconnection (hookup) wire for chassis and electronic subsystems for long-term high-reliability applications. Note that for space applications, the outgassing properties of wire fabricated to this specification are not controlled and must be evaluated for compliance with program requirements. Because of TFE cold-flow characteristics, care must be exercised in the design to assure that cable lacings or wire routing will not cut or bring pressure to bear on the insulation.

18.2.3 Part number designation. The complete part number designation is as follows.



As an example, an M22759/12 wire, 260 AWG size, white with green stripe would have a complete number, M22759/12-26-95.

18.2.4 Physical construction.

18.2.4.1 Conductor configuration. The conductors of wire sizes 28 through 10 AWG consist of a central strand surrounded by one or more concentric layers of helically wound strands as shown in Figure 1. Conductors of sizes 8 through 00 AWG are in a rope lay configuration concentrically wound with a central member surrounded by one or more layers of helically wound members as shown in Figure 2.

18.2.4.2 Conductor materials. Two conductor materials are available, pure copper and high-strength copper alloy. The pure copper conductors are used for larger sizes of wire and for applications where the wire resistance is an important consideration. The advantage of high-strength copper alloy is that it allows a smaller diameter with the same breaking strength for applications where weight is critical and where the increase in conductor resistance

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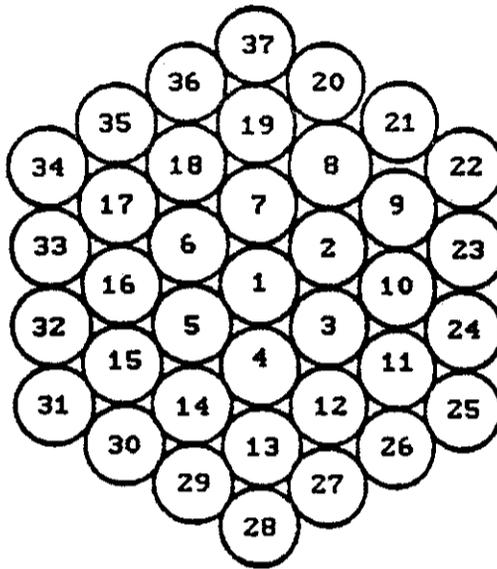


FIGURE 1. Stranded wire configuration for 7, 19, and 37 strands, for respective sizes 28 through 10.

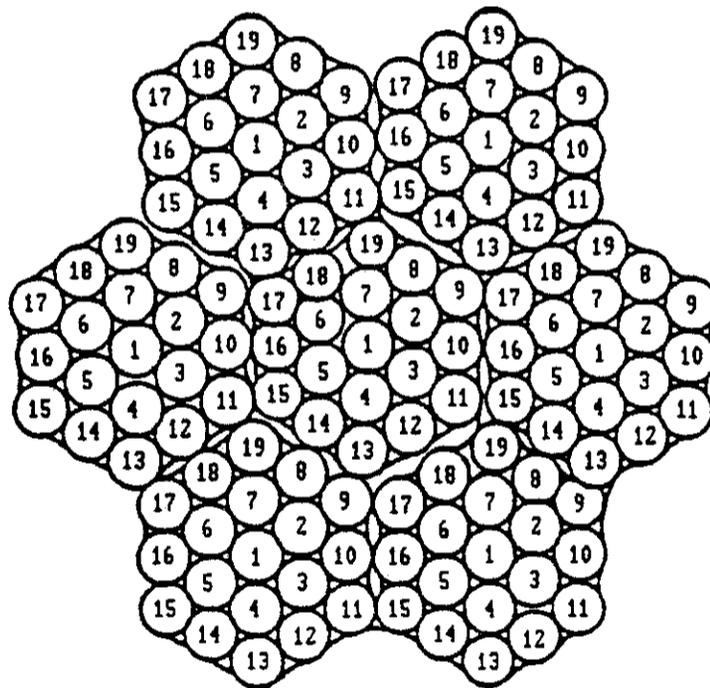


FIGURE 2. Typical rope-lay configuration, wire sizes 8 through 4 (sizes 2 through 0000 use additional bundles).

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**18.2 WIRE AND CABLE, FLUOROPOLYMER-
INSULATED, MIL-W-22759**

is not significant to the operation of the electronic equipment. Specific applications are signal cable harnesses. As an example, size 26 AWG high-strength copper alloy wire has approximately the same breaking strength as size 24 AWG pure copper wire. The "breaking strength" of pure copper is not given in military specifications because of its extreme ductility and the difficulty of determining when a break occurs. To assist the designer in selection, Table IV compares the tensile strengths of solid copper, stranded copper, and high-strength copper alloy of the same size. These values are calculated from the cross-sectional areas, assuming a pure copper ultimate tensile strength of 36,000 psi and copper alloy of 55,000 psi.

18.2.4.3 Conductor coatings/platings. Three conductor coatings are available in the specification slash sheets listed. Nickel-coated wire is intended mainly for crimp type termination because of the difficulty in soldering nickel. Nickel-coated wire can only be soldered when high temperature equipment and active fluxes are used. Silver-coated or tin-coated wire can be terminated by either crimping or soldering.

Active fluxes are not recommended for use. Corrosive fluxes should be avoided.

18.2.4.4 Insulation configurations. With the exception of M22759/3, the insulation materials of TFE or ETFE are applied over the conductor by an extrusion process. As shown in Figure 3, the first layer against the conductor consists of either TFE-skived tapes or TFE-coated glass tapes, or a combination of both, spirally wrapped on the conductor. The intermediate coat consists of TFE-coated glass braid and TFE finisher followed by a TFE tape jacket consisting of two layers cross-lapped. The layers are fused or sintered to eliminate the possibility of the jackets unwrapping. Extruded TFE over the conductor is more common and should have fewer voids than wound layers. Elimination of voids becomes important in applications such as high voltage to prevent corona.

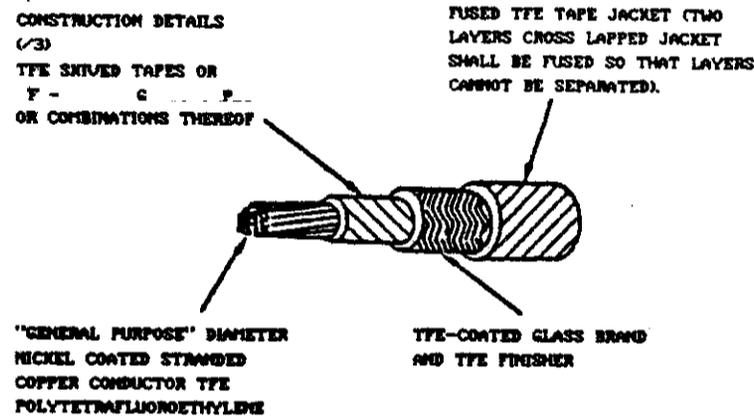
18.2.5 Electrical characteristics.

