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NASA SP-3067

RADIO FREQUENCY INTERFERENCE HANDBOOK

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

NASA SP-3067

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Prepared at NASA Goddard Space Flight Center

Scientific and Technical Information Office 1971
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
Washington, D.C.

For sale by the National Technical Information Service
Springfield, Va. 22151
Price $3.00.
Library of Congress Catalog Card No. 74-174579

FOREWORD

The problem of preventing radio-frequency interference (RFI) from adversely affecting field operations at ground receiving stations in a complex aerospace communication network is a continual battle. RFI problems involving transmitter harmonic generation, receiver spurious response, shielding, and grounding originate in the equipment design phase. Such problems can become serious in the field when an assemblage of receivers, transmitters, computers, time standards, and numerous other electronic equipment are interconnected to form a data acquisition radio link for spacecraft. Furthermore, the environment in which such equipment operates introduces its own problems. Worldwide field stations, such as the Rosman, North Carolina, Space Tracking and Data Acquisition Network (STADAN) facility, are subject to lightning discharges which pose problems of personnel hazard and equipment burn-out.

For several years the Headquarters Office of Tracking and Data Acquisition, Supporting Research and Technology (SRT), National Aeronautics and Space Administration, has supported a STADAN RFI Reduction Program (523-150-18-02-51) in the Advanced Development Division at Goddard Space Flight Center, Greenbelt, Maryland. An objective of the SRT program has been to define the various mechanisms that produce RFI, to provide effective solutions to minimize the degrading effects of RFI, and to ensure electromagnetic compatibility (EMC) in the performance of systems. The information contained in this handbook addresses such topics.

The handbook has been divided into three sections:

Section I—Electromagnetic Compatibility Fundamentals Applied to Spacecraft Radio Communication Systems.

Section II—Electromagnetic Compatibility Design Guideline for STADAN.

Section III—Lightning Protection Practices Applied to Field Station Installations.

Each of the sections, respectively, summarizes activities on NASA contracts: NAS5-9896 (Section I) with the Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia, Pennsylvania; NAS5-10017 (Section II) with Genisco Technology Corporation, Genistron Washington Facility; and NAS5-10572 (Section III) with the General Electric Company, High Voltage Laboratory, Pittsfield, Massachusetts. Furthermore, each section is essentially a separate entity, although some overlap does exist between Sections I and II. An extensive set of references is included in each section.

Furthermore, the three sections in the handbook contain a substantial amount of new information not generated in the above NASA contracts. I would especially like to thank the respective authors of each section for their assistance in this regard.

Also, I would like to thank the Electromagnetic Compatibility (EMC) Group of the Institute of Electrical and Electronics Engineers (IEEE), Inc., for providing the academic atmosphere that substantially influenced the preparation of this handbook. For instance, seven officials and members of the IEEE EMC Group directly contributed to the preparation of the three sections; their respective names appear therein.

Briefly, the scope of the three sections is summarized as follows:

Section I presents theoretical considerations involving the generation of unwanted output of transmitters, the unintentional generation of noise, and linear and nonlinear intrusion noise mechanisms in receivers; it also considers problems associated with the shielding, grounding, and filtering of equipment. Recommendations for control of interference and for the selection of sites for ground stations are also included.

Section II addresses desirable practices to achieve electromagnetic compatibility through proper design of equipment for field stations. Techniques discussed for preventing undesired electromagnetic interference under normal operating conditions include shielding to obtain desired reflection and absorption losses, grounding, filtering, and wiring and cabling. Various conversion charts related to RF interference have been included in Section II as helpful calculation aids to the design engineer.

Section III provides a discussion of lightning protection techniques applied to field station installations. Lightning phenomena and the effects of inductive, capacitive, and resistive components of lightning-induced voltages are discussed. The Rosman, North Carolina, STADAN installation is analyzed from the standpoint of lightning protection. Specific recommendations include the installation of voltage clippers of high surge current rating for the protection of sensitive circuits of solid-state electrical and electronic equipment.

The handbook brings together in one convenient place many topics in theory and practice that have proven to be useful. Related documents published by the Department of Defense contain certain material discussed here. It is hoped that the handbook may offer a concept, a curve, or a problem stated in a slightly different way and so prove valuable to the design or field installation engineer.

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SECTION I

ELECTROMAGNETIC COMPATIBILITY FUNDAMENTALS APPLIED TO SPACECRAFT RADIO COMMUNICATION SYSTEMS

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1. INTRODUCTION

This section is intended as a design guide to minimize electromagnetic interference in aerospace communication equipment for ground stations such as the National Aeronautics and Space Administration's (NASA) space tracking and data acquisition network (STADAN). Specifically treated are the mechanisms of generating unwanted radio emissions that may affect station operations as well as other communications services, the mechanisms by which sensitive receivers become susceptible to interference, means for reducing interference, standard methods of measurement, and the problems of site selection. The sources of interference are viewed primarily as originating from communications transmitters aboard spacecraft and aircraft, ground transmitters within and outside the ground stations, and other electrical sources on the ground that are not intended to radiate.

In this section the term radio frequency interference (RFI) noise is any electromagnetic interference, periodic or random, that may have a disturbing influence on devices exposed to it. RFI noise is distinguished from thermal noise and may arise from natural or man-made sources; the category of man-made RFI noise includes transmitters and similar devices producing intended radiation, as well as electrical devices that radiate incidentally. The RFI noise may be transmitted through a conductive path, or it may be propagated through space. RFI noise may result in a uniform reduction of normal output, it may take the form of a fluctuation superimposed on the normal output, or it may show other deleterious effects.

We shall make use of the term "susceptibility" to quantify the degree to which a device is sensitive to RFI noise. For example, the level of input RFI noise required to give a specified degradation of the output quality of a receiver is a measure of susceptibility. The degree of susceptibility will vary with the type of input RFI noise, with the path through which the RFI noise enters the device, and with the characteristics of the device itself.

Terms referring to specialized phenomena will be identified and defined where pertinent in the text.

Given below is a list of the known RFI noise sources and a summary of the intrusion mechanisms which play a role in ground stations. The order in which the RFI noise sources are presented corresponds to their significance.

1.1 SPACECRAFT EMISSIONS

Two satellites that are simultaneously present in the main beam of a ground antenna that have co-channel frequency assignments can inject RFI noise in a ground receiver.

1.2 AIRCRAFT EMISSIONS

Airborne transmitters above the horizon of the ground stations can inject RFI noise from adjacent channels. Because of large amounts of transmitter power at small distances, aircraft interference may become significant even if received by the antenna sidelobes. Ordinarily the magnitudes of received RFI noise will not be so high as to stimulate nonlinear effects at the ground receiver, except when an aircraft is in the main beam of the ground antenna and flying at a range of less than 10 miles from the antenna.

1.3 GROUND STATION TRANSMITTER EMISSIONS

Self-interference arises from station transmitters. Although not concentrated at the frequency to which the ground receiver is tuned, large magnitudes of power output from local transmitters may penetrate the receiver by a nonlinear mechanism or may circumvent the shield to enter the receiver at points other than the antenna.

Possible nonlinear mechanisms by which such RFI noise can enter a receiver are spurious responses, intermodulation, cross-modulation, and desensitization. Mechanisms involving inadvertent coupling through an indirect input often include conduction through power and control cables, conduction and induction through ground loops, and field-induced pickup through equipment and cable shields that are inadequate because of poor design or because of imperfections that have developed with use. It is also possible that the transmitter itself may generate at the frequency to which the receiver is tuned. RFI noise-producing mechanisms here include sideband noise, sideband splatter, harmonics of the output signal, leakage of signals intended for internal use, and intermodulation and cross-modulation with an external signal.

1.4 UNINTENTIONAL SITE EMISSIONS

Interference is also produced by electrical and electronic devices that are part of the ground installation but which are not intended to act as radiators. Conducted and/or radiated RFI noise may arise from rotating electrical machinery, power switches and relays, power control circuits such as thyratrons and silicon controlled rectifiers, pulse devices such as computers, and automotive ignition systems.

1.5 UNINTENTIONAL OFF-SITE EMISSIONS

Significant off-site RFI noise includes electrical noise generated by the many sources operating simultaneously in nearby populated areas (urban noise); noise from automobile ignition systems, particularly if major highways are nearby; discharges from nearby high voltage power transmission lines (corona noise); and emission from ground transmitters.

The nature of these phenomena has been under study for many years, and numerous publications can be found which treat one or more of these topics. Where more detail is required on interference mechanisms and reduction techniques References 1 through 3 are recommended.

Since quantitative prediction of RFI noise emission and of susceptibility levels depends upon the design of the electronic devices and circuits used, emphasis is placed here on the principle of evaluating RFI noise. The view is taken that, in the design of equipment, awareness of the RFI noise mechanism and knowledge of possible cures for RFI noise are essential. Measurement should be relied upon to give a quantitative assessment of the efficacy of the design.

Ordinarily, it will not be enough to know only the mechanism of entry and the level of unwanted signal admitted in the passband. The degree to which the unwanted signal influences the output is the real question. Typically, a desired signal level of +20 dB, relative to the undesired signal, will be satisfactory.

2. GENERATION OF UNWANTED OUTPUT IN TRANSMITTERS

A transmitter will radiate a certain amount of unwanted energy outside of its assigned frequency band. In some instances, this is a result of the normal nature of the modulation process. The mechanisms by which spurious signals are generated are, however, of equal or even greater importance. The most significant cause of spurious signals in the transmitter output is undesired nonlinearity in some portion of the transmitter. The two most serious offenders are the transmitter modulator and the final amplifier. The former, which must be a time-varying device, is usually a nonlinear element. Deviations from the required response law can cause spectrum broadening or "sideband splatter." The final amplifier is required to be linear over a wide range of input levels. Deviations from linearity can give rise to the emission of transmitter harmonics.

Less common sources of unwanted outputs may be found at intermediate points in a transmitting system. The audio or video signal, prior to modulation, may find a leakage path out of the transmitter. In systems wherein a low-frequency sinusoid is generated initially and then multiplied to the required output frequency, the subharmonics may leak out of the transmitter. The possibility of parasitic oscillations also exists at various points. These are not common in competently designed, low-frequency transmitters but are common in microwave transmitters. Another mildly troublesome source is oscillator noise. The effect of such noise is similar to that of sideband splatter, but the level is generally lower.

There are times when the transmitter is not, by itself, responsible for the spurious output. The output of nearby transmitters may enter a transmitter, usually through its antenna, and mix with the desired signal to form an unwanted output component. Again, the effect depends on nonlinearity in the final amplifier, in the transmission system between the final amplifier and the antenna, or in the antenna itself. The effect produced is either cross-modulation, wherein the information sidebands of the undesired signal appear with the desired signal, or intermodulation, wherein a third signal, containing some version of the information sidebands of both signals, is formed.

The nature of these mechanisms will be developed in greater detail in the following sections.

2.1 SIDEBAND SPLATTER

The unsavory sounding term "sideband splatter" identifies spectral components formed immediately outside the assigned frequency band. Such components arise in the modulator as the result of nonlinearity beyond that required for the modulation process. The problem is of particular importance in narrow-band modulation systems, such as in amplitude modulation (AM), single sideband (SSB), and double sideband (DSB) systems. To some extent phenomena similar to sideband splatter may be found in wide-band modulation systems, e.g., in frequency-modulation (FM) systems. Furthermore, wide-band modulation systems, such as pulse-modulation and frequency-modulation systems, contain energy throughout a wide range of frequencies. This does not constitute splatter but it is considered here because of the potential adjacent channel interference. In practice, filters are used to restrict the energy to a limited band in such a way that the capability of the system is not significantly impaired. A point should be made here concerning the measurement of sideband splatter. Because low-amplitude unwanted sidebands are to be measured in the presence of the large amplitude of wanted components, instruments that effectively filter out the desired components are necessary.

1. Spectral Properties of Amplitude Modulation and Related Systems

The fundamental process involved in generating AM, SSB, or DSB signals is multiplication. A low-frequency information waveform $x_1(t)$ is multiplied by a sinusoidal carrier frequency $x_2(t) = \cos(\omega_c t + \phi_c)$. As shown in Figure 1.1, the vehicle for generating the product is virtually always a nonlinear device. This is followed by a narrow-band filter to eliminate undesired components. Ideal nonlinear elements are the square-law device and the linear diode (Reference 4, pp. 97 to 100, or Reference 5, pp. 187 to 192). To obtain the required result without distorting the desired sidebands, the amplitude of the sinusoidal carrier applied to the linear diode must be much larger than that of the information waveform. The ideal characteristics are, however, only approximately realizable. A typical nonlinear device can be described by the output-input characteristic

$$y = \sum_{n=0}^{N} a_n x^n, \tag{1.1}$$

where y is the instantaneous output and x is the instantaneous input. The degree of significant nonlinearity is assumed here not to exceed that represented by N. The output of the nonlinear device for the input $[x_1(t) + x_2(t)]$ is

$$\begin{aligned} y &= \sum_{n=0}^{N} a_n (x_1 + x_2)^n \\ &= \sum_{n=0}^{N} a_n \sum_{k=0}^{n} \binom{n}{k} x_1^{(n-k)} \cos^k(\omega_c t + \phi_c). \end{aligned} \tag{1.2}$$

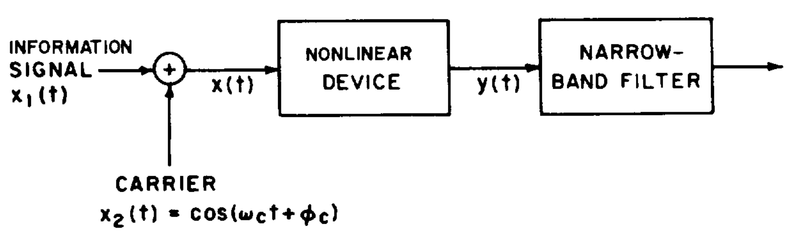

Figure 1.1.—Basic elements of product modulators.

The second form in (1.2) is obtained from the first, using the binomial expansion theorem. Assuming that the nonlinear element is followed by a narrow-band filter centered at the carrier frequency ω_c, only terms containing $\cos(\omega_c t + \phi_c)$ are passed. By expanding* $\cos^k(\omega_c t + \phi_c)$ and retaining only these fundamental terms, there is obtained

$$\sum_{n=1}^{N}\sum_{k=1}^{n} a_n \binom{n}{k} x_1^{(n-k)} \frac{k!}{\left(\frac{k-1}{2}\right)!\left(\frac{k+1}{2}\right)!} \cos(\omega_c t + \phi_c). \qquad (1.3)$$

The x_1 terms higher than the first degree are distortion terms which extend the sidebands to $(N-1)$ times the frequency band of the modulating signal. For instance, if $n = 7$,

$$\begin{aligned} y_a(t) = \Big[&(a_1 + 3a_3 + 10a_5 + 35a_7) \\ &+ (2a_2 + 12a_4 + 60a_6)x_1(t) \\ &+ (3a_3 + 30a_5 + 210a_7)x_1^2(t) \\ &+ (4a_4 + 60a_6)x_1^3(t) \\ &+ (5a_5 + 105a_7)x_1^4(t) \\ &+ 6a_6x_1^5(t) + 7a_7x_1^6(t)\Big]\cos(\omega_c t + \varphi_c). \end{aligned} \qquad (1.4)$$

If coefficients a_n are zero for $n \geqslant 3$, that is, if the device is purely square law, the output is exactly as required.

The component $x_1^2(t)\cos(\omega_c t + \phi_c)$ in (1.4) contributes sideband energy covering a band equal to four times the modulating frequency rather than two times, as in the ideal case; the $x_1^6(t)\cos(\omega_c t + \phi_c)$ component contributes energy in a band twelve times the original modulating frequency.

*The expansion of $\cos^p x$ in a terminating Fourier series is given by

$$\cos^p x = \frac{1}{2^{p-1}}\left[\cos px + p\cos(p-2)x + \frac{p(p-1)}{2!}\cos(p-4)x + \dots + \frac{1}{2}\frac{p!}{\left(\frac{p}{2}!\right)^2}\right],$$

when p is even. When p is odd, the sum is the same except for the last term, which is

$$\frac{p!}{\left(\frac{p-1}{2}\right)!\left(\frac{p+1}{2}\right)!}\cos x.$$

The exact nature of the resultant spectrum depends upon the quantities $x_1^n(t)$. If $x_1(t)$ is a pure sinusoid of frequency ω_m, then $x_1^n(t)$ can readily be determined, and (1.4) can be expanded in a harmonic series involving discrete sidebands of the form cos $[(\omega_c - n\omega_m)t + \phi_c]$. Furthermore, the coefficients a_n will be required. These are not ordinarily available and will have to be found by test (Reference 6). A set of typical values abstracted from Reference 1 is as follows (obtained by polynomial least squares fit to a triode characteristic operated at a bias point of −10 V on the grid and at a plate current of 3 mA):

$a_1 = 0.52 \times 10^{-3}$ A/V	$a_6 = -2 \times 10^{-8}$ A/V^6
$a_2 = 2.6 \times 10^{-5}$ A/V^2	$a_7 = -1 \times 10^{-8}$ A/V^7
$a_3 = -8 \times 10^{-7}$ A/V^3	$a_8 = 9 \times 10^{-10}$ A/V^8
$a_4 = -6.2 \times 10^{-8}$ A/V^4	$a_9 = 9 \times 10^{-11}$ A/V^9
$a_5 = 3.5 \times 10^{-7}$ A/V^5	$a_{10} = 8.2 \times 10^{-12}$ A/V^{10}

A word should be said here concerning overmodulation. The general form for an AM signal is

$$y(t) = A_a [1 + x(t)] \cos(\omega_c t + \phi_c), \tag{1.5}$$

where A_a is the peak amplitude of the unmodulated carrier and $x(t)$ is the information signal. A non-overmodulated waveform is one in which the factor

$$[1 + x(t)] > 0 \tag{1.6}$$

for all values of time. In AM transmitters, when this factor goes negative, the transmitter can be cut off. In such cases the amplitude of the splatter components becomes quite high. Correct practice requires that peaks of $x(t)$ be limited in such a way that (1.6) is always satisfied and that the modulating signal be filtered after amplitude limiting to eliminate the high-frequency components generated in the limiting process.

To minimize distortion, it is desirable to choose nonlinear elements with characteristics as nearly ideal as possible. For a given device intended to be used as a square-law modulator, it will usually be possible to find empirical operating conditions which result in a characteristic that is nearly square law so that the higher order coefficients are small. Furthermore, the final tuned circuits in the transmitter will also act to reduce the level of the unwanted components. As a rule, however, the circuit Q cannot be made high enough to completely eliminate the splatter components. An effective technique to eliminate the unwanted sidebands associated with even powers of $x_1(t)$ in (1.3) is to use a balanced modulator* (Reference 4, p. 104).

Prediction of the magnitude of the unwanted sidebands will ordinarily be difficult. The coefficients a_n in (1.1) depend on the bias conditions and the magnitude of the input signal; they depend on the curvature of the voltage-current characteristics of the modulator and even vary among individual modulating devices of the same kind. To determine the actual level of undesired sidebands, measurements are necessary.

Measured and calculated results of excessive sideband output are reported by Firestone, et al. (Reference 7). In an AM transmitter having an audio modulating signal of 3 kHz and 54.7% modulation, the spectrum distribution shown in Figure 1.2, was obtained. It was pointed out by the authors that the total harmonic distortion amounted to 3.3%, which is not an excessive figure. A receiver operating in an adjacent channel 10 to 15 kHz from the interfering signal will receive splatter components of the order of 60 dB below the level of the carrier component. If, as may readily happen with a nearby transmitter, the carrier power of the transmitter produces an input voltage of about 100 mV at the receiver, the level of unwanted sidebands will correspond to about 100 μV. This will often be far greater than the level of a desired signal.

2. Spectral Properties of Angle Modulation Systems

As is well known, a large number of sidebands is produced by an FM transmitter, even with a perfect modulation technique and with a sine-wave modulating signal. An FM signal containing an information waveform $x(t)$ is written

$$y(t) = A_f \cos\left[\omega_c t + \phi(t)\right] , \tag{1.7}$$

where

$$\phi(t) = \int x(t)dt . \tag{1.8}$$

The instantaneous frequency of $y(t)$ is defined as the derivative of the argument of the cosine, or,

$$\omega_i(t) = \frac{d}{dt}\left[\omega_c t + \phi(t)\right] = \omega_c + \frac{d\phi}{dt} = \omega_c + x(t) . \tag{1.9}$$

*The balanced modulator will be described further in Section 2.2 in connection with harmonic reduction.

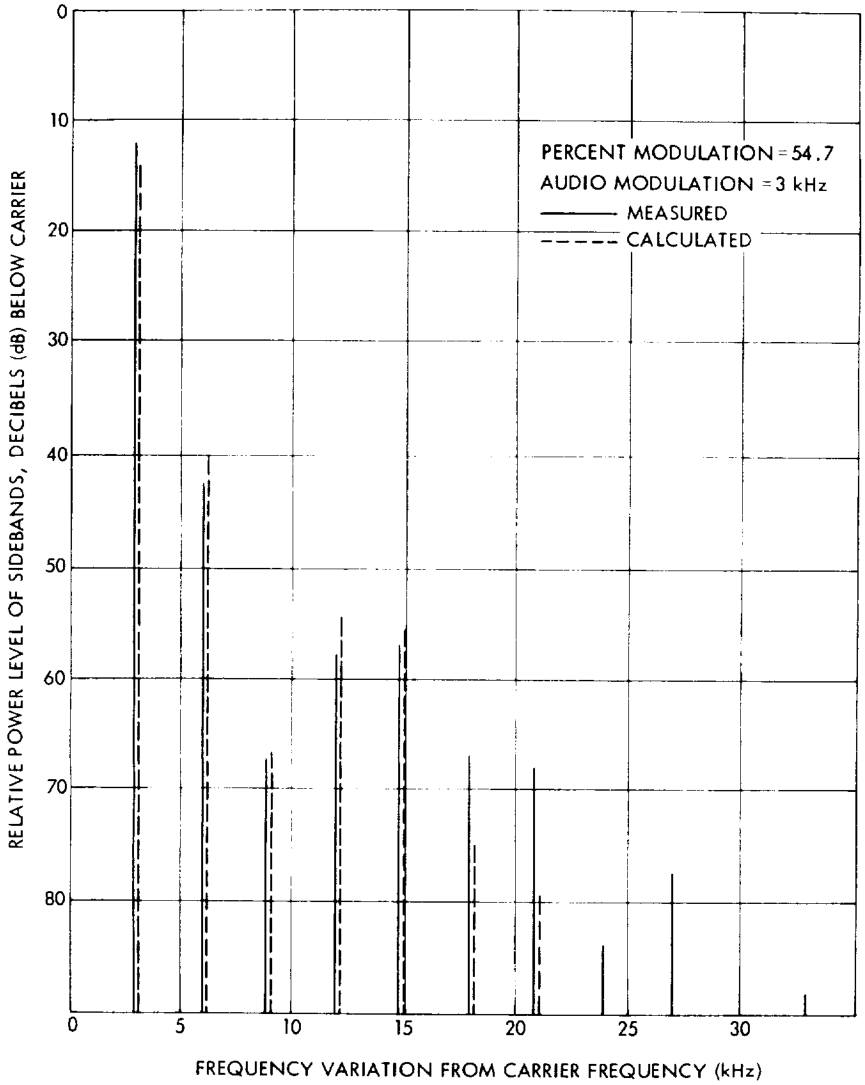

Figure 1.2.–Calculated and measured results of the typical spectrum distribution of sidebands produced by amplitude-modulated transmitters.

If $x(t)$ were a pure cosine wave of the form

$$x(t) = a \cos \omega_m t\,, \tag{1.10}$$

then a would be a peak frequency deviation in radians per second and

$$y(t) = A_f \left\{ J_0(\beta) \cos \omega_c t + \sum_{n=1}^{\infty} J_n(\beta) \left[\cos (\omega_c + n\omega_m)t + (-1)^n \cos (\omega_c - n\omega_m)t \right] \right\} \tag{1.11}$$

where $J_n(\beta)$ is the Bessel function of the first kind and of order n and $\beta \equiv a/\omega_m$ is the modulation index. Though in principle the result contains components to infinite frequency, the amplitudes of the sidebands fall off very rapidly beyond a certain point. A rule of thumb frequently used is that all significant frequency components lie inside the range

$$[f_c - (\beta + 2)f_m] < f < [f_c + (\beta + 2)f_m] \, . \tag{1.12}$$

The magnitudes of the components in this range are shown in Figure 1.3 as a function of n for several representative values of β. For example, for $\beta = 10$ and $n = 12$, Figure 1.3 shows the amplitude of the component $(\beta + 2)f_m$ to be $J_{12}(10) = 0.06$. For values of n larger than $(\beta + 2)$, the components will be relatively small but they may be of importance in some instances. To estimate the magnitude of components in this range, use may be made of the approximation

$$J_n(\beta) \approx \frac{1}{\sqrt{2\pi n}} \left(\frac{\epsilon\beta}{2n} \right)^n . \tag{1.13}$$

where ϵ is the base of the natural logarithm and is equal to 2.718. . . . For instance, if $f_m = 3$ kHz and $\beta = 1$, then according to (1.13), the component $n = 6$, corresponding to a frequency of $nf_m = 6 \times 3 = 18$ kHz away from the center frequency, has an amplitude $J_6(1)$ of about −94 dB, relative to the level of an unmodulated carrier of the same power.

As in the AM case, it is also possible for the modulation process to create a form of sideband splatter, though the extent of the increase in bandwidth for wide-band modulation will not be as great as for AM. In the case of narrow-band FM, however, modulation may give rise to a more significant effect. The origin of this phenomena is, once again, nonlinearity in the modulation process. To generate FM waves it is necessary to vary the instantaneous frequency, or phase, of the carrier in exact correspondence with the information waveform; that is, the characteristic relating applied signal voltage and frequency of carrier must be perfectly linear. Departures from linearity have the effect of distorting the information waveform and increasing the bandwidth. The waveform of the modulated output will therefore have a broader bandwidth than is necessary.

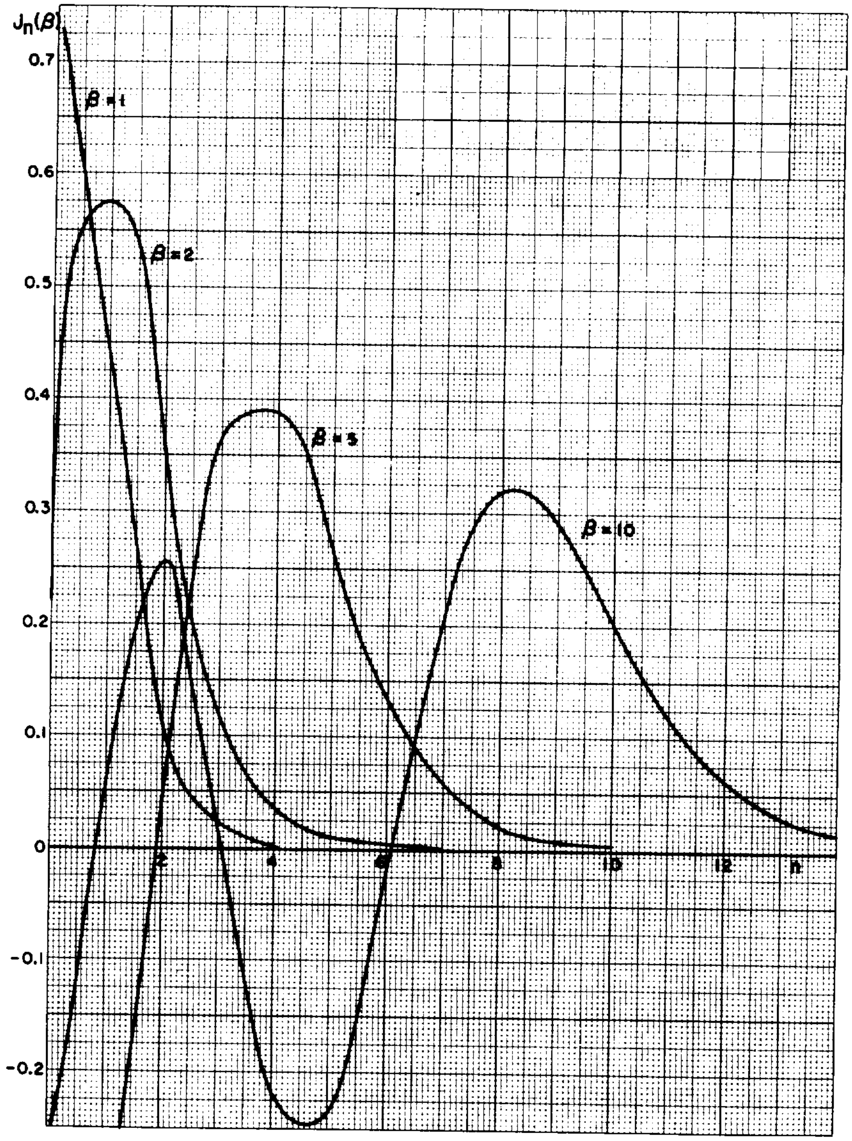

Figure 1.3.–Bessel function of the first kind, J_n (β), plotted as a function of n for various values of β.

As pointed out above, the bandwidth of wide-band FM waveforms is determined mainly by the peak deviation. In this case, for a fixed deviation, the bandwidth of the information signal will not significantly affect the overall bandwidth. For narrow-band FM signals, however, the bandwidth is determined largely by the frequency of the modulating signal, and the effects may be as great as in the AM case. An example involving an indirect FM generation technique will demonstrate some of these points.

FM waveforms are sometimes developed by generating, initially, a low-modulation index signal using adders and product modulators (Reference 4, pp. 116 to 120). If the audio modulation signal is a sinusoid, the form developed is

$$y(t) = A_f[\cos \omega_c t - \beta \sin \omega_m t \sin \omega_c t] \, , \tag{1.14}$$

where β is limited to a maximum value of 0.5. This can also be written

$$y(t) = A_f\left[1 + \beta^2 \sin^2 \omega_m t\right]^{1/2} \cos\left[\omega_c t + \tan^{-1}(\beta \sin \omega_m t)\right] . \tag{1.15}$$

When $\beta \ll 1$ this becomes

$$y(t) \approx A_f \cos\left[\omega_c t + \beta \sin \omega_m t\right] \, , \tag{1.16}$$

which is an FM wave with modulation index β. When high deviation FM signals are required the signal initially generated is of the form given by (1.14), with β maintained sufficiently small. The narrow-band signal is then converted to a wide-band signal by frequency multiplication. Even if β in (1.14) is not small, the amplitude factor in (1.15) can be made constant by the use of a limiter. However, if β is not small, the phase factor will carry some distortion components. For illustrative purposes, assume the amplitude factor A_f has been made constant by limiting and use the first two terms of an expansion for the arc tangent. Then (1.15) takes the form

$$\begin{aligned} y(t) &\approx A_f \cos\left[\omega_c t + \beta \sin \omega_m t - \frac{\beta^3}{3} \sin^3 \omega_m t\right] \\ &= A_f \cos\left[\omega_c t + \left(\beta - \frac{\beta^3}{4}\right) \sin \omega_m t + \frac{\beta^3}{12} \sin 3\omega_m t\right] . \end{aligned} \tag{1.17}$$

The phase, which represents the modulation, now contains a third harmonic term which, for $\beta = 1$, is about 10% of the desired fundamental term. When an FM wave of low modulation index is developed, β is kept less than 0.5; for $\beta = 0.5$, the third harmonic distortion is in the order of 1%. Thus, the modulation index used in the initial narrow-band FM signal should be made as low as possible if excessive bandwidth is not to be consumed.

A recent study (Reference 8) may be found useful in the control of nonlinearity. Two methods recommended here are: predistortion, wherein the input signals are distorted initially in a manner that compensates for the distortion introduced by the modulator, and feedback to attenuate unwanted components.

3. Spectral Properties of Pulse Modulation Systems

Information transmitted by perfectly rectangular pulses will require a larger spectrum than is ordinarily tolerable. It is well known (Reference 5, p. 41) that the Fourier transform of a single rectangular pulse* of width τ_r and height A_r, is given by

$$X_r(f) = A_r \tau_r \operatorname{sinc}(\pi\tau_r f), \tag{1.18}$$

where

$$\operatorname{sinc}(\pi\tau_r f) \equiv \frac{\sin(\pi\tau_r f)}{\pi\tau_r f}.$$

The envelope of the sinc function falls off as $1/f$ for higher frequencies; that is, it falls off at the rate of 6 dB per octave. This rather slow fall-off rate implies that signals of relatively large levels spread into adjacent channels. In practice, however, pulses of slower rise-time will be obtained either by forming such pulses initially or by passing rectangular pulses through a band-limiting filter. Filters with very sharp cutoff will cause some crosstalk between adjacent pulses. In general, the requirements of low crosstalk and low-level spectral tails are opposed to one another. It is possible, however, to find a pulse shape for which both are adequately small. An example is the pulse of a cosine cycle with width τ_c and height A_c, as are given by

$$\begin{aligned} x(t) &= \frac{A_c}{2}\left(1 + \cos\frac{2\pi t}{\tau_c}\right), \quad && -\frac{\tau_c}{2} < t < \frac{\tau_c}{2}, \\ &= 0 \quad && , \quad \text{elsewhere}. \end{aligned} \tag{1.19}$$

This pulse has a spectrum given by the Fourier transform

$$X_c(f) = A_c \tau_c \frac{\operatorname{sinc}(\pi\tau_c f)}{2\left(1 - \tau_c^2 f^2\right)}, \tag{1.20}$$

where

$$\operatorname{sinc}(\pi\tau_c f) = \frac{\sin(\pi\tau_c f)}{\pi\tau_c f}.$$

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*A rectangular pulse of midband frequency f_0 will have a spectrum $\frac{1}{2}[X(f - f_0) + X(f + f_0)]$.

For frequencies significantly greater than $1/\tau_c$ (i.e., for $(\tau_c f)^2 >> 1$), the amplitude falls off as $1/f^3$ or 18 dB per octave. A comparison of the spectra of a square pulse and of cosine pulses of two different widths is shown in Figure 1.4. The value of A_c has been chosen so that the energy is the same for all of the pulses being compared. For equal energy it turns out that $A_c = 1.635\ A_r \sqrt{\tau_r/\tau_c}$. The rectangular pulse is shown at (a) in Figure 1.4 and its Fourier spectrum is shown in normalized form as $X_r(f)/A_r\tau_r$ in the figure. Cosine pulse (b) in Figure 1.4 is chosen to have the same width as the rectangular pulse ($\tau_c = \tau_r$). Its spectrum is defined as $X_{c1}(f)$. Its value is given by (1.20) and it is plotted in a normalized form as $X_{c1}(f)/A_r\tau_r$. It will be noted that the spectrum of the cosine pulse (b) is concentrated at low frequencies, and has a greater bandwidth* than the rectangular pulse. At higher frequencies, however (i.e., $f > 2/\tau_r$), the amplitudes of spectral components of the cosine pulse are comparatively very small. In practice the duration of the cosine pulse might be made longer than the duration of the square pulse. Accordingly, a cosine pulse having twice the duration of the rectangular pulse ($\tau_c = 2\tau_r$) is chosen for illustration and is shown as pulse (c) in Figure 1.4. The spectrum of this pulse is defined as $X_{c2}(f)$ and is plotted as the normalized form $X_{c2}(f)/A_r\tau_r$. The amplitude of the spectrum in this case is comparatively quite small for $f > 1/\tau_r$.

4. Spectral Properties of Radar Pulses

The off-band emission generated by a narrow microwave radar pulse can cause serious interference, particularly in on-site, adjacent-channel receiving equipment. The following analysis gives a method for determining the power level of off-band interference, assuming that an ideal, rectangular radar pulse is transmitted.

The train of radar pulses shown in Figure 1.5 has an average power level P_a which is determined by the peak-pulse power level P_t, pulse width τ, and pulse repetition period T. The average power level is

$$P_a = \left(\frac{\tau}{T}\right) P_t \,. \tag{1.21}$$

As shown in Figure 1.6, the Fourier power spectrum of a single pulse is a $[(\sin x)/x]^2$ function, where $x = n\pi\tau/T$ and n is the number of the spectral line measured from the center frequency f_0. The fractional power P_n of the nth

*Bandwidth is defined as the frequency at which the amplitude is down $1/\sqrt{2}$ from the peak amplitude at origin.

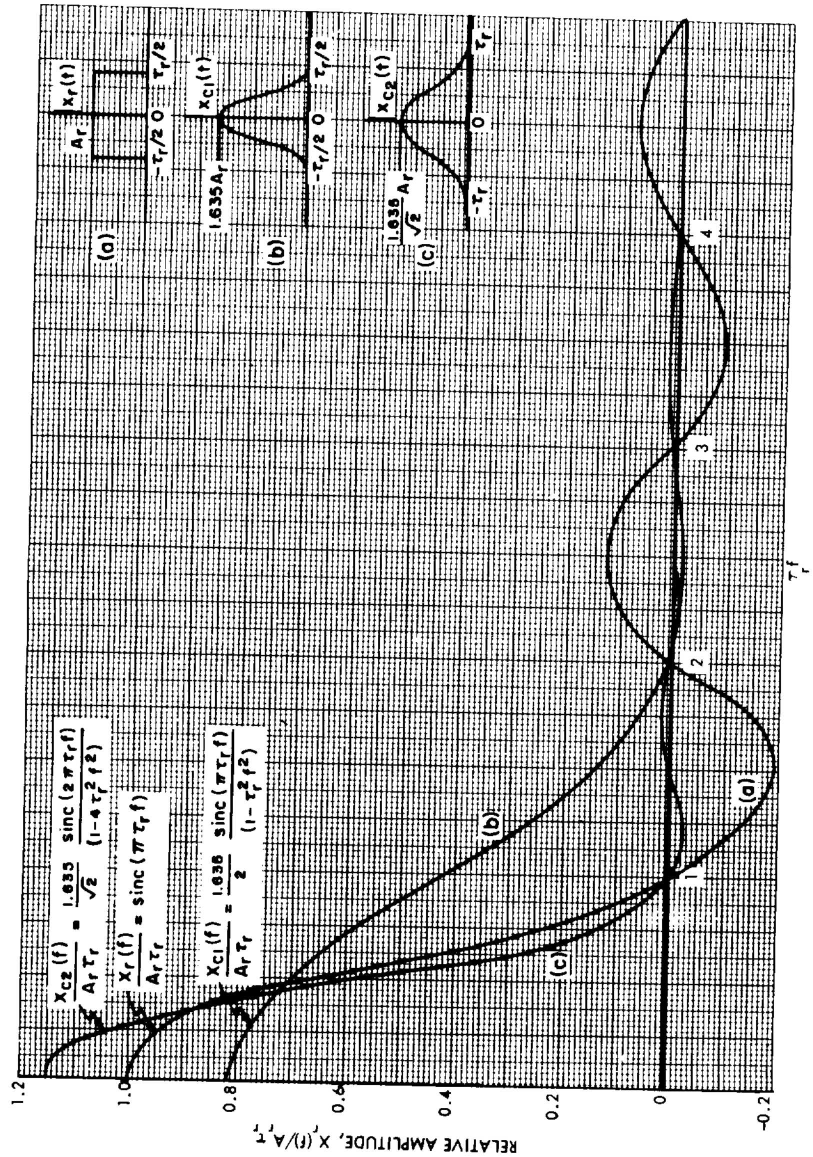


Figure 1.4.—Spectra of rectangular and cosinusoidal pulses.

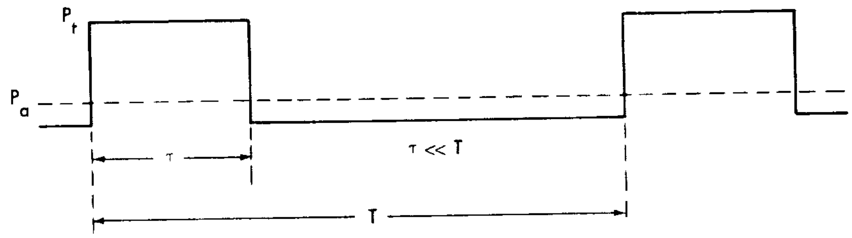


Figure 1.5.–Radar pulse train.

spectral line, normalized with respect to P_a, is determined from the relation (Reference 9)

$$P_n = \frac{T}{\tau}\left[\frac{\sin(n\pi\tau/T)}{n\pi}\right]^2 . \tag{1.22}$$

The power $P_n(0)$ at the carrier frequency f_0 may be expressed in terms of the average transmitted power P_a by using (1.21) and applying L'Hospital's rule to (1.22) to obtain

$$P_n(0) = \left(\frac{\tau}{T}\right)P_a , \tag{1.23}$$

where P_a is given by (1.21). By substituting (1.21) into (1.23), the dc component (for $n = 0$) is

$$P_n(0) = \left(\frac{\tau}{T}\right)^2 P_t . \tag{1.24}$$

When $N \neq 0$, the maximum power in the Nth spectral lobe is given by the relation (Reference 4, p. 25)

$$P_N(N) \approx \left(\frac{\tau}{T}\right)^2 P_t\left[\frac{2}{\pi(2N+1)}\right]^2 , \tag{1.25}$$

and when $2N \gg 1$,

$$P_N(N) \approx \left(\frac{\tau}{T}\right)^2 P_t\left(\frac{1}{\pi N}\right)^2 . \tag{1.26}$$

A typical pulse radar has the following characteristics:

$P_t = 10^6$ W (1 MW) peak power ,

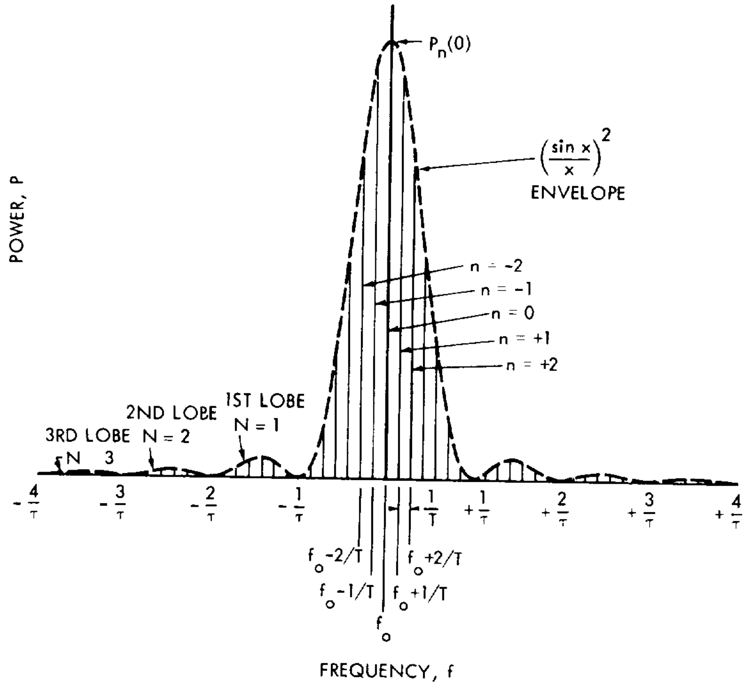


Figure 1.6.—Fourier power spectrum of an ideal pulse.

$$f_0 = 5500 \text{ MHz},$$

$$\tau = 0.25\ \mu\text{s pulse width},$$

and

$$T = 1/640 \text{ second}.$$

A typical problem is to determine the level of off-band interference at a frequency 100 MHz from f_0. With an ideal transmitter pulse and with $N = 100/[1/0.25 \times 10^{-6}] = 25$, the peak power level of the resulting 25th lobe is determined from (1.25) as

$$P_N(25) = \left(\frac{0.25 \times 10^{-6}}{1/640}\right)^2 \times 10^6 \times \left[\frac{2}{3.14(2 \times 25 + 1)}\right]^2$$

$$= 4.0 \times 10^{-6} \text{ W}$$

$$= -24 \text{ dBm}.$$

Communication receivers will normally respond to an interference of this level.