18.2.5.1 Conductor resistance. The electrical resistance of wire is specified as ohms per 1,000 feet measured at 20 °C. Resistance values for uncoated solid copper wire and coated stranded wire are compared in Table I. The resistance value as given is essentially that of pure copper, since the coating has a minor effect on the overall resistance. Silver-coated wire has slightly lower resistance than the tin-coated or nickel-coated wires. Specifications M22759/22 and /23 have silver-coated high-strength copper alloy and nickel-coated high-strength copper alloy conductors, respectively, and include only five wire sizes, from 28 to 20 AWG.

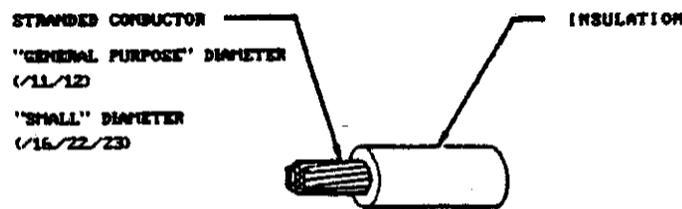
18.2.5.2 Working voltage. Although the working voltage of all the wires in this section is 600 V rms at sea level at the highest temperature rating of the insulation, it is customary to apply derating factors to all maximum ratings as a safety factor in high reliability applications (refer to Appendix A of MIL-STD-975). Even though the wire is rated at 600 V, it is required to pass an impulse dielectric test of from 6.5 to 8 KV at room temperature.

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18.2 WIRE AND CABLE, FLUOROPOLYMER-INSULATED, MIL-W-22759



A. TFE-glass-TFE



B. Extruded TFE

FIGURE 3. Typical construction, MIL-W-22759 wire.

18.2.5.3 Insulation resistance. The insulation resistance is a measure of the insulation integrity of the wire conductor coating. Depending on the wire specification, the range of insulation resistance values for 1,000 ft is from 5,000 MΩ to 50,000 MΩ minimum. Usually, the insulation resistance is not of significance in electronic circuits unless the wire is very long and is used in very high impedance applications where the insulation shunting effect might affect the signal voltage. The insulation resistance goes up as the lengths become shorter.

18.2.5.4 Current rating. The current ratings of wires are not usually considered unless significant current is involved which will cause heating of the wire due to the I^2R loss. Such a temperature rise must be added to the maximum environmental temperature to determine the actual temperature of the wire and, in any case, must not exceed the temperature rating of the wire itself. Currents flowing in wires in a bundle have a cumulative heating effect that must be taken into account. Refer to derating in Appendix A of MIL-STD-975.

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**18.2 WIRE AND CABLE, FLUOROPOLYMER-
INSULATED, MIL-W-22759**

18.2.6 Environmental considerations. In selecting a wire for specific applications, all the environments which the wire will encounter must be taken into consideration. For example, humidity is not a problem for a spacecraft in space, but might be at a launch center if the space vehicle were there for any length of time.

The outgassing properties of these wires are not controlled and must be evaluated for compliance to project outgassing requirements.

Silver-coated wire is susceptible to cuprous oxide corrosion (red plague) when stored or used in a moist or high-humidity environment.

18.2.7 Reliability considerations.

18.2.7.1 TFE, ETFE, and FEP insulating material. Since the conductor configurations are essentially the same for all military specification wires, considerations of reliability will be applied to the insulation materials. These insulation materials are chemically inert and have reasonable dielectric strength, resistance to high temperature, and flexibility at low temperatures. The chemical inertness property of these materials introduces a problem in applications where bonding to encapsulation materials is required such as encapsulation of the wiring ends of connectors. To achieve a reliable bond, an etchant material must be used. Another problem with TFE and FEP is cold flow under pressure or cut-through strength of the material. A bundle of TFE- or FEP-insulated wires held together with a very tight lacing (cord) could be cut through the insulation material down to the conductors. A similar problem is created where the wire passes over a corner or edge. Excessive stress applied to the wire may cause the corner to gradually penetrate the insulation all the way to the conductor.

18.2.7.2 Voltage. All of the wires in this section have a voltage rating of 600 Vrms at the maximum operating temperature at sea level. The intent of this rating is to establish a reference point for which this wire insulation is capable of withstanding indefinitely with the outer surface of the insulation at ground potential or zero voltage.

Other waveforms or higher frequencies may result in a reduction of the voltage rating. If the wires are encapsulated or otherwise separated from the ground plane, then the stress on the wire insulation itself is reduced with the same voltage applied. This rating does not apply to wire operating in the critical pressure region where corona discharge may take place from the surface of the wire to the ground plane. To prevent corona, the voltage rating should be 190 Vrms at 60 Hz, which gives a peak voltage of 270 V, the limiting value for breakdown in air at any pressure. If operation of the wire in the critical air pressure region at higher ac voltages up to 600 Vrms is desired, then additional steps are required to eliminate corona or dielectric breakdown.

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**18.2 WIRE AND CABLE, FLUOROPOLYMER-
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Because of the effects of all these conditions, it is mandatory for the designer to apply derating factors to all maximum ratings in high reliability applications to take into account unanticipated problems. The derating factors are presented in Appendix A of MIL-STD-975.

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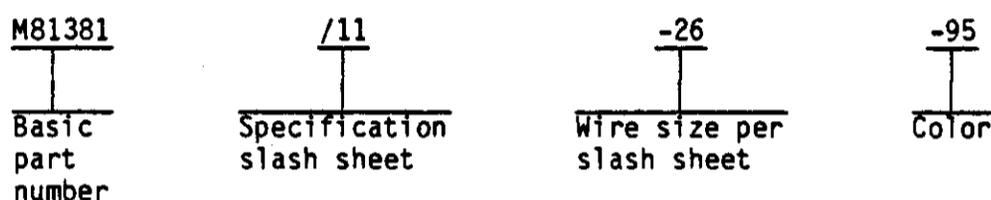
18.3 WIRE AND CABLE, FLUOROCARBON/
POLYIMIDE-INSULATED, MIL-W-81381

18.3 Wire, fluorocarbon/polyimide-insulated, MIL-W-81381.

18.3.1 Introduction. This specification covers light weight fluorocarbon/polyimide-insulated single-conductor electric wires made with tin-coated, silver-coated, or nickel-coated conductors of copper or high-strength copper alloy. The wire sizes range from 30 to 10 AWG depending on the slash sheet.

18.3.2 Usual applications. Wires fabricated to the basic specification and applicable slash sheets may be used in ground support, aircraft, and space applications in accordance with their performance characteristics. Specifically, this includes cable harnesses and interconnection (hookup) wire for chassis and electronic subsystems for long-term high-reliability applications. Note that for space applications, the outgassing properties of wire fabricated to this specification are not controlled and must be evaluated for compliance with program outgassing requirements. Generally, this wire should not come into contact with liquid missile propellants.