5. Reduction of Sideband Splatter

Reduction of sideband splatter requires that attention be given to the modulator to ensure linear variation of the modulation parameter (the amplitude or angle). In AM and related systems, balanced modulators will be useful in eliminating the sideband components arising from even-harmonic distortion of the modulating signal. To minimize splatter in such systems it is generally necessary to use modulators that act as square-law devices as nearly as possible. Perfect rectifier modulators with very large RF driving voltage can also be shown to give ideal linear modulation. For modulators operating at low levels it should be possible to come reasonably close to the ideal situation.

In systems of angle modulation, where the angle or frequency deviation is obtained by varying a reactive component in an oscillator or where frequency deviation is obtained by the indirect method discussed in Section 2.2, linearity is better for small initial phase deviation. Large values of initial frequency deviation will give rise to nonlinear modulation unless predistortion or feedback is used. It would appear that the methods of angle modulation in which a carrier is modulated by a pulse-position signal developed from the information waveform (Reference 5, p. 388) can be made linear over a wide deviation.

In addition to these methods of controlling spectral broadening at the source, filters will be useful for suppressing undesired components after they are generated. When the modulation process is carried out at low level, filtering can be very effective. There will usually be a number of tuned circuits between the output circuit of the modulator and that of the final amplifier. If we assume that all tuned circuits are parallel-RLC resonant circuits, then the amplitude response of each relative to its response at the center frequency is given approximately by

$$H(f) \approx \frac{B}{\left[B^2 + 4(f - f_0)^2\right]^{1/2}}, \tag{1.27}$$

where $B = 1/2\pi RC$ is the half-power bandwidth at which the response is 3 dB below that at the resonant frequency, and $f_0 = 1/2\pi\sqrt{LC}$ is the center or resonant frequency.

If the splatter frequency f is separated from the resonant frequency by several half-power bandwidths, (1.27) can be reduced to

$$H(f) \approx \frac{B}{2|f - f_0|} \tag{1.28}$$

If there are k stages, each with identical tuned circuits, the relative frequency response at the transmitter output is

$$H_k(f) \approx \left[\frac{B}{2|f-f_0|}\right]^k . \tag{1.29}$$

For instance, if a command transmitter operating at 148 MHz has three tuned circuits following the modulator each of 2-MHz bandwidth, the effect of the tuned circuits is to reduce any splatter components at 136 MHz by a factor

$$H_3(f) \approx \left[\frac{2}{2(148-136)}\right]^3 \approx \frac{1}{1728}$$

(corresponding to approximately −65 dB). Where high-level modulation is used there will be fewer tuned circuits to attenuate the splatter components unless special attention is given to this matter. In the example given above, note that a single tuned circuit attenuates any undesired 136-MHz sideband by only 21.6 dB.

To complete the estimate of the magnitude of undesired splatter that would reach a receiver requires knowledge of the power level of the splatter component at the point of generation and the attenuation in transmission. The former was discussed earlier in this section. While the effect of unwanted splatter may be calculated in principle it is better practice to measure the levels at the point of generation, especially for components far from the center frequency. In this way the elusive parameters which are required for calculation do not have to be determined explicitly; to determine them would require measurement, too. As shown in the following example, rough estimates using representative figures would be useful in preliminary work, however.

Suppose a rectangular RF pulse centered at f_0 is generated as described earlier. The pulse has a spectrum given by

$$X(f) = \frac{1}{2}[X_r(f-f_0) + X_r(f+f_0)] , \tag{1.30}$$

where $X_r(f)$ is given in (1.18). The energy of the pulse is

$$E = \frac{A_r^2 \tau_r}{2} . \tag{1.31}$$

If a narrow-band receiver of bandwidth* B is tuned to a frequency f_c in the vicinity of one of the peaks far from the center of the spectrum, we may approximate the magnitude of the spectrum by

$$|X(f_c)| \approx \frac{1}{2} \frac{A_r}{\pi |f_c - f_0|} . \tag{1.32}$$

*As used here, B is the effective noise bandwidth as defined in (1.47). A receiver with rectangular bandpass of width B has an effective noise bandwidth equal to B.

The energy of the pulse components selected by the receiver is

$$E_r = \frac{1}{2}\frac{A_r^2 B}{\pi^2 [|f_c - f_0|]^2}. \tag{1.33}$$

This energy relative to the total energy in the pulse is

$$\frac{E_r}{E} = \frac{B}{2\pi^2 \tau_r [|f_c - f_0|]^2}. \tag{1.34}$$

If, for instance, $B = 0.25$ MHz, $\tau_r = 10^{-6}$ second, and $|f_c - f_0| = 12$ MHz, then

$$\frac{E_r}{E} = \frac{1}{8\pi^2 (12)^2} \approx 0.875 \times 10^{-4},$$

or

$$\frac{E_r}{E}\,(\mathrm{dB}) = 10 \log (0.875 \times 10^{-4}) = -40.6\ \mathrm{dB};$$

that is, the unwanted sideband energy is about 41 dB below the total energy contained in the pulse.

If the rectangular pulses were fed through a succession of three single-tuned circuits prior to transmission, as described above, and if the tuned circuits produce no significant effect on the total pulse power, the sideband components in the 0.25 MHz band would be 65 dB + 41 dB = 106 dB below the pulse power transmitted.

At this point it is well to point out that, in effect, we are shaping the pulse in the filters to reduce the unwanted sidebands.

Finally, the path attenuation must be taken in account. This factor is not considered here except to point out that the problem of splatter is important when transmitter and receiver antennas are relatively close, or when they are, in fact, the same antenna. In the latter case, separation of functions is accomplished in a diplexer and one must determine the degree of isolation afforded by this device.

2.2 GENERATION AND SUPPRESSION OF HARMONICS

Harmonic generation is also associated with nonlinearity. The effect is a generation of a new frequency which is an integral multiple of the original frequency. When the original frequency is that of a modulated carrier, the sidebands may remain unchanged or they may become distorted. In some devices the nonlinearity is incidental to normal linear amplification as in class A tube and transistor amplifiers. In other devices nonlinearity is a consequence of linear amplification by impulsive re-enforcement of a wave. RF amplifiers, of the class B or C type, pulse a tank circuit throughout a portion of the sinusoidal cycle and are therefore in this category. Klystrons and magnetrons are also in this class. Oscillators continue to build up until the output is limited by saturation and cutoff of the active element; hence, they also function in this manner. In other cases the harmonics are an unwanted byproduct of a desired nonlinear function. It was pointed out earlier that the modulator may distort the modulating signal; it may also distort the RF signal. The ideal frequency multiplier will yield only the desired harmonic but this is unusual. Many unwanted harmonics will be present as well as the original input which becomes an unwanted component.

It will often be possible to find modes of operation for the active devices which minimize harmonic amplitudes, but this occurs usually at the cost of efficiency. The extent to which such efforts are successful depends on the amount of filtering used. Low-level circuits in the transmitter, incorporating many frequency selective circuits, are unlikely to result in significant harmonic output from the final stage. Some mechanisms for minimizing harmonic content will now be described in more detail.

As pointed out earlier, modulators for AM signals and signals related to AM are multipliers. The modulation function is performed perfectly when the sum of the low-frequency signal and a carrier is delivered to a square-law device. If $x_1(t)$ is the low-frequency signal and $\cos \omega_c t$ is the carrier, then the output of a square-law device is the squared sum of these input signals:

$$[x_1(t) + \cos \omega_c t]^2 = x_1^2(t) + 2x_1(t) \cos \omega_c t + \frac{1}{2} + \frac{1}{2} \cos 2\omega_c t. \quad (1.35)$$

The desired term is $2x_1(t) \cos \omega_c t$. The term $x_1^2(t)$ is a distorted low-frequency signal which is readily rejected in the RF filter following the modulator; the dc term, 1/2, is obviously also rejected. Some energy at the second harmonic of the carrier may trickle through, especially if the output circuit is not highly selective. The use of a properly designed balanced modulator shown in Figure 1.7 will eliminate transmission of second harmonics. Assume that each nonlinear device has an output-input characteristic given by the power series

$$y = a_0 + a_1 x + a_2 x^2, \quad (1.36)$$

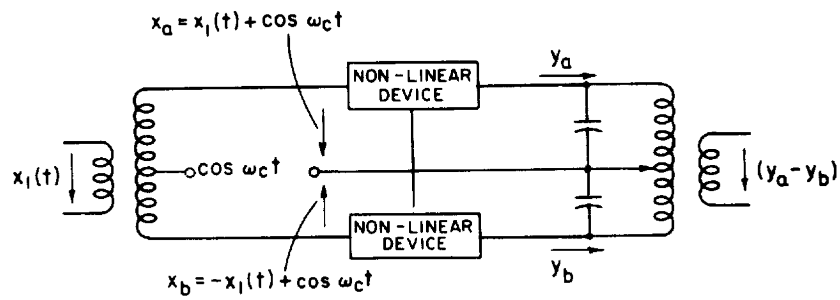


Figure 1.7.–Schematic diagram of a balanced modulator.

where y is the output and x is the input. When x_a and x_b have values shown in Figure 1.7, the total output may be written

$$\begin{aligned} y_a - y_b &= a_1(x_a - x_b) + a_2\left(x_a^2 - x_b^2\right) \\ &= 2a_1x_1(t) + a_2[x_1(t) + \cos\omega_c t]^2 - a_2[x_1(t) - \cos\omega_c t]^2 \\ &= 2a_1x_1(t) + 4a_2x_1(t)\cos\omega_c t\,. \end{aligned} \tag{1.37}$$

No second harmonic term is present, assuming perfect balance, and the term involving $x_1(t)$ alone* is rejected in the output tank circuit. This arrangement will, however, leave harmonics of odd order in the output if the degree of nonlinearity is not limited to square law. If, for instance, a term a_3x^3 were added to (1.36), the output would contain a term

$$2a_3\cos^3\omega_c t = \frac{3}{2}a_3\cos\omega_c t + \frac{1}{2}a_3\cos 3\omega_c t \tag{1.38}$$

which has a third harmonic component. [The rejection of splatter components involving even powers of $x_1(t)$ can be illustrated also by adding a_3x^3 and expanding as in (1.37).]

It is common practice to use crystal controlled oscillators operating at moderate frequencies (i.e., 30 MHz and below) and to employ frequency multiplication to obtain the needed output carrier frequency. In such systems electronic devices biased at or below cutoff (e.g., in class B and C amplifiers) are the source of the carrier output, which is rich in harmonics. The current

*Note that $x_1(t)$ is a low-frequency term; however, $x_1(t)\cos\omega_c t$ is a modulated RF signal.

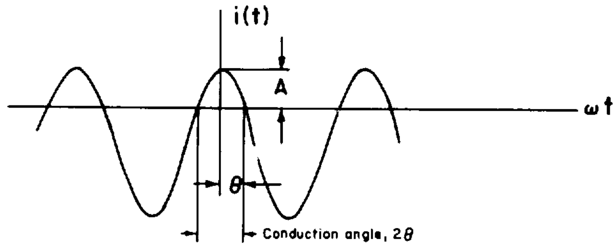


Figure 1.8.–Cosinusoidal cap waveform.

pulse is approximately of the form of the cap of a sinusoid as shown in Figure 1.8 and given by the expression

$$i(t) = \begin{cases} a[\cos \omega t - \cos \theta] , & \text{for } (2n\pi - \theta) \leqslant \omega t \leqslant (2n\pi + \theta) , \\ & n = 0, \pm 1, \pm 2, \ldots , \\ 0 , & \text{otherwise} . \end{cases} \tag{1.39}$$

The Fourier expansion of such a wave is given by

$$i(t) = \frac{A}{\pi(1 - \cos \theta)} \left\{ \sin \theta - \theta \cos \theta + \sum_{n=1}^{\infty} \left[\frac{\sin (n+1)\theta}{n+1} + \frac{\sin (n-1)\theta}{n-1} + \frac{2 \sin n\theta \cos \theta}{n} \right] \cos n\omega t \right\} . \tag{1.40}$$

It will often be found that the sinusoidal cap is itself somewhat distorted in passing through the amplifier, so that higher levels of harmonics appear. Squared sinusoidal caps, given by the expression

$$i(t) = \begin{cases} A^2 [\cos \omega t - \cos \theta]^2 , & \text{for } (2n\pi - \theta) \leqslant \omega t \leqslant (2n\pi + \theta) , \\ & n = 0, \pm 1, \pm 2, \ldots , \\ 0 , & \text{otherwise} , \end{cases} \tag{1.41}$$

may be a better approximation of the magnitude of current pulses from some electronic devices than the undistorted caps illustrated in Figure 1.8 (Reference 10). The relative magnitudes of the first three harmonics for the undistorted and square sinusoidal caps are shown in Figure 1.9 as a function of conduction angle 2θ. In general, the larger the conduction angle, the smaller will be the amplitude of the harmonic in the output.

It is, in principle, possible to estimate the harmonic output of the transmitter using the levels determined by (1.40) or Figure 1.9 together with the response characteristic of the tuned circuits following the source of harmonics. Such a procedure is described in Reference 7, but the equations of Section 2.1 provide the necessary forms. As a rough approximation, a single-tuned circuit with a Q=10 will attenuate the second, third, and fourth harmonics approximately 24, 30, and 33 dB respectively. Doubling the Q will increase the attenuation in each case by about 6 dB.

In the usual case, the final amplifier will contribute the major amount of harmonic output. The reasons for this are that ordinarily the final amplifier is driven hard in order to get as much efficiency as is possible and the tuned circuits following the final amplifier have limited selectivity, particularly at the lower frequencies.

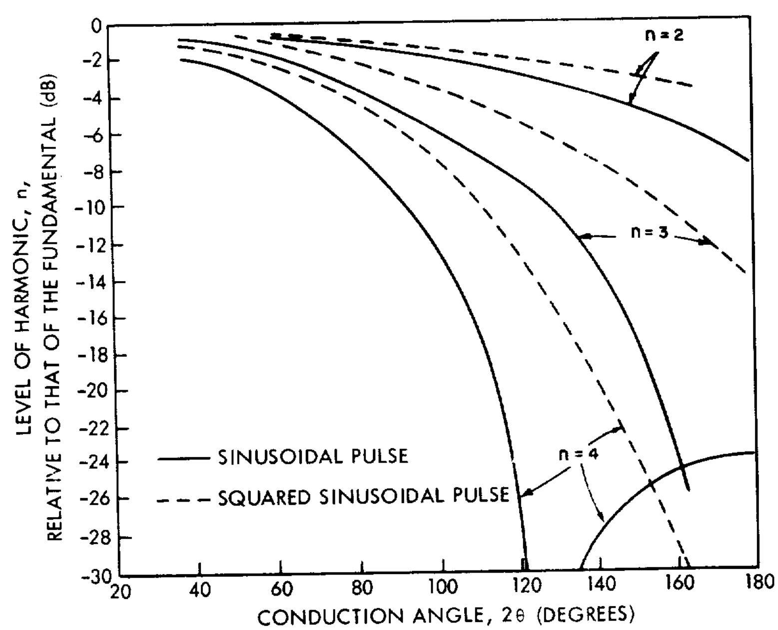


Figure 1.9.—Relative intensity of harmonics for two pulse shapes as a function of conduction angle.

As a rule tube-type final amplifiers are operated as class C amplifiers. In principle, the harmonic output from class C amplifiers is not different from that of the frequency multipliers discussed above. As the conduction angle is increased, the harmonic output decreases, but so does the efficiency. To reduce interference and obtain the effect of a linear amplifier, it is obviously advantageous to use a push-pull class B final amplifier. It may even be advantageous to use class C amplifiers in push-pull since this would tend to cancel the even harmonics. The amplitudes of odd harmonics relative to that of the fundamental would be unaffected by tubes in push-pull, but these harmonics are readily attenuated by tuned circuits in the output.

The harmonic outputs of 14 representative communication transmitters are shown in Figure 1.10 (Reference 11). The legend of Figure 1.10 gives the FCC limits for harmonic output. Note that some transmitters produce harmonics whose levels are within 30 dB of that of the fundamental. For such harmonic content, a 50-kW transmitter will emit a 50-W signal at the harmonic frequency—in this case hardly insignificant.

Amplifiers for microwave frequencies also generate significant harmonic output, though the mechanisms are different. The klystron, which is an important high-power amplifying device, is inherently a source of harmonics. The bunching process develops concentrations of moving charge which pass the catcher in the form of sharp bursts. The current induced in the catcher cavity is highly nonsinusoidal. It has been shown (Reference 12) that the induced current has a harmonic intensity given approximately by

$$I_c(n) = 2I_0 J_n(nx), \tag{1.42}$$

where

n = the order of the harmonic,

I_0 = the direct current in the tube,

J_n = the nth order Bessel function of the first kind,

and

$$x = \frac{\omega s V_1}{(e/m)^{1/2}(2V_0)^{3/2}}, \text{ bunching parameter,}$$

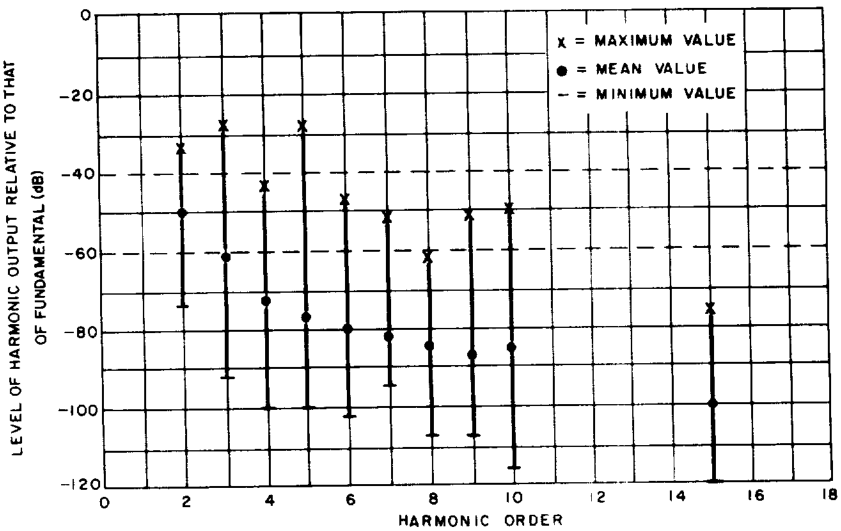
in which

ω = frequency of excitation,

s = spacing between input and output cavities,

V_1 = peak gap voltage at buncher,

e = electronic charge,



FCC LIMITS ON HARMONIC EMISSIONS			
FREQUENCY RANGE	LEVEL OF HARMONIC RELATIVE TO THAT OF FUNDAMENTAL	MAXIMUM POWER OF EMITTED HARMONIC	FOR TRANSMITTER POWER
10 kHz to 30 MHz	-40 dB	50 mW	
30 MHz to 235 MHz	-60 dB	1 mW	$P_o > 25$ W
30 MHz to 235 MHz	-40 dB	25 μW	$P_o < 25$ W

Figure 1.10.–Typical values of harmonic emissions determined from measurements on 14 radio transmitters.

m = electronic mass,

and

V_0 = beam accelerating voltage.

The bunching parameter value of $x = 1.84$ yields maximum efficiency. At maximum efficiency, the pulses of induced current are very nonsinusoidal and the harmonic-current content is very high. To avoid large harmonic outputs, the cavity Q must be high.

For loaded circuits, values of Q ranging from 500 to 1000 are obtainable. Using such values of Q, it should be possible to attenuate harmonic outputs more than 60 dB below the fundamental, even with equal fundamental and second harmonic current in the tube. Measurements reported on a particular pulsed klystron (Reference 13) are reproduced in Figure 1.11. Variation of beam voltage, which alters the bunching parameters, is seen to alter the second-harmonic content markedly. However, at the value of beam voltage for which the second harmonic is minimum, the second harmonic power is only about 40 dB below the fundamental output.

The traveling wave tube (TWT) appears to be a good choice as a microwave power amplifier, generating little interference. The operation of the TWT depends upon uniform re-enforçement of a traveling wave and does not involve impulsive re-enforcement of a field, as in a klystron. Yet, reports (Reference 12) indicate that, in particular cases, the second-harmonic output in TWT's may range from 20 to 40 dB below the fundamental. It is possible that these conditions exist only at maximum efficiency when the tube is driven heavily. The input-output characteristic for specific traveling wave tubes was found to be linear over a given range of inputs; beyond the linear range, saturation was slowly reached. Operating the tube into its region of saturation gives rise to maximum output; but, unfortunately, the harmonic content also increases.

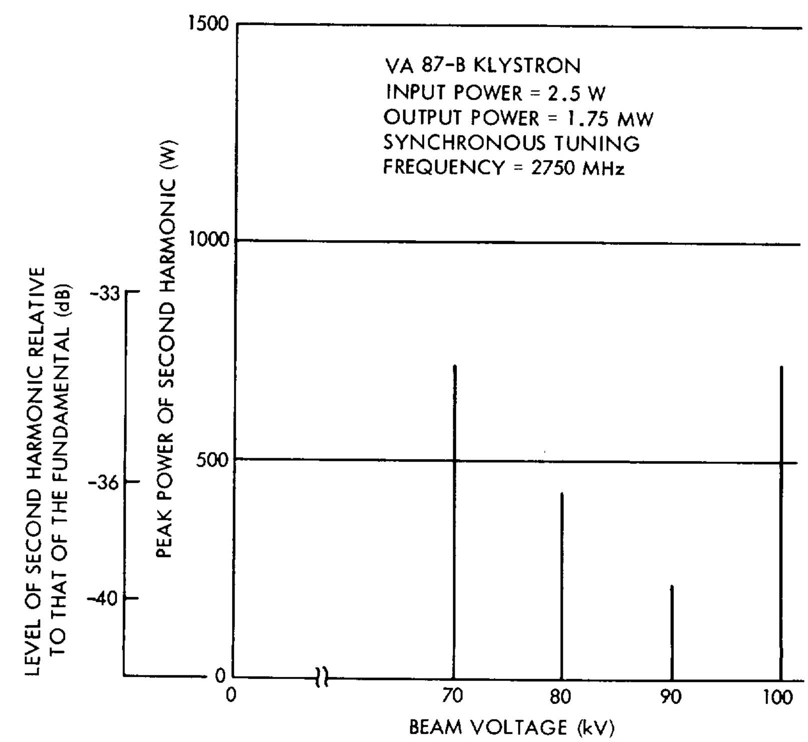


Figure 1.11.–Harmonic output as a function of beam voltage for VA 87-B Klystron.

High-power magnetrons are generally used as sources of pulsed RF energy, coupled directly to the antenna. In the magnetron the energy of a rotating cloud of electrons is imparted to the field in tuned cavities on the periphery of rotation. The mechanism which selects electrons emitted from the cathode, where they are roughly in proper phase, tends to maintain them in bunches as in the klystron. The spatial distribution of charge is not sinusoidal, and, again, pulses rather than sinusoids are induced in the cavities. The magnetron cavities can also be excited in a harmonic mode so that a large number of output components is possible in addition to the easily identifiable harmonics. Harmonic level and frequency are both difficult to predict. A value of -40 dB with respect to the fundamental has been quoted (Reference 13) for the second harmonic, but the third harmonic was said to reach -20 dB.

From the foregoing, it is evident that harmonic intensities can be reduced by operating oscillators and amplifiers over linear regions at reduced efficiencies, canceling nonlinear components in balanced circuits, and filtering. The first of these involves an increase in power consumption and heat generation. As a result, more effort must be spent on heat removal. Balanced circuits and filtering involve additional circuitry; the balanced push-pull circuit eliminates only even-order harmonics. Filtering is, by far, the most practical method for suppressing harmonics; techniques for suppressing harmonics are discussed below.

One way of suppressing harmonics is by utilizing the wave trap or bypass principle. In the case of a transmission line system, the filter elements may take the form of shunt or series stubs. One application of such a filter to remove second harmonics is shown in Figure 1.12.

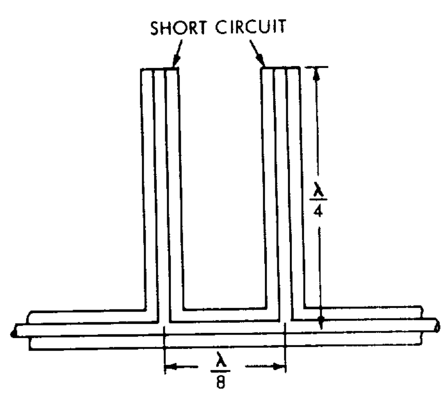


Figure 1.12.–Diagram of two short-circuited stubs (each one-quarter-wavelength long and spaced one-eighth wavelength at the fundamental frequency) arranged to attenuate the second harmonic.

The difficulty with wave-trap techniques is that power at the harmonic frequency is reflected back to the generator, causing an impedance mismatch. In theory, one way of avoiding this difficulty is to use an isolator between the generator and the filter. Unfortunately, most isolators are designed to operate at the fundamental frequency, and their performance at harmonic frequencies is usually unknown. This is especially true at the higher microwave frequencies. If an isolator is used, it must be able to dissipate the power of all reflected components (fundamental plus harmonics) without exceeding its rating.

It is possible to utilize a ferrite circulator to suppress harmonics. For example, the configuration in Figure 1.13 could be used to divert the reflected energy into a resistive load. Unfortunately, the circulator has bandwidth problems as does the isolator. Another harmonic filtering-absorbing technique directionally couples the unwanted signals into a second waveguide, where they are absorbed.

It should be pointed out that a simple method for harmonic absorption, which also affords some reduction of the level of harmonics in the main load, can be achieved merely by coupling to the main waveguide a smaller waveguide whose cutoff frequency is below the fundamental but above that of all harmonics.

Additional filtering techniques are given in References 1 and 14.

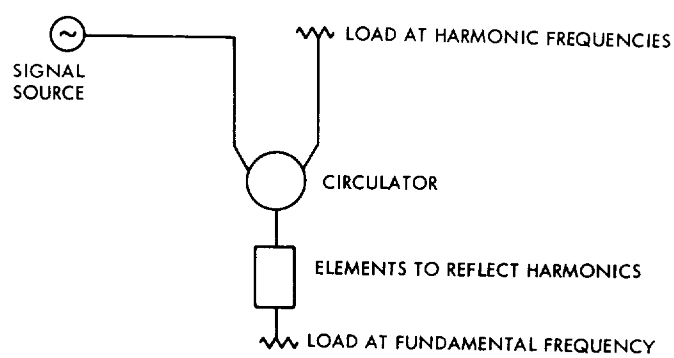


Figure 1.13.–Diagram illustrating the use of a circulator to separate harmonics.

2.3 TRANSMITTER NOISE

The spurious sidebands discussed in Section 2.1 can be reduced through careful design of the modulator, but it is frequently found that a significant background interference level still exists around the transmitter carrier center frequency. This results from oscillator noise modulating the carrier in amplitude, frequency, or phase. The noise is mainly associated with the oscillator itself, which is quite noisy compared to amplifiers, but some noise is produced in the amplifiers as well. It is possible for corona to form, or for arcing to occur, at high-voltage points in the final amplifier. Subsequent filtering in the output tank circuit results in a pair of noise sidebands on either side of the carrier frequency; spurious levels are difficult to estimate accurately.

Measurements of noise power are more appropriate though they are difficult to make because of the presence of the much larger power in the carrier. Some measurements have been reported on VHF transmitters (Reference 15), and the results of tests on a 45-MHz and a 160-MHz transmitter are shown in Figure 1.14. The basic crystal frequencies were 3.75 MHz and 6.67 MHz, respectively, and three or four frequency-multiplier stages were utilized before the final amplifier. The tests were performed on an unmodulated transmitter, but the effect of modulation on the noise sidebands was not determined. Data, not presented here, were also obtained for the power amplifier alone, without the exciter and its multipliers; the resulting noise was generally about 20 dB below that shown in Figure 1.14. The noise levels are quite low in these cases, but the noise interference injected into a receiver can be quite significant when transmitter and receiver antennas are very close to each other.

2.4 INTERMODULATION AND CROSSMODULATION

The phenomena of intermodulation and crossmodulation imply the mixing of two or more signals in a nonlinear element in such a way that multiplicative mixtures of the two signals result. This occurs in receiver input circuits, as well as in transmitters, and sometimes in a nonlinear element in the channel. As far as transmitters are concerned, the process involves the reception of an unwanted signal by the transmitting antenna, which conducts it back to the output plate of the final amplifier, where it is mixed with the transmitted signal. The process is therefore of greatest significance when both the unwanted signal and the nonlinear product are within the passband of the final amplifier. This can occur if the nonlinear device has a characteristic of odd degree. For instance, suppose the output signal has a component that depends on the third power of the output voltage. Two components of voltage at the output are the output signal voltage itself,

$$x_1(t) = v_1(t) \cos [\omega_1 t + \phi_1(t)] , \tag{1.43}$$

and the unwanted signal voltage,

$$x_2(t) = v_2(t) \cos [\omega_2 t + \phi_2(t)] . \tag{1.44}$$

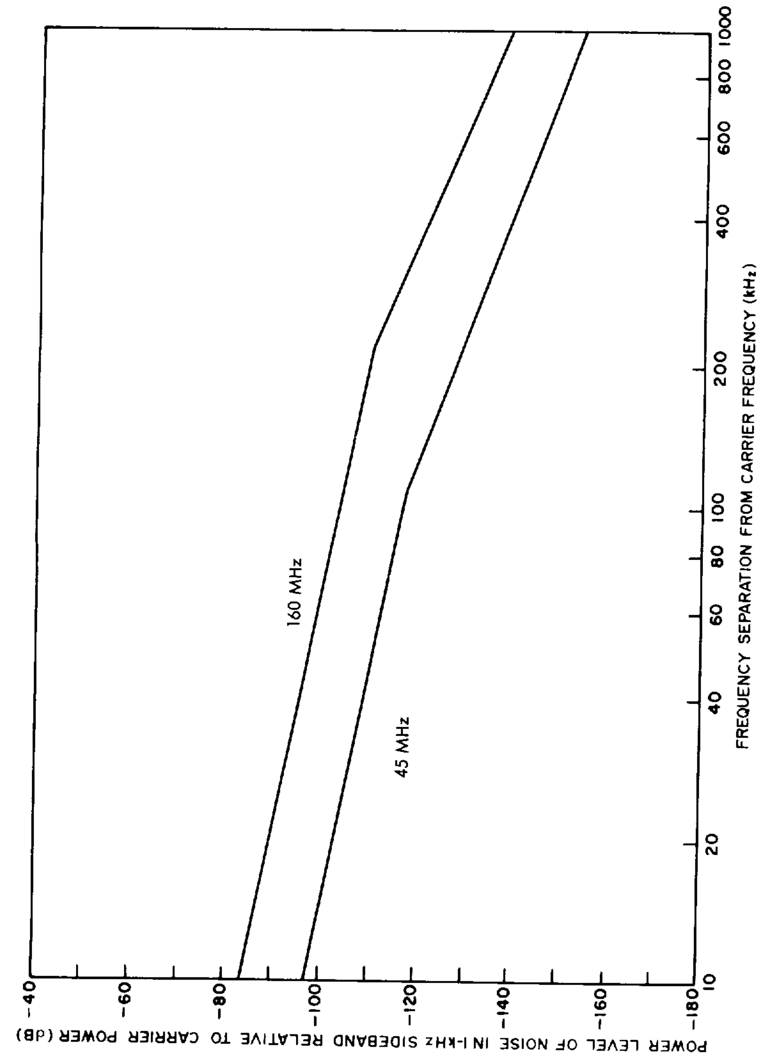


Figure 1.14.–Measured values of the sideband noise level of unmodulated 45- and 160-MHz transmitters as a function of frequency separation from the carrier frequency.

The sum $x_1(t) + x_2(t)$, when applied to a nonlinear device having a cubic term, will result in an intermodulation component,

$$y_a(t) = v_1^2(t) \cos \left[(2\omega_1 - \omega_2)t + 2\phi_1(t) - \phi_2(t)\right], \tag{1.45}$$

and a crossmodulation component,

$$y_b(t) = v_1^2(t) v_2(t) [\cos \omega_2 t + \phi_2(t)] . \tag{1.46}$$

The center frequency of the intermodulation component differs from either ω_1 or ω_2; but, if the two frequencies are close to one another, the frequency of the new term is not too different from ω_1. The modulation on the term whose frequency is $(2\omega_1 - \omega_2)$ is a combination of the modulations on each of the causative signals. The crossmodulation component has a frequency equal to one or the other of the original signals, and its amplitude is a combination of the original amplitude coefficients of both mixed signals. The intermodulation term is usually more serious since it occurs in an entirely new frequency band. The crossmodulation term, given in (1.46), implies that the unwanted signal is re-radiated with a distorted envelope. As a rule, the amplitude of the re-radiated signal is much smaller than that of the original unwanted signal, defined by (1.44), and the re-radiated signal might be viewed as a small new splatter or noise component.

2.5 OTHER SPURIOUS OUTPUTS

Parasitic oscillations are known in both low- and high-frequency amplifiers. In low-frequency devices they result from stray external and internal capacitances and inductances that form spurious feedback loops. The cures are often very simple, sometimes involving inserting resistors in the leads to the grid and plate of the amplifier to increase losses of the spurious tuned circuit. The methods of handling these situations are well described in standard reference books (Reference 16).

At microwave frequencies, corresponding phenomena exist. Often, the phenomena are completely internal to the tube. At some frequencies, transit-time effects are sources of negative resistance. If a tuned circuit somewhere in the structure is coupled to the negative resistance, parasitic oscillations occur. Various mechanisms for eliminating parasitic oscillations are described more fully in Reference 10.

3. UNINTENTIONAL NOISE GENERATION

Man-made electrical noise reaches significant proportions at VHF and UHF. Unwanted noise may be generated by any of the following: vehicle ignition systems, corona discharge and leakage from high power transmission lines, rotating electrical machinery, switching devices, rectifiers, arc welders, discharge lamps, industrial heating equipment, and medical and diathermy devices.

The majority of these noise sources are naturally concentrated in urban areas of population and industrial centers. Locating ground stations away from urban population centers eliminates most of the noise originating from such sources. However, there is usually vehicular traffic, as well as power lines, in the immediate vicinity of a ground station. Consequently, noise from these sources is of greater significance to station operations.

Noise sources located on the ground do not significantly affect the fixed interferometer antenna arrays used by NASA Minitrack interferometers because the main beam of these arrays is fixed and points vertically. Of course, this will not be true for any of the mechanically movable antennas such as the satellite automatic tracking antenna (SATAN) array, located at Minitrack sites, or the 85-ft parabolic dish located at data acquisition facilities. These antennas may be affected by noise sources on the ground when the mainlobe, backlobe, or sidelobes are pointed at the horizon. However, the energy entering the antenna sidelobes is lower than that collected by the mainlobe or backlobe.

3.1 NOISE FROM VEHICLE IGNITION SYSTEMS

Automobile and truck ignition systems undoubtedly make the largest contribution to the overall man-made noise level at VHF and UHF. A number of sources (References 17 through 25) describe quantitative measurements on the noise levels radiated by automobiles and trucks; these sources describe comprehensive theories for the explanation of radiation from ignition systems. Results of measurements made by George (Reference 19) at 180 MHz and 450 MHz are reproduced in Figures 1.15 and 1.16. Although these data were obtained in 1940, they are still applicable to modern-day vehicles.

For example, recent measurements reported by Rosa (Reference 25) determined average levels of peak ignition noise for the noisiest 20 percent of over 3000 vehicles measured. Rosa reported that one out of five vehicles

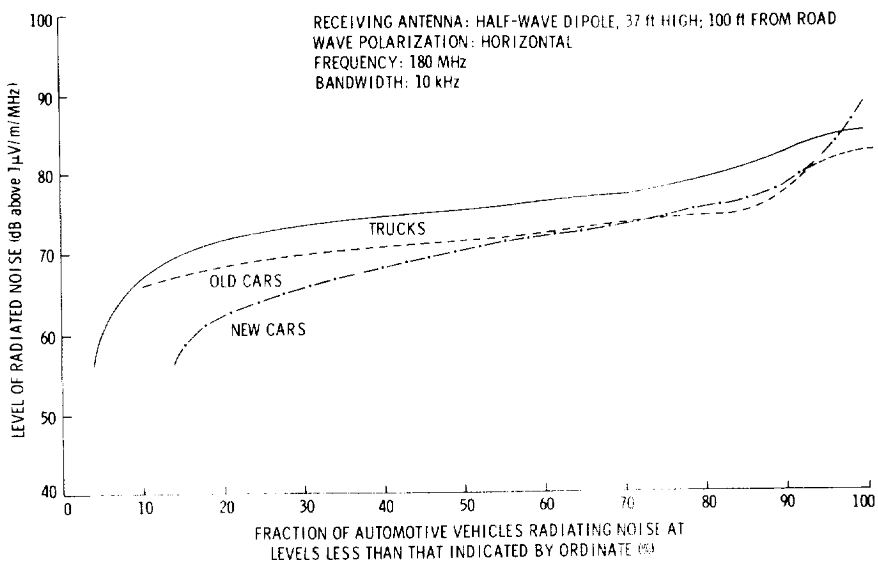


Figure 1.15.—Levels of noise radiated at 180 MHz by vehicular traffic ignition.

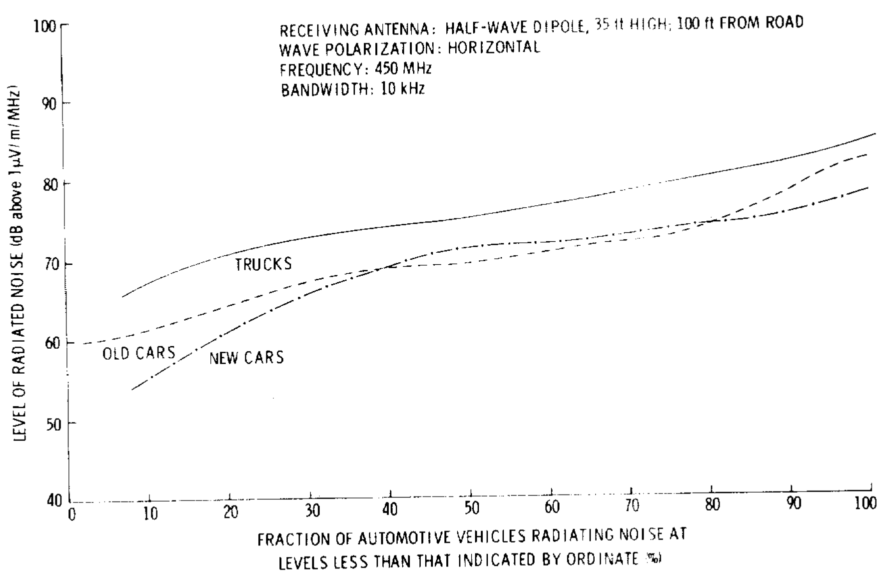


Figure 1.16.—Levels of noise radiated at 450 MHz by vehicular traffic ignition.

radiated a constant level of 84 dB above 1 μV/m/MHz from 40 to 300 MHz, almost fully covering the entire VHF band. By way of comparison, the upper 20 percent of vehicles in Figures 1.15 and 1.16 averaged about 80 dB above 1 μV/m/MHz, which closely agrees with Rosa's data.

Automobile ignition noise is, in general, impulsive in nature. As a result, the peak voltage response of a receiver to ignition noise [see Equation (1.52)] will vary in a manner directly proportional to the receiver predetection bandwidth. For instance, the data in Figures 1.15 and 1.16 may be translated to other receiver bandwidths from the relationship

$$E_x = E_r + 20 \log (\Delta f),$$

where

E_x = new effective radiated ignition noise level in dB above 1 μV/m/unit-bandwidth,

E_r = level of radiated ignition noise given in Figures 1.15 and 1.16, in dB above 1 μV/m/MHz,

and

Δf = receiver bandwidth in megahertz.

For example, in Figure 1.16, the 80-percent point for trucks corresponds to E_r = 80 dB above 1 μV/m/MHz. For a receiver bandwidth Δf = 0.1 MHz, the new effective radiated ignition noise level becomes

$$\begin{aligned} E_x &= 80 + 20 \log(0.1) \\ &= 60 \text{ dB above } 1\ \mu\text{V/m/100-kHz bandwidth}. \end{aligned}$$

From George's data, noise emanating from trucks is seen to be the most serious problem, and this is still true today. Figure 1.17 shows the theoretical propagation curves for horizontally polarized waves over level ground. Such curves serve as the basis for extending data of Figures 1.15 and 1.16 to larger distances. The field strength (in microvolts per meter) expected at a given distance d (that is, d = 100 ft for George's data) is obtained by multiplying the measured field strength at a known distance by the factor K. In practice, noise levels will probably fall off with distance at a faster rate than is shown in Figure 1.17, because of absorption and screening by vegetation, foliage, and trees.

Rosa (Reference 25) reported the space attenuation for ignition noise for distances up to 1000 ft from a highway. For instance, at 10, 70, 160, 200, and 300 MHz, the space attenuation A_s, in dB, for a distance D from a fixed reference point is determined according to the relationship

$$A_s = 10 \log(1/D^3).$$

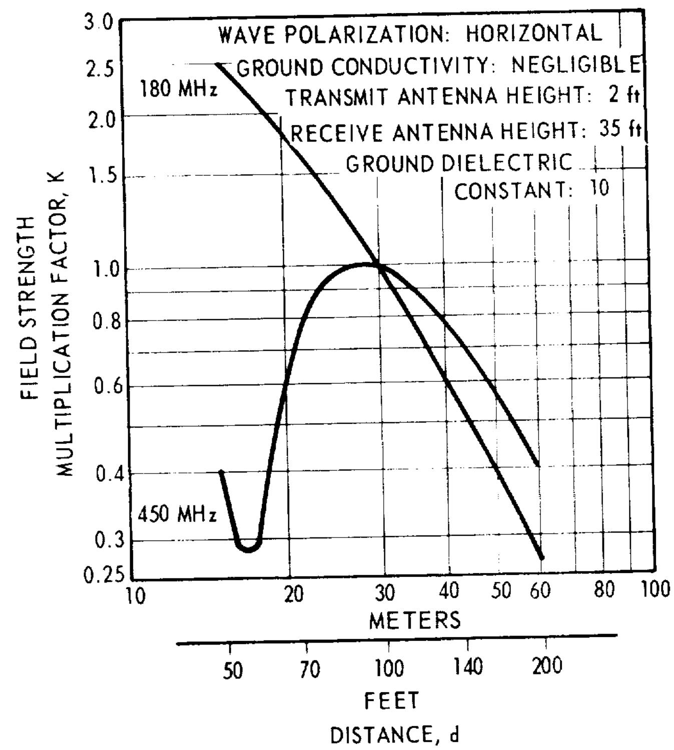


Figure 1.17.–Theoretical 180- and 450-MHz propagation field strength over level ground for horizontally polarized waves.

By way of comparison, this space attenuation increases at a greater rate than the 6 dB per octave rate ($1/D^2$) for free-space propagation.

In addition to the distance factor in assessing the noisiness of a station site due to ignition noise from passing vehicles, the frequency with which vehicles pass by the site must be taken into account. Consequently, the traffic density around a station site must also be considered when determining the degrading effects of ignition noise on the received data.

Ignition noise is created by the oscillatory current that is set up in the high-tension leads when the energy stored in the self-capacitance of the spark plug, the high-tension cables, and the ignition coil, is rapidly discharged through the spark plug gap. The rapidly changing current then radiates energy from the high-tension cables and also from coupled low-voltage circuits. This current, called the "capacitive component" of the ignition current, has an oscillatory wave shape and lasts for only a short time (of the order of 4 ns for a 1-mm spark gap). This creates a continuous frequency spectrum distributed throughout a very wide part of the VHF/UHF spectrum. The amplitude of the spectrum remains relatively flat up to at least 600 MHz, although some energy is usually still measurable at 1000 MHz. The spectral energy distribution radiated by an automobile will depend on the duration of each spark discharge, the layout of the electrical equipment and wiring within the engine compartment, and the shielding effectiveness of the hood and body frame that surrounds the engine. For this reason, the levels of interfering noise radiated by an ignition system vary widely with different makes and models of automobiles and it is very difficult to accurately predict the effective noise levels from a particular vehicle. The hood covering the automobile engine has been found to be especially important in the reduction of ignition interference.

A number of relatively effective means of suppressing ignition interference have been developed. The most effective method is the complete shielding of the engine compartment of an automobile. However, this method is generally considered to be impractical and too expensive. The most widely used system makes use of suppressor resistors which may be either lumped or distributed. The suppression resistance effectively reduces the amplitude and frequency of the oscillatory current. In the system using lumped resistors, a high-temperature, 10-W carbon resistor of 5 to 15 kΩ is placed close to the distributor end of the high-tension cable between the coil and distributor. Additional suppression can be obtained by the addition of similar resistors at the spark plugs. Special spark plugs available today incorporate the resistor into the body of the plug. In systems using distributed resistance, the metallic conductor in the high-tension leads is replaced by braided textile fibres that have a total resistance of 4 to 10 kΩ. Such cables are more effective than lumped resistors and have now been adopted by all United States automobile manufacturers for use as standard ignition cables.

Unfortunately, although suppressive devices are fitted to nearly all new cars, they frequently are removed during normal servicing and maintenance and generally are not replaced. This practice results from general ignorance of automobile mechanics and also from the fallacy that suppressive devices degrade engine performance. Any vehicles or gasoline engines stationed at or near ground sites should be fitted with effective and adequate suppressive devices. Care must also be taken to see that these devices are not removed during any maintenance work on such vehicles or engines.

3.2 NOISE FROM HIGH-POWER TRANSMISSION LINES

High-voltage transmission lines are a source of broadband noise that is produced by corona discharges at various points along the line. Noise may also be caused by arcing over dirty or wet insulators. The noise propagated by power lines is very quickly attenuated with distance: at a frequency of about 1

MHz, the corona noise from a 200-kV power line is undetectable by a receiver with a half-wave dipole located more than 200 ft from the line. Corona noise levels are relatively significant in the broadcast and HF bands but decrease rapidly at frequencies above about 10 MHz. At frequencies above 100 MHz, the corona noise levels appear to be virtually insignificant. Most of the measurements taken during on-site tests have been at frequencies in the broadcast band and below. These tests have shown that the levels of corona noise can also fluctuate greatly from day to day. Conditions of humidity and wind can cause large and rapid fluctuations of noise levels. These levels also tend to change over long periods of time as the transmission line ages; newly installed lines tend to be noisier than older lines.

The corona phenomenon occurs when the dielectric medium surrounding the high-voltage conductor breaks down and electrical discharges take place. Corona will always emanate from local points of sufficiently high field intensity; that is, from points of surface discontinuity at which a high potential gradient exists. For this reason, corona discharges always take place at the sharpest point on the surface of a conductor. The mechanism of discharge may be explained as follows. Incidental ionization, which is always present, provides a supply of electrons in the vicinity of the conductor. At negatively charged points of high-potential gradient, these electrons are accelerated away from the conductor by the strong field and, in fact, possess sufficient energy to ionize the surrounding air molecules. An avalanche effect then results. However, the region of ionization remains confined to the space surrounding the discharge point. The positive ions created by the discharge are attracted to the negatively charged corona point, thereby reducing the potential gradient and quickly quenching the corona discharge. The discharge current is therefore pulse shaped and lasts only for a short time, ranging between 0.1 and 0.5 μs.

The discharge from a corona point tends to repeat itself such that a continuous series of recurrent pulses is generated. The repetition rate of these pulses is a function of the potential gradient at the point where the corona is formed. Generally, pulse repetition rates are of the order of 1 MHz or greater, depending on the line voltage. Whereas the discharge current from a single corona point is roughly periodic, the total discharge current on a line containing numerous corona spots must be treated as a set of randomly occurring pulses. To this extent, the noise generated by a power line is similar in effect to shot noise.

Corona discharge from points that are at a high positive potential have been found to be somewhat different from those at a negative potential. The potential gradient required for the formation of corona is somewhat higher, and the current is not impulsive in nature. The pulses obtained are of greater amplitude and lower repetition frequency than those for negative corona.

Noise from corona discharges has been investigated (References 26 through 31) in both the laboratory and the field. Because of the large number of variables involved, it is difficult to correlate measurements obtained at one field site with those obtained at another site. Figure 1.18 shows some data

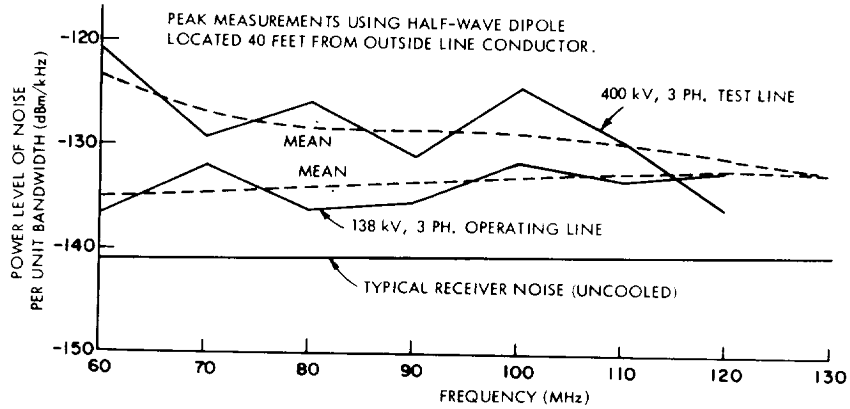


Figure 1.18.–Variation of corona noise level with frequency.

extracted from Reference 28. As can be seen, the power radiated at VHF is very low (only about 10 dB above the receiver threshold level). It should be remembered that these measurements were made with a half-wave dipole antenna located only 40 ft from the power lines; normally, receiving antennas would not be located so close to a power line. As would be expected, the 400-kV line radiates more noise than does the 138-kV line. However, part of the noise measured at these frequencies probably includes other man-made background noise, which is continuously present at VHF and UHF.

No data have been found on measurements of corona noise at 400 MHz; it can probably be concluded that corona noise may be ignored at UHF.

In summary, it may be concluded that corona noise from power lines does not represent a serious problem for aerospace ground-site operations, since the noise levels generated cease to be significant at the frequencies used in telemetry data acquisition.

It should be emphasized that, although corona noise radiated from a good power line is not expected to cause serious problems under normal conditions, a faulty power line can cause serious noise problems. In this case, the noise is usually caused by arcing somewhere along the line and is not due to corona. Such effects occur infrequently and can be prevented by regular maintenance of the power line.

3.3 OTHER SOURCES OF NOISE

The remaining sources of noise, which may exist at ground sites, include fluorescent lamps, electric motors, rectifiers, switching devices, and so forth. It can be safely assumed that all these noise sources will be housed inside the buildings and therefore will not be located in very close proximity to the receiving antennas. None of the sources mentioned radiate noise at very significant levels at VHF or UHF, and the additional attenuation provided by the buildings that surround these sources will probably cause such noise levels to be undetectable at the receiving antennas. Any slight disturbances that are noticed will occur when the antennas point straight at the station buildings. Many potential noise sources, such as rectifiers and switching devices, are found in the antenna control circuits. However, these are presumably housed in metal cabinets within the building, so the overall shielding should be more than adequate.

Of the different noise sources mentioned, the most serious is the fluorescent lamp. These lamps, as is common with all plasma-operated devices, generate radio noise. The noise originating from plasma devices is caused by the presence of periodic fluctuations in the ion or electron density in the area immediately surrounding the electrodes and by the periodic switching of the polarity on ac-operated devices. Figure 1.19, which is taken from Reference

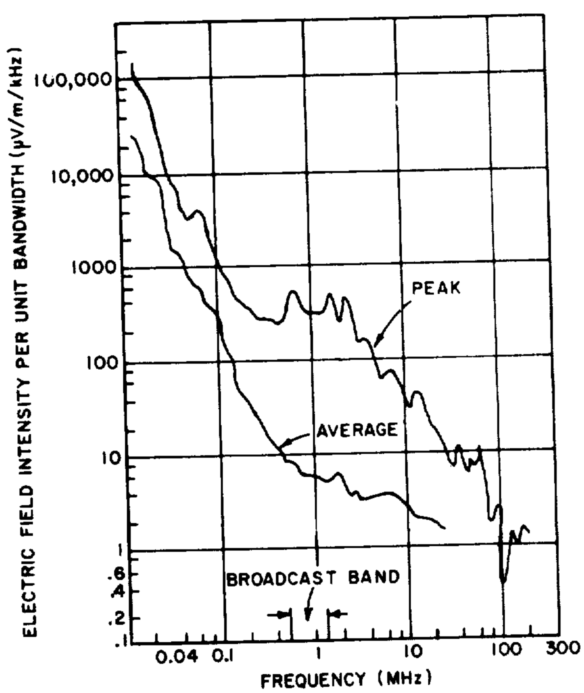


Figure 1.19.--Peak and average levels of noise radiated at a distance of 1 ft from a standard, low-noise, 40-W fluorescent lamp. (After H. R. Steele, Illuminating Engineering, 1954)

32, shows a typical spectrum of the field radiated by a fluorescent lamp. As can be seen, the level of noise radiated at frequencies of 100 MHz and above is virtually insignificant. The major part of the radiated energy occurs in the LF and VLF regions, below 100 kHz. Also, as with all these noise sources, the noise levels fall off rapidly with distance (References 32 through 34).

A particularly noisy device that disrupts signal reception will probably be relatively easy to suppress by effective shielding and by the addition of interference suppression filters. On the other hand, it should be emphasized that any of these devices being used outside the operations buildings and close to the receiving antennas may well cause some difficulties. Such devices might include hand tools, such as electric drills or saws for outside construction work, and arc welders. Of course, interference from such noise sources can be avoided by halting construction work and outside activities during scheduled satellite passes.

4. ADMISSION VIA INPUT TERMINALS

Receiver susceptibility to undesired inputs is classified according to the mechanism of unwanted signal intrusion, as follows:

(1) Linear intrusion via normal input terminals.
(2) Nonlinear intrusion via normal input terminals.
(3) Intrusion through ports not intended as signal inputs.

In this section the first two items are treated in detail. The third mechanism is covered in Section 5.1.

The block diagram of Figure 1.20 shows the essential elements of a receiver. The acceptance band of the receiver and the spectrum of an adjacent-channel signal are both shown in Figure 1.21. In the linear intrusion mode the receiver acts as a normal bandpass filter that accepts any input containing frequency components in the receiver passband, as indicated in Figure 1.21. Unwanted inputs, whose spectrum is centered at or near the frequency to which the RF filter is tuned, originate from communication systems or from other noise sources that cause interference. In the second mode, called the nonlinear intrusion mode, unwanted signal energy that lies outside the normal passband of the receiver acts on a nonlinear element in such a way as to enable the receiver to accept undesired signals. The RF filter in Figure 1.20 is a preselector network which limits the frequency band of energy passing through the succeeding active elements in the receiver. The latter nearly always have some residual nonlinear properties that play a significant role when the input amplitudes are large. When the RF filter is inadequate to limit large out-of-band inputs to a satisfactory low level, the nonlinear devices (vacuum tubes, transistors, and diodes) will generate frequency components not originally present. Interference can occur when these new frequency components are within the passband of the portion of the receiver following the nonlinear device. Phenomena typical of nonlinear intrusion include single spurious response, multiple spurious response (intermodulation), and sideband transfer (crossmodulation).

Ultimately, it will be necessary to know the degree to which the undesired signals interact with the desired signal. This is a difficult problem for several reasons. First, the original properties of unwanted signals are altered after the signals have been operated on by the linear and nonlinear receiver circuits. Second, the determination of the response of the detector to the combination of desired signal and the undesired components is an involved process. Third, it is difficult to evaluate the signal information lost through such interactions.

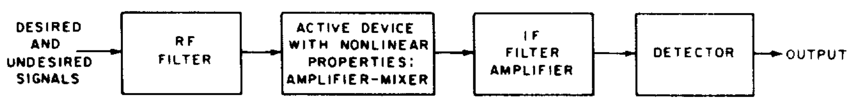


Figure 1.20.–Block diagram of basic receiver elements.

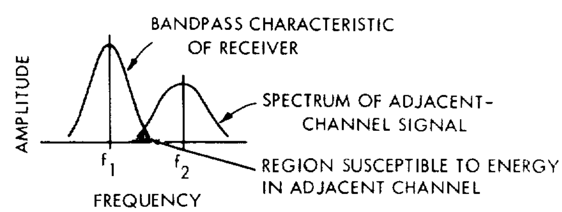


Figure 1.21.–Interference produced by a signal in an adjacent channel.

4.1 LINEAR INTRUSION

Linear admission of signals depends in a straightforward manner on the response of the receiver bandpass filter. The unwanted signals fall into one of the following categories:

(1) Broadband noise arising from natural or man-made sources that radiate inadvertently, or which are intended to radiate. (Over the bandpass of a typical receiver, the spectral density of such noise energy is essentially flat.)

(2) Signals from communication sources assigned to a frequency at or near the center frequency to which the receiver is tuned (i.e., co-channel interference; more particularly, when the center frequencies are separated by an amount less than the receiver bandwidth, the signals are said to be co-channel).

(3) Signals from communication sources assigned to operate at frequencies within one of the passbands to which the receiver is susceptible.

(4) Signals from communication sources assigned to operate on a frequency that differs from the receiver center frequency by more than the signal bandwidth (adjacent-channel interference).

A receiver usually contains one or more mixers for translating the input frequency to more convenient intermediate frequencies for further amplification. In this discussion, it is assumed that no fundamental change takes place in the nature of the signal with this translation (i.e., spectral content and phase are preserved). Once they are admitted, both wanted and unwanted signals are therefore viewed as being at or near the frequency at which the IF amplifier operates.

1. Broadband Noise

Broadband interference noise sources of the nonperiodic variety, sometimes referred to as noncoherent sources, are characterized by unpredictable and irregular waveforms. Typical natural noise sources are thermal noise, shot noise, galactic noise, solar noise, and atmospheric noise. Typical man-made sources of noise include discharges on high-voltage lines and electrical devices, noise from automobile ignition systems, commutator noise, noise in complex switching systems, and noise generated by fluorescent lamps. However, in some of these sources, a certain amount of waveform regularity exists. Atmospheric noise bursts sometimes are found to have some coherence because of multiple propagation paths (Reference 35); furthermore, there are long-time fluctuations, depending on the time of day, season, and sunspot cycle, that are roughly predictable (Reference 36). Corona noise on high-voltage lines and noise from fluorescent lamps is usually modulated at the power-line frequency (References 29 and 30). Noise from a single, stationary, internal-combustion vehicle is more or less periodic; however, noise from one or more passing vehicles or from several stationary vehicles is not. A more detailed discussion of some of these sources also appears in Section 5.1.

Nonperiodic broadband interference sources are best characterized by their power spectral densities (Reference 37). The mean-square value of the noise admitted through the IF amplifier of a radio receiver is

$$\langle e_0^2 \rangle = \int_0^\infty N_1(f)|G(f)|^2\,df\,, \tag{1.47}$$

where

$N_1(f)$ = the one-sided power spectral density at the receiver input in V^2/Hz

and

$G(f)$ = the amplitude versus frequency transfer characteristic from the receiver input to the IF amplifier output.

If the noise spectral density is flat over the receiver passband and if the center frequency of the IF amplifier is written f_1, the mean-square value of the IF amplifier output is

$$\langle e_0^2 \rangle = N_1(f_1)|G(f_1)|^2 \int_0^\infty \frac{|G(f)|^2}{|G(f_1)|^2}\, df = NG^2 B_p\,, \tag{1.48}$$

where

$N = N_1(f_1)$ is the spectral density of the receiver input noise power at the band center,

$G^2 = |G(f_1)|^2$ is the power gain at the band center,

and

$B_p = \int_0^\infty \frac{|G(f)|^2}{|G(f_1)|^2}\, df$ is the effective noise-power bandwidth of the IF amplifier.

Periodic broadband noise sources generate regularly spaced and constant-amplitude short pulses (or pulses which are nearly so). Two cases can be distinguished. In the first, the pulses are sufficiently spaced in time so that each pulse acts as a separate transient exciting the receiver. In the second, the pulses follow one another so rapidly that they tend to overlap at the filter output. The overlapping case arises when pulse spacing is of the order of the inverse of the receiver bandwidth or less. Since the more common impulse sources (e.g., radar transmitters and modulators) have relatively low repetition frequencies compared to receiver bandwidths, attention is directed exclusively to the nonoverlapping case.

The instantaneous output voltage of an IF amplifier to an input that has a Fourier spectrum given by $S(f)$ is

$$e_0(t) = \int_{-\infty}^{\infty} S(f)G(f)e^{j2\pi ft}\, df\,, \tag{1.49}$$

where

$G(f) = |G(f)| \exp[j\phi(f)]$,

$|G(f)|$ = the amplitude-versus-frequency characteristic of the IF amplifier,

and

$\phi(f)$ = the phase-versus-frequency characteristic of the IF amplifier.

For broadband inputs and narrow-band receivers, $S(f)$ will be virtually constant over the passband at a value $S(f_1)$, so

$$e_0(t) = S(f_1) \int_{-\infty}^{\infty} |G(f)| e^{j[2\pi ft + \phi(f)]} df . \tag{1.50}$$

Because $|G(f)|$ is an even function of frequency and $\phi(f)$ is an odd function, this reduces to

$$e_0(t) = 2S(f_1) \int_0^{\infty} |G(f)| \cos [2\pi ft + \phi(f)] \, df . \tag{1.51}$$

Frequently, the important output quantity is the peak of the resulting output waveform. It may be shown that for $|G(f)|$ with even symmetry around the center frequency f_1 and for $\phi(f)$ with odd symmetry and linear around f_1, the output maximum is

$$e_{om} = 2S(f_1)G(f_1) \int_0^{\infty} \frac{|G(f)|}{|G(f_1)|} df = 2SGB_e , \tag{1.52}$$

where

$S = S(f_1)$ is the Fourier spectrum of the input at the band center

and

$B_e = \int_0^{\infty} \frac{|G(f)|}{|G(f_1)|} df$ is the effective impulse bandwidth of the IF amplifier.

For nonoverlapping, irregularly spaced, nonconstant amplitude pulses, the peak value obtained with each pulse can be determined with the formulas above. Ignition noise typically falls into this category.

When random or impulse inputs are applied, the foregoing may be used to determine the mean-square value and peak value, respectively, of the output of the IF amplifier. How the normal function of the receiver is impaired then depends on the detector and the use made of the receiver output.

Remedies for broadband noise that overlaps the receiver band must take advantage of differences that are known to exist between the form of the signal and the noise in the time domain. A review of methods dealing with noise of a discrete impulse nature appears in Reference 37. The most common methods used are limiting and blanking, both done before the broadband pulses have been filtered in the IF amplifier. The principle here is that this sort of noise, having short duration and large peak value, can either be limited above the level of the desired signal or be totally blanked out for its brief duration. In either case, the signal is eliminated for the duration of the impulse, but this hardly ever causes any loss of information. In the case of the limiter, the original large-amplitude pulse is replaced by a smaller one, thereby reducing the impulsive spectral magnitude in the desired passband. It is evident that a filter preceding the limiter or blanker will widen the interference pulse and make these processes less effective as noise reducers.

Successful operation of a blanking system requires sensing of a precursor to trigger the blanking process. Systems for eliminating the periodic pulses of a nearby radar system may make use of direct synchronization from the radar source. Where there is no access to the source, the receiver itself must sense the pulse in one channel in order to blank it in a second channel.

The limiter is much simpler; frequently an operator adjusts the clipping level to achieve best reception. A technique cited in Reference 38 (Ch. 11, p. 19) for counteracting impulse noise is to pre-smear the information prior to transmission and to reassemble it after detection. The reassembling process smears the impulse over a large interval of time, thus reducing its effect.

An advantage can be realized by the use of a modulation process that builds distinguishable characteristics into the signal, which the detector then uses to help in identification. A matched-filter detector, for instance, uses *a priori* information about the shape of the signal pulses to discriminate between signal and noise (Reference 38). In general, broadband modulation schemes that enhance a signal in the presence of Gaussian noise will frequently result in signal enhancement in the presence of non-Gaussian interference.

2. Co-Channel Interference

The term "co-channel interference" designates interference that involves communications systems that have been assigned equal, or nearly equal, carrier frequencies. Co-channel frequency assignments are ordinarily made when the probability of the simultaneous encounter of signals from the two systems is insignificant. Such systems are separated physically by large distances or do not operate at the same time. Sometimes, co-channel interference will arise because of unusual propagation conditions or because co-channel sources operate under conditions for which they are not intended to operate.

The term "adjacent-channel interference" designates interference between communication systems that have been assigned neighboring channels. Channel-spacing policy varies, but the term "adjacent" will be used to mean channel separation by a frequency difference greater than the average of the two signal bandwidths. Figure 1.21 illustrates interference of this kind; energy on the skirt of the adjacent-channel signal spectrum is shown overlapping the bandpass characteristic of the receiver. Although, in the typical case, the receiver skirt sensitivity is low, compared to the in-band sensitivity, receivers located close to an adjacent-channel transmitter can be exposed to very large magnitudes of unwanted signals.

Estimates of interference arising from linear intrusion can be made in several ways without excessive numerical complexity: by treating the unwanted signal as a pure sinusoid, as a broadband waveform perfectly centered in the band, as a broadband waveform whose center frequency is sufficiently removed from the frequency to which the receiver is tuned so that the unwanted spectrum is nearly constant over the receiver band, or as a band-limited waveform falling on a small portion of the receiver selectivity curve. The first of these will be useful for the estimation of both co-channel and adjacent-channel effects. The second is appropriate for the evaluation of co-channel interference, and the third is appropriate for the determination of adjacent-channel interference. In the third case, the bandpass filter is exposed to a portion of the one sideband of the unwanted signal. The effect is not much different from that produced by thermal noise of equal mean-square value. Therefore, the spectral density of the unwanted signal is estimated at the center of the reception band and, as in the case of nonperiodic broadband noise discussed above, the mean-square value of the IF amplifier output is determined by use of (1.48). The fourth class will be illustrated in Section 4.1.4.

3. Receiver IF Channel Interference

One somewhat different mechanism involving only linear phenomena is the penetration of unwanted signals that are centered at one of the IF channels within the receiver. For instance, a large-amplitude signal centered at the frequency of one of the IF amplifiers may manage to pass through the input selective RF circuits to the IF amplifier in question. Once there, it proceeds down the rest of the receiver in a normal manner. To overcome this sort of difficulty, the selectivity of the input RF circuit and/or stray paths to the sensitive circuits must be controlled. As would be expected, the most susceptible frequency is that of the first IF amplifier, but consideration needs to be given to all succeeding IF amplifiers.

4. Adjacent-Channel Interference

Signals from two satellites with closely spaced adjacent-channel frequency assignments can produce interference within a given receiver. Two interference conditions will be analyzed, with the interfering signal unmodulated and then modulated with broadband random noise.

Two satellites, A and B, are assumed to lie within the antenna beam of a given ground receiving station. At the receiver, the average power in the unmodulated signal from satellite A is the mean-squared voltage, $\overline{X_a^2}$, and the average power in the unmodulated signal from satellite B is $\overline{X_b^2}$. At the ground station, the carrier frequencies received from the two satellites are f_a and f_b.

These frequencies are given by

$$f_a = f_1 + \Delta f_1$$

and

$$f_b = f_2 + \Delta f_2 ,$$

where

f_1 = carrier frequency of the unmodulated signal transmitted by satellite A,

f_2 = carrier frequency of the unmodulated signal transmitted by satellite B,

Δf_1 = Doppler shift of signal transmitted by A,

and

Δf_2 = Doppler shift of signal transmitted by B.

The receiver's selective circuits are assumed to consist of ***n*** single-tuned parallel *RLC* type circuits. From (1.27), the relative response of these circuits is written

$$H_n(f) = \left[1 + \frac{4(f - f_0)^2}{B^2}\right]^{-n/2} ,$$

where* B is the bandwidth of one single-tuned circuit at the half-power point, or 3 dB below maximum response. In terms of the 3-dB bandwidth, $B_{3\ \mathrm{dB}}$ of the receiver passband, $H_n(f)$ becomes

$$H_n(f) = \left[1 + \frac{4(f - f_0)^2[2^{1/n} - 1]}{B^2_{3\ \mathrm{dB}}}\right]^{-n/2} .$$

*The 3-dB bandwidth of n identical, single-tuned circuits in cascade is given by

$$B_{3\ \mathrm{dB}} = B\sqrt{2^{1/n} - 1} ,$$

where B is the 3-dB bandwidth of one single-tuned circuit (Reference 39). Similarly, the 6-dB bandwidth is given by

$$B_{6\ \mathrm{dB}} = B\sqrt{4^{1/n} - 1} .$$

When the receiver is tuned to the frequency $f_0 = f_a$, that is, when the station is tracking satellite A, the power P_i of the unmodulated interfering signal from satellite B in the receiver passband is given by

$$P_i = |H_n(f_b)|^2 \overline{X_b^2}.$$

To give a numerical example, the parameters are assumed as follows:

$$n = 3,$$

$$f_a = f_1 + \Delta f_1 = 136{,}200 \text{ kHz} + 2.0 \text{ kHz} = 136{,}202 \text{ kHz},$$

$$f_b = f_2 - \Delta f_2 = 136{,}230 \text{ kHz} - 3.0 \text{ kHz} = 136{,}227 \text{ kHz},$$

$$B_{3\,\text{dB}} = 30 \text{ kHz},$$

$$\overline{X_a^2} = 10^{-16} \text{ W} = -130 \text{ dBm},$$

and

$$\overline{X_b^2} = 3 \times 10^{-15} \text{ W} = -115 \text{ dBm}.$$

In this case, $(f - f_0) = (f_b - f_a)$ and P_i is given by

$$P_i = \left[1 + \frac{4(136{,}227 - 136{,}202)^2[2^{1/3} - 1]}{(30)^2}\right]^{-3} \times 3 \times 10^{-15},$$

or

$$P_i \approx 5.9 \times 10^{-16} \text{ W}.$$

For reception from satellite A, the signal-to-interference power ratio is

$$S/I = \frac{\overline{X_a^2}}{P_i} = \frac{10^{-16}}{5.9 \times 10^{-16}} \approx 0.17.$$

The signal has a level of -7.7 dB relative to the interference; such a low signal-to-interference ratio will usually represent an interfering situation.

The interfering signal is now assumed single-sideband modulated with broadband noise, and the power $\overline{X_b^2}$ is assumed to be uniformly distributed over an ideal bandwidth, ΔB, centered about the frequency f_b. The power spectrum of the undesired signal and the receiver response curve are shown in Figure 1.22, where the power spectral density of the interfering signal is $P(f) = \Phi_i$ W/kHz of bandwidth. When the receiver is tuned to the frequency f_a,

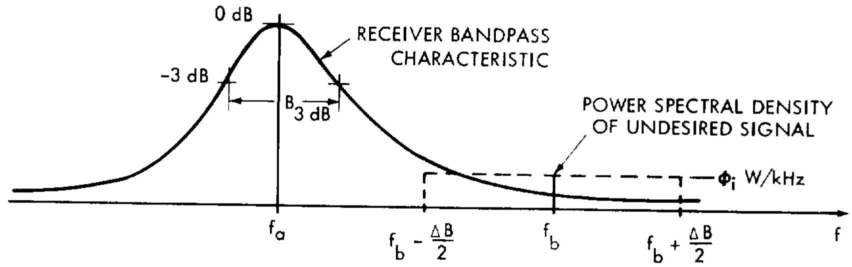


Figure 1.22.–Unwanted signal spectrum and the receiver bandpass characteristic.

the power of the unwanted interference that gets into the receiver passband selectivity skirt is given by

$$P_i = \int_0^{\infty} \Phi_i |H_n(f)|^2 \, df .$$

Assuming that Φ_i is constant over the interval*

$$\left(f_b - \frac{\Delta B}{2}\right) < f < \left(f_b + \frac{\Delta B}{2}\right),$$

and equal to zero elsewhere, the interfering power P_i is then

$$P_i = \Phi_i \int_{f_b - \frac{\Delta B}{2}}^{f_b + \frac{\Delta B}{2}} \left[1 + \frac{4(f - f_a)^2 (2^{1/n} - 1)}{B_{3\,\mathrm{dB}}^2} \right]^{-n} df .$$

By an appropriate change of variable, the integral above can be reduced to a standard integral (e.g., Reference 40, p. 1068, eq. 13) of the form

$$c \int_{a_f}^{b_f} [ay^2 + b]^{-n} \, dy ,$$

*It will sometimes be sufficiently correct to assume that the entire unwanted signal band is uniformly attenuated by the relatively flat portion of the tail of the receiver bandpass characteristic; that is, it will be possible to let

$$H_n(f) = H_n(f_b)$$

for all values of f in the band ΔB.

where

$$c = \Phi_i B_{3\,\mathrm{dB}},$$

$$a = 4[2^{1/n} - 1],$$

$$b = 1,$$

$$y^2 = \frac{(f - f_a)^2}{B^2_{3\,\mathrm{dB}}},$$

$$dy = \frac{df}{B_{3\,\mathrm{dB}}},$$

$$a_f = \frac{f_b + \frac{\Delta B}{2} - f_a}{B_{3\,\mathrm{dB}}} = \text{a constant},$$

and

$$b_f = \frac{f_b - \frac{\Delta B}{2} - f_a}{B_{3\,\mathrm{dB}}} = \text{a constant}.$$

The selective circuits of the receiver are assumed to consist of three single-tuned stages, i.e., $n = 3$, which makes $a \approx 1.0$. For $n = 3$, integrating the expression for P_i, changing the variable y back to f, and inserting the upper and lower integration frequency limits, the interfering power is

$$P_i = \Phi_i B_{3\,\mathrm{dB}} \left\{ \frac{a_f}{4\left(a_f^2 + 1\right)^2} + \frac{3}{4}\left[\frac{a_f}{2\left(a_f^2 + 1\right)} + \frac{1}{2}\tan^{-1}(a_f)\right] - \frac{b_f}{4\left(b_f^2 + 1\right)^2} - \frac{3}{4}\left[\frac{b_f}{2\left(b_f^2 + 1\right)} + \frac{1}{2}\tan^{-1}(b_f)\right] \right\},$$

or

$$P_i = \Phi_i B_{3\,\mathrm{dB}} K.$$

where K corresponds to the terms in the braces.

In the example below, the parameters of the first example are assumed and the unwanted signal power $\overline{X_b^2}$ is assumed uniformly spread in the band $\Delta B = 30$ kHz. From their definitions above, the following values can be written:

$$\Phi_i = \frac{\overline{X_b^2}}{\Delta B} = \frac{3 \times 10^{-15}}{30} = 10^{-16} \text{ W/kHz},$$

$$a_f = 1.33,$$

$$b_f = 0.33,$$

and

$$K \cong 0.267.$$

Therefore, inserting numerical values,

$$P_i = 10^{-16} \times 30 \times 0.267 = 8 \times 10^{-16} \text{ W} = -121 \text{ dBm}.$$

Assuming that the desired signal power $\overline{X_a^2}$ gets into the receiver, the resulting signal-to-interference power ratio after the selective circuits is

$$S/I = \frac{\overline{X_a^2}}{P_i} = \frac{10^{-16}}{8 \times 10^{-16}} = 0.125 .$$

Thus, the desired signal is 9 dB below the interference. This, too, will usually represent an interfering situation that is typical of satellite-to-satellite interference experienced in the STADAN network using the 136.0- to 137.0-MHz band.

4.2 NONLINEAR INTRUSION

Nonlinear effects can arise as a result of inadequate rejection of the unwanted signal in the input filter circuits of the receiver that is followed by some nonlinear process in an electronic device. Less common nonlinear admission mechanisms may include imperfect joints between conductors prior to the receiver filter circuits, which give rise to nonlinear junction effects. High-powered communication transmitters are usually the only significant sources. However, the spurious outputs of the transmitter are rarely large enough to be significant; therefore, only the main signal needs to be considered. Also, natural noise sources and incidental sources of man-made noise are usually insignificant.

1. Spurious Responses

The spurious responses of a receiver result from nonlinearity in an early stage, which gives rise to harmonics of incoming signals; nonlinearity in the mixer, which results in oscillator and signal harmonics; and frequency multiplication in the local oscillator and its related circuits. At each frequency to which the receiver is tuned, another set of possible spurious response frequencies exists, and each of these sets has its own level of significance (Reference 41).

Spurious responses arising in mixers can often be explained in terms of the following mechanism. The nonlinear device, such as a transistor, diode, or vacuum tube, has an input-output characteristic that may be specified by the power series

$$y = \sum_{n=0}^{N} a_n x^n \,. \tag{1.53}$$

The mixing operation occurs with the simultaneous application of a signal, $x_s(t)$, and oscillator voltage, $x_0(t)$, where

$$x_s(t) = v_s(t) \cos(\omega_s t + \phi_s) \tag{1.54}$$

and

$$x_0(t) = A \cos \omega_0 t \,, \tag{1.55}$$

for which

$v_s(t)$ = signal amplitude modulation function,
ω_s = angular frequency ($2\pi f_s$) of signal carrier,
f_s = signal carrier frequency,
ϕ_s = phase angle,
ω_0 = local oscillator frequency,
and
A = local oscillator amplitude.

Then, $x = x_s + x_0$ and

$$y(t) = \sum_{n=0}^{N} a_n [x_s(t) + x_0(t)]^n$$

$$= \sum_{n=0}^{N} a_n \sum_{k=0}^{n} \binom{n}{k} v_s^k(t) \cos^k(\omega_s t + \phi_s) A^{(n-k)} \cos^{(n-k)} \omega_0 t \,. \tag{1.56}$$

The periodic function $\cos^p x$ (see Section 2.1.1) can be expanded (Reference 42) in a Fourier series so that, when (1.56) is written as a sum of individual cosine terms, the result contains all frequencies

$$|m\omega_s \pm n\omega_0| , \tag{1.57}$$

where

$m = 0, 1, 2, \ldots, N,$

and

$n = 0, 1, 2, \ldots, N.$

Whenever one of these frequencies coincides with the intermediate frequency, a potential spurious response is said to exist. That is, any input frequency ω_s that satisfies the equation

$$\omega_s = \frac{(\pm n\omega_0 \pm \omega_{if})}{m} \tag{1.58}$$

with any combination of the signs leads to a potential spurious frequency.

Frequently, the amplitude of the signal component $x_s(t)$ is small compared to that of the oscillator component $x_0(t)$, and terms involving $x_s^k(t)$ can be ignored when $k > 1$. The significant portion of (1.56) is, then,

$$y_1(t) = \sum_{n=1}^{N} na_n x_0^{(n-1)}(t) x_s(t) = g(t) x_s(t) . \tag{1.59}$$

The quantity $g(t)$ is the transconductance as a function of time when an oscillator voltage, $x_0(t)$ is applied. That is, the transconductance is the derivative of (1.53), or

$$g = \frac{dy}{dx} = \sum_{n=1}^{N} na_n x^{(n-1)} . \tag{1.60}$$

Then, writing $x = x_0(t) = A \cos \omega_0 t$, we obtain

$$g(t) = \sum_{n=1}^{N} na_n A^{(n-1)} \cos^{(n-1)} \omega_0 t = \sum_{n=0}^{N-1} g_n \cos n\omega_0 t . \tag{1.61}$$

The last term on the right of (1.61) is the form that would be obtained if $\cos^{(n-1)}\omega_0 t$ were expanded as was done for $\cos^k x$ in (1.2) and all terms of like harmonic were collected. Thus, from (1.59),

$$y_1(t) = \frac{v_s(t)}{2} \sum_{n=0}^{N-1} g_n \Big\{ \cos\left[(\omega_s - n\omega_0)t + \phi_s\right] + \cos\left[(\omega_s + n\omega_0)t + \phi_s\right] \Big\}. \tag{1.62}$$

That is, frequencies $\omega_s \pm n\omega_0$ will be obtained. The quantity $g_n/2$ is the conversion transconductance corresponding to the nth oscillator harmonic. It may be noted here that if $g(t)$ is a pure cosine wave at the frequency ω_0 (that is, if g versus x is a straight line over the region of oscillator swing), then the only output frequencies are $\omega_s \pm \omega_0$. Some electronic devices do, in fact, come fairly close to this ideal over a portion of their operating range, and from the viewpoint of minimizing interference, operation ought to be restricted to this range. However, designers frequently do not, or cannot easily, control the level of oscillator voltage applied to the mixer. Maximum conversion transconductance at the fundamental frequency (at radian frequency $\omega_s \pm \omega_0$) is obtained with large oscillator input, and this often results in more than a proportionate increase in the conversion transconductance at harmonic frequencies. Then, too, the output of a variable-frequency oscillator is rarely constant over an appreciable range of frequencies; the conversion gain generally varies over the band.

When the mixer is a diode, as it often is in microwave receivers, the mixing of the signal with a harmonic of the local oscillator ordinarily cannot be avoided. An ideal diode acts, in effect, like a switch that is turned off and on by the local oscillator at the oscillator frequency. The signal voltage is therefore multiplied by a square-wave switching function. The square wave contains all odd harmonics of the oscillator frequency so that harmonic mixing with all odd oscillator harmonics is unavoidable. In a real diode the performance is somewhat modified, but the principle is essentially the same. A more precise evaluation of the frequency conversion for a crystal mixer is obtained by use of the current-voltage characteristic of the diode. This is given by

$$i = I_s(\epsilon^{av} - 1), \tag{1.63}$$

where

I_s = the reverse saturation current,

a = a constant which, in theory, is e/KT (≈ 40),

v = the voltage at the quiescent operating point,

$e = 1.602 \times 10^{-19}$ C, the electronic charge,

$K = 1.38 \times 10^{-23}$ J/°K, Boltzmann's constant,

and

T = temperature in degrees Kelvin.

Therefore,

$$g = \frac{di}{dv} = aI_s\epsilon^{av} . \tag{1.64}$$

When

$$v = A \cos \omega_0 t ,$$

then

$$g(t) = aI_s\epsilon^{(aA \cos \omega_0 t)} = aI_s\left[I_0(aA) + 2\sum_{n=1}^{\infty} I_n(aA) \cos n\omega_0 t\right] . \tag{1.65}$$

In this expression, $I_n(aA)$ is the modified Bessel function of the first kind of order n ($n = 0, 1, 2, \ldots$) and of argument (aA) (Reference 42). The conversion transconductance, as defined in (1.61), is

$$g_n = 2aI_sI_n(aA) . \tag{1.66}$$

From Bessel function theory it will be found that for values of (aA) apt to be used here (ranging around 10), values of $I_n(aA)$ up to about $n = 6$ are of the same order of magnitude. That is, harmonic conversion will be very significant for the sixth harmonic of the oscillator. For higher values of the argument (aA), the value of $I_n(aA)$ becomes significant for even greater values of n.

The interference-to-signal ratios are not easily calculated because the gain or loss of amplitude of the undesired signal between the input and the point at which the nonlinear effect takes place is not ordinarily known. Furthermore, the oscillator level and harmonic and subharmonic content, as a function of the frequency to which the receiver is tuned, are not ordinarily known. It is more common to measure the intensity of the spurious responses than to calculate them. The usual procedure is to set the tuning control to three frequencies in each band, one at the center and the others in the vicinity of the band extremes. With the receiver fixed at each frequency, an input signal is applied from a generator that is tuned through the frequencies of potential response. A desired signal, modulated or unmodulated, may be applied simultaneously. The observed quantity is the ratio of signal input voltage to interference voltage required to give a stated output. The ratio may depend on the input level. The intensity of the spurious responses, nevertheless, can be calculated in many cases, especially when only approximate values are needed. An example of calculated spurious response intensities follows.