18.3.3 Part number designation. The complete part number designation is as follows.



As an example, an M81381/11 wire, 26 AWG size, white with green stripe would have a complete number of M81381/11-26-95.

18.3.4 Physical construction.

18.3.4.1 Conductor configuration. The conductors of sizes 30 through 10 AWG consist of a central strand surrounded by one or more concentric layers of helically wound strands, as shown in Figure 4.

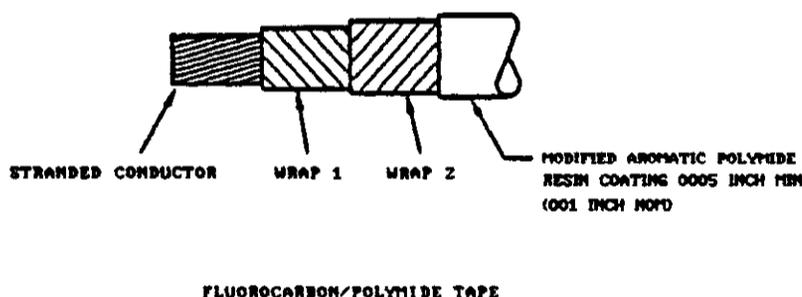


FIGURE 4. Typical construction, MIL-W-81381 wire.

MIL-HDBK-978-B (NASA)

**18.3 WIRE AND CABLE, FLUOROCARBON/
POLYIMIDE-INSULATED, MIL-W-81381**

18.3.4.2 Conductor materials. Two conductor materials are available, pure copper and high-strength copper alloy. Pure copper conductors are used for larger sizes of wire and for applications where the wire resistance is an unimportant consideration. The advantage of high-strength copper alloy is that it allows a smaller diameter with the same breaking strength for applications where weight is critical and the increase in conductor resistance is not significant to the operation of the electronic equipment. Specific applications are signal cable harnesses. As an example, size 26 AWG high-strength alloy copper wire has approximately the same breaking strength as size 24 AWG pure copper wire. The breaking strength of pure copper is not given in military specifications because of its extreme ductility and the difficulty of determining when a break occurs. To assist the designer in selection, Table I compares the tensile strengths of solid copper, stranded copper, and high-strength copper alloy of the same size. These values are calculated from the cross-sectional areas, assuming a pure copper ultimate breaking strength of 36,000 psi and copper alloy of 55,000 psi.

18.3.4.3 Conductor coatings/platings. Three conductor coatings are available in the specification slash sheets listed. A selection of the coating material is dependent on the method of termination and maximum temperature to which the wire will be exposed. Of the wires listed in this specification, nickel- and silver-coated wire have the highest temperature rating of 200 °C. Because of the difficulty of soldering nickel, nickel-coated wire is intended mainly for crimp-type terminations.

18.3.4.4 Insulation configuration. The fluorocarbon/polyimide insulation is in the form of a tape which is spirally wound with a 50-percent minimum overlap around the conductor. A second tape of the same configuration is wound over the first with an opposite lay. Each tape is specified by a tape code as .1/1/.1, which means 0.1 mil FEP fluorocarbon resin/1 mil polyimide film /0.1 mil FEP fluorocarbon resin. "FEP" is fluorinated ethylene propylene. A sintering process fuses the tapes together to form a continuous coating.

18.3.5 Electrical characteristics.

18.3.5.1 Conductor resistance. The electrical resistance of wire is specified as ohms per 1,000 ft measured at 20 °C. Resistance values for uncoated solid copper and coated stranded wire are compared in Table I. The resistance value as given is essentially that of the pure copper, since the coating has a minor effect on the overall resistance. Silver-coated wire has slightly less resistance than the tin- or nickel-coated wires.

18.3.5.2 Working voltage. Although the working voltage of all the wires in this section is 600 V rms at sea level at the highest temperature rating of the insulation, it is customary to apply derating factors to all maximum ratings as a safety factor in high reliability applications. Even though the wire is rated at 600 V, it is required to pass an impulse dielectric test of 8 KV at room temperature. See Appendix A of MIL-STD-975.

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**18.3 WIRE AND CABLE, FLUOROCARBON/
POLYIMIDE-INSULATED, MIL-W-81381**

18.3.5.3 Insulation resistance. The insulation resistance is a measure of the insulation integrity of the wire conductor coating. The insulation resistance is 2500 M Ω minimum. Usually, the insulation resistance is not of significance in electronic circuits unless the wire is very long and is used in very high impedance applications where the insulation resistance shunting effect might affect the signal voltage. The insulation resistance goes up as the length becomes shorter.

18.3.5.4 Current rating. The current ratings of wires are not usually considered unless appreciable current is involved which will cause heating of the wire due to the I^2R loss. Such temperature rises must be added to the maximum environmental temperature to determine the actual temperature of the wire and must not exceed the temperature rating of the wire itself. Currents flowing in wires in a bundle have a cumulative heating effect that must be taken into account. See derating of wires in Appendix A of MIL-STD-975.

18.3.6 Environmental considerations. In the selection of the wire for a specific application, all the environments which the wire will encounter must be taken into consideration. For example, humidity is not a problem for a spacecraft in space, but might be at a launch center if the space vehicle were there for any length of time.

The outgassing properties of these wires are not controlled and must be evaluated for compliance to project outgassing requirements.

Silver-coated wire is susceptible to cuprous oxide corrosion (red plague) when stored or used in a moist or high-humidity environment.

18.3.7 Reliability considerations.

18.3.7.1 Polyimide insulating material. Because the conductor configurations are essentially the same for all military specification wires, considerations of reliability will be applied to the insulation materials. Polyimide insulating materials are difficult to strip and are less flexible than TFE and FEP. The hygroscopic nature of polyimide insulation should be considered when used for high-humidity or moist environments. Liquid oxygen compatibility is inferior to that of TFE and FEP.

18.3.7.2 Voltage. All of the wires in this section have a voltage rating of 600 Vrms at the maximum operating temperature at sea level. The intent of this rating is to establish a reference point for which the wire insulation is capable of withstanding indefinitely with the outer surface of the insulation at ground potential or zero voltage. Different waveforms or higher frequencies may result in a reduction of the voltage rating. If the wires are encapsulated or otherwise separated from the ground plane, then the stress on the wire insulation itself is reduced with the same voltage applied. This rating does not apply to wire operating in the critical pressure region where corona discharge may take place from the surface of the wire to the ground plane. To prevent corona, the voltage rating should be 190 Vrms at 60 Hz, which gives a peak voltage of 270 V, the limiting value for breakdown in air at any pressure.