Let $N = 3$ in (1.53). The spectrum of signals produced in the nonlinear device is found by expanding (1.56) to obtain

$$
\begin{aligned}
y(t) = {} & a_0 + a_1 v_s(t) \cos(\omega_s t + \phi_s) + a_1 A \cos \omega_0 t \\
& + \frac{1}{2} a_2 [v_s^2(t) + A^2] + \frac{1}{2} a_2 v_s^2(t) \cos 2(\omega_s t + \phi_s) \\
& + \frac{1}{2} a_2 A^2 \cos 2\omega_0 t + a_2 v_s(t) A \cos [(\omega_s - \omega_0)t + \phi_s] \\
& + a_2 v_s(t) A \cos [(\omega_s + \omega_0)t + \phi_s] \\
& + \left[\frac{3}{4} a_3 A^3 + \frac{3}{2} a_3 A v_s^2(t)\right] \cos \omega_0 t \\
& + \left[\frac{3}{4} a_3 v_s^3(t) + \frac{3}{2} a_3 A^2 v_s(t)\right] \cos(\omega_s t + \phi_s) \\
& + \frac{1}{4} a_3 v_s^3(t) \cos 3(\omega_s t + \phi_s) + \frac{1}{4} a_3 A^3 \cos 3\omega_0 t \\
& + \frac{3}{4} a_3 v_s^2(t) A \cos [(2\omega_s - \omega_0)t + 2\phi_s] \\
& + \frac{3}{4} a_3 v_s^2(t) A \cos [(2\omega_s + \omega_0)t + 2\phi_s] \\
& + \frac{3}{4} a_3 v_s(t) A^2 \cos [(\omega_s - 2\omega_0)t + \phi_s] \\
& + \frac{3}{4} a_3 v_s(t) A^2 \cos [(\omega_s + 2\omega_0)t + \phi_s] \,.
\end{aligned}
\tag{1.67}
$$

From the collection of components in (1.67) we consider the terms for which the frequency of the interfering or undesired signal, now written ω_{SI}, is [see (1.58)]

$$\omega_{SI} - 2\omega_0 = \pm \omega_{if} \,.$$

The receiver is tuned to the desired component, whose frequency is now written ω_{SD},

$$\omega_{SD} - \omega_0 = \omega_{if} .$$

The intensity of a spurious response at frequency ω_{SI} when receiving a desired signal of frequency ω_{SD} is determined from the coefficients of the appropriate frequency terms given in (1.67). Suppose for instance that ω_{SD} = 140 MHz, ω_0 = 110 MHz, and ω_{if} = 30 MHz. Then, the frequency of the interfering signal $\omega_{SI} = 2\omega_0 \pm \omega_{if}$ has the two values of 250 MHz and 190 MHz. From (1.67), the desired component at the output of the mixer is of angular frequency $\omega_s - \omega_0$ and is given by

$$D = a_2 v_s(t) A \cos [(\omega_s - \omega_0)t + \phi_s] .$$

The peak voltage of this component is

$$D = a_2 A V_{SD} ,$$

where V_{SD} is the amplitude value of $v_s(t)$ and the signal input is the desired signal. The undesired or interfering component has angular frequency satisfying $\omega_{if} = \pm\omega_s \pm 2\omega_0$ and is given by

$$I = \frac{3}{4} a_3 v_s(t) A^2 \cos [(\omega_s \pm 2\omega_0)t + \phi_s] .$$

Its peak value is

$$I = \frac{3}{4} a_3 V_{SI} A^2 ,$$

where V_{SI} is the peak of $v_s(t)$ and the signal input is the undesired signal. At the output of the mixer, the ratio of the level of the desired signal to that of the interfering or undesired signal is, therefore,

$$\frac{D}{I} = \frac{4a_2 V_{SD}}{3a_3 V_{SI} A} , \tag{1.68}$$

where V_{SD} and V_{SI} are the peak voltage levels of the desired and undesired or interfering signals, respectively, at the input of the mixer. However, the quantity of interest is the ratio of the voltage of the desired signal to that of the interfering signal at the receiver RF input rather than at the input to the mixer.

The voltages given in (1.68) may be converted to voltages at the input of the receiver by taking account of the RF gain of the receiver. Let K be the voltage amplification of the interfering signal relative to that of the desired signal as a result of gain in the RF circuits preceding the mixer. Then, if V_D (in place of V_{SD}) is written to designate the peak voltage of the desired signal at the input of the receiver, and if V_I (in place of V_{SI}) is written to designate the voltage of the interfering signal at the input of the receiver, then $V_{SI} = KV_I$ and the required ratio becomes

$$\left(\frac{D}{I}\right)_{\text{input}} = \frac{4a_2 V_D}{3a_3 AKV_I} . \tag{1.69}$$

For instance, if two single-tuned circuits each of bandwidth $B = 15$ MHz are used in the RF stage, then, from (1.29) for the 190 MHz interference component ($k = 2$),

$$K = \left[\frac{B}{2(f_{SI} - f_{SD})}\right]^2 = 0.0225 \text{ voltage ratio.}$$

If now we use the values of a_2 and a_3 given in Section 2.1.1 and assume the local oscillator level $A = 10$ V, we obtain from (1.69)

$$\frac{D}{I} = \frac{4 \times 2.6 \times 10^{-5}}{3 \times 8 \times 10^{-7} \times 10 \times 2.25 \times 10^{-2}} \frac{V_D}{V_I} = 192.5 \frac{V_D}{V_I} .$$

For equal values of desired and undesired components, $D/I = 1$, and the input ratio must be

$$\frac{V_I}{V_D} = 192.5 .$$

That is, for equal signal and interference at the mixer output, the interfering signal at the receiver input has to be 10 log(192.5) = 45.7 dB above the level of the desired signal.

Figure 1.23 shows a plot of equation (1.58) relating the tuned frequency f_{SD} and frequency of potential interference f_{SI} for several values of m and n for a receiver covering a range from 100 to 200 MHz and having an oscillator frequency f_0 set 30 MHz below the tuned frequency. Measured or computed values of strength of response are indicated on the diagram at the appropriate points, as shown on the line labeled $f_{SI} = 2f_0 - f_{if}$ at $f_{SD} = 140$ MHz. Or, for each spurious response line on Figure 1.23, a corresponding curve can be plotted as shown in Figure 1.24, showing the relative response at each tuned frequency.

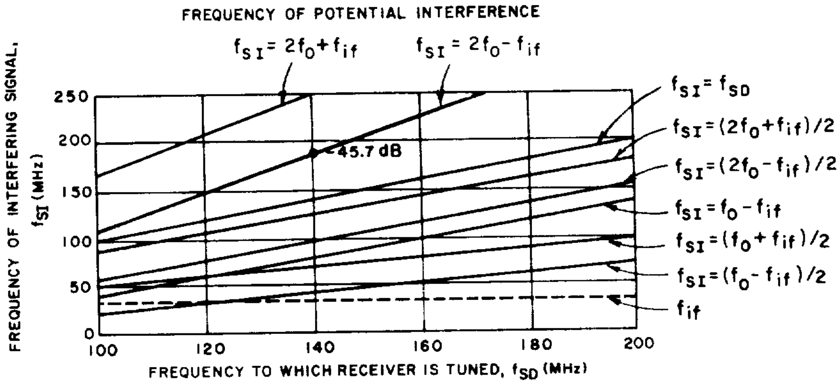


Figure 1.23.–Possible spurious responses in a high-frequency receiver with local oscillator frequency f_0 set 30 MHz below desired frequency.

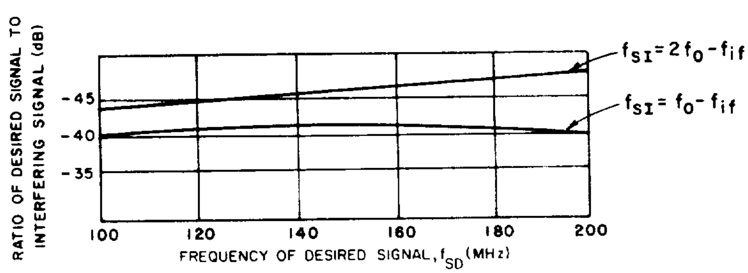


Figure 1.24.–Relative signal-to-interference response for equal outputs.

The intensity of spurious signals introduced by nonlinear mechanisms involving the generation and admission of signals may be reduced by filtering in appropriate places and reducing the levels of the input signals. For instance, the spurious response of a receiver is increased when the local oscillator is made to drive the mixer over a wider range of its voltage-current characteristic. Although the desired conversion efficiency might be increased somewhat, the "harmonic conversion" efficiency is also increased and, beyond a certain point, becomes proportionately more than the desired effect. To keep the spurious response low, additional filtering can be used prior to the mixer. As a general rule, a trade-off between the cost of additional filtering and that for increased gain ought to be considered. In addition, the region of the characteristic over which the mixer is operated should be as nearly square-law as possible.

2. Intermodulation and Crossmodulation

Intermodulation and crossmodulation, discussed for transmitter mechanisms in Section 2.4, are also important in receivers. The mechanisms are essentially the same. Intermodulation in receivers, however, results when two or more unwanted signals are present simultaneously at the input. Crossmodulation is the transfer of information from an undesired carrier onto the desired one. In either case, nonlinearity in a circuit near the receiver input is usually the cause.

Intermodulation is the more important of these mechanisms. It becomes especially important when a range of frequencies is subdivided into separate communication channels and when a number of closely spaced channels must be used simultaneously. Then, two unwanted signals of the form in (1.43) and (1.44) of Section 2.4 give rise to a component of the form

$$y_I(t) = v_1^2(t)v_2(t)\cos\left[(2\omega_1 - \omega_2) + 2\phi_1(t) - \phi_2(t)\right] \tag{1.70}$$

when the nonlinearity is of the third degree. As was pointed out in Section 2.4, this component is significant because $2\omega_1 - \omega_2$ is not too different from either frequency if ω_1 and ω_2 are not far apart. It should be clear that the component at frequency $2\omega_2 - \omega_1$ is significant, too, for the same reason.

It will similarly be found that with three channels at frequencies f_1, f_2, and f_3, intermodulation products having frequencies near to but not coincident with the original generating frequencies are

$$f_1 + f_2 - f_3,$$

$$f_1 - f_2 + f_3,$$

$$2f_1 - f_2,$$

$$2f_1 - f_3,$$

$$2f_2 - f_1,$$

$$2f_2 - f_3,$$

$$2f_3 - f_1,$$

and

$$2f_3 - f_2.$$

The even-degree terms in Taylor's series expansion for the output-input characteristic of the nonlinear element also give rise to intermodulation components, but they are all far from the range of frequencies in question. Though the third-degree term is generally the most important, the fifth-degree term may have to be accounted for also. Possible interference components due to fifth-degree nonlinearity are of the form

$$f_1 + f_2 + f_3 - f_4 - f_5 ,$$

$$2f_1 + f_2 - f_3 - f_4 ,$$

$$f_1 + f_2 + f_3 - 2f_4 ,$$

$$2f_1 + f_2 - 2f_3 ,$$

$$3f_1 - f_2 - f_3 ,$$

and

$$3f_1 - 2f_2 .$$

These are only representative forms; the subscript on the frequencies above may be permuted in any way among the assigned frequencies. Thus, if there are 10 frequencies, $3f_7 - 2f_{10}$ or $f_1 - f_2 + f_3 - f_4 + f_5$ are frequencies that can be significant. Techniques for channel selection to avoid interference are given by Babcock (Reference 43) and also by Beauchamp (Reference 44).

Tests of susceptibility to intermodulation in actual receivers have been described in detail. McLenon (Reference 45) applied signals to commercial-grade receivers to give potential intermodulation at 5.1 MHz. He obtained a resultant equivalent interference carrier level of 0.5 μV for inputs ranging from 0.01 V to 0.1 V. The highest input was required in a receiver that had two tuned circuits before the first amplifier tube.

A sample calculation of the magnitude of intermodulation interference will now be given. Third-degree nonlinearity is assumed. Carrying out an expansion similar to that given in (1.56) but with $x_s(t)$ and $x_0(t)$ replaced by two incoming signals $x_1(t)$ and $x_2(t)$ as given by (1.43) and (1.44) and with $n = 3$, an output interference component

$$y_I(t) = \frac{3a_3}{4} v_1^2(t) v_2(t) \cos (2\omega_1 - \omega_2)t \tag{1.71}$$

is obtained. The tuned frequency of the receiver is $2\omega_1 - \omega_2 = \omega_0$. A desired signal, $x_s(t) \cos \omega_0 t$, entering the receiver at the same time will result in an output term determined by the first-degree terms (with coefficient a_1) of the Taylor series. Thus, $y_s(t) = a_1 v_s(t) \cos \omega_0 t$.

The signal-to-interference voltage ratio S/I is defined as the ratio of the coefficients of these two components, or

$$S/I = \frac{4a_1 v_s(t)}{3a_3 v_1^2(t) v_2(t)} . \tag{1.72}$$

If, for simplicity, $v_1(t)$, $v_2(t)$, and $v_s(t)$ are taken to be the constants v_1, v_2, and v_s, respectively, and the two unwanted signal amplitudes are assumed equal so that $v_1 = v_2$, then the signal-to-interference ratio is unity when

$$v_1 = \left(\frac{4a_1 v_s}{3a_3}\right)^{1/3} . \tag{1.73}$$

For instance, if

$$v_s = 10 \times 10^{-6}\ \text{V} ,$$

$$a_1 = 5 \times 10^{-3}\ \text{mho} ,$$

and

$$a_3 = 5 \times 10^{-5}\ \text{A/V}^3 ,$$

then $v_1 = 0.11$ V. At VHF, an interfering signal of this magnitude could be produced by a 50-W transmitter with a spacing of about 150 ft between the transmitting source and the receiving antenna.

The intermodulation interference has amplitude v_1 at the input to the nonlinear element, but its amplitude at the antenna terminals can be greater than this value. However, if the selectivity of the input circuit is not sufficient to cause much attenuation to the unwanted signals, the unwanted input signal voltage thus can be about 0.11 V, also.

In the case of crossmodulation arising from third-degree nonlinearity, the interference component I [again using (1.43) and (1.44) and expanding in a form such as (1.56) with $v_2(t) \cos [\omega_2 t + \phi_2(t)]$ viewed as the desired signal] is

$$y_I = \frac{3a_3 v_1^2(t) v_2(t)}{2} \cos [\omega_2 t + \phi(t)] . \tag{1.74}$$

This component contains a mixture of sidebands from the unwanted v_1 and the wanted v_2 signals. Since the desired component, in this case, is

$$y_s(t) = a_1 v_2(t) \cos [\omega_2 t + \phi_2(t)] ,$$

the signal-to-interference voltage ratio, defined as the ratio of the coefficients of $y_s(t)$ and $y_I(t)$, with $v_1(t)$ and $v_2(t)$ constant at v_1 and v_2, respectively, is

$$S/I = \frac{2a_1}{3a_3 v_1{}^2} . \tag{1.75}$$

Only unwanted signals of large amplitude will make this ratio significant. When the signal-to-interference ratio is unity,

$$v_1 = \left[\frac{2a_1}{3a_3}\right]^{1/2}$$

for crossmodulation. For example, if the values of a_1 and a_3 given previously are used,

$$v_1 = 8.1 \text{ V} .$$

3. Desensitization

Desensitization refers to a reduction in overall receiver gain, or sensitivity, or both, when a large unwanted signal enters the receiver. The interference signal alone may not even be observed if it is either unmodulated or modulated in such a way that the receiver is nonresponsive. Typical mechanisms of desensitization are discussed in the following paragraphs.

A large-amplitude unwanted signal at carrier frequency passing through a receiver having automatic gain control (AGC) will sometimes depress the receiver gain. The AGC voltage is determined by the carrier level at the detector input, and any signals present there will affect it. In envelope detector systems, in FM receivers, and in receivers using frequency tracking, a large undesired signal will tend to "capture" the detector.

Desensitization is also encountered when unwanted signals have sufficient amplitude to overload one of the early stages of the receiver. This effect may be found to occur even with unwanted signals at frequencies relatively far from that to which the receiver is tuned, because of the large bandwidth of the initial stages. The mechanism varies according to the circuit. The unwanted signal may overload the first active device and cause the desired signal to be suppressed during periods of saturation and cutoff; this usually occurs due to a lowering of the effective Q of the tuned circuit. In systems having RC networks for bias generation or in those having AGC filter networks in the early stages, overload will cause a change in bias and a reduction in gain. The bias is sustained for an interval of time, depending on the RC time constant and the peak value of the undesired signal.

Pulsed signals of low duty cycle, such as radar emissions, can be especially troublesome because of their large peak amplitudes. Microwave radar interference to low-frequency communications systems, by overload of an early receiver stage, is not uncommon, largely because the lumped tuned circuits at the input are virtually useless as filters of microwave energy. Once the unwanted pulse appears at the input amplifying device, it will act according to one of the mechanisms described in the previous paragraph. Even when charging networks are not present, unwanted pulse signals of low duty cycle may be a source of interference though they do not fall strictly into the category of desensitization. Two mechanisms are described below.

Broadband signals, such as pulsed radar signals, if admitted at least as far as the first amplifier stage of low-frequency receivers, may be detected in the amplifier if they are sufficiently large. Their sidebands may contain energy in the receiver passband at the point of detection. Furthermore, distortion in the detection process will sometimes increase the effective bandwidth of the sidebands. Broadband signals of the kind mentioned above may have bandwidths to about 15 MHz or even more, depending upon the radar parameters. Taking into account possible sideband distortion, this mechanism should be considered potentially significant through the HF band.

An unwanted signal that would ordinarily be rejected by the receiver will sometimes cause interference by transferring its information sidebands to the carrier of the desired signal. Large-amplitude pulsed signals that are admitted to the first amplifier stage and overload the input circuit create, essentially, a short circuit across the tuned circuit for the period of the pulse. The desired signal is, therefore, altered at the pulse rate; this is, in effect, a modulation of the desired carrier by the pulse information. No charging networks need be involved here. When unwanted signals are not so large as to cause overload directly at the input, the nonlinear input-output characteristic of the active device may still be a cause of crossmodulation..

Under "Spurious Response," it was pointed out that diode mixers act naturally as harmonic mixers to create spurious responses. Diode mixers are also subject to desensitization effects (Reference 46). The effect is found to arise in microwave receivers (e.g., radar receivers) where the mixer is the first electronic device following the input terminals. It can be shown that the conversion transconductance [defined by $g(t)$ in (1.65)] is altered by the presence of a large unwanted signal. A more important effect, however, appears to be connected with impedance mismatch; the effective output impedance of the mixer at IF is altered by the unwanted signal. If the input impedance of the IF amplifier is matched to the impedance of the mixer in the absence of unwanted signals, it will become unmatched when the unwanted signal appears. Tests reported (Reference 45, pp. 28 to 29) show a drop in conversion efficiency by approximately 3 dB for an unwanted sinusoid whose amplitude is equal to that of the local oscillator signal (see Figure 1.25); the greater the local oscillator power fed to the mixer, the larger will be the magnitude of the unwanted signal

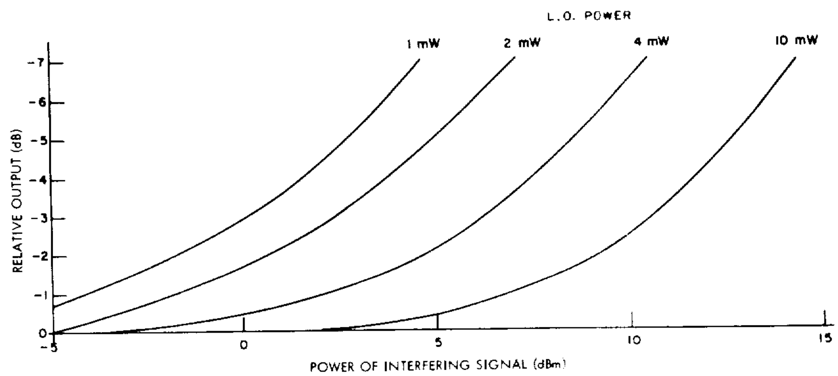


Figure 1.25.—Test results of conversion loss in a crystal mixer.

that can be tolerated. It was pointed out above, however, that with increasing local oscillator inputs, the harmonic conversion transconductance g_n becomes significant for higher values of n. A compromise is therefore needed between high local oscillator power (to minimize desensitization potential) and low local oscillator power (to minimize spurious response potential).

It is evident that with adequate filtering prior to the active elements in the receiver, the effects of nonlinearity in these elements can be reduced. Ideally, the bandwidth of circuits ahead of a potentially nonlinear element should be equal to the bandwidth of the IF amplifier, but this will generally be impractical and difficult to accomplish. Unwanted signals whose frequency is relatively near that of the desired band will therefore not always be easy to reject in the RF amplifier. Where such interference is expected, as would seem to be the case when aircraft interference signals located in the 118.00- to 135.95-MHz band are encountered, it is desirable to use input circuits with large dynamic ranges to avoid such effects as overload and desensitization. However, sharp rejection filters (wave traps) have been devised (References 1 and 13) especially for rejecting fixed-frequency unwanted signals in an adjacent channel.

4. Miscellaneous Interference Reduction Techniques

In addition to interference reduction by filtering and by proper choice of the electronic device and its operating point, certain additional techniques should be considered. Electromagnetic signals can be isolated by the use of different frequencies, time sharing, codes that make different signals separable, spatial separation, shielding, and different wave polarizations. For the case of potential interference to a ground station receiving signals from multiple satellites operating in a given band, studies have been completed for establishing methods of frequency assignment to minimize interference (Reference 47).

5. SPURIOUS PATH ADMISSION MECHANISMS

The antenna represents the sensor in a receiving system and is obviously the most important point of energy pickup. However, significant amounts of unwanted RF energy may penetrate the system through other paths.

The energy may penetrate in one of two ways: as conducted interference along the power lines and control cables of the system or as radiated interference in which the energy is able to penetrate the system directly because of poor and inadequate shielding.

The two basic forms of energy penetration will be considered separately, and methods of reducing the unwanted noise will be discussed. These reduction methods include the use of filters in the case of conducted noise on cables and the use of adequate shielding in the case of radiated noise.

5.1 PENETRATION THROUGH CABLES

The basic mechanisms by which noise may penetrate into a receiver via the input cables are illustrated in Figures 1.26 to 1.28. As shown in Figure 1.26, noise from a motor is able to enter the receiver directly through a conductive path provided by power leads. In Figures 1.27 and 1.28, however, the noise is coupled into the input cable through impedance and inductive coupling, respectively. The mechanism of Figure 1.27 is referred to as "common-mode" interference, and those of Figures 1.26 and 1.28 are referred to as "differential-mode" interference. These basic mechanisms will now be further discussed.

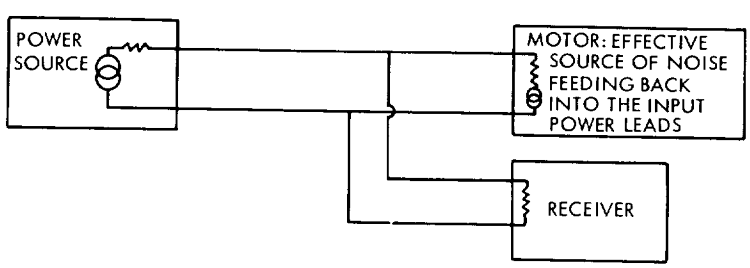


Figure 1.26.–Simplified circuit illustrating how noise may be conducted between two units sharing a common power source.

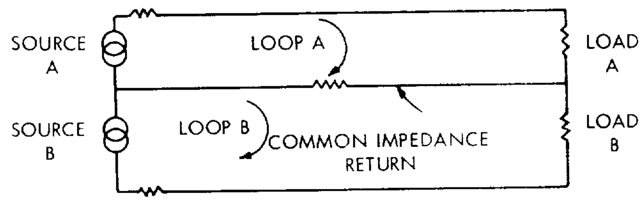


Figure 1.27.—Simplified circuit of a common-mode interference path providing noise transfer through a common impedance.

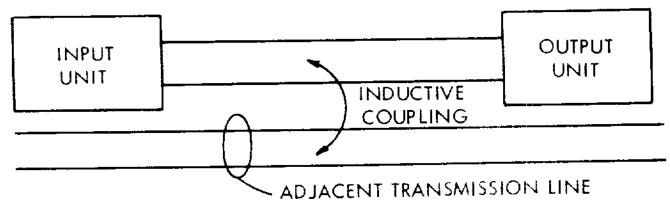


Figure 1.28.--Simplified circuit showing noise introduced by a differential-mode path (inductive coupling).

1. Conductive Path

Figure 1.26 illustrates an example of the transfer of conducted noise from one unit to another via a common power supply. Since the power source has a finite internal impedance, the noise existing in one unit is transferred to the other unit. The magnitude of such an effect is difficult to predict since a knowledge is required of the noise transfer characteristics of the system as well as of the susceptibility of the unit being interfered with. It is a common practice in government or industry to set maximum limits on the allowable conducted noise output from any electrical device or equipment connected to a power line.*

2. Common-Mode Path

One means by which common-mode interference can be produced is illustrated in Figure 1.27. The ground return is seen to be common to both loops and also to contain some impedance. Consequently, the two loops are coupled to each other through their common impedance. If the impedance of the common branch were zero, there would obviously be no noise transfer. The

*See for example, Mil. Spec. MIL-I-6181D, "Interference Control Requirements, Aircraft Equipment."

obvious remedy to this problem is to avoid the use of common ground returns. A designer will frequently use separate grounds, tied to a single common point; or he will use a common ground bus made of low-impedance material. It should be noted that, although resistance in a common ground return is a very important factor in common-mode coupling, inductive reactance in the ground return is equally important at high frequencies. In such a situation, the coupling, once it occurs, can be remedied only by the use of filters.

3. Inductive Coupling

Inductive mode coupling is shown in Figure 1.28. Because of inductive coupling, noise is induced in series with the desired input.

At low frequencies (below 50 kHz) the coupling between a pair of identical cables can be accurately determined. The types of cable commonly used include the coaxial and the shielded and unshielded pair. At low frequencies, electrostatic shielding, when it exists, will usually eliminate any electric coupling between cables. Magnetic coupling depends upon and is inversely proportional to the square of the distance between conductors. Therefore, inductive coupling can be reduced to tolerable levels by maintaining an adequate spacing between cables or by twisting the paired conductors of open-wire unshielded lines.

A coaxial cable with a solid outer conductor and a perfectly concentric cross section will not exhibit any coupling. However, in practice, coaxial cables contain braided outer conductors, so some coupling generally does exist. A typical value of mutual impedance is $10^{-6}\Omega$, so 1 μV is induced in a line for each 1 A of current.

At the higher frequencies, quantitative formulas for coupling between cables are less well known, and those that exist are quite complex. One of the difficulties is that coaxial cables usually consist of braided outer conductors; there is no generally accepted theory for coupling in this case. Indications are that when good quality coaxial cables are properly grounded along their length and are properly terminated, one should experience almost negligible interaction. On the other hand, inadequate cable terminations or discontinuities in the shields will permit signal voltages and currents to travel on the outside surface of a coaxial cable. This can be a serious source of coupling, particularly if the operating frequency and length of the cable produce resonant effects so that substantial standing waves can exist. In fact, the incidental imperfections in joining and grounding cables are usually the main sources of trouble with interconnections between system components. Awareness and care are obviously necessary.

5.2 FILTERING

The isolation and removal of conducted noise is largely achieved by means of RF filters. A large number of commercially made filters are presently available for use in radio interference applications. As a rule, specific kinds of filters are used in certain applications particularly susceptible to interference. Among the different types of filters are power line filters, bypass (or feed-through) filters, harmonic suppressors, and more complicated networks for use on control and output leads.

The function of such filters is to limit the frequency bandwidth of the various leads entering and leaving the equipment to that which is required for undistorted transmission of the desired waveforms. However, this is difficult to achieve in practice; uniformly high rejection of all frequencies outside a wanted band is virtually impossible to achieve. There are two basic filter techniques: reflective filtering and absorptive filtering. In the former, rejection of the unwanted signals is achieved by completely mismatching the impedances of interfacing circuits at the undesired signal frequencies. The unwanted signals "see" either open-circuit or short-circuit paths at the appropriate frequencies. In the absorptive technique, the undesired signals are actually separated from the wanted signals and are diverted into a separate channel where they are absorbed in a resistive circuit. The second technique is used in those cases where some resemblance of "matching" is to be preserved at all frequencies, e.g., in generators.

1. Power Line Filters

Power line filters are used to prevent unwanted high-frequency signals from being coupled between equipment by a common power connection. A common technique is to utilize a "brute force" filter, i.e., a filter having several ladder or recurrent sections in which the choice of the values of circuit parameters is not critical. Sometimes it is necessary for such filters to have extremely broad band performance, e.g., filters for screened rooms. In these cases, multiple sections are used, each being designed to reject a different portion of the spectrum. For stringent requirements, multiple-section filters combining constant k- and m-derived sections may be required. Filters of this kind may be useful in other applications where lumped parameter filters are needed. An example of the design of such a filter will be found in Reference 40, p. 184.

A number of general precautions must be followed in constructing all rejection filters, and some particular precautions must be observed for power line filters. Since substantial amounts of current are taken from the power line and must pass through the filter, its inductors may require wire of rather large size. This places severe limitations on the amount of inductive reactance that can be realized. (Section 11 discusses these aspects.)

2. Bypassing and Feedthrough Capacitors

The objective in bypassing is to provide a low-impedance shunt path to effectively short circuit unwanted signals. Typically, a capacitor whose impedance at the frequency of interest is less than one-tenth that of the circuit being bypassed is used. It is to be noted that a 10:1 impedance ratio may not be enough in some cases. Unfortunately, it does not always suffice merely to use a unit of larger capacitance, since capacitor elements have internal and lead inductances that provide a frequency limit above which their capacitive properties cannot be realized. (For the resonance properties of capacitors, see, for instance, Reference 14, Figure 6.27.) The self-resonant property is sometimes utilized to obtain nearly perfect bypassing in a narrow band of frequencies. Bypassing of input and output leads can be accomplished effectively by using feedthrough capacitors, which provide a conductive path from one terminal to the other and at the same time offer high capacitance (up to 2300 pF) to a bulkhead or mounting panel. Feedthrough capacitors are available in such small sizes that they can be incorporated into electrical connectors. In special cases, transmission line sections may be used instead by making use of the impedance-transforming property of a quarter-wavelength line.

Reference 14 provides useful data on modern synthesis and design of electrical filters. Many examples are given of the design of low-pass, high-pass, and band-reject filters. Details of filter design at VHF and higher frequencies are given in Sections 7.5.4.2 and 7.5.4.3 of Reference 1.

5.3 GROUNDING

Conducted noise on signal leads and cables is created frequently by poor grounding practices. Problems arising in such cases can be avoided only by very careful design of the grounding system (see References 48 and 49).

The grounding system usually refers to the network of conductors that tie the various parts of a system to some common reference point. This point, designated as "ground," represents a reference potential to which all signal and power voltages are established. The reference potential is often represented physically by a surface, such as a metal sheet, provided that the various points on this surface are connected by sufficiently low-resistance paths. All points on the surface can then be considered to be approximately at the same potential. The extent to which this approximation holds depends, in turn, on the electrical system involved. If, for example, the metal sheet is used as a power supply ground, then this ground may carry a current of the order of several amperes. Any small resistance in the path of this current will create a small potential difference across the ground conductor (see Figure 1.27). This voltage will be insignificant when compared to the power supply voltage. However, if the same metal sheet is also used as a ground return for some signal voltage, this voltage now becomes very significant in comparison with the signal voltage since noise signals of even a few microvolts are quite intolerable in signal circuits. This example illustrates how noise is created by the use of a common path for both power and signal return currents.

Problems arising from coupling through a common ground can obviously be avoided by completely separating the various networks, such as the signal circuit, power circuit, and control circuit, that together make up the complete electrical system. These networks should not be interconnected through common impedances and common ground returns. Furthermore, care must be taken to see that, even though a direct connection does not exist, voltages do not become inductively or capacitively coupled into signal circuits from the power or control circuits. Generally this requires that signal circuits and power circuits be put into separate compartments or drawers with adequate shielding between compartments.

1. Single-Point vs. Multiple-Point Grounds

Generally, multiple-point grounds are preferred by electronic designers for two reasons: convenience and circuit efficiency. Circuit construction can be simplified by returning circuit elements to the nearest appropriate multiple-point ground. At the higher frequencies, the lead lengths (connections to components) must be kept short in order to keep the self-inductance of the leads small, and this is facilitated by multiple-point grounds.

In a multiple-point ground system, the various ground points must be connected together so that the possibility of common current paths exists; this may produce interaction. In order to avoid this, a designer will frequently use a single-point ground system, particularly at the lower, audio frequencies. In a single-point system, ground returns must be kept as short as possible, so the ground point should be centrally located inside the circuit layout.

When physical separation of the different parts of a circuit is necessary, the long ground leads in a single-point system become objectionable because of the possibility of mutual inductive coupling between leads. In this case, the single-point system becomes impractical, and the designer must revert to a multiple-point ground system.

When circuits are separated by long distances, cables are generally used to interconnect the system. The use of cables can create noise problems. Furthermore, the length of the cable can be such as to cause resonance. If so, the cable can act as an efficient radiator of energy; it also can receive energy radiated from some other source. The possibility of such radiation can be reduced substantially by connecting the cable to earth grounds at a large number of points. The ground connection of the cable provides a common-mode path for the coupling of currents that may flow in the earth or reference grounds. The solution to this problem of coupling is to use balanced signal circuits so that undesired voltages or currents are nulled out.

2. Balanced Circuits

Figure 1.29 shows the arrangement of a balanced coupling circuit. Current flowing in the loop formed by the cable shield and the ground will induce voltages across the twisted pair of signal leads. If these leads are perfectly balanced, the induced voltage will be canceled at the input of the receiver. The use of a balanced cable has therefore eliminated the noise pickup. In particularly sensitive circuits, the balance may be made adjustable. Current in the shield of the cable can be minimized by connecting only one end of the shield to the ground, as shown, but then a substantial potential may appear between the end of the shield and the receiver chassis. As a rule, the cable ends are connected metallically to the receiver case, so that a continuous ground loop exists. The dotted line in Figure 1.29 represents such a connection. An alternative procedure to minimize the current in the ground loop is to operate either the source or the receiver without its own ground connection.

The above principle is most effective when conductive coupling is not required in the original circuit, such as when an ac signal and transformer coupling are used. Even with dc, a balanced system can be used, but it may be less satisfactory. In some cases, dc-to-ac inverters have been inserted in signal circuits to reduce pickup in the balanced cable.

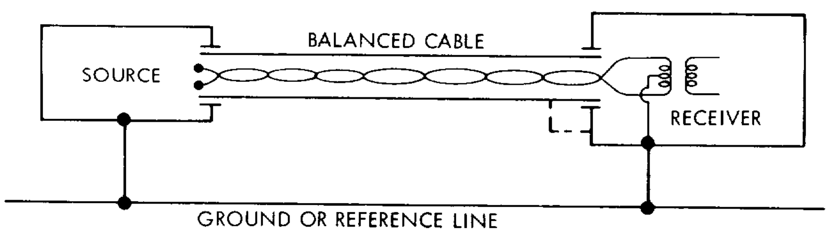


Figure 1.29.—Balanced coupling circuit.

3. Ground Loops

Ground loops arise where parallel grounds are necessary. For example, where coaxial cable is required for low-loss or distortionless signal transmission, the outer conductor may be in parallel with the power supply ground. A magnetic field in the loop formed by the parallel grounds may result in an induced current at the frequency of the magnetic field. A magnetic field arising from switching or pulsed currents may have a broad frequency spectrum. To reduce the effects of such loops, the grounds for power circuits should be run close to the signal cable. Triaxial cable may also be used so that a portion of the cable may be insulated from such loop currents.

5.4 PENETRATION THROUGH SHIELDS

Radiated energy may be coupled into a susceptible device through a shield of inadequate thickness, holes provided for ventilation and other purposes, and imperfectly joined shield sections. Precise calculation of shielding effectiveness, even for perfectly joined, solid shields, depends on the form of the shield and the type of field for which the shielding is meant. Both electric and magnetic coupling can occur; but normally it is easy to provide electric shielding, so that is usually not a serious problem. Magnetic shielding is more difficult to provide, particularly at frequencies below 100 kHz. To avoid uncertainties in critical situations, tests should be run to check shielding effectiveness. Such tests require the establishment of a known field and a measurement of the insertion loss introduced by the shielding.

Interference may also be created by certain nonlinear characteristics in the metal that forms the equipment shield. Nonlinearity gives rise to cross-modulation components that are manifested in the receiver output. Such effects usually are created only when the equipment is in a relatively strong RF field, such as that which exists in the vicinity of broadcast and radar transmitters. Nonlinear effects are known to arise in magnetic materials, such as steel, nickel, and mu-metal, and in corroded metals and corroded joints (especially loose joints). Rough or oxidized surfaces on steel tend to increase the nonlinear effects. Consequently, if the surface is coated with a nonmagnetic conducting material, such as copper, nonlinear effects become significantly reduced. Welded joints also have been found to be superior to riveted or bolted joints. Reference 50 gives further details of these effects.

5.5 SHIELDING AND BONDING

Shielding of VHF equipment against stray fields requires adequate bonding at the seams and maintenance of adequate shielding at the ventilation louvres. The thickness of the shielding material is, as a rule, no problem; if the shield is thick enough to be adequate mechanically, it will usually be adequate electrically at VHF.

The function of a shielded enclosure is to terminate the RF fields that exist both inside and outside the shield. The terminated fields give rise to surface currents that flow on the inside and outside surfaces of the metal shield. In a perfect situation, these surface currents are able to circulate in an uninterrupted manner on both surfaces of the shield.

However, if there are conductive imperfections at a seam that joins sections of the shield, then part of the inside surface current will be able to flow to the outside surface; similarly, part of the outside surface current will be able to flow to the inside surface. Consequently, part of the field inside the equipment becomes propagated outside the shield; in the same manner, part of the outside field is set up inside the shield.

Perforations, louvres, and so forth are obviously essential for adequate ventilation in the equipment. Consequently, it is generally not possible to have an ideal solid shield. Small perforations in the metal or sections of conductive screening usually provide satisfactory shielding at ventilation louvres since the surface currents will flow around the openings without appreciable penetration. Significant surface currents are forced onto the opposite surface mainly at seams, at long louvres, and at imperfectly joined shielded cables. Whenever conducting seams are required, soldered or welded joints are usually preferred. Pressure joints must be clean; abrasive gasket material that is also conductive may be useful. Further details on shielding in general are given in Section II and in Chapter 5 of Reference 3.

5.6 SHIELDING AND FILTERING

Duct filters are sometimes used to improve the shielding inside ventilation louvres. They consist of an array of closely spaced parallel tubes. The tubes are conductive and have a narrow cross section. Consequently, the tubes act as waveguides below cutoff frequency and present a relatively high attenuation to any fields incident on the filter. When the tubes are operated below cutoff frequency ($\lambda >> \lambda_c$), the attenuation of the tubes in dB per unit length is given approximately by

$$a = \frac{54.6}{\lambda_c} \quad , \tag{1.76}$$

where length is measured in the same units as wavelength for cutoff, λ_c.

Figure 1.30 shows the attenuation properties of a circular waveguide operating well below cutoff. It can be seen that the smaller the cross section of the tube, the greater is its attenuation. For example, suppose that a tube attenuation of 30 dB/cm is required in a certain ventilation assembly. From Figure 1.30, it can be seen that a tube of about 4-mm radius is needed, for the TE_{11} mode taken to represent the "worst case" (lowest attenuation). The corresponding cutoff frequency of the tube is 20 GHz. Therefore, the desired attenuation can be obtained at frequencies up to 10 GHz. A tube length of 4 cm will therefore provide about 120 dB of total shielding effectiveness in the filter, which is generally considered to be more than adequate for most applications.

The effective attenuation is much more readily calculable when ventilation is provided by tube assemblies than when holes or louvres in a panel are used. Edge effects tend to introduce large errors in the calculation of louvre attenuation.

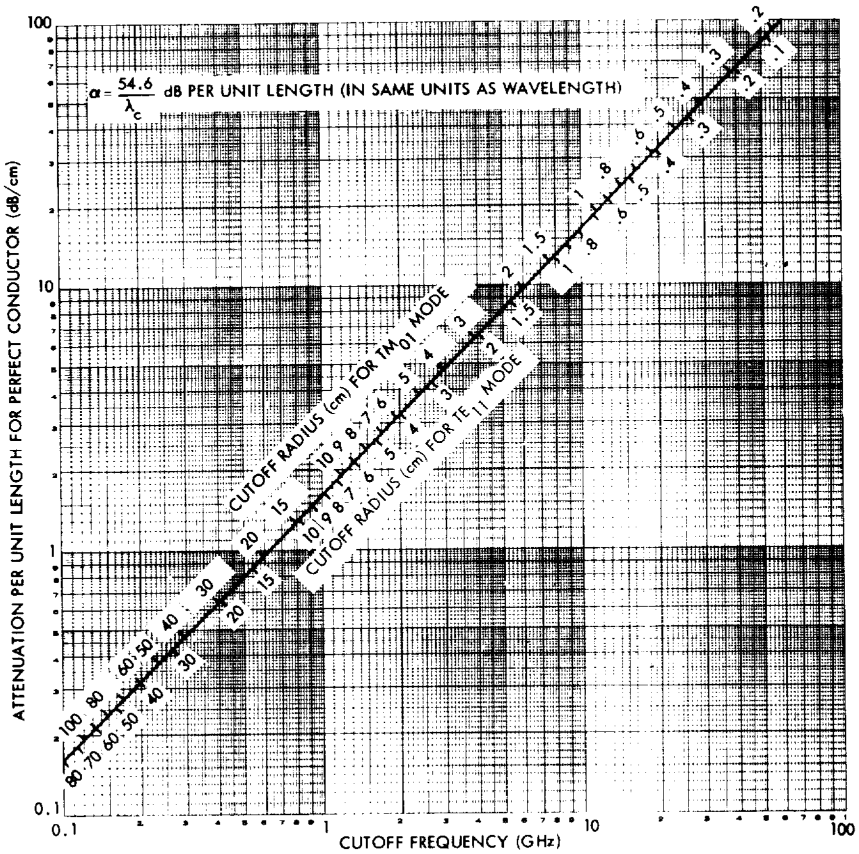


Figure 1.30.—Attenuation properties of circular tubes used in air ducting assembly.

6. INTERFERENCE CONTROL

6.1 INSTRUMENTS AND MEASURING METHODS

The measurement of RFI noise and the susceptibility of equipment to RFI noise is a vast field that in a brief account is apt to be oversimplified. The following United States Government military specifications will be found to be germane:

MIL-I-6181D, "Interference Control Requirements, Aircraft Equipment."

MIL-I-11748B (Sig. C), "Interference Reduction for Electrical and Electronic Equipment."

MIL-I-16910B (Ships), "Interference Measurement, Radio Methods and Limits, 14 kc to 1000 Mc."

MIL-I-26600 (USAF), "Interference Control Requirements, Aeronautical Equipment."

MIL-STD-826 (USAF), "Electromagnetic Interference Test Requirements and Test Methods."

Industrial standards that are pertinent include the following:

C63.2-1963, "Radio-Noise and Field Strength Meters, 0.015 to 30 Mc/s," American Standards Association.

C63.3-1964, "Radio-Noise and Field Strength Meters, 20 to 1000 Mc/s," American Standards Association.

C63.4-1963, "Radio-Noise Voltage and Radio-Noise Field Strength, 0.015 to 25 Mc/s," American Standards Association.

No. 107-1964, "Methods of Measurement of Radio Influence Voltage of High Voltage Apparatus," National Electrical Manufacturers Association.

It is proper to distinguish between three different kinds of testing methods according to the purpose of the measurement. In the first, the measurements are part of the design procedure. The designer will perform various tests with laboratory apparatus to measure the interference properties of the equipment being developed. In the second, measurements are made for the purpose of determining the detailed nature of RFI noise and how it affects systems. These measurements are tailored to fit specialized requirements, are done in the analysis laboratory or in the field under conditions simulating actual use, and frequently involve statistical analysis. In the third, the measurements are made on the final equipment as part of a series of tests to determine general acceptability of equipment. Measurements of the third kind follow standard procedures and involve interference limits that virtually guarantee interference-free operation in normal use. The military and civilian specifications mentioned above describe tests and test equipment for use in connection with measurements falling into the third category. Following is a discussion of the third class of measurements in terms of their intent, applicability, and limitations.

The tests are intended to reveal both the susceptibility of sensitive devices and the RFI noise-generating properties of electrical equipment—in radiation or induction modes and in conduction modes. Well-established standards have been set by the military for application at frequencies up to 1000 MHz; beyond this frequency the procedures and limits are still tentative.

6.2 SUSCEPTIBILITY TESTS

The conventional methods for the evaluation of component susceptibility make use of two types of waveforms: the sine wave and the recurrent impulse. The first concentrates energy at one frequency; the second disperses energy over a wide band of frequencies but concentrates it in time. Circuits containing tuned elements are sensitive in a narrow band and will register a large response when the input energy is concentrated within the passband. Sine waves are most effective in this case. Impulses, on the other hand, are more effectively used with broadband circuits because such circuits yield a larger peak response to impulses of a given energy than to sine waves of the same energy. Quantitative relations were discussed in Section 4.1. The large peak response will reveal nonlinear mechanisms that might otherwise go undetected. Where overdesign against interference is too costly, attention should be given to the expected environment, and the tests should be tailored to the need.

For testing purposes, typical impulse sources generate uni-directional, short impulses lasting anywhere from 10^{-9} to 10^{-10} s. Correspondingly, the energy is spread from near-zero frequency to the 1 to 10 GHz range. An RF generator producing pulses of about 10^{-7}-s duration and with a peak amplitude two orders of magnitude less than the impulse peak will produce approximately the same spectral density as the impulse over a band of about 10 MHz. Since many potentially interfering pulse sources are, in fact, better simulated by pulses, RF pulse testing is in many cases advisable.

Tests with sine waves are also carried out to determine certain nonlinear behavior. Susceptibility to subharmonics of tuned frequencies will utilize a single sine wave of large enough amplitude to create the effect. Some kinds of tests for nonlinear mechanisms (e.g., intermodulation and crossmodulation) require more than one sine wave. Finally, it is customary to use a modulated sine wave, particularly in testing receivers, in order to get an identifiable output.

1. Conducted Susceptibility Tests

Conducted susceptibility tests are carried out on electronic components connected to power sources or to other system components. Ideally, such tests are made on all external leads. For conducted noise on power lines, standard procedures have been devised (see the specifications listed above). Typically, for measuring susceptibility to audio voltages in series with the power line, a setup as shown in Figure 1.31 is used (see MIL-I-6181D). The capacitor C acts to ensure low impedance at the power supply at all frequencies of the test, and the test is carried out over the entire audio range (50 to 15,000 Hz). The purpose of the test is to insert, by means of a source of certain impedance, an audio voltage of a given magnitude in series with the power supply. The various

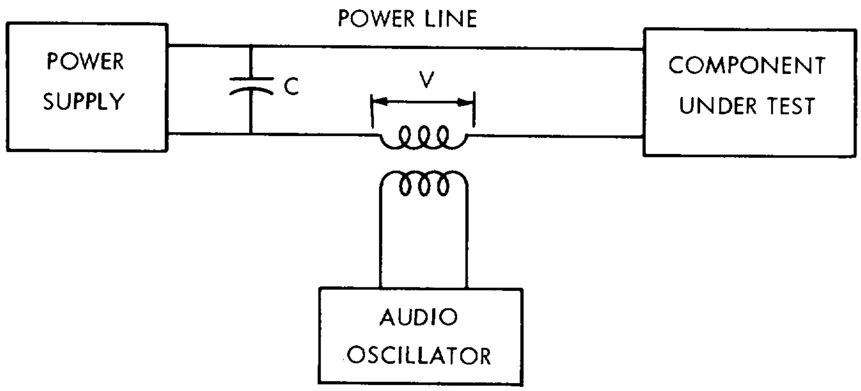


Figure 1.31.–Typical setup for performing audio susceptibility test on a power line.

standard specifications invoke slightly different test conditions. For instance, one requires that the effective open-circuit voltage inserted in this way be 3 V and that the impedance be less than 0.6 Ω. Under these conditions "no change in indication, malfunctioning, or degradation of performance shall be produced."

Sine wave tests of RF susceptibility on power lines are also fairly well standardized. Each ungrounded line is tested, sometimes with the arrangement shown in Figure 1.32. In effect, the noise voltage from the RF signal generator is applied between a "hot lead" on the power line and ground. The "line stabilization network" is used to standardize the impedance looking back from the unit under test into the power source. The impedance looking into terminals 1-3 on the network varies from about 5 Ω to about 50 Ω for the conditions shown over the range of frequencies for which the device is used. Specifications vary on the conditions to be employed. Typically (see MIL-I-6181D), the RF sine wave is amplitude modulated at 400 to 1000 Hz, and the modulated signal is applied with a carrier level of 0.1 V applied over a range of frequencies from 0.15 kHz to 10 GHz. The unit under test must not malfunction under these conditions. Various deviations from this procedure are quoted. In one procedure (MIL-STD-826), output from the signal generator is applied between a line and ground using any suitable device to isolate the power line.

Similar procedures are sometimes prescribed for impulse testing. One specification (MIL-I-11748B) requires that the circuit of Figure 1-32 be used with a standard impulse generator replacing the RF signal generator. Proper functioning is required when the impulse spectral density is 90 dB above 1 μV/MHz.

When leads other than the power leads are to be examined, similar methods are used. The specifications listed above do not prescribe any fixed

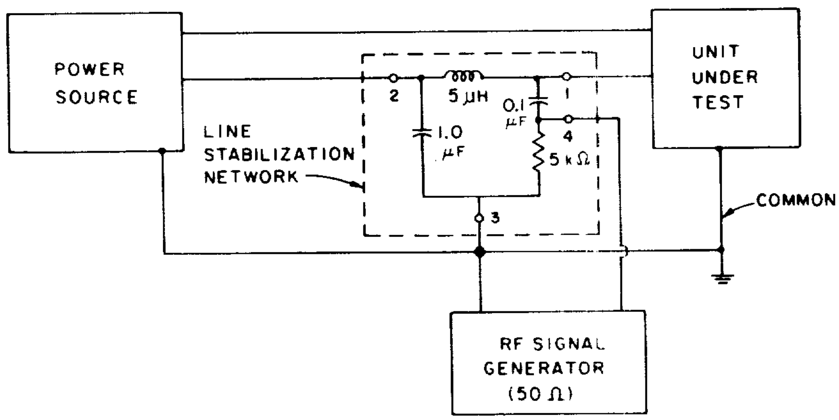


Figure 1.32.– Typical setup for performing RF susceptibility test on a power line.

methods and procedures; these should be tailored to the requirements of the equipment and to the environment. Broadband inputs to control leads and to input leads will ordinarily result in some measurable effect since the circuits to which the leads are connected are meant to accept some of the spectral components in the broadband source. Either levels of input RFI noise must be appropriately limited or filters must be used to eliminate RFI noise in the band to which the device is normally receptive.

The term "proper functioning" is not precisely defined in the specifications mentioned; it is a matter of choice on the part of the engineer to determine whether or not the equipment is adversely affected. A criterion sometimes used is that the RFI noise injected by the test source shall cause no effect beyond that produced by the internal noise of the system.

2. Radiated Susceptibility Tests

Examples of procedures, which again generally differ from one another only in detail, are described in the previously listed military specifications. All of these specifications specify a sinusoidal test frequency, audio modulated; and the range of testing extends from 14 kHz to 20 GHz. Impulse testing has been proposed in some quarters, but such methods are not common practice. All of the accepted procedures involve the placement of the component under test in a known field, or in a prescribed location with respect to a standard source, and evaluating the effect. Normal signal inputs, such as the antenna input on a receiver, are terminated in suitable, shielded, dummy sources. Tests are carried out with the largest amplitude expected and, as in the case of conducted susceptibility, the effect is required to be "insignificant."

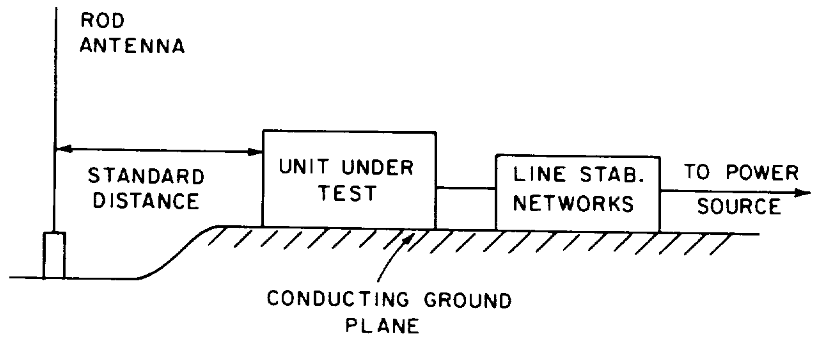


Figure 1.33.–Typical equipment arrangement for making radiated susceptibility measurements.

A typical test setup, illustrated in Figure 1.33, shows the use of a rod antenna placed some standard distance from the equipment under test. This distance is specified as 1 ft in MIL-I-6181D, 1 m or 7.6 m (depending on the equipment class) in MIL-STD-826, and 25 ft in MIL-I-11748B. The latter specification allows reduction in the distance as frequency is increased; the distance quoted is that specified for low frequencies and is apropos for measurements with a rod antenna.

As frequency is increased, the rod is replaced by a dipole. The point at which the dipole is used is either 25 or 50 MHz, depending on the specification employed. An untuned dipole is required in two of the specifications, but tuning is used in all cases above 50 MHz. As a rule, tests can be carried out with the component oriented in a number of ways; the four sides facing the source is specified by one test.

The tuned dipole is specified up to 1000 MHz. Beyond this, and sometimes at lower frequencies, directive antennas (horns, parabolic dishes, discones, and so forth) are used. At frequencies greater than 50 MHz, tests with both horizontal and vertical polarization are required by MIL-I-11748B.

3. Other Tests

The mechanisms of spurious signal susceptibility, described in Section 4.2, are examined experimentally with various standard test procedures. A set of such procedures is fully described in MIL-STD-826 and will therefore not be repeated here. Tests are specified in the aforementioned specification for such mechanisms as two-signal intermodulation, spurious response to a single unwanted signal, and cross-modulation. In each case, the methods simply involve the use of standard signal generators as calibrated sources representing both wanted and unwanted signals. Generally, however, a great deal of care is required in setting up and performing the tests since most signal generators themselves produce spurious outputs that can cause confusion in interpreting test results. For tunable receivers, it may be necessary to make tests at many points of tuning since the sensitivity to unwanted signals is not constant throughout a given frequency band. Some standard procedures compromise by requiring tests at a few points in each band of tuning; e.g., points near the edge of each continuously tuned band and one point near the middle of the band. Specification MIL-STD-826 requires scanning throughout a frequency range. This is a time-consuming operation. As a rule, the spurious responses will be found where they are expected; it may save much time, particularly for testing at the production stage, to examine only those points at which such responses are expected.

When testing for such mechanisms as spurious response and cross-modulation, a point to be borne in mind is that the magnitude of the response will not always be linearly related to the input signal level. The mechanisms are nonlinear and, as a consequence, the interference response is nonlinear. Tests should be carried out using the maximum input level of the unwanted signal for which interference protection is required.

6.3 TESTS OF RFI NOISE OUTPUT

Tests that measure RFI noise output are performed with setups that are not too different from those used for susceptibility testing. The important distinction is that an RFI measuring instrument replaces the signal (or impulse) generator. Test set instruments for measuring RFI are available for frequencies ranging from low audio into the microwave range. Though manually tunable instruments are used most often, there is a tendency to utilize either mechanically tuned devices or electronically swept spectrum analyzers to speed data collection.

The two major problems of measurement are the following. First, measurements of electromagnetic fields at distances close to the source are difficult to reproduce because they are a function of antenna location and reflections from surrounding objects. Furthermore, unlike measurements taken in the far field of a source, measurements in the near field are almost impossible to extrapolate to give the field level at other distances. However, if measurements are attempted at a greater distance from the source, the measured value is naturally smaller and the overriding effects of other RFI sources in the environment begin to be felt. Though shielded enclosures for tests are not always required, they are often beneficial. The maximum distance is naturally restricted in such instances. The second problem concerns the measure of the RFI noise itself (Reference 51). The instruments under discussion all have a moderately narrow passband; typically, bandwidths are of the order of 0.1 to 1.0 percent of the center frequency. For the measurement of pure sinusoids, all the instruments can be calibrated to give the same reading. For broadband inputs, the readings will depend on the detailed nature of the input, on the bandwidth, and on the function performed by the detectors following the RF-intermediate-frequency amplifier.

Three types of detectors will be found on standard instruments: the average-of-the-envelope detector, the peak detector, and the quasi-peak detector. Some instruments contain all three, and some omit the quasi-peak detector. As the name indicates, the average-of-the-envelope detector reads the average of the envelope of the output of the IF amplifier. The peak detector reads the peak value of the IF amplifier output. The quasi-peak detector, shown schematically in Figure 1.34, reads less than the peak value and is, in effect, defined by its charging and discharging time constants. The time constants are measured by applying a unit RF step input and determining the time to reach 63 percent of the final output or by suddenly removing the input and determining the time required to fall 63 percent of the way to zero. The accepted charging and discharging times, according to United States standards, are 1 ms and 600 ms, respectively.

The response of standard RFI test instruments to various inputs is shown in Table 1.1 (Reference 52); all instruments are assumed to have been calibrated to read the root-mean-square (RMS) value E of an applied sine wave. The symbols used in Table 1.1 are defined in Section 4.1.1. The input quantities given in Table 1.1 have been measured with an RMS detector, too, though standard instruments are not ordinarily equipped with such a device. The impulse input is a periodic pulse that occurs with a period not so high that successive pulses will overlap one another in the RF or IF amplifiers. The quantity S_i is defined as the impulse strength and is twice the Fourier spectral intensity at the tuned frequency (see Section 4.1). For a pulse of duration $\tau(1/\tau >>$ tuned frequency) and height A,

$$S_i = 2A\tau \ . \tag{1.77}$$

The response of quasi-peak detectors to periodic impulses depends on the recurrence frequency of the pulses, on the bandwidth, and on the time constants. This relationship is shown in Figure 1.35.

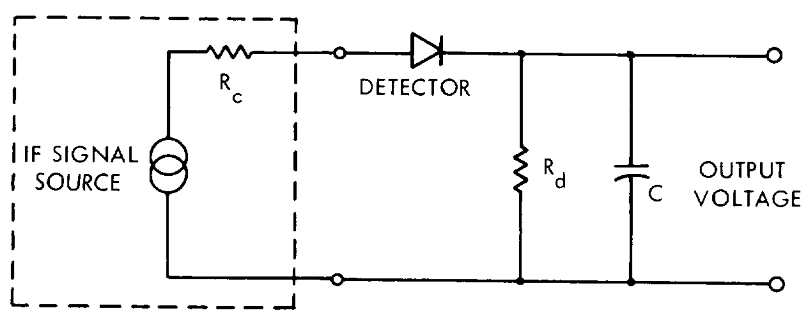


Figure 1.34.–Circuit of a quasi-peak detector.

Table 1.1.—Response of an RFI meter.*

Input	Type of Detector			
	Peak	Average	Quasi Peak**	RMS
Sine wave	E	E	E	E
Periodic impulses (rate = f_p/s)	$\frac{S_i B_e}{\sqrt{2}}$	$\frac{S_i f_p}{\sqrt{2}}$	$\frac{S_i P(a) B_e}{\sqrt{2}}$	$\frac{S_i \sqrt{f_p B_p}}{\sqrt{2}}$
Random noise	—	$0.884\sqrt{B_p N}$	$1.820\sqrt{B_p N}$	$\sqrt{B_p N}$

*N, B_e, and B_p are defined in Section 4.1.1.

**Charge time = 1 ms; Discharge time = 600 ms.

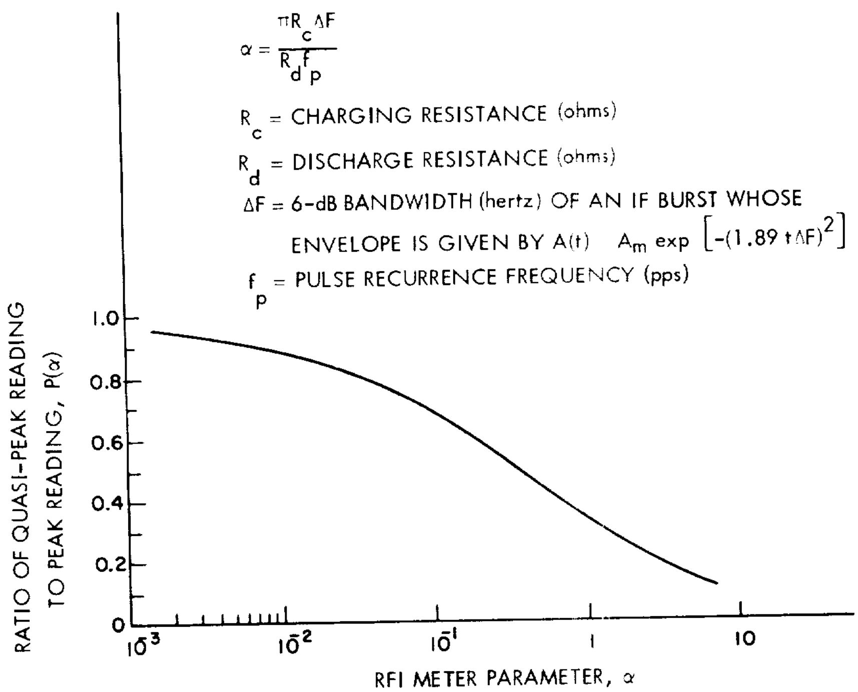


Figure 1.35.—Response of a quasi-peak detector as a function of RFI meter parameters.

1. Conducted RFI Noise Measurements

Measurements of the RFI noise level at the output of an electrical apparatus operating from the power input line are frequently made with the setup of Figure 1.32. The RF signal generator is replaced by an RFI measuring instrument. The line stabilization network prevents variability of the power-line impedance from affecting the readings. In some cases, particularly in instances where the power drawn by the equipment under test is high, alternative methods that use a capacitive connection to the line can be used. Another test involves making an inductive measurement of the RFI noise current on the line through the use of a clamp-on toroidal pickup coil. In the latter method, the RF impedances of the power source and the line do affect the readings. These tests also can be performed on other interconnecting cables, such as control lines.

2. Radiated RFI Noise Measurements

Measurements of radiated RFI noise field intensity in the vicinity of an electrical device are made with a setup (essentially that of Figure 1.33) in which the antenna feeds a standard RFI measuring instrument. As in the case of susceptibility measurement, the distance from the source and the details of the setup depend on the particular application. At low frequencies, loops and rods are used; at higher frequencies, dipoles and horns are used, as discussed in Section 6.2. Because the radiated field often depends on location, it is common practice that the test sample be turned until the point of maximum emission is found.

7. SITE SELECTION

The factors influencing site selection for ground stations are discussed here in brief. The material presented herein is a summary of the interference study program reported in Reference 53 (Section 2).

Operational requirements largely dictate the selection of the geographic region in which the station is to be located. However, topography criteria and RFI-environment criteria also influence site selection.

In order to ensure that the RF communication link between satellite and ground stations is as reliable as possible, it is essential to establish a ground station in noise-free areas where the level of local ambient RFI noise is as low as possible and where no nearby high-powered ground or airborne transmitters operate on the same or adjacent frequencies.

In the past, the most suitable topography for a data acquisition site has been a very shallow "hollow" composed of a relatively flat plain surrounded by distant hills with an elevation angle not exceeding 10 degrees. The hills effectively shield the site from unwanted RFI noise being propagated along the ground.

7.1 FACTORS THAT INFLUENCE THE RFI ENVIRONMENT OF A SITE

The factors upon which the RFI environment of a site depends may be summarized briefly as follows:

(1) Natural and man-made ambient noise background at a site.

(2) Proximity of fixed and mobile transmitters operating on the same frequency or on frequencies adjacent to those used for the satellite-to-ground communications.

(3) Propagation characteristics of the area in which the site is to be located. (This includes the soil conductivity of the region, the roughness of the terrain, and the vegetation of the terrain.)

(4) The natural shielding that a site possesses.

Many of these factors are obviously closely interrelated. For example, the ambient RFI noise level will be closely dependent on the natural shielding of a site and also, to a certain extent, on the propagation conditions of the region. These factors will now be considered in detail.

7.2 NATURAL AND MAN-MADE AMBIENT NOISE LEVELS

At VHF and UHF, man-made noise generally represents by far the largest part of the overall background noise level at any particular site. For comparison, Figure 1.36 shows that noise levels from celestial sources are lower than the man-made noise levels; however, celestial sources can become significant sources of interference, especially for narrow-beam, high-gain antennas (Reference 54). Figure 1.36 shows the relative ambient noise levels expected from different sources, over the 100 to 600 MHz frequency range, when a half-wave dipole antenna is used as described by Skomal (Reference 55). Most sources of man-made noise are naturally concentrated in heavily industrialized urban population centers. Such sources include automobile ignition systems, corona discharge from power lines, fluorescent lights, switching systems, motors, rectifiers, and arc welders. Figure 1.36 presents Skomal's data for the noise background in a typical city and at points approximately 10 mi and 25 mi from the city center. The RFI noise levels within a 10-mi radius of the city center are prohibitively high, thus making it undesirable to locate a sensitive receiving system within or very close to a city. Figure 1.37 shows the average fall-off of the RFI noise levels with distance from the urban center, at 200 and 400 MHz.

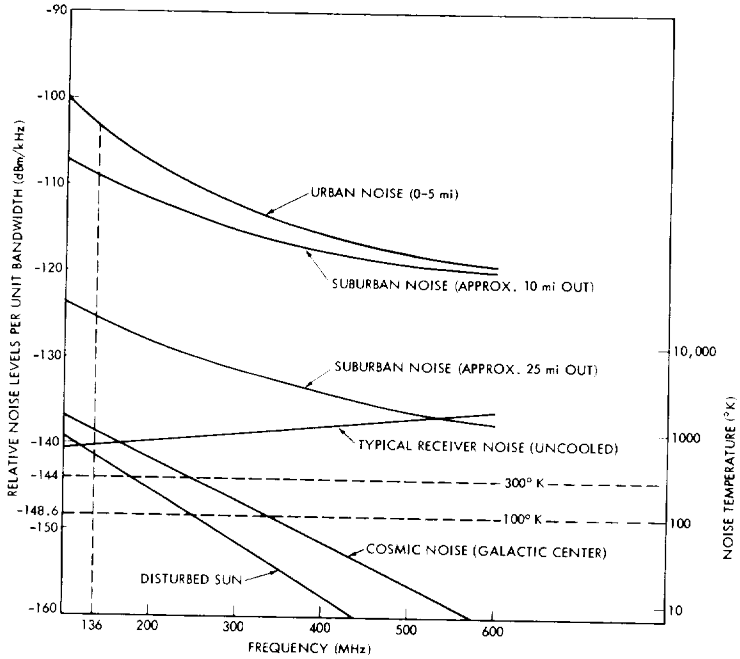


Figure 1.36.—Relative levels of natural and man-made noise measured with a half-wave dipole antenna.

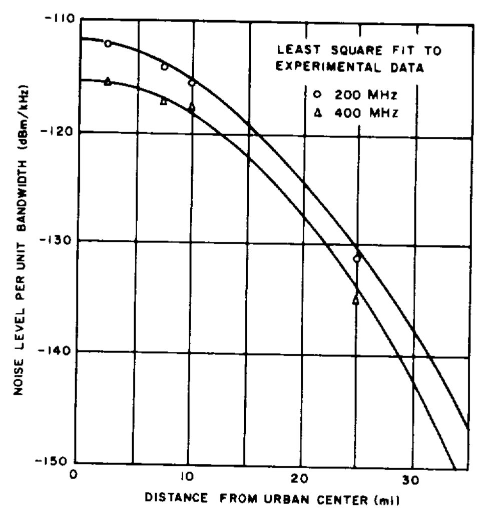


Figure 1.37.—Attenuation of man-made noise with distance from an urban center for 200 and 400 MHz. (After E. N. Skomal, IEEE Trans., 1965)

Various authors have attempted to correlate the level of man-made RFI noise existing in an urban environment with the population of the town or city. However, because of the general lack of sufficient data, it has not been possible to obtain meaningful correlations. Skomal has concluded that the data shown in Figures 1.36 and 1.37 are applicable to any town or city with a population of 50,000 or greater. Using experimental evidence, Skomal furthermore concludes that the major source of man-made RFI noise at VHF and UHF is automotive ignition interference. Satellite data acquisition ground stations should be located away from major highways to ensure that the sensitive receiver systems are not disturbed by ignition noise.

There are usually many factors that influence the selection of a specific site for a given ground station. Some of these factors are discussed in the following. Since there are obvious practical and economic advantages to locating a site in proximity to towns or cities, the question that remains is, how close to a town can such a site be located and still continue to operate without interference? From the data already presented in this section, it may be safely concluded that stations of the Minitrack type (Reference 56) that employ a vertically directed fan beam can be located as close as 30 to 35 mi to a town of 50,000 population or greater. However, a site that effectively shields the receiving system from urban noise could actually be even closer to the town. Sites for data acquisition facilities, which use high-gain steerable antennas, should be located further away to eliminate the possibility of serious interference when the antenna is pointing directly at the town; a minimum distance of about 50 mi is usually a good compromise. However, this distance will be modified by the height of the antenna above ground and by the irregularity of the terrain.

Finally, an RFI noise-measurement survey, at the frequency bands of interest, should be made at several alternative sites before a specific site selection is made.

7.3 FIXED AND MOBILE TRANSMITTERS OPERATING ON THE SAME OR ADJACENT FREQUENCIES

Certain frequency bands have been allocated by the International Telecommunications Union (ITU) for the transmission of telemetry data from scientific satellites. Those frequency bands, which are presently of concern in data acquisition ground station operations*, are as follows:

(1) 136 to 138 MHz.
(2) 400.05 to 402 MHz.
(3) 1700 to 1710 MHz.
(4) 2200 to 2300 MHz.

*Other frequencies are being used by NASA for such purposes as communication satellite and deep space probe down links (see Reference 57).

The location data for fixed transmitters in these bands, throughout the world, are available from the International Frequency List compiled by the International Frequency Registration Board in Geneva (Reference 58). Worldwide maps showing fixed transmitter locations for the 136-, 400-, and 1700-MHz space research bands are given in Reference 53.

Although the exclusive allocation of these frequency bands to the space telemetry service has been realized within the United States, this is not the case throughout the rest of the world. In many nations, these frequency bands are, in effect, shared by both the space service and fixed and mobile services (or other services). Consequently, low-power transmitters continue to operate in these bands.

The fact that fixed and mobile services continue to operate in frequency bands adjacent to the four bands assigned for space research has created the possibility of adjacent-channel interference to ground station operations, such as interference from the United States aeronautical mobile service, which operates at frequencies immediately below 136 MHz. Some sites are particularly susceptible to interference from airborne transmitters, because of the lack of path attenuation between transmitter and site. This problem is further discussed in Section 7.3.3.

1. Radio Regulations of the International Telecommunication Union

The first provisional allocations of certain bands of the radio spectrum to the space telemetry service were made at the Administrative Radio Conference held at Geneva in 1959 (Reference 59). These allocations were later confirmed and expanded at the Extraordinary Administrative Radio Conference (EARC) held at Geneva in October, 1963. The revised radio regulations, which were agreed to at the EARC, were put into effect in the United States on January 1, 1965.

The details of the four space research allocations that were made at the EARC, as well as the existing allocations in the adjacent bands above and below the four bands, are given in Section 7.5 (Reference 60).

2. Interference From Airborne Transmitters

The lack of sufficient path attenuation between an airborne transmitter and a site can cause interference because of the consequent high power levels arriving at the site. The expression for the power (in watts) received at a 136-MHz telemetry receiver input is given by

$$P_r = \frac{P\lambda^2}{8\pi^2 \times 1600^2} \frac{G}{R^2}, \tag{1.78}$$

assuming a transmitter antenna gain of 3 dB, where

P = the output of the airborne transmitter in watts,

λ = the wavelength in meters,

G = receiving antenna power gain (above isotropic),

and

R = distance in miles between transmitter and site.

This expression has been plotted as a function of distance in Figure 1.38 for an aircraft transmitter operating in the range of 135.0 to 135.95 MHz with a 25-W output and an antenna gain of 3 dB (hemispheric coverage). Curves are shown for three values of G, the receiving antenna power gain, corresponding to three of the different types of receiving antennas used at STADAN sites.

It can be seen from Figure 1.38 that the levels of power received from aircraft transmitters can be prohibitively high. The signal level received from an aircraft 100 mi away may be as much as 60 dB greater than that received from a satellite. Even if the aircraft is flying in a region outside the main lobe of the ground receiving antenna, there can be sufficient unwanted energy entering through the antenna sidelobes or backlobes to exceed the level of the satellite signal by 30 to 40 dB.

Whether or not interference is actually created depends on a number of factors, including the general density of air traffic in the area, the height at which aircraft fly, and the amount of time the aircraft are in communication with the ground. From Figure 1.38 it is evident that an aircraft flying at an altitude of 40,000 ft will be "visible" to a ground station as far away as 250 mi, if smooth earth conditions are assumed. Obviously, this distance also depends on the height of the receiving antennas above ground; an aircraft will be "visible" to the 85-ft parabolic dish antenna from much farther away than it will be to a Minitrack array since the dish antenna is mounted about 100 ft above the ground. Consequently, the higher an aircraft flies, the greater is the probability of interference. Also, we see that the data acquisition facilities, which employ the directional dish antennas, are more susceptible to aircraft interference than are Minitrack stations. Certain STADAN sites in the eastern United States have experienced more aircraft interference because they are located in regions of high air traffic density. Figure 1.39 (Reference 61) is a map of the peak density of air traffic in the United States for the busiest day of fiscal year 1964. This map shows the traffic flow for aircraft flying under instrument flight rules.

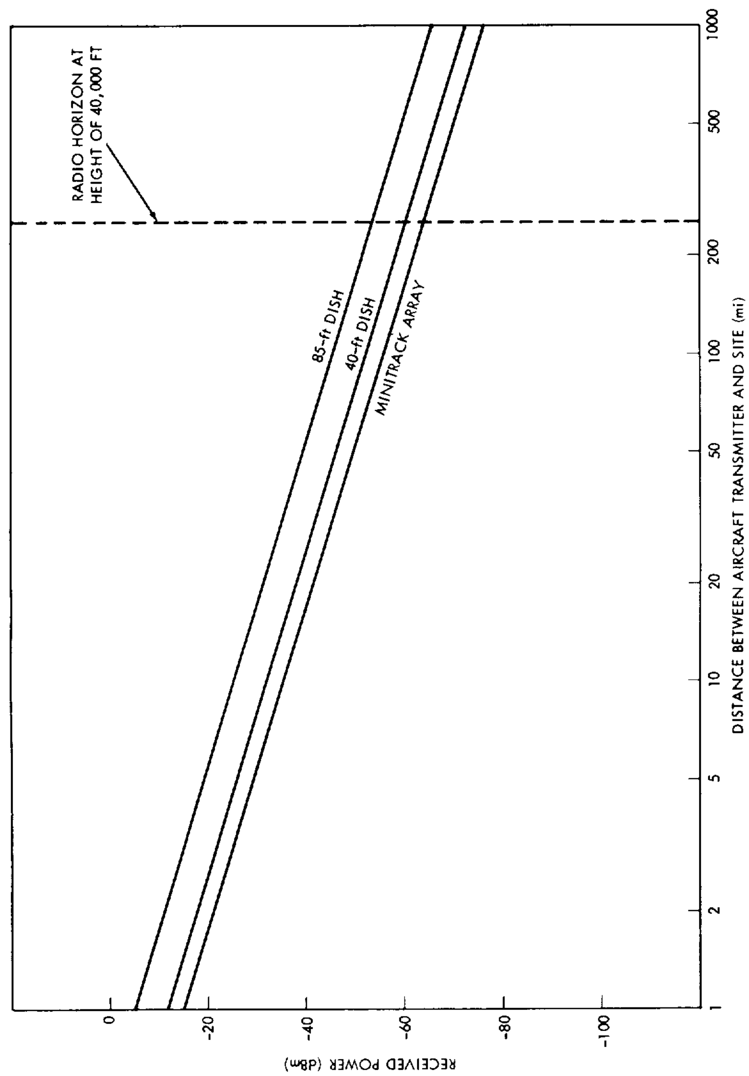


Figure 1.38.–Power level at input of a 136-MHz receiver from an interfering 25-W airborne transmitter.

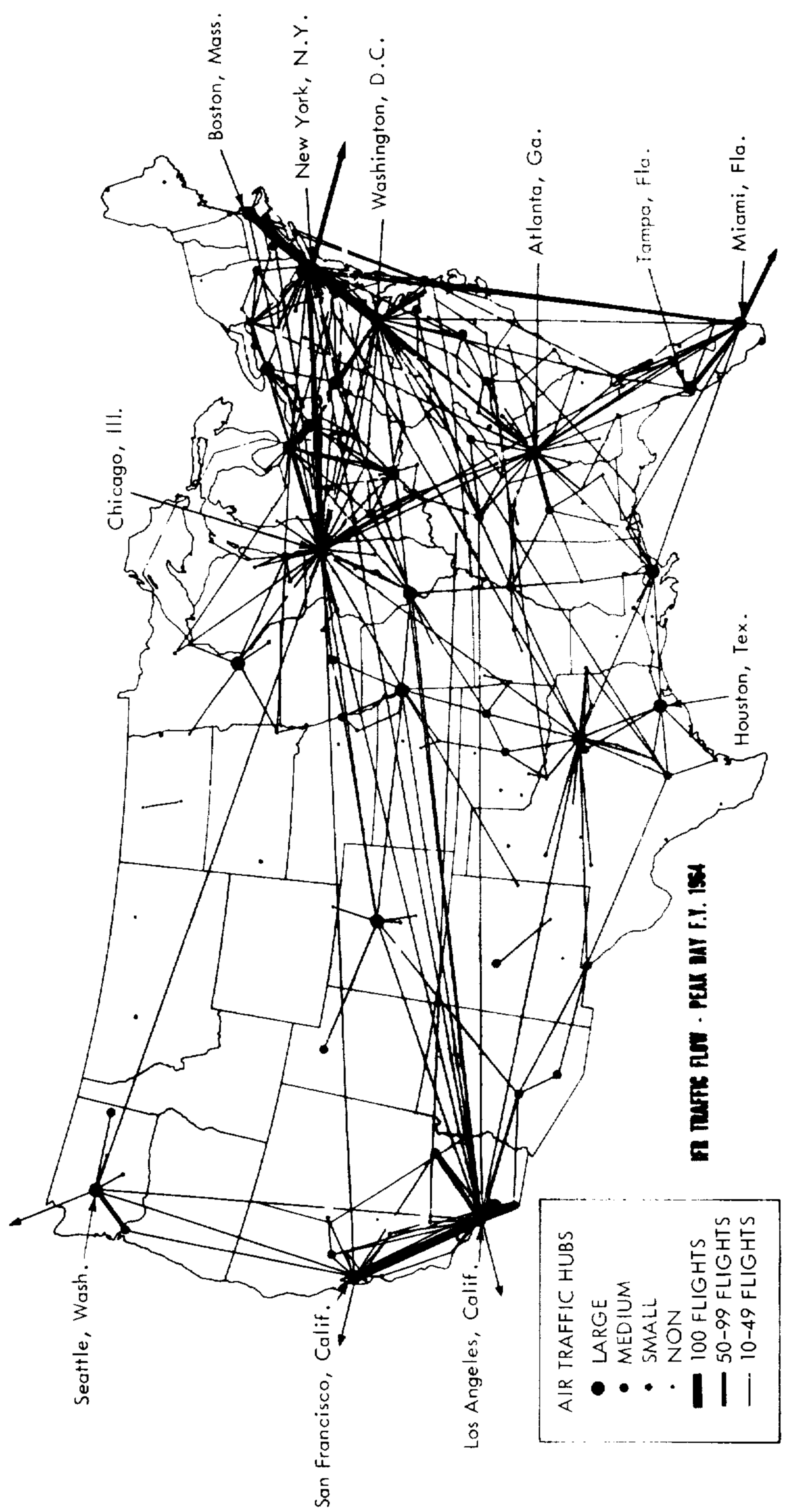


Figure 1.39.–Flow of air traffic within the United States for aircraft flying under instrument flying rules.

7.4 VHF PROPAGATION CHARACTERISTICS

At the frequencies presently used in the STADAN system, the ground or surface wave that predominates at the lower frequencies is rapidly attenuated so that propagation is restricted to the space-wave, which consists of a direct wave and a ground-reflected wave. For relatively short distances between the transmitting and receiving antennas (20 mi or less), line-of-sight conditions are said to exist and, to a good approximation, the earth may be considered to be flat. In this region, the direct wave and reflected wave create interference effects, so that the resulting change of field strength with distance contains many deep nulls, assuming a perfectly smooth reflecting earth (see Figure 1.40). Radio propagation above 40 MHz over irregular terrain is given in Reference 62. In practice, the earth has a rough surface and is not a perfect conductor. Consequently, the ground-reflected wave is both attenuated and altered in phase relative to the wave reflected from the smooth earth. The solid curve of Figure 1.40 (Reference 63) will therefore be modified to that shown by the dotted curve for actual rough earth conditions. The form of both solid and dotted curves is dependent on the height of the transmitting and receiving antennas and on the frequency used; increasing these parameters will greatly increase the number of nulls.

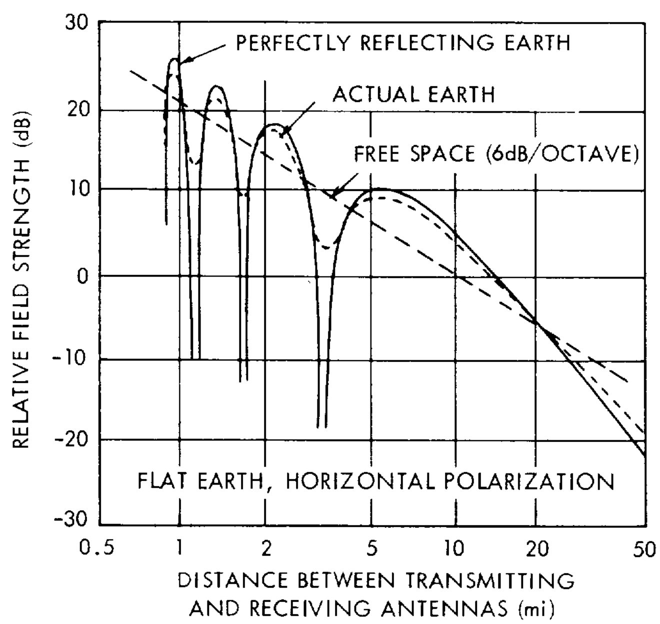


Figure 1.40.–Field strength as a function of distance for a smooth reflecting earth. (After F. E. Terman, Electronic and Radio Engineering, 1955)

For distances beyond the line-of-sight region, refraction and diffraction effects become very significant. The earth's curvature must now be considered, and in order to account for the change in refractive index of the earth's atmosphere with height, the effective radius equal to 4/3 times the true radius of the earth is generally used in propagation calculations. The first theory developed for wave behavior in this region was the Sommerfeld "smooth earth theory" (Reference 63), which computed the average bending of a wave over a perfectly smooth and perfectly conducting earth. However, experimental data accumulated over many years did not agree with the theoretical predictions of the smooth earth theory; in practice, signal strengths were found to be considerably greater than those predicted in theory. The increase in signal strength is largely attributed to two effects: diffraction of the wave (caused by surface irregularities) and scattering of energy in the troposphere.

Diffraction of a wave by a rough surface is difficult to evaluate theoretically since it depends very much on the type of terrain being considered, including its roughness and its conductivity. The results of some work done on knife-edge diffraction are occasionally applicable in the case where a single large mountain or hill is located between transmitting and receiving antennas. However, this situation is not commonly found in practice.

The scattering of energy that takes place in the troposphere (that part of the earth's atmosphere closest to the earth's surface) is thought to be caused by small discontinuities in the refractive index of the troposphere. The amount of energy scattered depends largely on the meteorological conditions, the time of day, the season of the year, and the latitude on the earth's surface at which scattering occurs. There is, therefore, a very wide hourly variation in the energy scattered by the troposphere.

Figure 1.41 (Reference 64) shows the change of both theoretical and measured field intensities with distance at a frequency of about 50 MHz. The curve for the measured data represents an average curve, since the experimental data vary over wide limits. It can be seen that the measured curve falls roughly between the curve derived from the smooth-earth theory and that predicted from knife-edge diffraction theory. The experimental curve probably falls close to that which can be predicted by rough-surface diffraction theory (References 65 and 66).

7.5 ALLOCATED SPACE RESEARCH FREQUENCY BANDS

Intelligent solutions to problems of radio frequency interference to, or arising from, spacecraft communications require knowledge of the frequencies on which spacecraft communication and data transmission may be carried out. The data in this section apply to the frequency bands that have been allocated to the space research service. The adjacent band allocations also have been

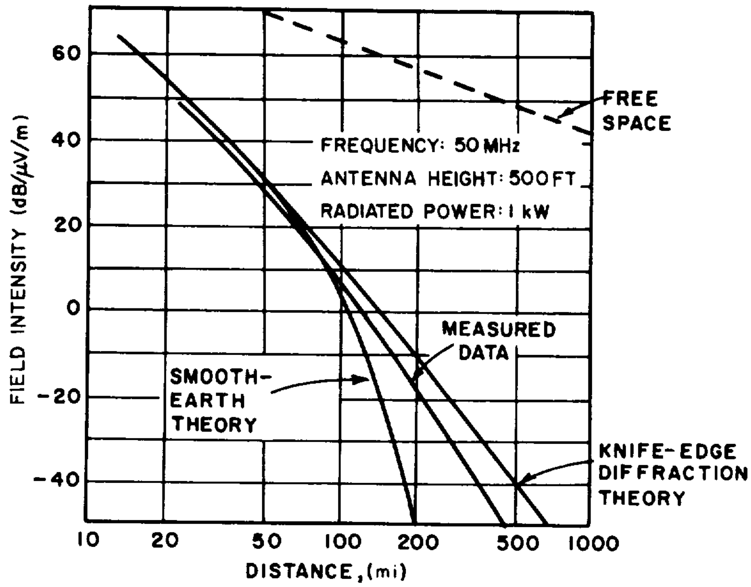


Figure 1.41.—Measured and theoretical field intensities as function of distance between transmitting and receiving antennas. (After K. Bullington, Proc. IRE, 1950)

included, however. The frequency allocation information presented in Table 1.2 represents a partial revision of the Geneva 1959 radio regulations adopted by the EARC, convened in Geneva in October 1963.

The allocations listed in Table 1.2 differ according to the different parts of the world. For purposes of frequency allocations, the ITU has divided the world into three different regions; the three columns of regional data in Table 1.2 correspond to the ITU worldwide divisions.

The names of services printed in capital letters (e.g., FIXED) represent "primary" services. If only one primary service is allocated to a particular frequency band, then that allocation is an exclusive one. If more than one primary service has been allocated to a particular band, then the band is shared by the services listed.