MIL-HDBK-978-B (NASA)

**18.3 WIRE AND CABLE, FLUOROCARBON/
POLYIMIDE-INSULATED, MIL-W-81381**

If operation of the wire in the critical air pressure region at higher ac voltages up to 600 Vrms is desired, then additional steps are required to eliminate corona or dielectric breakdown.

Because of the effects of all these conditions, it is mandatory for the designer to apply derating factors to all maximum ratings in high-reliability applications to take into account unanticipated problems. The derating factors are presented in Appendix A of MIL-STD-975.

MIL-HDBK-978-B (NASA)

**18.4 WIRE AND CABLE, POLYVINYLCHLORIDE
(PVC) INSULATED COPPER, MIL-W-5086**

18.4 Wire, polyvinylchloride (PVC) insulated copper, MIL-W-5086.

18.4.1 Introduction. This specification covers PVC-insulated, single-conductor electric wire made with tin-coated copper conductors listed as "general purpose" diameters in the applicable military specification sheets. For M5086/1 and M5086/2, the wires called out in MIL-STD-975 employ PVC insulation in combination with other insulating and protective materials. MIL-W-5086 prohibits the usage for aerospace application.

18.4.2 Usual applications. NASA has restricted M5086/1 and M5086/2 wire to GSE applications. This wire can be used for internal wiring for electric and electronic equipment. The nylon jacket, which protects the primary insulation, has a slipperiness that facilitates its being pulled through conduits for cable harnesses. The wire sizes called out in MIL-STD-975 are the larger sizes which are intended for power and current applications rather than low-level signal wiring.

18.4.3 Part number designation. The complete part number designation is shown below.



As an example, an M5086/2 wire, 16 AWG size white with green stripe would have a complete part number of M5086/2-0016-95.

18.4.4 Physical construction.

18.4.4.1 Conductor materials. Although the MIL-W-5086 general specification calls out both copper and copper alloy conductors, /1 and /2 which are designated in MIL-STD-975 are for copper only. Figure 5 presents the typical configuration for this wire.

The "breaking strength" of pure copper is not given in military specifications because of its extreme ductility and the difficulty of determining when a break occurs. To assist the designer and enable comparison with wire listed in other sections, Table I lists the breaking strengths based on an assumed copper ultimate strength of 36,000 psi and 55,000 psi for copper alloy.

18.4.4.2 Insulation configuration. The insulation for the M5086/1 wire sizes 16 through 12 AWG has two layers consisting of an extruded polyvinyl chloride primary insulation over the conductor followed by a jacket of extruded nylon.

The M5086/2 insulation has three layers. The first is the same as M5086/1, the second is glass fiber braid with finisher. For size 10, the third layer is an extruded jacket of clean nylon which is the same as that in M5086/1. For the larger sizes, 8 through 0000, a nylon braid jacket with nylon finisher is used.

MIL-HDBK-978-B (NASA)

18.4 WIRE AND CABLE, POLYVINYLCHLORIDE
(PVC) INSULATED COPPER, MIL-W-5086

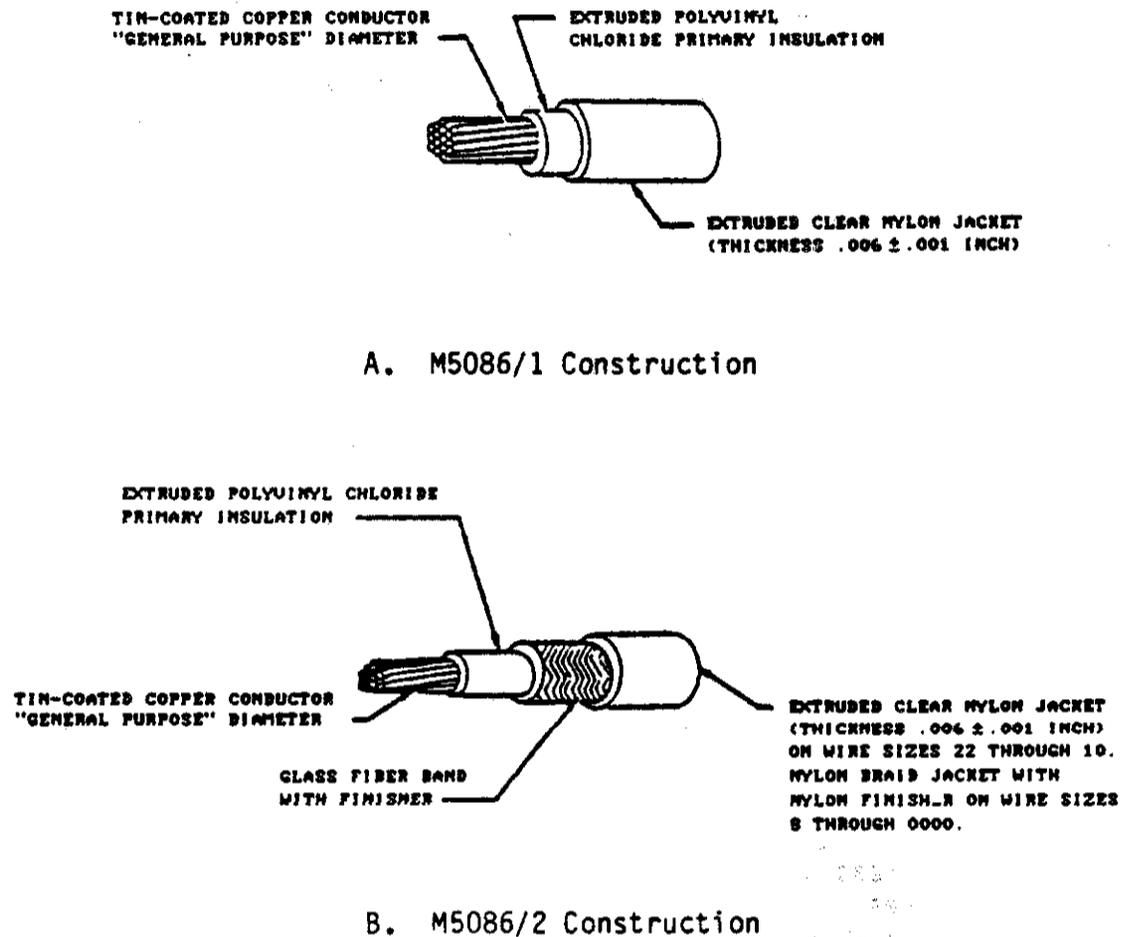


FIGURE 5. Typical construction, MIL-W-5086.