Permitted and primary services have equal rights, except that, in the preparation of frequency plans, the primary service has prior choice of frequencies over the permitted service.

Secondary service stations must not cause harmful interference to primary or permitted service stations that are already operating on assigned frequencies. Furthermore, they cannot claim protection from harmful interference from primary or permitted service stations.

Table 1.2 – Frequency allocations by international treaty.

Frequency Band (MHz)	Service	Notes and Comments Applying to Various Regions		
		Region 1	Region 2	Region 3
132.0 to 136.0	FIXED MOBILE	*Western Europe, Middle-East, North and East Africa*: AERONAUTICAL MOBILE *Western, Central, and Southern Africa*: FIXED, MOBILE	FIXED, MOBILE (AERONAUTICAL MOBILE is exclusive to the *U.S.A.*)	*All countries (except Australia and New Zealand)*: (Eventually to be exclusively allocated to AERONAUTICAL MOBILE) FIXED, MOBILE *Australia and New Zealand*: AERONAUTICAL MOBILE
136.0 to 137.0	SPACE RESEARCH (Telemetry and Tracking) The application of space communication techniques for use in Aeronautical Mobile service is permitted in this band. This is to be limited to satellite relay stations belonging to Aeronautical Mobile service.	FIXED, MOBILE, SPACE RESEARCH (Telemetry and Tracking)	SPACE RESEARCH (Telemetry and Tracking) This band also allocated to FIXED, MOBILE (indefinite in *Cuba*)	FIXED, MOBILE, SPACE RESEARCH (Telemetry and Tracking)
137.0 to 138.0	SPACE RESEARCH (Telemetry and Tracking) SPACE, OPERATIONAL (Telemetry and Tracking) METEOROLOGICAL-SATELLITE	*Norway, Switzerland, Turkey*: This band also allocated to Fixed, Mobile (except Aeronautical). *U.S.S.R. Bloc, Algeria, Morocco, U.A.R., Lebanon*: This band also allocated to Aeronautical Mobile indefinitely. *All other countries in Western Europe, Middle East, North and East Africa*: Until 1 Jan. 1969, was allocated to Aeronautical Mobile service. *Central and Western Africa*: This band also allocated to FIXED, MOBILE indefinitely.	This band also allocated to FIXED MOBILE until 1 Jan 1969 (indefinite in *Cuba*)	This band also allocated to FIXED, MOBILE until 1 Jan. 1969 (indefinite in *Malaysia, Pakistan, and Phillippines*) *Australia*: This band also allocated to BROADCASTING (TV)

Table 1.2 (Continued)

Frequency Band (MHz)	Service	Notes and Comments Applying to Various Regions		
		Region 1	Region 2	Region 3
138.0 to 143.6	FIXED MOBILE Radiolocation	*Austria, Denmark, Greece, Netherlands, Norway, Portugal, Sweden, Switzerland, Turkey, U.K., Western Germany*: This band will be allocated in the future to Fixed and Mobile (excluding Aeronautical) service. *Western, Central, and South Africa*: FIXED, MOBILE AERONAUTICAL MOBILE	FIXED, MOBILE Permitted Service: Radiolocation (In the U.S.A., this band is reserved for government use)	*All countries (except New Zealand)*: FIXED, MOBILE *New Zealand*: AERONAUTICAL MOBILE *Australia*: This band also allocated to Broadcasting (TV). *Taiwan*: This band also allocated to Radiolocation.
335.4 to 399.9	FIXED MOBILE		Reserved for government use in the U.S.A.	
399.9 to 400.05	RADIONAVIGATION-SATELLITE	*Bulgaria, Greece, Hungary, Lebanon, Morocco, U.A.R., Yugoslavia*: This band also allocated to Fixed and Mobile service.	*Cuba*: This band also allocated to Fixed and Mobile service.	*Iran, Kuwait*: This band also allocated to Fixed and Mobile service.
400.05 to 401.0	METEOROLOGICAL AIDS METEOROLOGICAL-SATELLITE (Maintenance Telemetry) SPACE RESEARCH (Telemetry and Tracking)	*U.S.S.R. bloc, Greece, U.A.R.*: This band also allocated indefinitely to Fixed and Mobile service. *U.K.*: This band also allocated on a secondary basis to Radiolocation.		
401.0 to 402.0	METEOROLOGICAL AIDS SPACE (Telemetry and Tracking) Secondary Services: Fixed Mobile (excluding Aviation)	*U.S.S.R. bloc, Greece, Norway, Sweden, Switzerland, Turkey*: This band also allocated on a primary basis to Fixed and Mobile (excluding Aviation) service. *France*: Meteorological Aids is the *only* primary allocation. *U.K.*: Secondary allocation to Radiolocation.		*Iran*: This band also allocated on a primary basis to Fixed and Mobile (excluding Aviation) service. *Australia*: The Space service operates in this band on a secondary basis only.
402.0 to 406.0	METEOROLOGICAL AIDS Secondary Services: Fixed Mobile (excluding Aviation)	All of the exceptions listed in the band 401 to 402 MHz also apply in this band (omit *Australia*).		

Table 1.2 (Continued)

Frequency Band (MHz)	Service	Notes and Comments Applying to Various Regions		
		Region 1	Region 2	Region 3
1690 to 1700	METEOROLOGICAL AIDS METEOROLOGICAL-SATELLITE	*Secondary Service*: Fixed Mobile (excluding Aviation) *U.S.S.R. bloc, Algeria, Lebanon, Morocco, U.A.R.*: This band also allocated on a primary basis to the Fixed and Mobile (excluding Aviation) service. *Austria, Finland*: Meteorological-Aids is the *only* primary allocation.	*Cuba*: Primary allocation also to Fixed and Mobile (excluding Aviation) service.	*Pakistan, Kuwait*: Primary allocation also to Fixed and Mobile (excluding Aviation) service. *Australia, Indonesia, New Zealand*: Secondary allocation to Fixed and Mobile (excluding Aviation) service.
1700 to 1710	SPACE RESEARCH (Telemetry and Tracking)	FIXED SPACE RESEARCH (Telemetry and Tracking) *Secondary Service*: Mobile	SPACE RESEARCH (Telemetry and Tracking) *Cuba*: Primary allocation to Fixed and Mobile services.	FIXED MOBILE SPACE RESEARCH (Telemetry and Tracking)
1710 to 1770	FIXED MOBILE	FIXED *Secondary Service*: Mobile *Switzerland*: Primary allocation to both Fixed and Mobile (excluding Aviation) services.	FIXED, MOBILE (Allocated for government use in the *U.S.A.*)	FIXED MOBILE
1790 to 2290	FIXED MOBILE Meteorological-Satellite	FIXED MOBILE *Switzerland*: Allocated to Fixed service and to Mobile, except Aeronautical mobile service. Band 2110 to 2120 MHz may be used for telecommand in conjunction with spacecraft engaged in deep-space research.	FIXED MOBILE Band 2110 to 2120 MHz may be used for telecommand in conjunction with spacecraft engaged in deep-space research.	
2200 to 2290	SPACE, RESEARCH Fixed Mobile	2200 to 2290 MHz per IRAC, Primary *U.S.A.* use for space research, down-data band. Shared with FIXED, low-power line-of-sight radio relay, and with MOBILE low-power line-of-sight systems and telemetry. Also authorized by IRAC: 2270.133 MHz 2270.933 MHz 2271.933 MHz Earth-to-satellite range data links.		

Table 1.2 (Concluded)

<table>
<tr><th rowspan="2">Frequency Band (MHz)</th><th rowspan="2">Service</th><th colspan="3">Notes and Comments Applying to Various Regions</th></tr>
<tr><th>Region 1</th><th>Region 2</th><th>Region 3</th></tr>
<tr><td>2290 to 2300</td><td>SPACE RESEARCH (Telemetry and Tracking in deep space)</td><td>FIXED
SPACE RESEARCH (Telemetry and Tracking in deep space)
Austria: Space research service in the 2290 to 2300 MHz band is a secondary service.
MOBILE</td><td>SPACE RESEARCH (Telemetry and Tracking in deep space)
Cuba uses 2290 to 2300 MHz band for Fixed and Mobile services.</td><td>FIXED
MOBILE
SPACE RESEARCH (Telemetry and Tracking in deep space)</td></tr>
<tr><td>2300 to 2450</td><td>RADIO LOCATION
Amateur
Fixed
Mobile</td><td>FIXED
Amateur
Mobile
Radiolocation
Frequency 2450 MHz is designated for industrial scientific and medical purposes. Emission is to be confined within ± 50 MHz of designated frequency. 2375 MHz is designated in certain European countries listed under Regions 2 and 3.
In United Kingdom, 2300 to 2450 MHz is allocated on primary basis to Radiolocation service and on secondary basis to Amateur, Fixed, and Mobile services.
In F. R. of Germany, 2300 to 2350 MHz band is allocated to Amateur service.</td><td colspan="2">RADIOLOCATION
AMATEUR
FIXED
MOBILE
Frequency 2450 MHz is designated for industrial, scientific, and medical purposes except in Albania, Bulgaria, Hungary, Poland, Roumania, Czechoslovakia, and the U.S.S.R., where 2375 MHz is used. Emissions must be confined within ± 50 MHz of designated frequency.
In India, Japan, and Pakistan, 2300 to 2450 MHz band is allocated on a primary basis to Fixed, Mobile, and Radiolocation services and on a secondary basis to Amateur service.</td></tr>
</table>

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SECTION II

ELECTROMAGNETIC COMPATIBILITY DESIGN GUIDELINE FOR STADAN

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1. INTRODUCTION

The purpose of this section is to provide a reference source for guidance in achieving electromagnetic compatibility (EMC) in the design of electronic and electrical equipment for aerospace ground stations. For example, the application of shielding theory to good design is presented and standards of good practice are outlined for bonding, grounding, wiring, and cabling. Some aspects of filter design are explained, and suggestions are given for the application of filters to electronic and electrical equipment. References and a bibliography are provided so that the user may be directed to sources of information more detailed than the presentation included in this handbook.

An example of an aerospace ground station network is the NASA Space Tracking and Data Acquisition Network (STADAN), which is comprised of three major functional systems: the Minitrack, the Data Acquistion Facilities (DAF), and the Goddard Range and Range Rate (RARR) systems.

Minitrack, historically the first of these systems, has been used for tracking United States satellites equipped with suitable radio beacons, as well as some foreign satellites. The RARR system complements Minitrack by providing improved tracking data at VHF and S-band for space probes, launch vehicles, and satellites in highly elliptical orbits. The DAF's, now operating at several locations, provide greater high-speed data handling capability than Minitrack and are equipped with high-gain parabolic antennas with frequency capability at 136, 400, and 1700 MHz.

The world-wide STADAN currently includes stations in the United States, Australia, South America, Africa, and Europe. The major functions of STADAN are tracking, command, data acquisition, and data transmission. For detailed descriptions of STADAN and its functions, see References 1 and 2.

2. SHIELDING

Data signals transmitted from satellites or orbital manned spacecraft are of low amplitude when they reach a ground station. The purpose of this section is to define shielding that prevents the penetration of undesired radiated electromagnetic energy through equipment interfaces. The undesired radiated interference can be generated within the ground station itself and/or be generated by external sources. Proper placement of equipment, and the judicious use of shielding, will minimize intrasystem interference.

In order to prevent malfunctioning of electronic equipment arising from undesired signals at frequencies from audio to X-band, it is necessary to enclose all electrical and electronic equipment within a metallic case or shielded enclosure.

A satisfactorily shielded enclosure should provide shielding effectivness of 50 to 100 dB, depending upon the intensity of the undesired signals that are present and the type of electromagnetic fields. In general, the design should call for maximum shielding effectiveness within the limitations of weight, size, mechanical rigidity, and cost. In the case of a telemetry receiver, for example, the receiver itself is not essentially a generator of undesired signals (except for low-level, local-oscillator radiation) and should not cause malfunctioning of adjacent equipment. The receiver can be very susceptible to external undesired signals, however, and therefore should be provided with a shielding enclosure having high shielding effectiveness. Cases for transmitter equipment are required to provide a shielding effectiveness of at least 100 dB in order that leakage through the case at harmonic frequencies may be reduced to acceptable levels.

In practice it is difficult to reduce the level of interfering signals to allowable limits by shielding against spurious radiation from transmitters and radiation of local oscillators in receivers. For this reason it becomes necessary to require some excess shielding effectiveness for the shielding enclosures of adjacent equipment even if high levels of undesired signals are not generated by such equipment. In the final analysis, the burden of shielding adjacent electronic equipment should be divided about equally among them.

An electromagnetic field is made up of two components: an electric field **E** and an magnetic field **H**. When an electromagnetic wave encounters a metal surface a shielding action occurs. Part of the wave energy is reflected and part of the energy penetrates the metal; the latter portion is partially dissipated as it passes out of the far surface of the metal. Essentially, the electric field component **E** induces a charge of equal but opposite polarity at the surface of the shield. The magnetic field component **H** induces a current flow whose field is equal in magnitude and opposite in direction to the incident field. (Reference 3.)

All metals exhibit a finite conductivity. Current flows on the surface and within the metal to a depth based on the magnitude of the skin effect. Only at temperatures near absolute zero does a metal approach the properties of a perfect conductor so that current flows at the surface only. Under these conditions all incident energy would be reflected. Since the actual fields at the metallic surface are not equal and opposite, only a portion of the incident energy is reflected. The remaining energy is dissipated as heat and as a transmitted wave on the other side of the shield.

The characteristic wave impedance is generally defined as the ratio **E/H**, the electric field component to the magnetic field component, both of which are transverse to the direction of propagation. Assuming that free space and air have the same characteristics, a plane wave has a free-space characteristic impedance, in ohms, of

$$Z_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} = 120\pi = 377\ \Omega \tag{2.1}$$

where

ϵ_0 = permittivity of free space (air), or 8.85×10^{-12} F/m,

and

μ_0 = permeability of free space (air), or 1.26×10^{-6} H/m.

When distance from the radiator is small in terms of wave length, **E** and **H** field impedance magnitude may be approximated using the following expressions:

$$Z_W = \frac{1}{\omega \epsilon_0 r} \tag{2.2}$$

for an electric field and

$$Z_W = \omega \mu_0 r \tag{2.3}$$

for a magnetic field. In these equations,

Z_W = impedance at distance r,

r = distance from source in meters,

and

ω = $2\pi f$ is the angular frequency in radians per second.

Shielding effectiveness describes the effectiveness of a given metal as a shield and is measured in dB. The equation expressing shielding loss in dB (References 3 and 4) is

$$S = R + A + B \tag{2.4}$$

where

R = total reflection loss in dB from both surfaces of the shield (neglecting the multiple reflections inside the barrier),

A = absorption loss in dB inside the barrier,

and

B = a positive or negative correction factor caused by the reflecting waves (secondary reflections) inside the barrier and is calculated in dB. When a metallic barrier has an absorption loss A of less than 15 dB, the shield is designated as being electrically thin. (The term B need not be taken into account when the absorption loss is more than 15 dB.)

In the determination of the total shielding effectiveness of a shield, values for R, A, and B are to be determined as indicated in the following sections. To simplify computations, a number of nomograms are provided.

The computation of reflection losses can be greatly simplified by considering the shielding effectiveness for incident electric fields as a problem separate from that for magnetic fields or plane waves. Thus, the determination of reflection loss becomes a problem of determining the reflection loss for an electric field R_e, a magnetic field R_h, or a plane wave R_p. Nomograms for determining values of R_e, R_h, or R_p are given in Figures 2.1, 2.2, and 2.3, respectively.

Secondary losses can be positive or negative and should be taken into account when the absorption loss A is less than 15 dB. To avoid computing B, it is desirable to maintain the absorption loss greater than 15 dB at frequencies where shielding effectiveness S is close to the required level.

2.1 REFLECTION AND ABSORPTION LOSSES

Reflection loss depends upon the type of field to be shielded and the distance between the source and the shield. Reflection loss is a maximum when the impedance of the incident wave is much higher or lower than that of the shield. Magnetic fields are low impedance in nature at all frequencies, but electric fields exhibit high impedance at all frequencies.

For conditions such that $fr \ll 2 \times 10^9$ the reflection loss in dB for a high impedance field is given (References 3 and 4) as

$$R_e = 354 - 10 \log_{10}\left(\frac{\mu f^3 r^2}{G}\right), \qquad (2.5)$$

where

r = distance between source and shield in inches,

μ = relative magnetic permeability of shield material,

G = conductivity of shield material relative to that of copper (for whicn $G = 1$),

and

f = frequency in hertz.

Reflection losses in dB for the magnetic field can be determined mathematically by the following equation (References 3 and 4)

$$R_h = 20 \log_{10}\left[\frac{0.462}{r\sqrt{fG/\mu}} + 0.136 r\sqrt{\frac{fG}{\mu}} + 0.354\right]. \qquad (2.6)$$

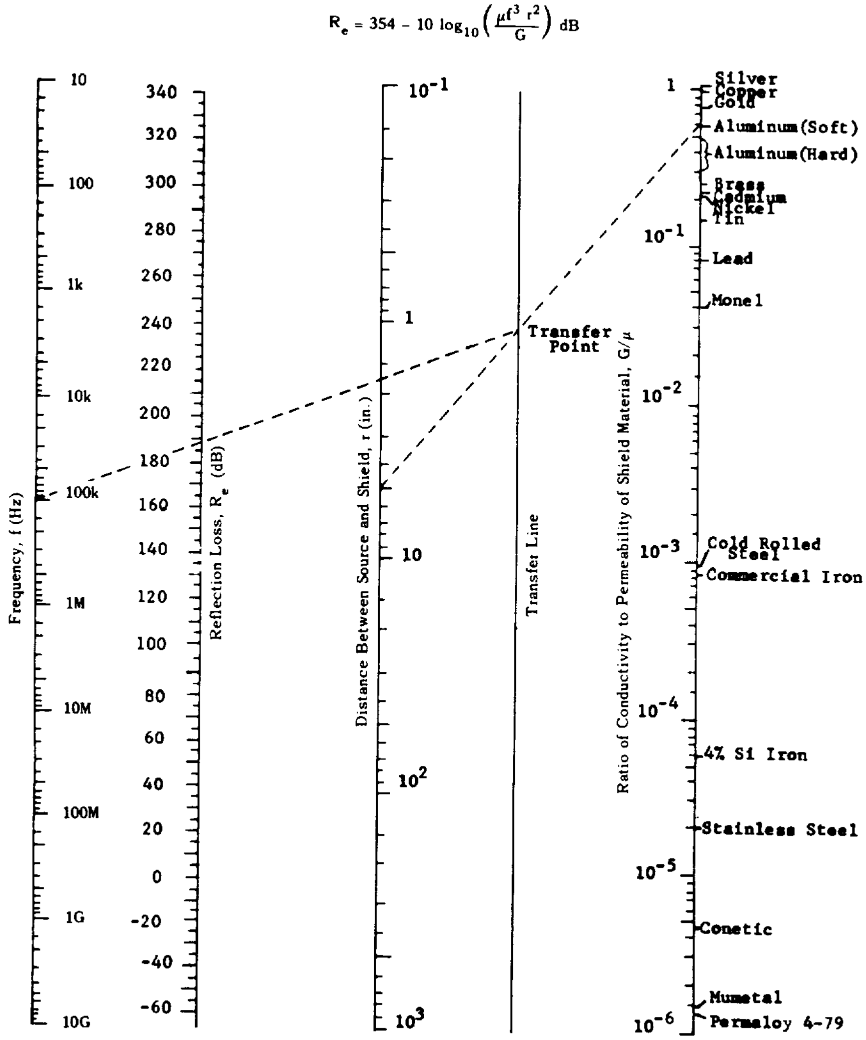


Figure 2.1.–Nomogram for determining electric field reflection loss. (After R. B. Cowdell, 1967 IEEE EMC Symposium Record)

The plane wave is characterized by an impedance of 377 Ω ($E/H = 377$). Mathematically, reflection losses in dB can be determined from a relation given in References 3 and 4:

$$R_p = 168 + 10 \log_{10}\left(\frac{G}{\mu f}\right). \tag{2.7}$$

$$R_h = 20 \log_{10}\left(\frac{0.462}{r}\sqrt{\frac{\mu}{fG}} + 0.136\, r\sqrt{\frac{fG}{\mu}} + 0.354\right) \text{ dB}$$

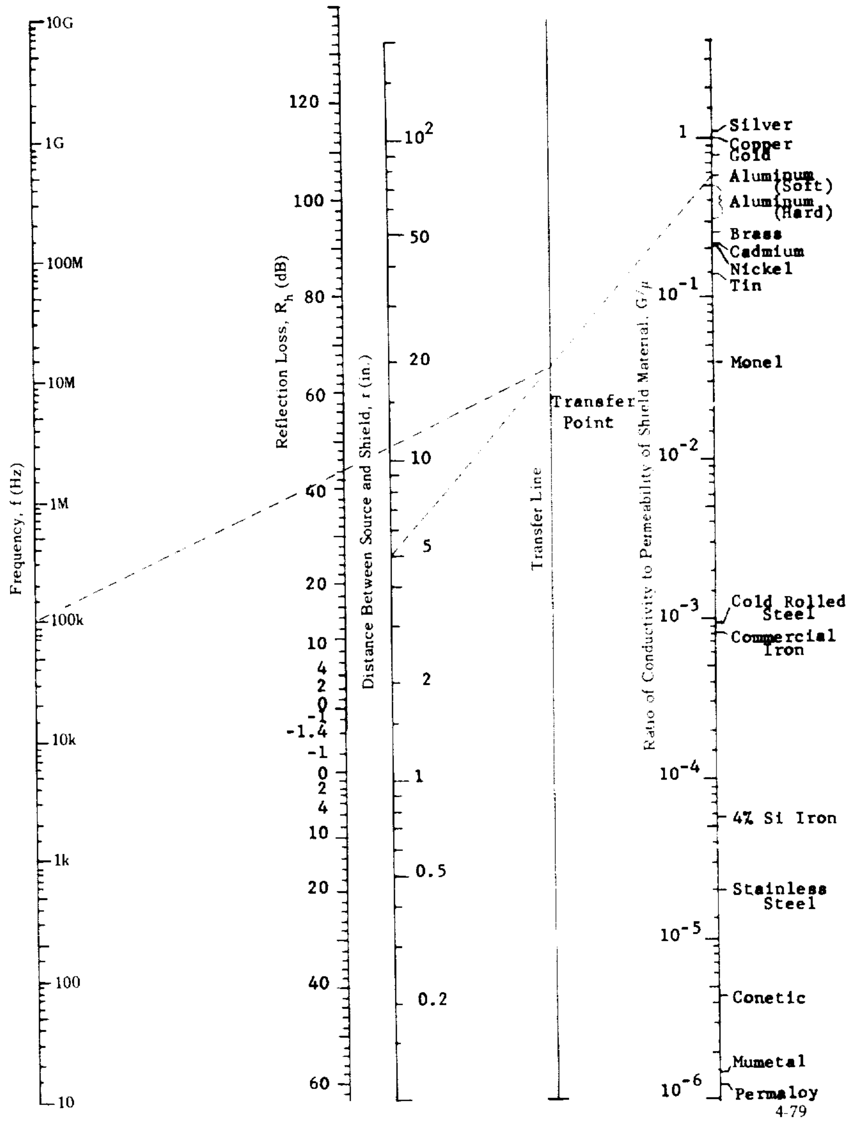


Figure 2.2.—Nomogram for determining magnetic field reflection loss. (After R. B. Cowdell, 1967 IEEE EMC Symposium Record)

$$R_p = 168 + 10 \log_{10} \left[\frac{G}{\mu f}\right] \text{ dB}$$

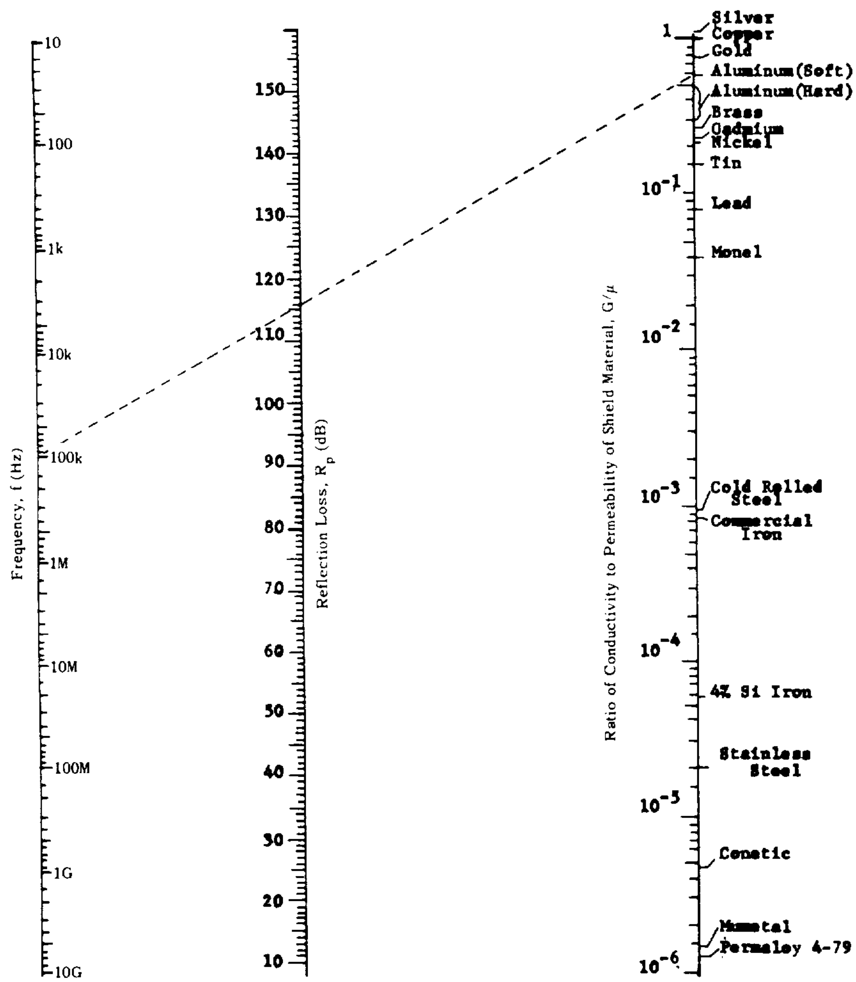


Figure 2.3.–Nomogram for determining plane wave reflection loss. (After R. B. Cowdell, 1967 IEEE EMC Symposium Record)

Absorption losses are a function of the physical properties of the shield material and are independent of the type of source field. The important parameters of the shield are its relative permeability, relative conductivity, and thickness. The absorption loss in dB of a shield is given (References 3 and 4) as

$$A = 3.34 \times 10^{-3} t \sqrt{\mu f G}\,, \tag{2.8}$$

where

t = thickness of shield material in mils (10^{-3} in.),

μ = relative permeability of shield material,

f = frequency in hertz,

and

G = conductivity of shield material relative to that of copper.

A close approximation to the basic shielding effectiveness for reflection loss and absorption loss can be determined with the use of the nomograms presented in Figures 2.1 through 2.4. Figure 2.1 is used for reflection loss where the field is generated by a high-impedance source or where the electric field predominates. If the source is of low-impedance (where the magnetic field is predominant), then Figure 2.2 is to be used to estimate the reflection loss.

A plane wave reflection loss may be obtained with Figure 2.3. The plane wave case applies when the shielding material is at least one wavelength distant from the source, or at a distance of at least D^2/λ, where D is the largest dimension of the field source or shield and λ is the wavelength.

Figure 2.4 is used to determine absorption loss. Absorption loss is independent of the field source impedance or the distance from the field source. The magnitude of the absorption loss is directly proportional to the thickness of the shield material.

Each nomogram is marked to show a sample calculation. In the case of reflection loss for the electric field and magnetic field, the example is based on the use of soft aluminum shield material at a distance of 5 in. from the source of the field having a frequency of 100 kHz. A line joining these points on the two appropriately marked scales intersects the unmarked line at a transfer point. From this point a line is drawn to the selected frequency, 100 kHz, on the frequency scale. The reflection loss is read off of the R_e scale of Figure 2.1 or the R_h scale of Figure 2.2.

The nomogram for plane wave reflection loss is used by simply drawing a line on Figure 2.3 joining appropriate points on the scales for the frequency and the selected shielding material. In the example, soft aluminum is used at a frequency of 100 kHz. The reflection loss of 116 dB is read from the R_p scale.

To use the absorption loss nomogram, follow the example shown. A soft aluminum shield of 0.003 in. (3 mils) thickness is used. The line joining these

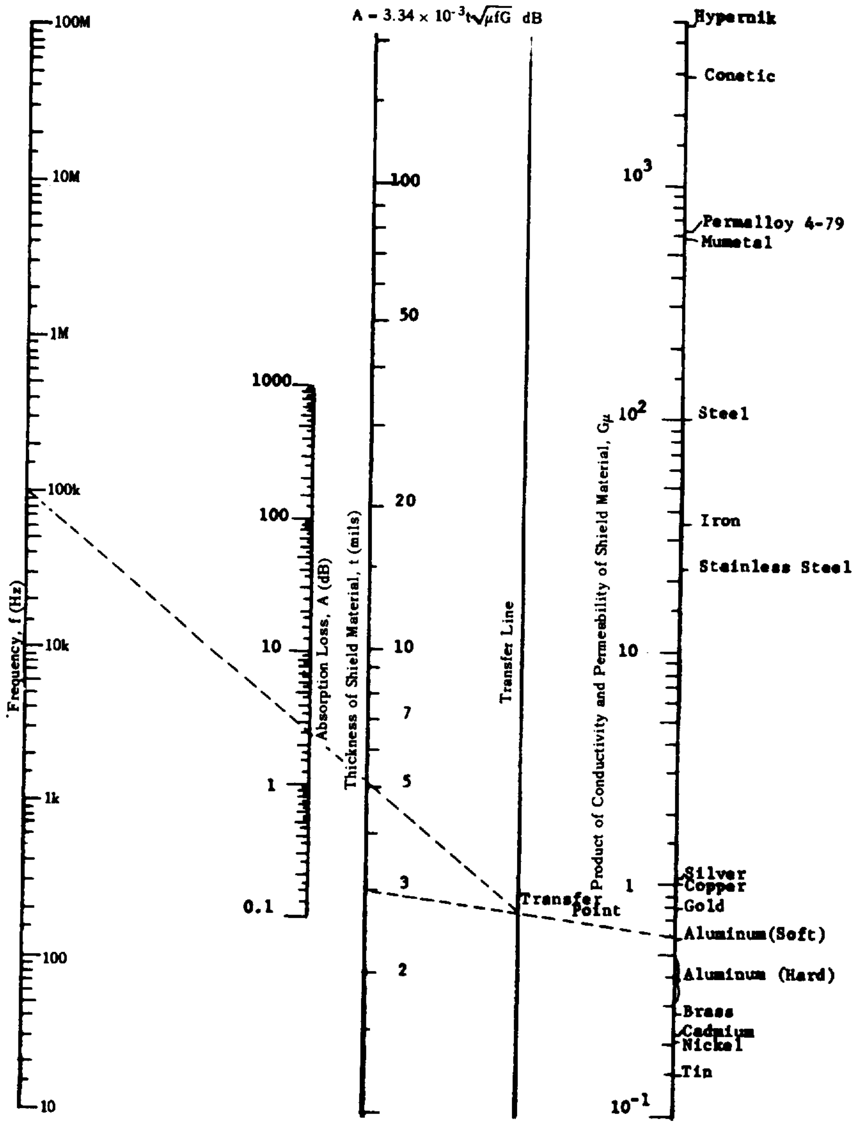


Figure 2.4.–Nomogram for determining absorption loss. (After R. B. Cowdell, 1967 IEEE EMC Symposium Record) two values intersects the unmarked line at the transfer point. The line from the transfer point to the frequency of 100 kHz intersects the A scale at 2.5 dB, which is the absorption loss.

If the shield is electrically thin (absorption loss <15 dB), internal (secondary) reflections become critical. Internal reflection loss B may be a positive or negative value. For a very thin shield, B could have a relatively high negative value.

When the absorption loss A is equal to or greater than 15 dB, the internal reflection loss B may be neglected. However, for $A < 15$ dB, a correction must be made for the multiple reflections within the shield. When a metallic barrier has sufficiently small absorption so that the value of A is less than 15 dB, it is designated as electrically thin. The internal reflection loss in dB may be computed from the following equation (obtained from Reference 5, pp. 2-47):

$$B = 20 \log_{10} |1 - a \cos 7.68 \times 10^{-4} t \sqrt{Gf\mu} + ja \sin 7.68 \times 10^{-4} t \sqrt{Gf\mu}| , \tag{2.9}$$

where

$$a = \left[\frac{(Z_s - Z_w)}{(Z_s + Z_w)}\right]^2 \times 10^{-0.1A} .$$

The intrinsic impedance of the metal, in ohms, is defined as

$$Z_s = (1 + j)\left(\frac{\mu f}{2G}\right)^{1/2} \times 3.69 \times 10^{-7} .$$

The impedance in ohms of the incident wave in free space for high-impedance (electric) fields is

$$Z_w = \frac{1}{c\epsilon_0}\left(\frac{1 + j\beta r - \beta^2 r^2}{j\beta r - \beta^2 r^2}\right),$$

and for low-impedance (magnetic) fields

$$Z_w = c\mu_0 \left(\frac{j\beta r - \beta^2 r^2}{1 + j\beta r - \beta^2 r^2}\right) .$$

In these equations, the symbols have the following meanings:

c = velocity of light (3×10^8 m/s),

Z_s = intrinsic impedance of metal in ohms,

Z_w = impedance of incident wave in free space,

μ_0 = permeability of free space (1.26×10^{-6} H/m),

t = thickness of barrier in mils,

μ = relative magnetic permeability referred to free space,

ϵ_0 = permittivity of free space (8.85×10^{-12} F/m),

G = relative conductivity of shield material referred to that of copper (e.g., $G = 1$ for copper, $G = 0.61$ for aluminum, and $G = 0.17$ for iron),

f = frequency in hertz,

β = $2\pi/\lambda$, where λ is the wavelength in meters,

and

r = distance from source to barrier in meters.

2.2 SELECTION OF SHIELDING MATERIALS

The selection of the material is determined by its ability to drain off induced electrical charge and to carry sufficient out-of-phase current to cancel the effects of the interfering field. Inherent characteristics that make the metal an effective shield are its conductivity G and permeability μ. A physical characteristic that enhances the effectiveness of a metal as a shield is its thickness; the shielding effectiveness of a metal is also dependent upon frequency.

The selection of proper materials for shielding is made in accordance with the following basic rules:

(1) At low frequencies (LF), only magnetic materials can furnish adequate shielding against magnetic fields. Thickness is an important factor.

(2) For a given material, magnetic fields require a greater shield thickness than do electric fields.

(3) At higher frequencies, smaller shield thickness is required for a given material.

(4) At sufficiently high frequencies, nonferrous materials, such as copper and aluminum, will give adequate shielding for either electric or magnetic fields.

(5) The electric field component for frequencies from 60 to 800 Hz (i.e., ac power) can readily be shielded with thin conducting materials such as iron, copper, aluminum, and brass.

Care must be used when adding a shield to a subsystem. For example, a shield placed too close to a circuit in which the circuit Q is a critical factor can cause degradation of performance because the losses in the shield will appear as an effective resistance in the critical circuit, thereby lowering the circuit Q (Reference 6).

The permeability of any magnetic material depends on the intensity of the field surrounding it. If this field is very strong, the shield may become magnetically saturated, thereby rendering the shield ineffective. A remedial solution to this problem is to use a shield constructed in two or more layers. The outer layer should have medium permeability and the innermost layer should have high permeability. Additional layers between the outer and inner layer may be required, depending on the intensity of the source.

Even a properly designed shield will be ineffective if it contains uncontrolled discontinuities. Most of the discontinuities are necessary to accommodate leads, such as those for input and output lines, power lines, antennas, control shafts, fuses, jacks, test receptacles, plug-in receptacles, indicator lights, meters, equipment covers, and door and ventilating holes (Reference 7). Provisions for such discontinuities can be made.

1. Penetration Holes in Shielding Materials

One effective method of neutralizing the shielding discontinuities created by planned holes (e.g., air ventilation and circuit adjustment) in a shield is to use cylindrical and rectangular waveguide-shaped openings. When properly designed, a waveguide-shaped opening will act like a high-pass filter with a cutoff frequency above the highest frequency of interest. The cutoff frequency is a function of the cross section of the waveguide.

For a cylindrical waveguide, the cutoff frequency is

$$f_c = \frac{6.92}{d} . \tag{2.10}$$

The cutoff frequency for a rectangular waveguide is

$$f_c = \frac{5.90}{b} . \tag{2.11}$$

In these equations,

f_c = cutoff frequency for the dominant mode in gigahertz,

d = inside diameter of a cylindrical waveguide in inches.

and

b = greatest dimension of rectangular waveguide in inches, At any frequency f_a considerably less than cutoff (i.e., $f_a \ll 0.1 f_c$), the attenuation in dB per inch for cylindrical waveguides is approximated by the relation

$$a \approx \frac{32}{d}. \tag{2.12}$$

For rectangular waveguides, the attenuation in dB per inch is

$$a \approx \frac{27.3}{b}. \tag{2.13}$$

The equations given above are valid for air-filled waveguides with length-to-width or length-to-diameter ratios of 3 or more (Reference 8).

Only insulating (nonconductive) material is allowed to extend through the waveguide opening; otherwise, if a conductive material is used, it will act like an antenna aborting the shielding effectiveness. A control (e.g., a variable resistor) can be adjusted by means of a nonconductive shaft through a cylindrical waveguide. Ventilation can be effected through the use of a honeycomb composed of a number of tiny waveguides clustered adjacent to each other; however, the shielding effectiveness is reduced in proportion to the area covered by honeycomb.

2. Finger Stock and Metallic Gaskets

Access doors and covers in shielded enclosures cause shield discontinuities at each junction. Effective shielding at these junctions can be achieved through the use of finger stock and/or resilient metallic gasketing. The junction surfaces should be completely free of oily film, corrosion, moisture, and paint. Gaskets should be mounted firmly to one side of the junction and should be composed of a combination of resilient and metallic material. The outer edge of the gasket which mates with the mating junction should have the ability to penetrate all surface film to ensure a tight junction at audio frequency (AF) and radio frequency (RF).

Finger stock is a strip of phosphor-bronze or similar material cut to form contact fingers under spring pressure. This type of contact strip is subject to damage unless it is installed in a recessed or inner lip; it should be handled with extreme care.

3. Screen and Conductive Glass Shields

Screen shields can be used effectively over pilot lights and sockets, meter faces and gauges. For example, the regular glass in meters and gauges can be replaced with conducting glass. As a further precaution, their junctions with the panel should be sealed with metallic gaskets. Plastic-cased pilot lights and meters should be avoided and metallic cased items used (References 5 and 9).

4. Additional Considerations

Fuses, jacks, and all receptacles should have metallic bodies and should be provided with screw-on, snap-on, or spring-loaded metal caps. Each junction with the panel should be sealed with metallic gaskets.

It is usually necessary to form a discontinuity in the shield by inserting power, control, and signal lines. The usual procedure for power and control lines is to mount the filter inside the shield. The filter input terminals should extend through the shield. The signal line poses more of a problem inasmuch as filtering to any great extent can cause signal degradation. A Faraday shield coupling device in conjunction with a shielded twisted line is one solution.

2.3 RECEIVERS

The following shielding guidelines should be employed during the development phase (Reference 10):

(1) All unremovable butted joints should be welded or brazed.

(2) Removable panels should be mated to adjoining contact surfaces through the use of metallic gasketing or finger stock. A sufficient number of binding devices should be spaced around the mating area to ensure the tight contact over the entire surface.

(3) Control shafts should penetrate the case via a circular waveguide by means of nonconducting shafts. Conducting shafts should use a grounding nut on the shaft.

(4) Any meter affixed to the case should be housed in a seamless metallic case. The glass covering the meter face should be a conducting glass or laminated glass incorporating a wire mesh.

(5) Ventilation through the case should be effected via honeycomb waveguide material, waveguides, or wire mesh.

(6) All connectors should be types providing shielding continuity. When connectors are not in use, metallic caps should be used to maintain the shielding continuity. Metallic gaskets should be used to ensure shielding continuity.

(7) Fuse holders and their removable caps should be metallic. Metallic gaskets should be used to ensure shielding continuity.

(8) The panel indicator lights should be housed within a metallic holder. The glass portion of the indicator should be a conducting-type glass or the inner surface should be covered with a wire mesh. Metallic gaskets should be used to ensure shielding continuity.

(9) The tuning dial window should be covered with conducting glass or wire mesh.

(10) Shield boots of metal mesh, conductive plastic, or rubber should be used on all toggle switches.

(11) The receiver subsystems should be individually shielded to prevent intrasystem interference.

2.4 TRANSMITTERS

The transmitter housing should reflect the basic shielding techniques given for receivers. Primary concern is with interference signals radiating from the transmitter to other equipment within the electromagnetic complex via the radiating interface of the housing.

Each stage of the transmitter, from the oscillator to the final amplifier, will require individual shielding. The diameter of the shield around a coil should be twice the diameter of the coil. The shield length should be equal to the length of the coil plus twice the diameter (Reference 11).

During the development stage, radiated signal levels should be obtained from breadboard transmitter subsystems. With these data, the type of material and the thickness needed to contain the interfering signals can be calculated.

2.5 ELECTRONIC SERVO SYSTEMS

The electronic servo systems should employ the same shielding techniques as given for receivers, where applicable. The servo subsystems should be individually shielded to prevent susceptibility from inter- and intra-system interference.

Type and thickness of the cabinet material may be selected on the basis of the shielding requirement. The requirements for mechanical strength will often dictate use of a thicker material than may be required for shielding alone.

2.6 COMPUTERS

The computer housing shielding should be in accordance with the guidelines set forth for receivers. Electronic subsystems should be individually shielded if susceptible to inter- and intra-system interference. Normally, the material selected to offer structural rigidity to the equipment will afford enough shielding if the cabinet penetrations are properly designed.

2.7 COMMAND ENCODERS

The shielding guidelines set forth for receivers are applicable. All low-level subsystems should be completely shielded in order to prevent susceptibility from inter- and intra-system interference. The conducting material selected to offer structural rigidity of the equipment will usually afford shielding adequate to meet requirements.

2.8 DATA PROCESSING

The section for receivers should be used as a guideline for shielding the data processing equipment cabinetry. All susceptible subsystems should be completely shielded from inter- and intra-system interference. The conducting material selected to achieve structural rigidity of the equipment will usually provide sufficient shielding.

2.9 TIME STANDARDS

The time standard should employ the same shielding techniques as for receivers, where applicable. The time standard subsystems should be individually shielded to prevent susceptibility from inter- and intra-system interference. Tests of susceptibility to radiated energy should be performed on a breadboard system during the development state. The data obtained can be used to calculate the type of shielding material needed and the thickness needed to protect the time standard.