18.4.5 Electrical characteristics.

18.4.5.1 Conductor resistance. Table I identifies the resistance in ohms per 1,000 feet at 20 °C for each size of stranded conductor listed. For comparison, the resistance per 1000 ft for solid wire is also given. The AWG sizes are exact only for the solid wire. Cross-sectional areas for stranded wire may vary as much as 20 percent of the nominal for solid wire.

18.4.5.2 Working and test voltages. The working voltage of all wires in this section is specified as 600 V ac rms at sea level at the highest temperature rating of the insulation. However, it is customary to apply derating factors to all maximum ratings as a safety factor in high reliability application. Refer to Appendix A of MIL-STD-975.

MIL-HDBK-978-B (NASA)

**18.4 WIRE AND CABLE, POLYVINYLCHLORIDE
(PVC) INSULATED COPPER, MIL-W-5086**

Since the insulation is applied in two steps, an intermediate spark or impulse dielectric test is required to prove the integrity of the extruded primary insulation before the nylon jacket is extruded over it. The specifications require that 100 percent of the wire must be tested, both the primary and completed insulation.

18.4.5.3 Insulation resistance. The insulation resistance is a measure of the insulation integrity at the wire conductor coating. The insulation resistance for M5086/1 wire is specified at 500 M Ω for 1,000 ft minimum, and for M5086/2 wire the insulation resistance for size 10 is specified at 40 M Ω for 1,000 ft. No requirements are given for sizes 8 through 0000 AWG.

18.4.5.4 Current rating. The sizes of copper conductors called out in MIL-STD-975 show that these wires are intended for high-current applications such as long runs, heavy power distribution for launch complexes, or similar installations. Since these applications require runs consisting of many cables together, the cumulative heating effects of the cable I²R losses must be taken into account. See derating of wires in Appendix A of MIL-STD-975.

18.4.6 Environmental considerations. The temperature rating of 105 °C represents the maximum permissible operating temperature of the conductor. The maximum allowable ambient temperature should be no more than 105 °C minus the operating rise in temperature of the conductor due to I²R loss.

18.4.7 Reliability considerations. Since the conductor configurations are essentially the same for all MIL-W-5086 wires, considerations of reliability will be applied to the insulation materials. The PVC insulation materials are characterized by reasonable dielectric strength, ease of stripping, flexibility at low temperatures, bondability to most encapsulation materials, and low cost. The problem of low abrasive resistance is overcome in MIL-W-5086 wire by the extruded nylon jacket. Since these wires are not approved for space applications, low outgassing and stability in a vacuum are not prerequisites. Resistance to cold flow and cut-through is very good. The wire is easy to strip and handle, which facilitates manufacturing operations.

CAUTION

PVC materials emit harmful fumes when overheated and should not be applied in confined areas in which there is inadequate ventilation or where crew and test personnel can ingest harmful gases.

MIL-HDBK-978-B (NASA)

**18.5 WIRE AND CABLE, WIRE, POLYVINYLCHLORIDE
(PVC INSULATED, MIL-W-16878)**

18.5 Wire, polyvinylchloride (PVC) insulated, MIL-W-16878.

18.5.1 Introduction. This specification covers polyvinylchloride (PVC) insulated wires made with tin-coated stranded copper conductors from sizes 32 to 18 AWG, rated for 600 V and limited by MIL-STD-975 to ground support equipment use only. MIL-STD-975 makes available to the designer a range of wire sizes from 32 to 0000 AWG with essentially the same characteristics. Each size in MIL-W-16878/1 calls out several stranding configurations. The stranding configurations authorized for use by MIL-STD-975 are 7 strands for sizes 32 through 28 AWG and 19 strands for sizes 26 through 18 AWG.

18.5.2 Usual applications. This wire is authorized by MIL-STD-975 only for ground support equipment. This is unshielded wire for internal hookup and lead wiring for electrical and electronic equipment and switchboards.

18.5.3 Part number designation. The part number designation for ordering or a drawing callout is as follows.

M16878	/1	-B	-C	-B	-96
Basic part number	Specification slash sheet (/1 only)	Conductor material (B = copper only)	Conductor size (A = 32 to H = 18)	Number of strands (B = 7, E = 19)	Base color (white with blue stripe)

As an example, a seven stranded copper wire, 18 AWG size, white with green stripe would have a complete part number of M16878/1-B-H-B-95.

18.5.4 Physical construction.

18.5.4.1 Conductor configuration. Specification MIL-W-16878/1 calls out either solid or stranding configuration, different conductor materials, and tin or silver coating for each wire size. MIL-STD-975 selects the stranded copper, tin-coated wire.

These conductors from sizes 32 to 18 AWG are the conventional configuration consisting of a central strand surrounded by one or more layers of helically wound strands. The strands are positioned in a geometric arrangement of concentric layers to produce a smooth and uniform conductor, circular in cross-section, free of any crossovers, high strands, or other irregularities. The number of strands of each helically wound layer are selected so as to completely fill the available space. The stranding is shown in Figure 6.

The conductor sizes and corresponding size designations are in accordance with established usage for stranded copper conductors for hookup wire in the electronic and aircraft industries. The cross-sectional areas of the stranded conductors in most sizes only roughly approximate those of solid conductors of the same numerical AWG designation.

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**18.5 WIRE AND CABLE, WIRE, POLYVINYLCHLORIDE
(PVC INSULATED, MIL-W-16878)**

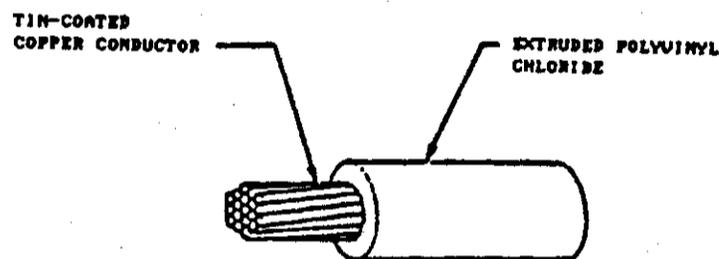


FIGURE 6. Typical construction, MIL-W-16878.

18.5.4.2 Conductor material. Annealed pure copper is the only material called out as the conductor material. Practical experience has shown that copper wire smaller than size 24 is fragile and difficult to handle; consequently its use is not recommended for high reliability applications. Because the "breaking strength" of pure copper conductors is not given in military specifications, calculated values based on an assumed ultimate strength for copper of 36,000 psi are listed in Table I. The designer can make comparisons with the copper alloy wire of the same AWG sizes also listed in Table I.

18.5.4.3 Conductor coating. Only a tin coating is specified.

18.5.4.4 Insulation configuration. The insulation for this wire consists of a single layer of polyvinyl chloride extruded over the conductor.

18.5.5 Electrical characteristics.

18.5.5.1 Conductor resistance. Table I shows the resistance in ohms per 1000 ft at 20 °C for stranded wire and for comparison, solid wire of the same AWG. AWG sizes are exact only for the solid wire. The cross-sectional area for stranded wire is from 5 to 17 percent larger than the equivalent solid-wire areas over this range. Since the cross-sectional area of the stranded wire is greater than that for the equivalent AWG solid wire, a lower resistance would be expected for the stranded; this is not the case for the six sizes, 32 through 22, shown in Table I. Note that the resistance for the solid wire is nominal, whereas that for stranded is a maximum that is set for wire vendors as a "not-to-exceed" specification. Also, the solid wire is uncoated and the stranded is coated, affecting resistance differently. In short, no simple relationship exists between the stranded and solid-wire resistances.

18.5.5.2 Working voltages. The maximum recommended working voltage at sea level and at the maximum temperature is specified as 600 Vac rms between conductor and ground for continuous operation. For dc voltages, this value may be 40 percent higher. However, it is customary to apply derating factors to all maximum ratings as a safety factor in high reliability applications. Refer to Appendix A of MIL-STD-975.

**18.5 WIRE AND CABLE, WIRE, POLYVINYLCHLORIDE
(PVC INSULATED, MIL-W-16878)**

18.5.5.3 Insulation resistance. The insulation resistance is a measure of the insulation integrity of the wire conductor coating, in this case polyvinyl chloride (PVC). An equation is given in M16878/1 to calculate the insulation resistance for each AWG wire size as a function of the insulation volume between the conductor and the outer insulation surface.

18.5.5.4 Current ratings. The current ratings of wires are not considered significant unless significant power is involved. Significant current passing through the wire will cause heating of the wire due to the I^2R loss. Such a temperature rise must be added to the maximum environmental temperature to determine the actual temperature of the wire and in any case must not exceed the temperature rating of the wire itself. Currents flowing in wires in a bundle have a cumulative heating effect that must be taken into account. Refer to derating of wire in Appendix A of MIL-STD-975.

18.5.6 Environmental considerations. The temperature rating of 105 °C represents the maximum permissible operating temperature of the conductor. The maximum allowable ambient temperature should be no more than 105 °C minus the operating rise in temperature of the conductor due to I^2R loss.

18.5.7 Reliability considerations. Since the conductor configurations are essentially the same for all wires called out in MIL-STD-975, considerations of reliability will be applied to the insulation materials. Polyvinyl chloride (PVC) has an operating temperature range of -54 to +105 °C maximum and is characterized by good abrasion resistance, flexibility, high strength, excellent electrical properties (including low water absorption), high cut-through strength, easy strippability, and low cost. The PVC materials have been used extensively in electrical and electronic wiring applications. The material is easily bondable with various encapsulation and potting compounds in applications such as potted connectors.

For such applications as pulling multiple wires through a conduit, wire types with a nylon jacket over the PVC are recommended. This is because the "slipperiness" of the nylon permits easier pulling and protects the PVC insulation from mechanical damage.

The PVC wire shows good tolerance to nuclear and high-energy-particle radiation. However, as previously stated, PVC-insulated wire is recommended by MIL-STD-975 only for ground support applications because of toxic smoke when burned.

This wire is intended for use as a single conductor in the internal wiring of electrical and electronic equipment and switchboards. The general specification points out that wire sizes having a thin-wall insulation of 0.007-in. thickness or less, intended for limited, low-voltage applications, are relatively fragile and easily damaged, and should not be used where chemical stresses or abrasive environments may exist. Precautions should be taken using MIL-W-16878/1 wire, since the lower limits of the finished wire diameter, in comparison with the sizes of the conductors, result in a wall thickness of 0.007 in.

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**18.5 WIRE AND CABLE, WIRE, POLYVINYLCHLORIDE
(PVC INSULATED, MIL-W-16878)**

MIL-W-16878/1 further states that this wire may not be suitable in circuits requiring the highest degree of reliability. Care must be taken in installation to avoid damaging the dielectric with a hot soldering iron and to leave no residual physical strain on the dielectric wall because plastic flow may result in failures.

The conductor sizes on the corresponding size designations of MIL-W-16878/1 are in accordance with established usage for stranded-copper conductors or hook-up wire in the electronic and aircraft industries, although these wire-size designations are not identical with AWG sizes for solid wire. The diameters and cross-sectional areas of the stranded conductors of this specification are, in most sizes, only roughly approximate to those of AWG solid conductors in the same numerical size designation.

**.6 WIRE AND CABLE, CABLE, RADIO-
FREQUENCY, FLEXIBLE, COAXIAL, MIL-C-17**

18.6 Cable, radio-frequency, flexible, coaxial, MIL-C-17.

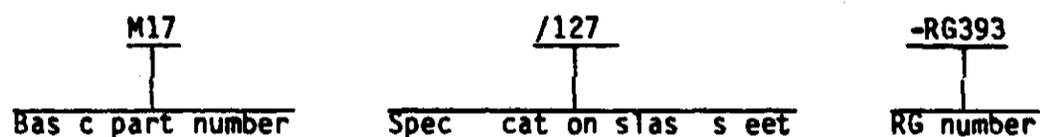
18.6.1 Introduction. This section covers cables with a coaxial configuration, consisting of a center conductor, concentric insulation material and an outside braided conductor. Coaxial cables are distinguished from single-conductor shielded wire by the requirement that the ratio of the outside diameter of the center conductor and inside diameter of the outer conductor must be maintained accurately to achieve optimum transmission line characteristics at radio frequencies. Maintenance of these dimensions also achieves a constant series inductance per foot and parallel capacitance per foot, which are characteristics of rf cable configurations.

Neglecting other factors for the moment, the selection of a coaxial cable over a single-wire shielded configuration is dictated when the wavelength of the highest-frequency signal being transmitted over the wire becomes significant with respect to the length of the cable. Quantitatively, this is usually when the length of cable is 5 to 10 percent of the wavelength of the highest frequency involved. Another factor which may dictate the selection of a coaxial cable is the availability of coaxial connectors. Coaxial cables can also be used as pulse forming lines and as capacitive or inductive impedance elements at high frequencies although cables are usually specifically designed for such applications.

Coaxial cables designated in MIL-STD-975 are of the flexible type, most commonly used in aerospace systems. These cables are characterized by a woven mesh outer conductor, in contrast to rigid and semirigid coaxial cables which have a solid aluminum or copper outer conductor that usually can be bent to a given configuration, although they are not considered flexible.

18.6.2 Usual applications. Coaxial cables are primarily intended for applications requiring the transfer of rf energy from one location to another. The rf can consist of a single frequency as in continuous wave (CW) or pulse wave forms requiring a wide band capability.