2.10 POWER GENERATION EQUIPMENT

Motor generators should be completely housed within a shielded enclosure. The guidelines set forth for receivers are applicable. Adequate ventilation is a necessity. Ventilation openings may be screened with the screen continuously grounded around the edge, or a honeycomb panel may be used.

The power distribution panel box should be completely shielded. Access plates should be gasketed with metallic gaskets.

2.11 TELETYPE EQUIPMENT

Standard, unmodified commercial teletype equipment such as the M28/Automatic Send and Receive unit is a generator of high-level implusive broadband noise. The three interference sources within the teletype equipment are the motors, the selector magnets, and the transmitter-distributor contacts. Fortunately, all of these source generators (e.g., M28/ASR unit) are housed within a single shell. This housing is not an effective shield so the guidelines set forth for receivers should be followed.

The following electrical subsystems should be shielded:

(1) Selector magnets.
(2) Transmitter distributor contact box.
(3) Typing reperforator contact box.

2.12 MOTORS

Motors should be completely shielded with the motor shaft connected to ground through a phosphor-bronze sleeve or a conductive grease. A conductive screen should be placed over openings on the inside of the end bells.

2.13 VENDING MACHINES

The housings of vending machines should be shielded in accordance with the shielding techniques given for receivers, when applicable. The electronic control mechanism that dispenses a selected item and makes the proper change should be completely shielded.

2.14 MOTOR VEHICLES

The following shielding guidelines should be followed:

(1) The ignition system should be completely shielded.

(2) The battery charging system should be completely shielded.

(3) The shields should be accessible through AN or other suitable connectors mounted on the wall of the shield.

2.15 ENCLOSURES FOR ELECTROMAGNETIC INTERFERENCE TESTS

Most tests for electromagnetic interference (EMI) should be conducted in a shielded laboratory enclosure. The purpose of the shielded enclosure is to reduce ambient electromagnetic fields to levels that will not affect the accuracy of the tests being performed. The guidelines given for shielding equipment enclosures are directly applicable to the design or procurement of laboratory shielded enclosures (i.e., shielded rooms).

Most users of shielded enclosures purchase the enclosures in disassembled modular form. The manufacturer's price for an enclosure normally includes cost for erection, filter installation, and certification of shielding effectiveness.

Unusually large shielded enclosures are normally custom built as an integral part of a new building. When this procedure is followed, great care must be taken during construction to verify complete continuity of all joints and proper installation of doors, ventilation panels, and filters. Upon completion, and regularly thereafter, any shielded enclosure should be tested to verify shielding integrity or to determine corrective maintenance required.

Commercial modular enclosures are available as single-wall or double-wall units. The most commonly used materials are galvanized and copper-clad steel. The single-wall structure usually has sufficiently simple joint construction to make it as effective, or almost so, as the more complicated double-wall structure. Double-wall construction is usually accomplished by bonding a metal sheet to each side of plywood or by welding stiffener channels to the inside surfaces of the two spaced sheets.

Proper assembly of modular shielded enclosures is critical. Every bolt of the enclosure must be uniformly and securely tightened to ensure electrical continuity at all points on each joint or seam. Proper initial assembly is best left to the experienced technicians employed by the enclosure manufacturer. Stresses introduced by temperature changes and movement of floor loads and oxidation of contacts will cause some deterioration of shielding effectiveness with time. Also, the contact fingers around door edges can become fatigued and introduce RF leakage. These problems can be minimized by periodic inspection and corrective maintenance. Maintenance usually includes a systematic routine for uniform retightening of all assembly bolts and replacement of worn finger stock at door openings.

Custom-built permanent enclosures should be inspected for cracks in welded or soldered seams and worn finger stock at doors. Any cracked joints should be carefully cleaned and repaired, and worn finger stock should be replaced.

Measurement procedures for shielded-enclosure shielding effectiveness are specified in Reference 12. Proposed specifications for built-in-place shielded enclosures and additional test procedures applicable to both built-in-place and modular enclosures are given in Reference 13.

3. BONDING

Bonding refers to the establishment of a low-impedance path between two metal surfaces. This path may be between two points on a system ground plane as well as between ground reference and a component, circuit, or structural element.

Poor bonding between equipment and ground reference plane will cause interference because it prevents methods of suppression, such as insertion of filters or shields, from being completely effective. Consider the installation of the poorly bonded filter in Figure 2.5. At low frequencies above the passband, interference currents will follow path 1 to ground. When the impedance of a bond ($Z_b = R_b + j\omega L_b$) becomes larger than the reactance of the capacitor ($Z_c = 1/j\omega C$) in series with the load impedance (R_L), interference current will follow path 2 into susceptible equipment (Reference 14).

3.1 GENERAL CONSIDERATIONS

Generally, the impedance of a bond is of greatest concern at high frequencies (HF) where the skin effect impedance is much higher than the dc resistance. Typical HF bonding design is shown in Figure 2.6.

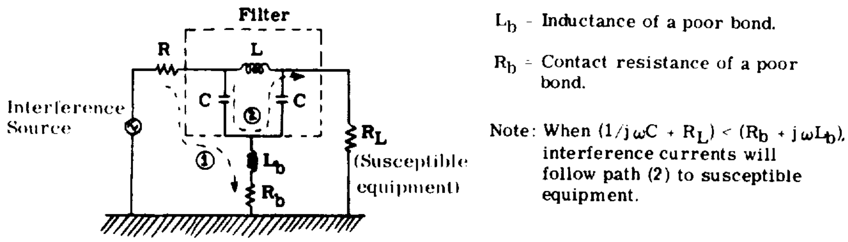


Figure 2.5.–Circuit representing poor bonding between a filter and ground.

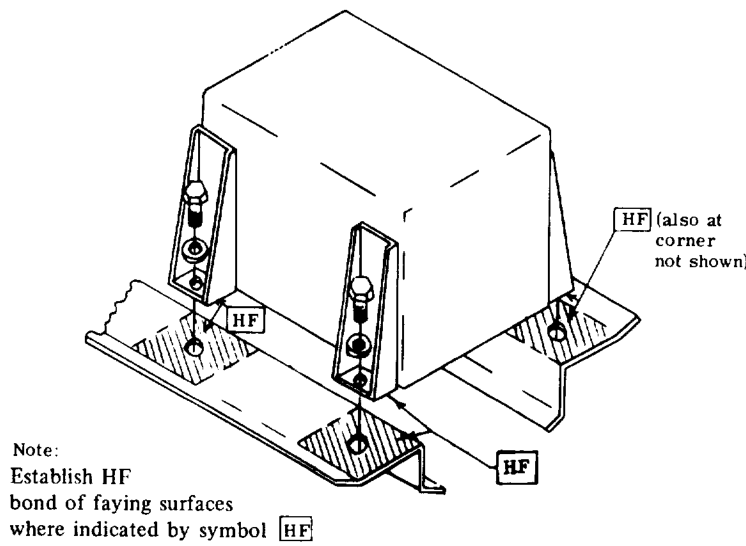


Figure 2.6.–Example of good HF bonding installation.

A dc bond is a connection that is effective in reducing dc potential between the bonded parts to a negligible level when full design currents are flowing (see Figure 2.6). The impedance of an HF bond may be very high at frequencies outside the range of interest (including dc and audio frequencies). Figure 2.7 shows the typical impedance of a solid copper bond strap for various values of its length-to-width ratio (References 15 and 16).

Electrical bonding serves other purposes in addition to the elimination of interference. Good bonding prevents the buildup of potential differences between points connected to a ground plane and in this manner eliminates potential ground loops. Bonding between components deters the buildup of static charges in normal equipment operation and also minimizes the damage that might be caused by lightning strikes (which produce high voltage buildup and heavy current flow). Good bonding will protect personnel from the shock

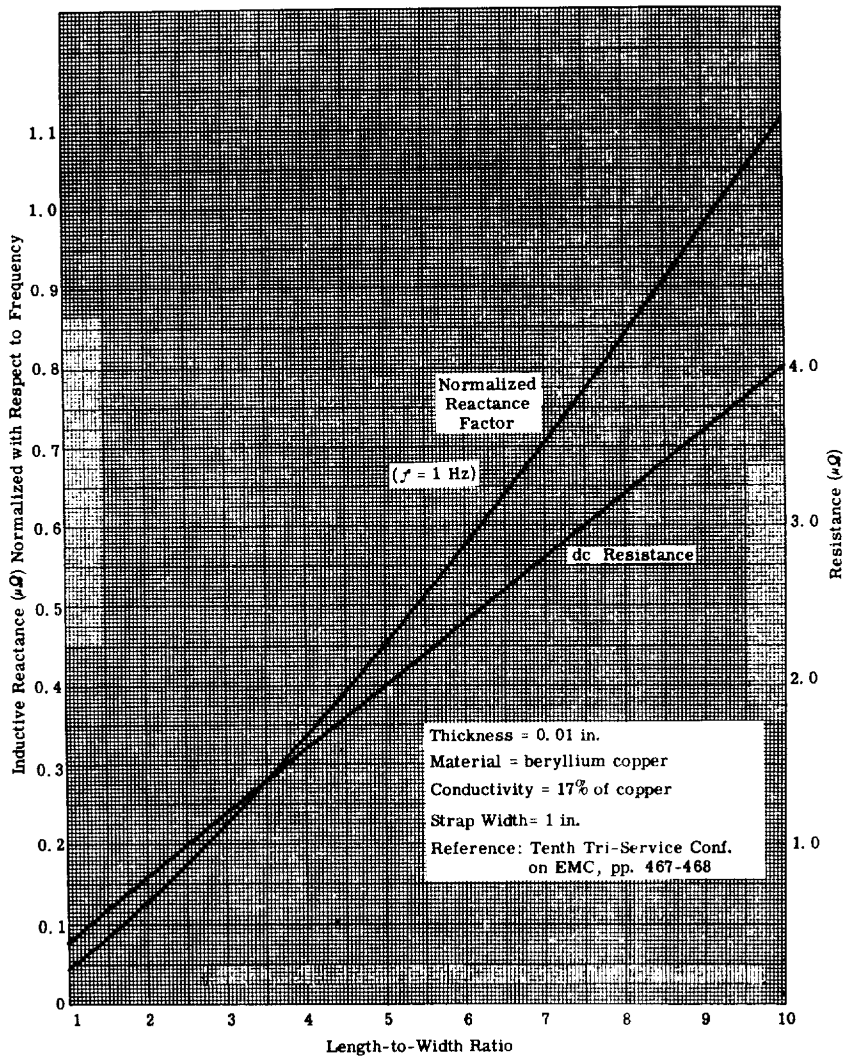


Figure 2.7.—Normalized impedance of a copper bonding strap for different ratios of length to width.

hazard that would result if power were inadvertently shorted to an enclosure. It is imperative that bonding practices receive the careful attention of design personnel to optimize system reliability and security, as well as to reduce problems of electromagnetic compatibility.

The designer must specify adequate bonding to ensure that the end product will require minimum use of suppression components. It is the designer's responsibility to determine bonding requirements and to call out and illustrate on drawings areas in need of bonding. The designer must also determine which surface areas are to remain unfinished or require a conductive finish in order to provide acceptable electrical continuity for bonding. Figure 2.6 illustrates a recommended bonding callout on a typical drawing (Reference 17).

Joints made by direct metal-to-metal contact fulfill bonding requirements. Such joints may be produced by welding, brazing, sweating, swaging, and in most cases by soldering. Semipermanent bonds, such as those provided by bolts or rivets, are acceptable when good electrical contact exists between bare metal surfaces. Star, or lock, washers may be used with threaded devices to ensure continued electrical contact and tightness. Star washers are very effective in cutting through nonconductive coatings such as those caused by corrosion. Joints that are press fitted or joined by self-tapping or sheet metal screws cannot be relied upon to provide a low-impedance bond at high frequencies. Riveted joints on 3/4-in. centers are acceptable if the rivet holes are bare. Clamped fittings that are permanent after installation are also acceptable, if they meet the requirements for maximum allowable resistance listed in Table 2.1. Direct bonds must always be made through continuous contact between bare, conductively finished metals.

An indirect bond (or bonding jumper) is an intermediate electrical conductor used to connect two isolated items. Because jumpers often have significant impedance at HF, their use must be avoided wherever possible (see Figure 2.8). Resonance between the inductance L of the jumper and the

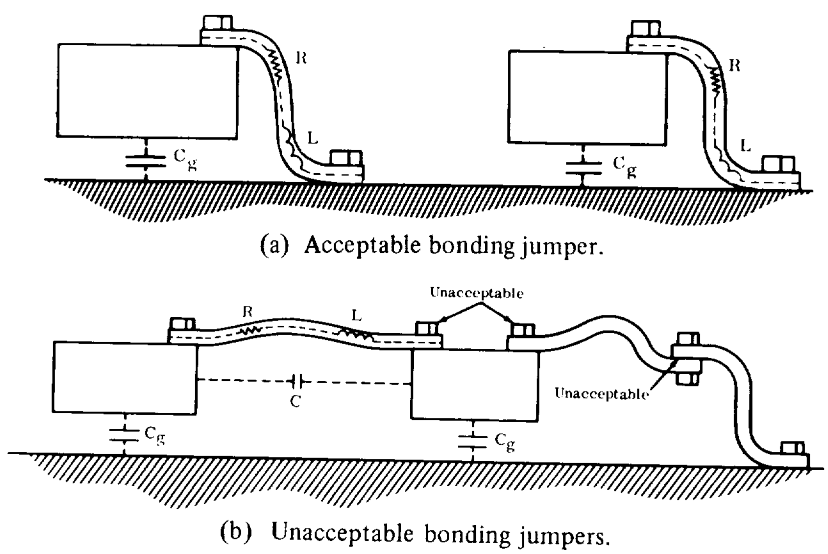


Figure 2.8.—Diagrams of acceptable and unacceptable bonding jumpers.

capacitance C between components results in a maximum bond impedance at a given frequency. The voltage buildup at resonance could result in the generation of arc discharges and strong electric-field interference at this frequency. The inductance can be minimized by using a metal jumper whose length is less than five times its width. In this way the resonant frequency may

Table 2.1—Resistance limits for dc bonding.

Component	Maximum dc Resistance to Structure (ohms)
Support brackets and electrical/electronic cabinets	0.0025
Access or inspection doors	0.005
Fuel, oil, hydraulic lines, and fittings	500 k
Power conduit	0.10
Conduit carrying signals or low currents	1.0
Filters	0.0025
Electrical motors, starters, generators, and attenuators	0.0025
Metal tanks with fuel filler provisions (no electrical installations)	0.01
Metal tanks containing no flammable material	500 k
Other electrical devices attached to enclosures	0.005
Switches, circuit breakers, and potentiometers in circuits exceeding 50 V	1.0
Metal instrument panels	
(a) Nonelectrical	500 k
(b) No rotating or vibrating electrical equipment	0.01
(c) With rotating or vibrating electrical equipment	0.0025
Radiators and heat exchangers	500 k
Metal ducts (nonelectrical, rigid, and flexible)	500 k
Engine supports	0.0025
Antennas	0.0025
Coaxial cables (HF)	0.0025
Structural joints or breaks	0.0025
Ground support equipment enclosures*	
(a) Cabinet seams	0.005
(b) Drawers	0.005
(c) Panels	0.005
(d) Access doors	0.1

*See paragraph 2.2 on shielding for information concerning prevention or containment of electromagnetic radiation.

often be raised so as to be above the range of concern. Braided jumpers may be objectionable at HF because braids tend to have higher inductance, but they may be preferable at LF because they offer greater flexibility than solid jumpers (see Figure 2.9). Direct metal-to-metal contact is essential in all cases. Figure 2.10 presents a typical arrangement for a bonding jumper.

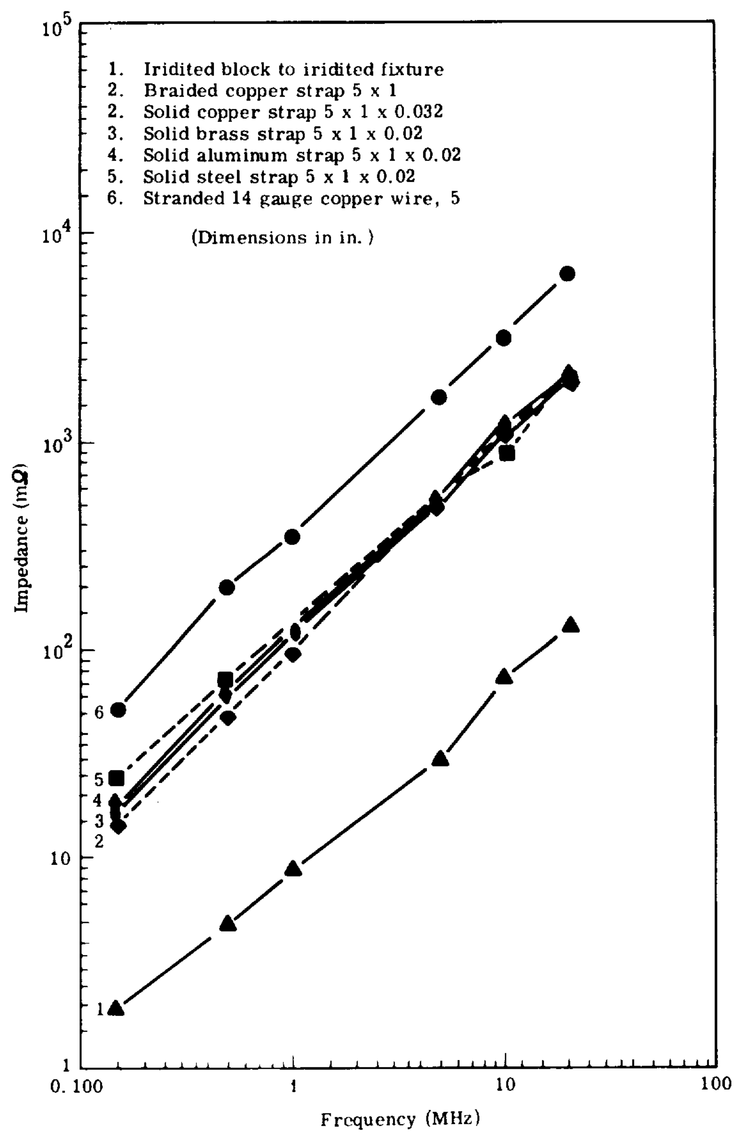


Figure 2.9.—Impedance of various bonding materials.

The following list defines the design guidelines for electrical bonding jumpers:

(1) Jumpers should be bonded directly to the basic structure rather than through an adjacent part (see Figure 2.8).

(2) Furthermore, jumpers should not be installed two or more in series (see Figure 2.8).

(3) Jumpers should be as short as possible.

(4) No more than several jumpers should be installed.

(5) Jumpers should not be connected with self-tapping screws.

(6) Jumpers should be installed so that vibration or motion will not affect the impedance of the bonding path.

(7) Jumpers should be made of tinned copper, cadmium-plated phosphor bronze, aluminum, or cadmium-plated steel. Mating metals should be selected to have properties of low corrosion.

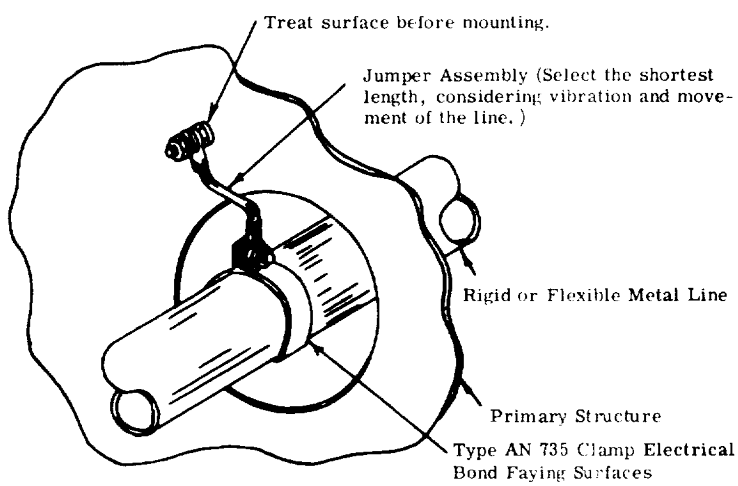


Figure 2.10.–Example of bonding jumper installation.

3.2 BONDING PRACTICES AND TECHNIQUES

1. Surface Treatment

Both direct and indirect bonding connections require metal-to-metal contact of bare surfaces. It is frequently necessary to remove protective coatings from metals to provide a satisfactory bond. The area cleaned for bonding should be only slightly larger than the area to be bonded. Ridges of paint around the periphery of the bonding area can prevent good metal-to-metal contact. Washers or fittings must fit inside the cleaned area. Immediately prior to bonding, all chips, paint, grease, or other foreign matter must be removed with the proper cleaning solution.

After bonding, the exposed areas should be refinished as soon as possible with the original finish; however, if the paints are too thin, refinishing paints may seep under the edges of bonded components and impair the quality of the bond.

A suitable conductive coating may be used when removable components must be provided with a protective finish. Where aluminum or its alloys are used, corrosion resistant finishes that offer low electrical resistance are available. Some of the many conductive coatings are alodine, iridite, oakite, turco, and bonderrite. These finishes need not be removed for bonding purposes. Figure 2.11 shows the degradation of shielding effectiveness of metal-to-metal joints caused by finishes on aluminum and magnesium.

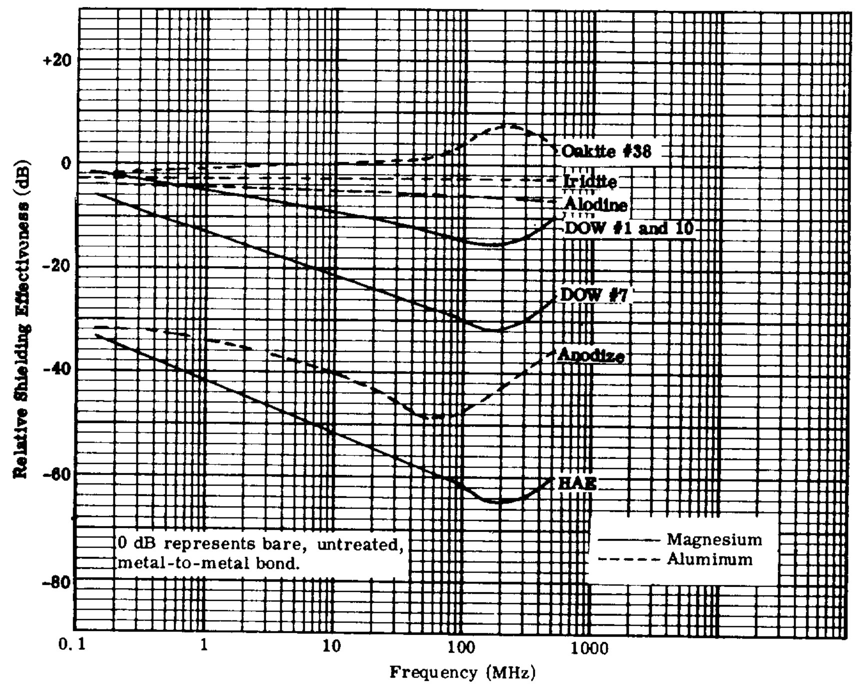


Figure 2.11.–Degradation of shielding effectiveness caused by finishes on metal. (After C. B. Pearlston, IRE Trans. RFI, October 1962)

2. Corrosion

Severe weight restrictions in systems often dictate the use of light metals (such as aluminum or magnesium) that have favorable strength-to-weight ratios. If their surfaces are untreated, such light metals are highly active chemically when placed in contact with other metals. For this reason, they are usually coated with nonconductive finishes to prevent corrosion. However, the impedance of the bonding joint is increased by such protective finishes, and it is often necessary to remove nonconducting portions of the finishes at the bonding interface. If the surfaces remain unfinished, more nonconductive corrosion products can be formed that also increase the impedance of the joint. Because of this, it is generally recommended that conducting protective finishes be applied to light metals at the bonding joints to obtain a low-impedance joint.

Corrosion occurs between two dissimilar metals in solution, or even in a moist atmosphere, since they form an electrochemical cell. The extent of corrosion depends upon the metals comprising the electrochemical cell and the conditions under which the dissimilar metals come into contact with each other. By properly modifying these two factors, the extent of corrosion can be reduced.

No appreciable corrosive action occurs between two metals in the same electrochemical or galvanic group. If they are in different groups, the metal coming first in the following list will form the anode and be relatively heavily corroded, whereas the metal coming later will form the cathode and be relatively free from corrosion or will be protected. According to their electrochemical activity, the common metals may then be arranged in the following groups of decreasing corrosion tendency (from anodic end to cathodic end):

Anodic End (most easily corroded)

Group I	Magnesium
Group II	Aluminum, aluminum alloys, zinc, cadmium
Group III	Carbon steel, iron, lead, tin, tin-lead solder
Group IV	Nickel, chromium, stainless steel
Group V	Copper, silver, gold, platinum, titanium

Cathodic End (least easily corroded)

The greatest degree of corrosion occurs when dissimilar metals are openly exposed to salt water, rain, or other liquids that may act as an electrolyte. Minimum corrosion occurs when metals are kept dry and completely free from exposure to moisture. The following three exposure conditions are defined:

Exposed:	Metal has an open, unprotected exposure to weather.
Sheltered:	Milder exposure than above and metal surfaces receive limited protection from direct action of weather. Such sheltered protection would be afforded by louvered housing, sheds, vehicles, aircraft. and boats.
Housed:	Metal surfaces of equipment are housed in weatherproof buildings.

3. Bonding Protection Code

For a given pair of dissimilar metals in contact under the three exposure conditions enumerated above, the extent of corrosion can be minimized with certain protective measures. The following bonding protection practices have proven to be useful:

(1) The joint of dissimilar metals should have a protective finish applied after the metal-to-metal contact has been established so that a liquid film cannot bridge the gap between the two dissimilar metals.

(2) The two dissimilar metals should be joined with bare metal exposed over an area slightly larger than that of the joint itself. The joint is untreated, but the remainder of the surface of the two metals has an appropriate protective finish.

By making use of the above grouping of anode and cathode materials, together with the exposure conditions under which equipment is to be used, it is possible to establish acceptable practices for the protection of equipment against corrosion or to minimize corrosion under a given set of circumstances. Suggestions for providing a protective bond between two dissimilar metals, identified in anode and cathode groupings, are given in Table 2.2

Table 2.2 may be used to minimize corrosion conditions when it becomes necessary to place dissimilar metals in contact with one another. The use of Table 2.2 is demonstrated in the following example: Equipment having an aluminum housing (Group II) is to be mounted on a stainless steel frame (Group IV) and used outside in continuously exposed conditions. Table 2.2 suggests that a protective finish should be applied after bonding. In most cases, it is also a good practice to interpose a tin- (Group III) or cadmium- (Group II) plated washer between the two metal surfaces. Thus, if the protective coating is chipped, the washer instead of the irreplaceable aluminum frame will be attacked by corrosion.

Table 2.2–Groups of materials recommended for providing protective bond between two dissimilar metals used as anode and cathode.

Condition of Exposure	Anode				Cathode
	I	II	III	IV	
Exposed	A	A			
Sheltered	A	A			II
Housed	A	A			
Exposed	C	A	B		
Sheltered	A	B	B		III
Housed	A	B	B		
Exposed	C	A	B	B	
Sheltered	A	A	B	B	IV
Housed	A	B	B	B	
Exposed	C	C	C	A	
Sheltered	A	A	A	B	V
Housed	A	A	B	B	

Notes: Bond protection code

A. The couple must have a protective finish after metal-to-metal contact has been established so that no liquid film can bridge the two elements of the couple.
B. The two metals may be joined with bare metal exposed at junction surfaces. The remainder must be given an appropriate protective finish.
C. This combination cannot be used except under very unusual circumstances where short life expectancy can be tolerated or when the equipment is normally stored and exposed for only short intervals. Protective coatings are mandatory.

3.3 BONDING OF STRUCTURAL ELEMENTS

The high-frequency impedance of a bond that can adversely affect system performance cannot be conveniently measured. As shown in Figure 2.7, the high-frequency impedance of a bond at the highest operational frequencies of the system may be reduced through the use of bonding straps having small ratios of length to width.

The limits for dc bonding in Table 2.1 may also be used to establish recommended limits of impedance of AF and HF bonds. Measurements of bonding resistance can be made with commercially available bonding meters.

The following examples of bonding techniques for typical items commonly included in many systems are listed to demonstrate the application and importance of good bonding practices.

The dc resistance measured across any individual bond from any point on a chassis across the bond to any point on the other side usually should not exceed 2.5 mΩ. The dc resistance between any equipment and its rack or between the rack and structure also should not exceed 2.5 mΩ. If these conditions are fulfilled, no component will be above the structure potential by more than the voltage drop across a resistance of 5 mΩ.

1. Connector Bonding

Standard AN-type connectors as well as coaxial connectors must be bonded to their respective panels over the entire mating surface as illustrated in Figure 2.12. Panel surfaces must be cleaned to the base metal for no less than 1/8 in. beyond the periphery of the mating connector. Connector bonding is essential for obtaining the appropriate shield wire grounding through a connector backshell.

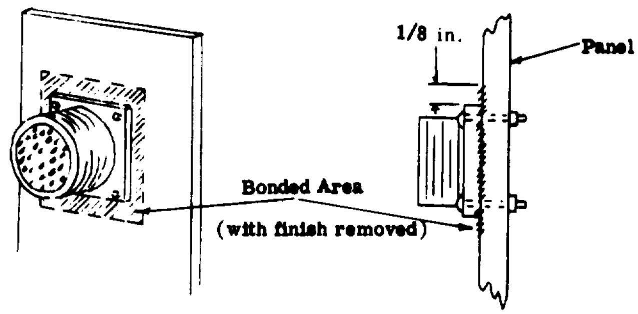


Figure 2.12.—Bonding of connector.

2. Bonding Enclosure Flange to Structure

Enclosures that are attached to frames or racks by means of flange-mounted quick disconnect fasteners must be bonded about the entire flange periphery. Both the flange surface and the mating rack surface must be cleaned and polished over the entire contact area, as shown in Figure 2.13. The direct bonding method of Figure 2.13a is far more desirable than the jumper bond shown in Figure 2.13b. The use of jumpers should be avoided if possible.

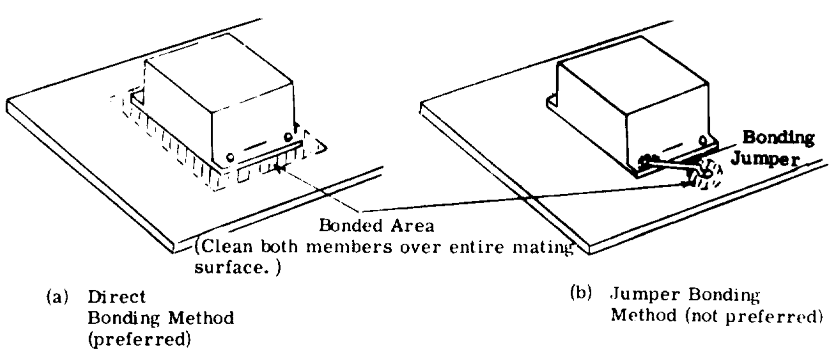


Figure 2.13.—Bonding of base member.

3. Bonding Structural Members

Bonding between various structural members and brackets must be accomplished through clean metal surfaces, as illustrated by the typical applications shown in Figures 2.14 and 2.15.

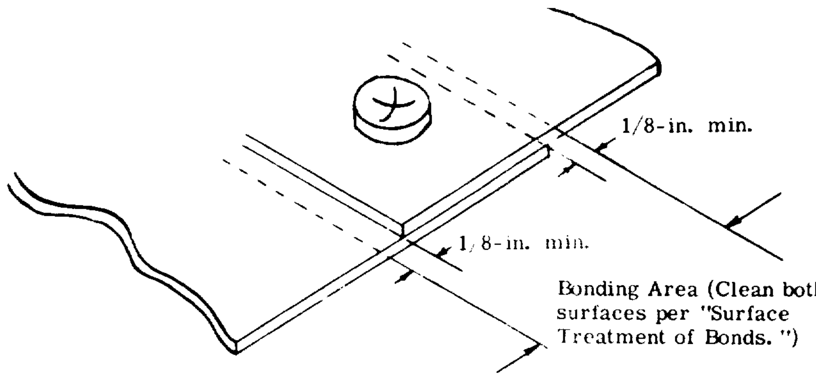


Figure 2.14.–Bonding of bolted members.

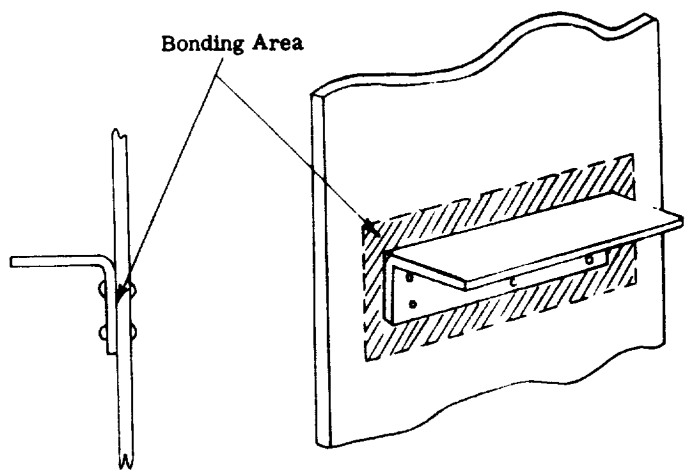


Figure 2.15.–Bonding for riveted or welded bracket.

4. Vibration Isolators

Bonding jumpers with properly cleaned surfaces should be installed on each vibration isolator, as shown in Figure 2.16. Jumpers for this application should have a maximum thickness of 0.025 in. so that damping efficiency is not impaired.

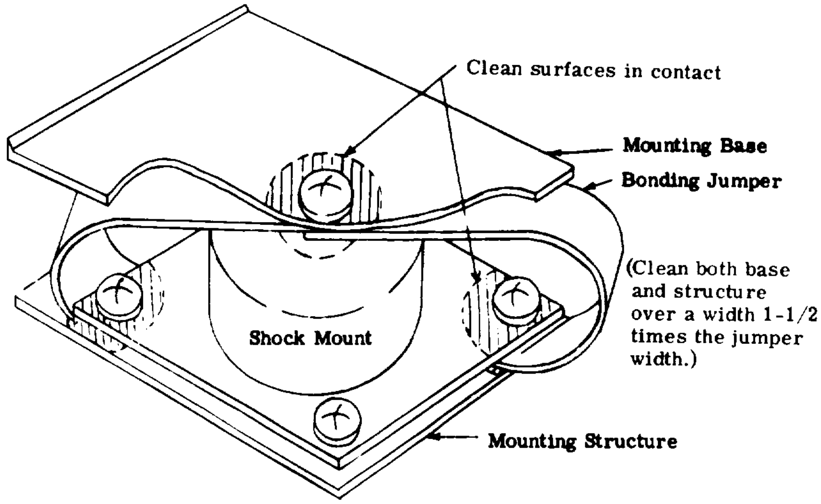


Figure 2.16.–Installation of vibration isolators.

5. Base-Mounted Components

Components mounted to their respective chassis through normal base-mounting methods should use the full mounting area of the component to provide a suitable electrical bond, as illustrated in Figure 2.13a. Separate jumpers (as shown in Figure 2.13b) should never be used for this purpose.

6. Rack-Mounted Components

Rack-mounted components should be bonded in the same way as flange-mounted equipment. The preferred method of bonding rack-mounted packages employing one or more shear pins is shown in Figure 2.17. Mating surfaces should be cleaned for 1/4 in. around the periphery of the bushing, as shown in Figure 2.17.

7. Knitted Wire Gaskets

Knitted wire gaskets may be employed to provide both bonding and shielding between removable parts and structure. Designs for inclusion of gaskets should be based on minimum plate-to-gasket pressure of 30 psi. Gaskets should be 1/8 in. in width. Additional information about gasket installation is given in the section on shielding (Reference 18).

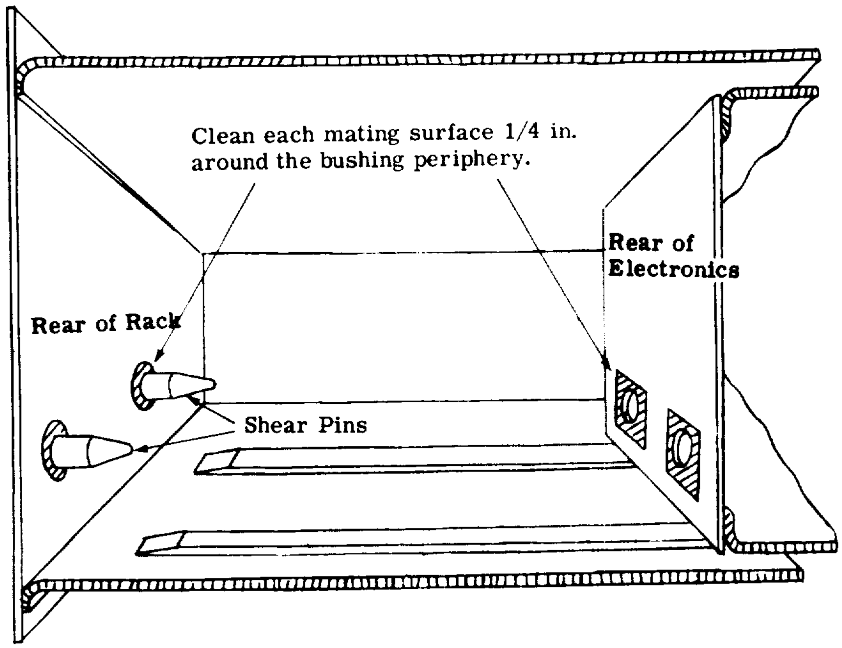


Figure 2.17.—Bonding for shear pins.

3.4 RECEIVERS AND TRANSMITTERS

During the development stage of receivers and transmitters, the following bonding guideline is recommended:

(1) Where feasible, hard bonding (direct) should be utilized. When indirect bonds must be used, the jumpers should not exceed 3 in. in length.

(2) All bonding surfaces should be thoroughly cleaned of any nonconductive material (Reference 18).

(3) All subsystem ground busses should be bonded to the intra-system ground plane.

(4) Base-mounted components should employ the base-mounting area to provide a suitable electrical bond.

(5) All panel-mounted components should be bonded directly to the panel.

(6) Removable panels that are attached to the case flange should be bonded about the entire flange periphery. Flange-mounted quick-disconnect fasteners may be used.

(7) Bonding between structural members and brackets should be effected by welding or brazing.

3.5 ANTENNA BONDING

Arcing between two pieces of metal that are not properly bonded together is a source of interference that can be both troublesome and difficult to locate. When the arcing metal joint is a part of the receiving antenna, the receiver may become saturated with electromagnetic interference from the arc. This type of interference may occur at any point where two pieces of metal make a high-resistance contact or where the resistance of the contact varies with vibration. The condition may be aggravated by dirt and corrosion. Structural members of a framework, meshed gears, and links of a chain have been documented as sources of interference. Frequently the situation is compounded by the proximity of a local high-power transmitter that induces fairly high currents in the metal joints and establishes a source of broadband noise.

Several types of ultrasonic devices are now available to aid in locating arc sources in metal structures, including antennas. Characteristically, a high proportion of the arc energy lies in the ultrasonic range of frequencies usually peaking around 150 kHz. A typical ultrasonic detector, used in finding arcs, corona, and gas and liquid leaks, has a sharply directional pickup pattern. The high directivity of a sensitive detector can be used to pinpoint a source of energy, such as an arc, at several hundred meters distance. When the source of the interference has been found, the pieces of metal across which the arc forms must be bonded together to form a low-impedance junction. Gear teeth and other rotating or sliding metal parts present special problems that may be dealt with by using brushes or slip rings (References 19 and 20).

The following guidelines for bonding antenna structures are recommended:

(1) All bonding surfaces should be thoroughly cleaned of any nonconductive material.

(2) All structural support members should be welded or brazed.

(3) After bonding, the exposed areas should be given a protective coating to prevent corrosion.

(4) Adjustable or movable parts of the antenna and structure must be bonded. Adjustable sectors of the parabolic reflector surface should be bonded to the frame. These bonds must have low impedance at all operating frequencies, including the operating frequency of other antennas in the immediate vicinity (Reference 21).

3.6 BONDING OF OTHER EQUIPMENT

Applicable guidelines for bonding electronic servo systems, computers, command encoders, data processing equipment, and time standards are given in Section 3.1. Also, the bonding procedures set forth in Section 3.1 are recommended for the bonding of motors, generators, alternators, and power systems; rotating shafts should be bonded through a phosphor-bronze ring or through conducting grease to reduce static discharges.

Where applicable, the bonding procedures guideline set forth in Section 3.1 is also recommended for teletype equipment, vending machines, and motor vehicles.

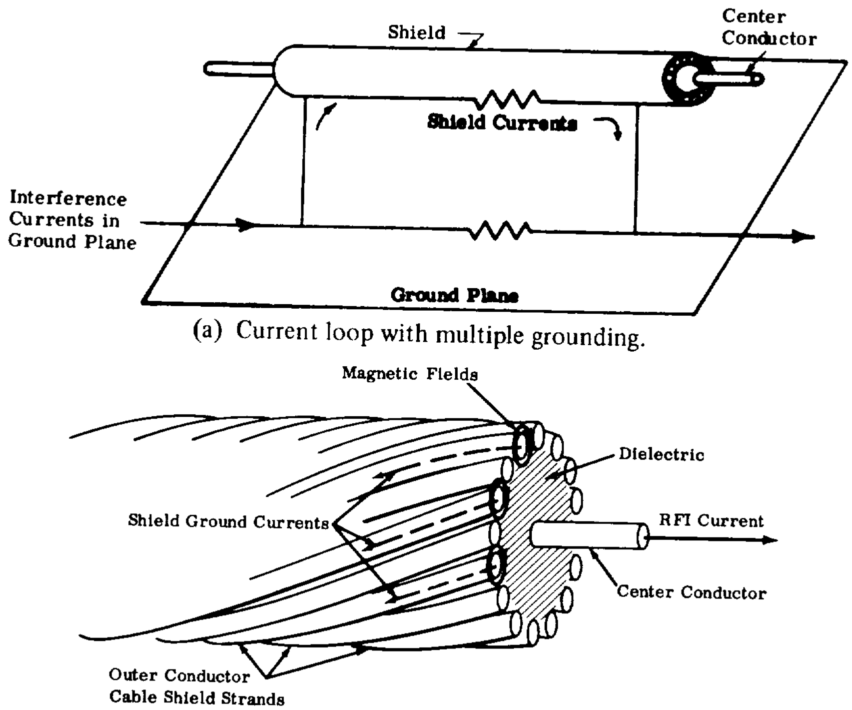
4. GROUNDING

The overall ground system must meet requirements for personnel safety as related to the electrical power system, lightning protection of personnel and property, and electromagnetic compatibility by providing a quiet, earth-potential common bus for the grounding of electronic equipment. The last requirement is met by the signal ground system as part of the overall ground system. The signal ground system must function as a ground over a wide frequency range, as wide or wider than the spectrum occupied by the various signals processed by the electronic equipment (Reference 22).