18.6.3 Part number designation. The part number designation is as follows:



18.6.4 Physical construction. The center conductor of the coaxial cables is either a single-strand steel wire, copper-covered and silver-coated, or a stranded copper conductor. Because of the small diameters involved, the steel wire is required for strength, whereas the copper covering is a base for the silver coating which is used to improve the conductivity of the center conductor at rf frequencies.

18.6 WIRE AND CABLE, CABLE, RADIO-FREQUENCY, FLEXIBLE, COAXIAL, MIL-C-17

The dielectric core is solid extruded TFE of the precise outside diameter to obtain the design nominal characteristic impedance. The outer conductor consists of one or two layers of a woven braid of silver-coated round copper wire. The braid coverage is usually specified as a percentage, approximately 90 percent (approximately 10 percent of the area is taken up by small holes and gaps in the weave of the braid). In some low-level applications where emi is a consideration, the percentage coverage becomes important. Double-braid helps to give a more complete coverage to the outer conductor. Typical construction shown in Figure 7.

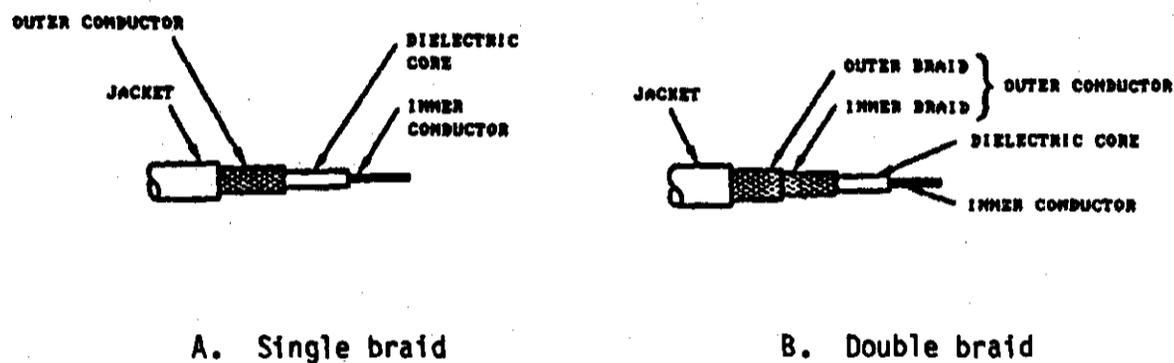


FIGURE 7. Typical construction, MIL-C-17 cable.

All cables have the outer conductor covered by an FEP jacket to protect the braid from damage in handling and to keep out moisture.

Eccentricity. Analysis shows that the eccentricity limit of MIL-C-17, which permits the center conductor to be displaced 10 percent of the OD from the center, reduces the characteristic impedance by 1 percent. The limits on the various characteristic impedances are given in ohms, \pm from the nominal value, which figures to be ± 4 percent. The eccentricity factor is among other various manufacturing tolerances, such as variations in dielectric constant, etc, which, added together, will be within the 4 percent allowed. If the design application requires closer tolerance for the characteristic impedance, then an investigation should be made to select the proper cable.

Conductor adhesion. The maximum and minimum adhesion requirements for the center conductor allow for handling and stripping. The minimum is set high enough to permit handling and flexing of the coaxial cable without separation of the dielectric from the center conductor, thus preventing the occurrence of voids which would generate electrical noise in low-level circuits and corona in high-voltage applications. The high limit is set to facilitate stripping without exerting an undue amount of force which might damage the cable.

18.6 WIRE AND CABLE, CABLE, RADIO-FREQUENCY, FLEXIBLE, COAXIAL, MIL-C-17

Elongation and breaking loads. Elongation minimums in percent and the ultimate tensile strength minimum in lbs per square inch are specified in the slash sheets, but the basic specification cites an ASTM specification rather than addressing the procedures to measure these parameters.

18.6.5 Electrical characteristics.

18.6.5.1 Characteristic impedance. For a coaxial configuration, the characteristic impedance of a cable is given by $Z_0 = \frac{138}{\sqrt{\epsilon}} \log_{10} \frac{D}{d}$, where D is the ID of the outside conductor, d is the OD of the inner conductor, and ϵ is the dielectric constant of the insulation material. This quantity Z_0 is the effective impedance (resistance) that would be encountered by an ac circuit connected to a coaxial cable infinitely long.

If the cable is cut and a resistor equal to the characteristic impedance is connected across the end, the effect on the sending end is unchanged. Impedances which have been selected as standard are 50 ohm, 75 ohm, and 95 ohm.

If the cable termination is either open circuited or short-circuited, this discontinuity causes a reflection in the applied signal as it propagates to the end. This reflection causes voltage standing waves and is expressed as a voltage standing wave ratio (VSWR) in power applications. A high VSWR in power systems dissipates power in the cable, causing localized heating and increased losses. For low level and signal pulse circuits, the termination in other than the characteristic impedance causes reflections of the pulses. It is a good practice to terminate a coaxial cable in its characteristic impedance to prevent these effects.

18.6.5.2 Maximum operating frequencies. Coaxial cables operate in the transverse electromagnetic (TEM) mode. The TEM mode has both the electric and the magnetic fields normal to the direction of propagation. The possibility of propagation in the higher modes limits the usefulness of the coaxial cable to below the lowest mode cutoff frequency. The maximum operating frequency given in the specification is derated sufficiently from the TEM cutoff frequency to prevent any possibility of operation of the cable in higher modes. Other factors contributing to a lower operating frequency are the elements of the construction of the cable and associated connectors. The recommended operating frequency takes into account all of these factors, but designers should always check the capabilities of cables, connectors, and assemblies before operating at any high frequency near cutoff.

18.6.5.3 Attenuation. The reduction or attenuation of the signal passing through the cable as a function of frequency is plotted in each slash sheet except /94 and /95, where single values are listed. Note that the attenuation is given in db per 100 ft, and increases with frequency. The db attenuation at cable lengths other than 100 ft can be calculated by the direct length ratio. Also note that the attenuation is for the cable only. Attenuation values for connectors must be added to the cable to obtain the overall maximum attenuation.

18.6 WIRE AND CABLE, CABLE, RADIO-FREQUENCY, FLEXIBLE, COAXIAL, MIL-C-17

18.6.5.4 Power ratings. The power rating is a maximum power handling capability, in watts, that a coaxial cable can safely transmit without overheating or developing dielectric breakdown throughout the useful frequency range. This maximum power as a function of frequency is plotted in each slash sheet.

Because of the increased dielectric loss as the frequency is increased, the power transferring capability of the cable goes down. In order to keep the cable from exceeding its maximum allowable temperature, in MIL-C-17 these power ratings assume a VSWR of 2 and an ambient temperature from 38 °C to +149 °C for TFE dielectric, and also a maximum inner conductor temperature of +200 °C for TFE.

18.6.5.5 Capacitance. The capacitance of a coaxial cable is determined by the ratio of the inner and outer conductor diameters and the dielectric constant of the insulation material. This quantity is usually specified in picofarads per foot. For low frequencies or dc applications, the total capacitance of an open circuit cable is the capacitance per foot times the number of feet. At higher rf frequencies the reactance of the series inductance becomes significant. The length of the cable becomes an appreciable fraction of the wavelength of the applied rf signal and the input impedance becomes capacitive or inductive, depending on the length of the cable and the frequency applied.

18.6.5.6 Maximum continuous working voltage. The maximum continuous working voltage is the ac rms voltage that can be continuously applied to the coaxial cable. This voltage is limited by the onset of corona discharge or partial breakdown. In MIL-C-17, this working voltage is 75 percent of the corona extinction voltage for sea-level atmospheric conditions and 60 Hz test voltages. Higher frequencies and operation at high altitudes will reduce the corona inception voltages. For high voltage applications, tests and evaluation of these parameters should be made before the design is finalized.

18.6.5.7 Insulation resistance. Because of the high performance requirements of the dielectric materials for high frequency applications, insulation resistance is not considered applicable to coaxial cables.

18.6.5.8 Current ratings. The current ratings of the designated coaxial cables are considered adequate for space applications so that overloading of the center conductor is not a problem. No requirement for current is stated in the basic MIL-C-17 specification or in the slash sheets. However, the I^2R loss for a high current application should be evaluated since a maximum resistance for the center conductor is designated for each of the coaxial cables. This value is intended as a check on the method or adequacy of termination to assure that there is a good joint between the connector and the cable.

18.6.6 Environmental considerations. Outgassing properties of these cables are not controlled and must be evaluated for compliance to project outgassing requirements.

18.6 WIRE AND CABLE, CABLE, RADIO-FREQUENCY, FLEXIBLE, COAXIAL, MIL-C-17

Silver-coated copper conductors, whether used as shielding or as a center conductor, are susceptible to cuprous oxide corrosion (red plaque) when stored or used in moist or high-humidity environments.

18.6.7 Reliability considerations. Good practice sets a minimum bend of 6 times the jacket diameter of the coaxial cable but 10 times is more desirable. Care should be taken in the forming and handling of any flexible coaxial cable to prevent wrinkling or cracking.

**18.7 WIRE AND CABLE, CABLE ELECTRIC
SHIELDED AND UNSHIELDED, MIL-C-27500**

18.7 Cables, electric, shielded and unshielded, MIL-C-27500.

18.7.1 Introduction. This specification covers the requirements for aerospace electrical cables fabricated from individual wires called out in two basic wire specifications, MIL-W-22759 and MIL-W-81381. Four classifications of cables are covered as follows:

Unshielded, unjacketed - From two to seven color-coded wires, spirally laid without an overall outer jacket

Unshielded, jacketed - From two to seven color-coded wires, spirally laid with an overall outer jacket

Shielded, unjacketed - A single wire, or from two to seven color-coded wires, spirally laid with one or two overall shields

Shielded and jacketed - A single wire, or from two to seven color-coded wires, spirally laid with one or two shields and jackets.

Note that the maximum number of wires in any one cable is seven and all are required to be the same AWG size.

18.7.2 Usual applications. These cables are intended for use in applications requiring wires in a cable configuration for additional versatility and protection. Cables covered by this specification can be used for Grade 1 or Grade 2 applications. Since the outgassing characteristic of the various wires and cables used is not controlled, it is necessary that this property be evaluated for compliance to project outgassing requirements.

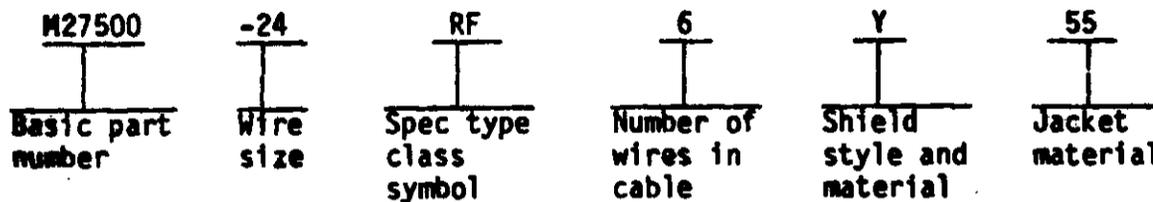
Cables obtained to this specification are intended for low noise or emi control. For example, a twisted pair, triplet, or quad with overall shielding provides excellent reduction of both electromagnetic (low frequency) and electrostatic interference. This is true both for "noisy" power circuits which generate emi and, thus, adversely affect nearby low-noise circuits, and for low-noise circuits themselves. To achieve the maximum noise attenuation, care must be taken to prevent ground loops or circulating currents in the shielding.

Even without shielding, some circuits are sensitive to the relative positions of certain wires in the harness. The relative locations of such wires are maintained when a jacket is shrunk over them in accordance with MIL-C-27500.

Further discussions of the applications of conductor materials and coatings is covered in subsections for MIL-W-22759 and for MIL-W-81381.

18.7.3 Part number designation. Following is the complete part number designation. Note that the color coding of the wires making up the cable must be designated separately because no provision exists for this in the cable part number.

**18.7 WIRE AND CABLE, CABLE ELECTRIC
SHIELDED AND UNSHIELDED, MIL-C-27500**



18.7.4 Physical construction. As noted in 18.7.1, there are four classifications of cables. The maximum number of wires in a cable (all the same AWG) is seven. Descriptions of the construction can be found in MIL-C-27500.

18.7.5 Voltage-withstanding considerations. The capabilities of the individual wires to withstand specified environments is given under the applicable wire specifications. The additional tests called out by this specification refer to the shielding braid, detection of jacket flaws, and dielectric voltage-withstanding capability of the jackets. When two shields are called out, double jackets are required. One jacket is between the shields, to insulate them from each other, and an outer jacket is usually required over the shield.

18.7.6 Environmental considerations. Silver-coated wire is susceptible to cuprous oxide corrosion (red plaque) when stored or used in a moist or high-humidity environment.

18.7.7 Reliability considerations. Because the cables covered by this specification are intended for use in applications which require cabled wire for additional versatility and protection, some additional precautions should be taken to obtain the highest reliability possible. Since a bundle of wires will not bend as sharply and is not as flexible as a single wire, the designer should take care in laying out the path and location of cable bends with proper clearances so that chafing or vibration will not cause a problem. Cable bends should be gradual without kinking. The cable, if it is intended to be flexed, should be clamped or spot bonded as required to prevent any motion during high shock or vibration environments. The selection of TFE, ETFE, FEP, or polyimide wire in shielded cable harnesses depends upon factors such as the environment and material compatibility. Refer to Appendix A of MIL-STD-975 for derating.