Grounding techniques for the power system and for lightning protection are well developed and have been standardized (see Bibliography). Signal ground systems, on the other hand, have not been well standardized (Reference 23).

The purpose of this section is to suggest a grounding method for the signal system. The uniform plan presented herein recommends single-point grounding for low-frequency and instrumentation circuits and multiple-point grounding for high-frequency applications.

The distinction between LF and HF grounding is based on the different interference frequencies to which circuits are likely to be susceptible and on differences in the performance of shields in the HF and LF ranges. Low-frequency circuits are affected primarily by the power frequency and its harmonics (60, 120, and 180 or 400, 800, and 1200 Hz) and are relatively insensitive to HF effects. Interference fields from nearby power lines are made up of both electric and magnetic field components. Shielding is effective against electric fields at power frequencies whether shields are grounded at one or both ends. For magnetic fields, however, conventional shields have no appreciable shielding effect unless impractically thick, solid iron shields are used. If more than one ground point is used, the shield and ground of the structure will form a large loop which may be closely coupled to audio leads of field station equipment. Circulating shield currents will inevitably make interference conditions worse than with no shield at all. Shield ground currents create magnetic fields which induce an RFI current in the center conductor (see Figure 2.18). These currents cannot flow when the shield is grounded only at one point (see Figure 2.19). If leads are grounded at more than one point, return currents will not balance sending currents and a net field will be generated to create potential interference. Shielding of the electric field can be provided by grounding the shield at a single point.



(a) Current loop with multiple grounding.

(b) RFI current induced in center conductor.

Figure 2.18.–Effect of circulating shield currents in producing interference.

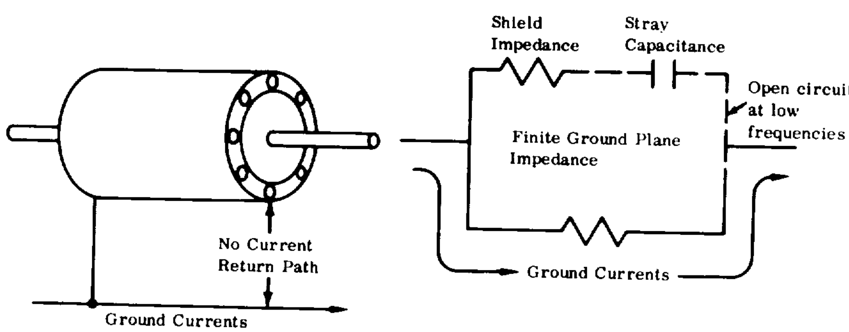


Figure 2.19.–Elimination of low-frequency loops between shield and ground by means of single-point grounding.

Leads carrying frequencies in excess of 1 MHz are far less sensitive to power-frequency interference than to HF disturbances. To effectively protect these leads from electric and magnetic field coupling, they must be completely enclosed by the shield. In completely closing the shield, the intent is to prevent HF currents of ambient fields from flowing on the shield interior. To ensure the proper contact, the cable shields should be grounded at both ends. Good electrical contact can be established by grounding the shield completely around the periphery of the connector shell. The use of pigtail grounding should be avoided on all HF cables. Where it becomes necessary to use a pigtail, the length should be minimized. Generally, it is not necessary to ground HF shields at points other than at both ends, unless the cable is unusually long or carries high-level signals.

4.1 SINGLE-POINT GROUNDING (SPG) SYSTEM

Data acquisition in areas of high ambient noise demands the use of differential or balanced circuits. Since a differential circuit responds only to the voltage applied between its input leads, the noise voltage at the source may be above ground potential by a considerable amount without degrading circuit performance. Figure 2.20 illustrates such a differential circuit. The input voltage, V_g, is the voltage to which the device responds. The noise voltage, V_n, is simultaneously impressed on both input leads but is balanced out in the input to the device because each input lead has the same impedance to ground. Thus, the device does not respond to the ambient noise. In theory, the ambient noise voltage is cancelled out, assuming the impedance of V_g is zero ohms. In practice, there is always some unbalance in the differential device or associated circuitry, and some part of V_n will appear as a difference voltage across an equivalent resistance R.

The noise voltage differential causes a reduced output signal-to-noise ratio. Figure 2.21 shows how the unbalance causes a portion of V_n, ΔV_n, to appear across the input terminals of the device.

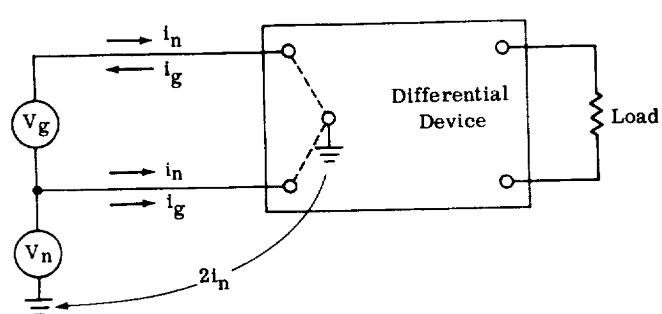


Figure 2.20.–Schematic diagram of a differential or balanced circuit.

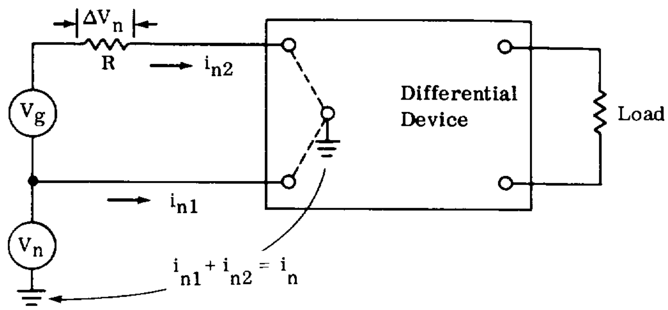


Figure 2.21.—Effect of unbalance in a differential circuit.

The noise generator shown in Figure 2.21 will produce a voltage ΔV_n between the two input leads that is caused by the noise current i_n flowing in the finitely conductive ground plane. The generation of this voltage is illustrated in Figure 2.22. If the ground connection to either the load or source is removed, thereby establishing a single-point ground (SPG), a ground loop is no longer formed, and no appreciable common-mode voltage V_n can develop. This is the advantage of an SPG. If the connection to one of the grounds is removed, an SPG is achieved only at dc and very low frequencies. At high frequencies, the ground loop is completed by capacitive coupling to the ground plane through C_d, as shown in Figure 2.23 (Reference 24).

The SPG system is practical and effective only in frequency ranges in which distributed and stray capacitances are not large enough to degrade it. The high-frequency equivalent of a nominally SPG system is shown in Figure 2.24. Only a few of the ground loops are shown, but these are sufficient to illustrate the point.

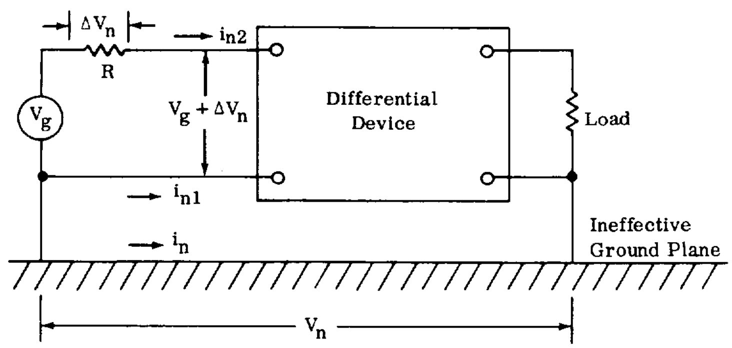


Figure 2.22.—Common-mode voltage generated by current flowing in a finitely conducting ground plane.

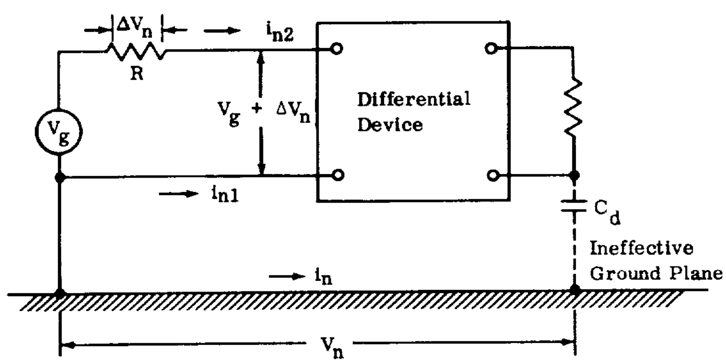


Figure 2.23.–Ground loop produced by distributed capacitance.

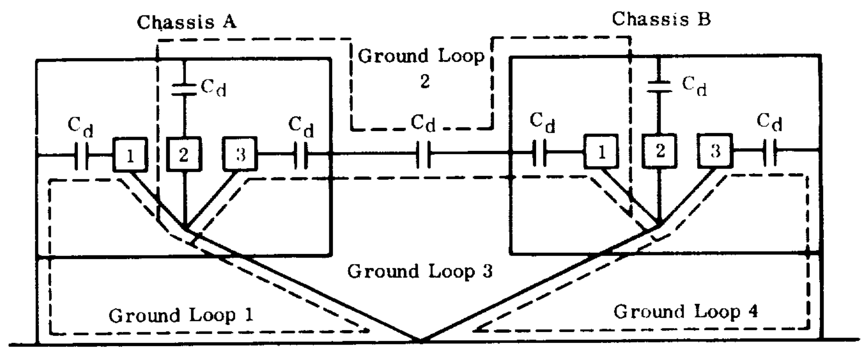


Figure 2.24.–Ground loops produced by several distributed capacitances.

There is a transitional frequency range in which the impedance of the stray ground loops is too small to be called single-point grounding and too large to be true multiple-point grounding. Grounding systems within this range are termed hybrids. Distributed capacitance and the filter capacitance to ground generally establish upper frequency limits for SPG requirements.

4.2 MULTIPLE-POINT GROUNDING (MPG) SYSTEM

The voltage and current of high-frequency signals are distributed along a conductor according to the spatial retardation effects of electromagnetic energy transmission. A conductor or shield grounded at only one point behaves like a grounded monopole antenna, and the potential developed at the ungrounded end must be considered because the conductor may then become a reasonably efficient antenna system even at physical lengths that are a relatively small fraction of the signal wavelength. Thus, it may radiate energy from a noise source and expose other equipment to the electromagnetic fields, or it may act as a receiving antenna.

The antenna effect may be largely overcome by grounding the conductor at its ends. In an equipment cabinet that cannot be solidly bonded to the ground plane, the diagonally opposite corners of the mounting surface may be grounded. If coaxial cables pass through a metal bulkhead, a bulkhead adapter with the outer shield grounded to the bulkhead should be used. The multiple-point ground (MPG) system must be employed in the frequency range above 1 MHz because of the degrading effects of ground loops caused by distributed capacitances in the high-frequency range.

4.3 THE ROLE OF STRUCTURE

The mechanical elements of the system are defined as its "structure." A structure may be composed of–

(1) Frames that hold circuit board receptacles.
(2) Racks for mounting frames.
(3) Ducts, trays, and supporting frame of the cabinet.
(4) Cabinet casework.
(5) Miscellaneous items that support or influence electrical currents.

Although structural design has traditionally been the responsibility of the mechanical engineer, it has become an area of increasing interest to the circuit designer because structures frequently function as a shield and a ground reference for circuitry. A highly conductive structure is desirable because it allows the circuit designer considerably more flexibility than an insulating structure (Reference 25). The circuit designer should negotiate with the structural designer to provide solid conductive paths between modules, power supplies, and peripheral equipment elements as an integral part of the structure. A recommended requirement of such negotiations is that every joint in the structure must exhibit good conductivity. To the structural designer this means that maximum areas must be provided for metal-to-metal contact and welding, brazing, or high pressure fasteners for joints.

For power frequencies and for dc, the inductance and capacitance of the structural elements may be disregarded; the effect of resistance on voltage drop is the primary consideration. At higher frequencies, inductance is a more important factor. This simply means that structural members stressed by RF energy should have low inherent inductance. Interpreted structurally, low-inductance design involves designing members with a high ratio of surface-to-cross-sectional area. Permeability can also have a significant bearing on the self-inductance of the structure. For example, by using iron instead of copper or aluminum, the inductance increases by a factor of 10 to 100.

4.4 DESIGN OBJECTIVES

A specific objective of good design practice is to bring all chassis of a system to a common potential, or failing that, to minimize the potential differences between them. A shift in the reference potential of a signal (which can be minimized but not completely eliminated) occurs when any signal is transmitted by a wire or cable over any distance. In addition to a shift in the reference potential in the cable, the chassis of the line termination equipment will not be at the same potential as that of the source equipment for other reasons. The structure may be "hot" or the ground bus connecting the chassis to the system reference may have excessive voltage drop. Thus, an unavoidable net difference in reference voltage occurs between the receiving chassis and the incoming circuit. Even a good structure cannot completely prevent cable flux linkages, but it will at least bring the two chassis to a common potential. A good design approach is to attempt to get all of the chassis within some arbitrary voltage difference tolerance and to identify cables that run between chassis that have a voltage difference. The design objective for cabling is to minimize the net cable current, i.e., to arrange cables so that the vector sum of all currents entering and leaving a specified volume is minimum.

Circuit designers should develop a scaled module layout showing all module-to-module interconnecting cables. A chart should be prepared to tabulate system waveforms, as illustrated in Table 2.3.

Table 2.3—Example of typical waveforms carried by cables interconnecting modules.

Signal Frequency (kHz)	Waveform	Amplitude Peak-Peak (V)	Rise Time (μs)	Pulse Width (μs)	Load Resist-ance (kΩ)	Location (card-des-ignation)
0.300	Pulse	6	1.5	400	1	CF-3
9.6	Square	6	0.5	5	8	CC-7
28.6	Pulse	40	0.5	10	2	AC-9
100	Sine wave	32	0.0	0.0	7	Oscillator
186	Square wave	6	0.1	2.7	1.5	CA-4

The next step is to use this information as a guide to establish the grounding and protection techniques for each lead or set of leads, as described in the following paragraphs (References 14 and 26).

4.5 REQUIREMENTS FOR FIELD FACILITIES

An SPG system is desirable for field station facilities. Each ground system should connect to the earth ground at only one point. If necessary, there should be a ground system for each type of power used. Separate ground systems should also be used for signal leads, structure or interference grounds, and signal shield grounds. All interference-producing devices, filters, and high-frequency shields should be connected to the chassis ground. Prior to the connection of equipment to the rack grounding bus, each ground system in a given rack should be electrically isolated from all others by a minimum resistance of 10 MΩ (Reference 27).

The connection to earth is an essential part of a power grounding system. If this connection is inadequate, even the most elaborate grounding system will be ineffective because there will be no low resistance path to earth.

A buried ground that establishes a good connection to earth must be—

(1) Made of good electrical conductors.

(2) Able to withstand mechanical abrasion.

(3) Able to withstand the effects of corrosion.

(4) Able to provide sufficient contact area with the soil to minimize the grounding resistance.

The resistivity of the earth ground should remain reasonably constant through the changing seasons of the year and be relatively unaffected by normal current flow in the system. To prevent excessive corrosion of the connection, copper or zinc coated (galvanized) iron electrodes should be used. Electrodes driven into the ground may be rods or pipes. If soil conditions permit, a few deep pipes are usually more effective than many short pipes.

The approximate resistance of a rod or pipe driven vertically into the earth as a function of the depth to which the rod is driven is shown in Figure 2.25. The decrease in resistance with increased rod diameter does not justify using rods with a diameter greater than that required to withstand the stresses of driving. Since an effective electrode consists of a rod and 6 to 10 ft of soil around the rod, it is advisable to drive rods at least 12 ft apart. The usual procedure is to drive a group of electrodes in a straight line (this configuration affords the lowest resistance). However, when several ground rods are connected together, their combined resistances may be higher than the calculated parallel resistances of the individual electrodes. This is because only a part of the resistance of each ground electrode connection parallels the resistances of the other electrode connections; the remainder is common. The relationship between the cumulative paralleling efficiency and the spacing between ground rods is shown in Figure 2.26 (Reference 28).

In cases where ground conductivity is very low, it may be advisable to treat the earth with a salt solution around the driven electrodes. The change in resistivity brought about by changing the salt content of sandy loam and red

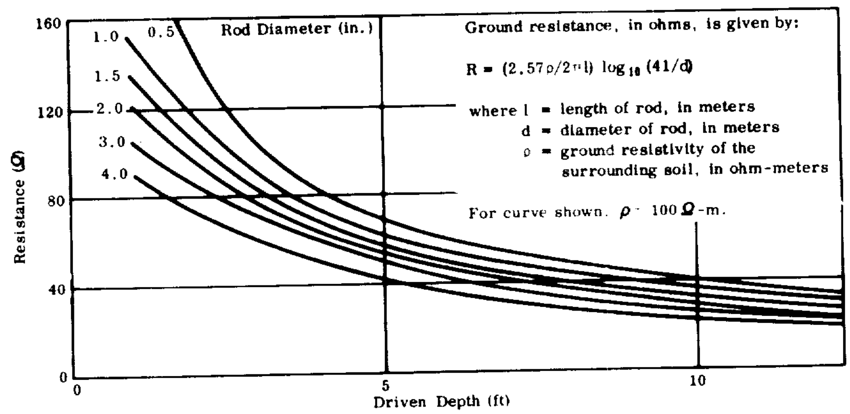


Figure 2.25.–Theoretical variation of ground resistance with depth of ground rod. (After "Electrical Interference," by Rocco Ficchi, 1964)

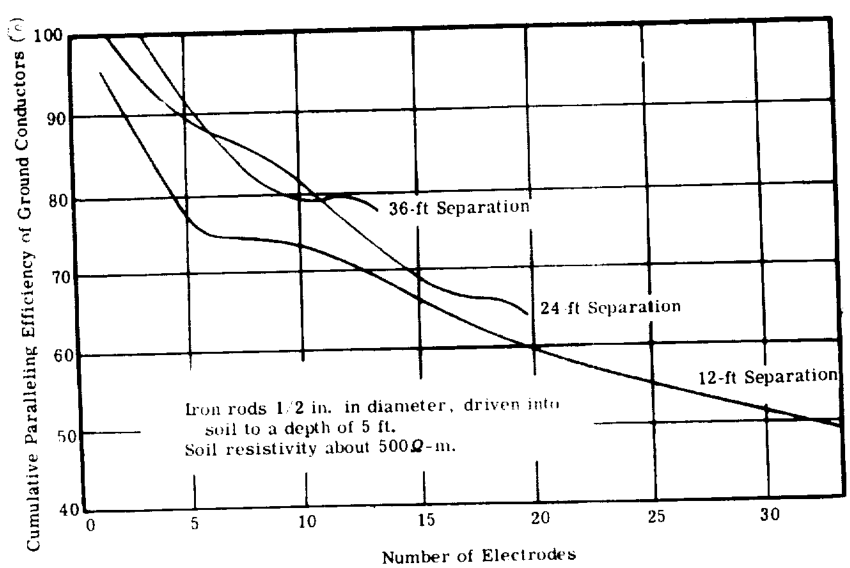


Figure 2.26.–Cumulative paralleling efficiency of rod electrodes in untreated soils for various separation of ground rods. (After "Electrical Interference," by Rocco Ficchi, 1964)

clay from 0.1% to 20% is shown in Figure 2.27. Also shown is the change in ground resistance with time as the salt is carried away by the water in the soil. After approximately a year it is advisable to retreat the soil.

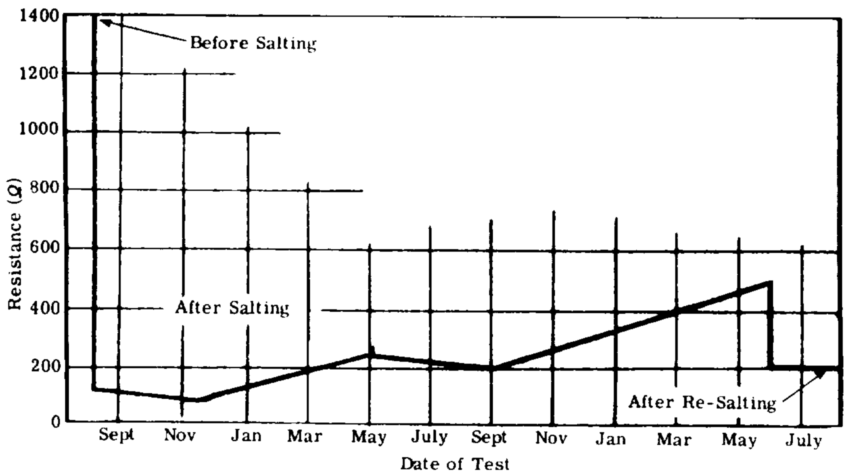


Figure 2.27.–Changes in resistance of a ground connection as the result of salting. (After "Electrical Interference," by Rocco Ficchi, 1964)

The resistivity of salted soil also varies with temperature. Resistivities increase rapidly as temperatures fall below freezing. Sandy loam soil with 20% moisture and 5% salt shows a variation of about ten to one over the temperature range from +20° to −13°C.

Specific earth ground requirements for STADAN stations are given in Reference 2. Additional grounding information is given in NASA-GSFC Specification S-523-P-8, November 20, 1967, "Lightning Protection Guideline for STADAN Ground Equipment." (Also, see Section III.)

4.6 RACK AND CABINET GROUND SYSTEM

Each ground station rack should be provided with a signal ground bus 1 in. wide by 1/4 in. thick, the bus length being determined by the cabinet height. Each chassis in a rack should be designed to have four independent ground systems that are electrically isolated except at one point. All ground systems should tie electrically at one point to the rack signal ground bus (see Figure 2.28). Each ground system becomes a conductive network that provides a reference voltage for circuits requiring the same reference potential. No ground currents should be intentionally allowed to flow in any of the following four independent ground systems:

(1) Signal ground system for instrumentation.
(2) Chassis or interference ground system.
(3) Signal shield ground system.
(4) Power ground system.

Ground loops are usually the result of unintentionally completed paths that permit the flow of currents in grounding circuits. They are most often caused by a high impedance that may develop between two points within the same ground system. A loop which was passive in an original circuit design may become active when corrosion causes a high-impedance joint to develop in the loop. Ground currents can then produce a difference in potential, and the active ground loop becomes a possible source of trouble. The area of a ground loop can vary from a fraction of a square inch in micro-miniature circuits to

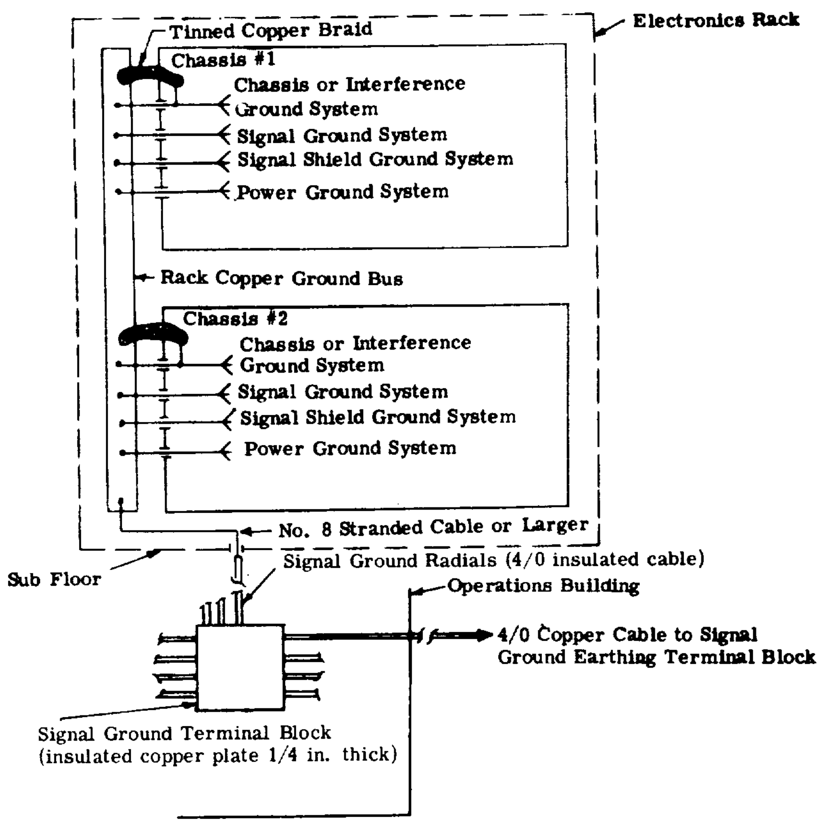


Figure 2.28.–Grounding scheme for a station signal system.

thousands of square feet in buildings and antenna assemblies. Use of SPG systems will reduce the potential in ground loops. Careful implementation of the principles of grounding for circuits and shields will prevent the formation of ground loops that might be formed within a chassis or between chassis or racks.

The structure should be highly conductive and, hence, be designed so that there is no significant impedance between any two members. To achieve this goal, RF bonding described in Section 3 should be used.

4.7 CIRCUIT AND SHIELD GROUNDS

Methods of circuit and shield grounding will be determined by the circuit frequency, sensitivity, and the presence of follow-on circuits. A follow-on is an extension of a circuit beyond its initial termination.

When the leads in a system are indexed and classified according to Section 4.4, the alternatives listed below can be applied. In all classifications, a careful effort should be made to group interconnecting wire bundles so that there will be zero net ground current flow in a bundle.

1. Terminated Circuits

If one end of a circuit is terminated in a load (no follow-on connections) the grounding method should be selected according to the circuit frequency.

For circuits operating at frequencies below 5 kHz, float the terminated end and use an SPG at the other end. Tightly twist high and return leads. If shields are required they should be grounded at a single point.

For circuits operating at frequencies between 5 kHz and 1 MHz, the most desirable practices depend upon circuit sensitivity. For circuits with a sensitivity threshold of 50 mV or less, balanced circuits should be used. If unbalanced circuits are used, the load end should be floated and triaxial shields (grounded at a single point as in Figure 2.29) should be used to reduce pickup of stray energy. For moderately sensitive circuits with a threshold of 0.05 to 1 V, the circuit may be grounded at both ends.

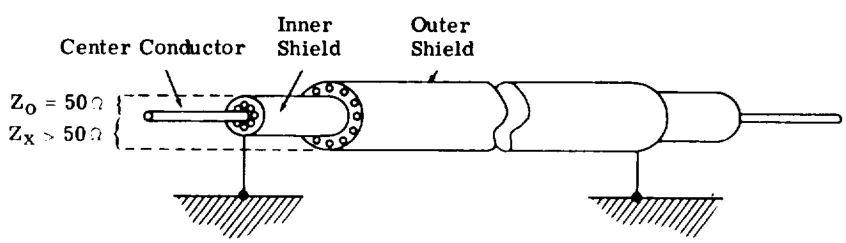


Figure 2.29.—Example of grounding a double-shielded (or triaxial) coaxial cable.

Circuits operating at frequencies above 1 MHz should be grounded at two or more locations. Shields should be grounded at frequent intervals if they are long. A high-quality ground path should be maintained between ground points.

2. Follow-On Circuits

When both ends of a circuit have follow-ons to other circuits, both ends should be grounded. Precautions should be taken in this case to obtain very-low-impedance bonding between frames of the two modules. The interconnection pairs (high and return leads) should be twisted and routed on adjacent connector pins as close together as possible. If the circuit is very sensitive, magnetic field pickup problems can be quite severe. To avoid grounding a circuit in two ground systems, isolation transformers may be used if an overlap in grounding system occurs.

3. Poor Structure Continuity

If the structure continuity is poor* and cannot be improved, the circuit must float at the free end. If both ends have follow-on ties, then some method of isolation is required. Isolation transformers with Faraday shields can be used in extreme cases. Under some conditions balanced circuits may provide the solution. Problems of poor structure continuity are best handled by checking for potential sources of trouble and by incorporating correct measures early in the design stage. The use of very sensitive or high-power circuits in the frequency range between 5 kHz and 1 MHz should be avoided if structure continuity is poor.

4.8 SHIELD GROUNDING FOR LOW-FREQUENCY CIRCUITS

Low-frequency (LF) lines are those intended only for transmission of signals (not power) at audio frequencies. Shields on LF leads should be grounded at one point only. The shields of sensitive instruments and balanced, or differential, circuits should be grounded at the receiving end to prevent excessive induction into cables at that point. In most other cases, shields should be grounded at the sending end to prevent the signals carried on cables from being induced into other circuits. All LF shields should be brought to a single point where they can be securely grounded with the LF common lead and the LF power bus.

*In this context, "poor" refers to loosely joined metal structures with high contact resistance between joints.

When cables are long, the grounding of LF shields becomes complex. Shields are often carried through connectors on leads to preserve the SPG concept. This has the disadvantage of using valuable connector pins and degrading the performance of the shield. When shields are combined and carried through on one pin, the resulting cross coupling can prove to be excessive. It is recommended that long shields be grounded at intermediate connectors, as shown in Figure 2.30. It is desirable to create several short cable shields that are single-point grounded in place of one long, inefficient shield. In addition, cable coverage by the shield is maximized using this scheme because shields can be run directly to the pin; otherwise, cable coverage is lost when shields are stripped back for pigtail leads.

The following rules summarize recommended practices for grounding LF circuits and their shields:

(1) Ground LF shields, LF power supply leads, and LF common leads together at one point only.

(2) Tightly twist all LF leads. Never run the supply and return leads separately or in separate shields.

(3) Never include LF signal leads and power leads in the same bundle, on the same connector, or on the same terminal board if it can be avoided.

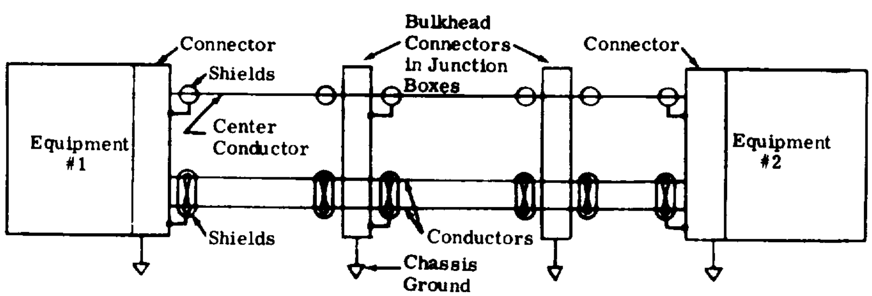


Figure 2.30.–Grounding of a long, low-frequency cable.

4.9 SHIELD GROUNDING FOR HIGH-FREQUENCY CIRCUITS

High-frequency leads are those intended for transmission of signals above the audio frequency range. All high-frequency shields should be grounded at more than one point. When a line is short and there are no intermediate breaks in the shield, grounding at both ends of the cable will suffice. In the case of very long runs where a cable must be routed through one or more intermediate connectors, it is preferable to ground cable shields at each connector (see Figure 2.31). The preferred method of grounding to connectors is illustrated in Figure 2.32.

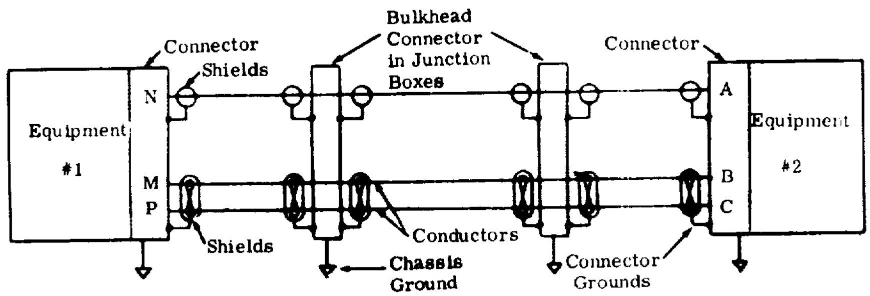


Figure 2.31.–Grounding of a long, HF cable.

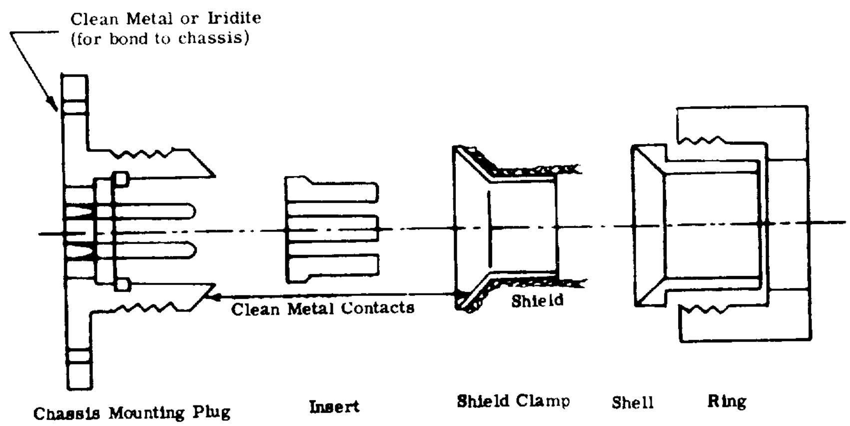


Figure 2.32.–Sectional diagram of a typical shielded plug showing preferred method of grounding to connectors.

Occasionally a circuit will be susceptible to both low- and high-frequency interference. An example of this could be a low-frequency analog circuit, which would normally be susceptible to radiation at power frequencies. In addition, it could be very susceptible to high frequencies modulated at an audio rate. To protect leads against this phenomenon, a hybrid method which offers the advantages of both single and double grounded cables is necessary.

This can be accomplished by grounding one end of the shield conductively and the other end through a capacitor. The value of capacitance will depend upon frequency; however, its value will typically range between 0.01 and 1.0 μF. The cable is thus grounded at a single point at low frequencies and grounded at two points for high frequencies.

4.10 SHIELD GROUNDS

At connectors or junction boxes, it may be desirable to group shield grounds together at one point. This practice may be hazardous when the same shields are used to prevent radiation and also to protect sensitive circuits. In this case, coupling into sensitive circuits is inevitable to some degree.

To reduce coupling into sensitive circuits by common shields carrying different kinds of current, it is advisable to group leads into bundles according to the power they carry. Lines may carry—

(1) Noisy power (to motors, solenoids, and so forth).
(2) Interference-free power.
(3) High-level signal or instrumentation power.
(4) Sensitive signal or instrumentation power.

Cable shields can be grounded together in any wiring bundle without danger of cross coupling when leads are routed in separate groups as classified above. If the grouping of shields and leads is not controlled early in a program, an incurable condition may result. It is unlikely that the cross-coupling problem on shields can be dealt with feasibly in a completed system.

4.11 LENGTH OF GROUND LEADS

Because shields are normally difficult to ground, the length of the ground lead tends to become excessive. The reactance of a short, straight ground lead is so low at low frequencies that it is inconsequential. This is especially true when the diameter of the wire is relatively large so that the ratio of length to diameter is small. However, the effect of this inductance becomes important at higher frequencies.

As an example, a 1-in. length of AWG 20 copper wire has an inductive reactance of 13.0 Ω at 100 MHz. It is evident that at high frequencies the inductance of short pigtail grounds is highly undesirable because it can effectively isolate a double grounded shield from its structure.

The standard equation for the self-inductance of straight, round wires in air is given in Reference 29 as

$$L_0 = 5.08 \times 10^{-3} l \left[-1.0 + \frac{\mu}{4} + 2.3 \log_{10}\left(\frac{4l}{d}\right)\right], \qquad (2.14)$$

where

L_0 = self inductance in microhenrys,

μ = permeability of wire relative to that of copper (taken to be 1.0),

l = length of the wire in inches,

and

d = wire diameter in inches.

Pigtail ground leads introduce undesirable reactance and should not be used for grounding HF shields. For LF shield grounds, it is permissible to allow pigtail leads of 1 to 2 in. The chart in Figure 2.33 is presented to assist in determining the permissible length of shield ground connections (Reference 30).

The problem of terminating shields at connectors becomes highly complex as the cable density of a bundle increases. To eliminate pigtail grounds and simultaneously simplify the problem of terminating shields in high density wire bundles, shields may be connected to chassis through the connector backshell as shown in Figure 2.32. Using this method, shields connect to ground through the backshell to the chassis-mounted plug and finally to chassis. Connectors must be finished with a conductive, noncorrosive finish and be securely bonded to ground to obtain acceptable HF grounding.

4.12 CIRCUIT AND POWER RETURN GROUNDS

A summary of recommended practice for power and return circuits is itemized as follows:

(1) Separate return conductors should be connected back to the source for delivering ac power, dc power, or power to interference-generating circuits as well as for instrumentation and signal circuits.

(2) An SPG should be established for each ground system. This point should be centrally located and permanently bonded to the structure.

(3) Returns of a given type should not share a common conductor when there is a possibility of circuit coupling.

(4) Circuits that produce large, abrupt current variations should have a separate grounding system or should be provided with a separate return lead to the SPG. This reduces transient pickup in other circuits.

(5) When returns of a given type are combined, as for a group of returns from a single frame, a return bus may be used. This bus should be of minimum length and be flat and of low impedance. The cross section of the return bus should be at least as great as that of the combined return conductors connected to it.

(6) All circuit returns should conform to the following principles:

- (a) The two leads of a given circuit should be routed together as close as possible.
- (b) Leads should be routed as close to the ground reference plane as possible.
- (c) Circuit returns should never be shielded separately or carried outside the shield used for the corresponding hot conductor.
- (d) Leads carrying audio frequencies should be tightly twisted with their returns (18 turns per foot is satisfactory) and carried in insulated shields grounded at one end only.

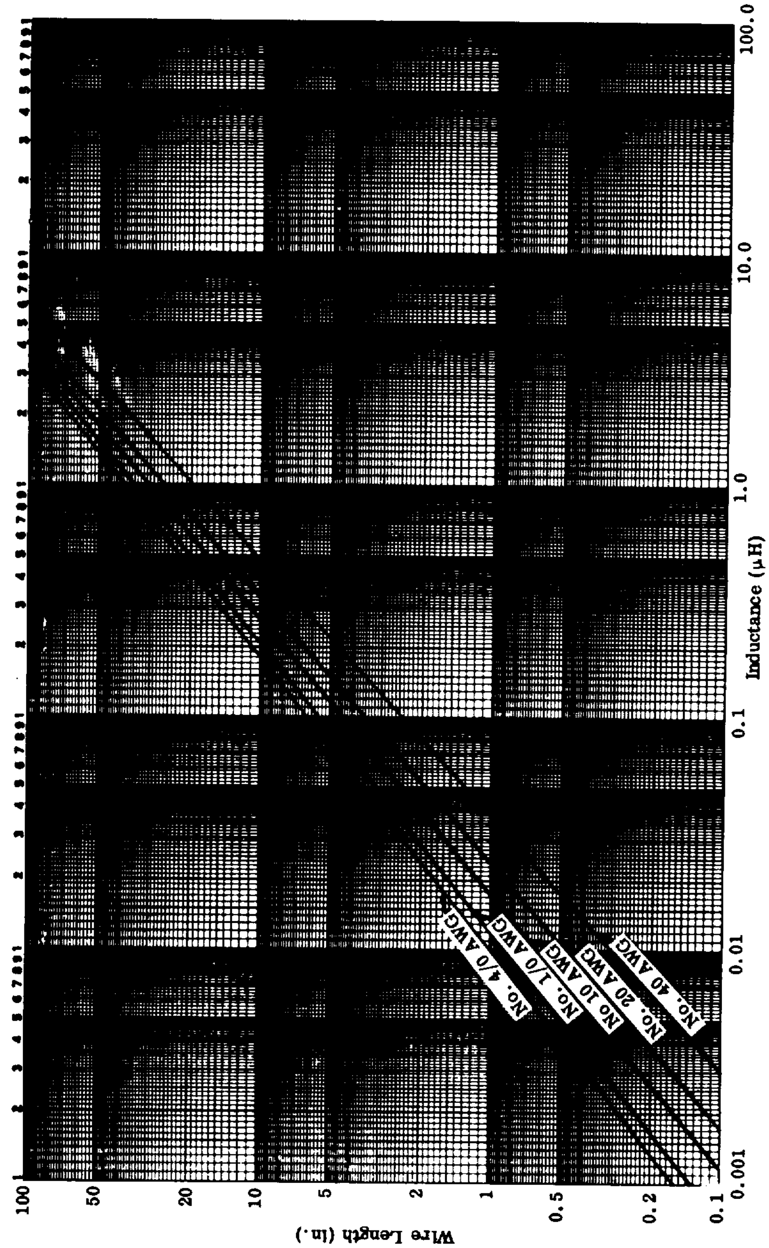


Figure 2.33.—Self-inductance of a straight, round, copper wire. (After "Radio Engineers Handbook," by F. E. Terman, 1943)

(7) The high-voltage and grounded leads of power circuits should be formed into a twisted pair to reduce magnetic field generation. The two leads should pass through adjacent connector pins to reduce capacitive coupling to other circuits. All power returns for modules must be isolated from the frame. They should be bunched together and routed as far as possible from signal circuits.

4.13 RECEIVERS AND TRANSMITTERS

Grounding minimizes coupling between circuits, establishes an equipotential plane, eliminates inductive loops, and provides low-impedance paths for return currents. The following guidelines should be employed to achieve this:

(1) At frequencies below 5 kHz, the SPG system should be used.

(2) At frequencies above 1 MHz, the MPG system should be used.

(3) A hybrid of both the SPG system and the MPG system may be necessary to achieve good grounding in the critical frequency range from 5 kHz to 1 MHz.

(4) Circuits that produce large, abrupt current variations should have a separate grounding system or should be provided with a separate return conductor to the SPG.

(5) A metal chassis or cabinet used as a ground return circuit should be made of metals having high electrical conductivity.

(6) Grounding impedances should be kept low by using the shortest possible grounding leads.

(7) Grounds for low-level signals should be isolated from all other grounds.

(8) Grounding straps should have the maximum practical surface area and conductivity.

(9) Grounding of the antenna structure must conform to the requirements for lightning protection and safety of personnel.

4.14 OTHER EQUIPMENT

The grounding guidelines of Section 4.8 are applicable. Personnel safety and electrostatic charge dissipation are additional important considerations. The grounding designed with short, low-inductance leads, as outlined in Section 4.8, should be sufficient.

5. FILTERING

This section is devoted to a discussion of the design and application of filters commonly used in EMI work. A complete mathematical treatment of the principles of interference reduction will be found in various texts listed in the bibliography which deal exclusively with theory.

Filters are normally used as a last resort whenever it becomes necessary to reduce interference currents flowing in a conductor. An initial effort should be made to design circuits that are inherently free of such currents. Failing this, filters should be utilized to limit the magnitude of EMI currents and to confine the currents to the smallest practical physical area. Properly installed filters can significantly suppress radiated and conducted EMI.

After the basic circuit design has been determined, it is necessary to establish a plan for dealing with any remaining EMI. In general, this plan will depend upon circuit layout, shielding, bonding, and filtering.

The effectiveness of any EMI filter is greatly influenced by the impedance of the noise source and the load impedance. Manufacturers of EMI suppression filters normally specify the filter insertion loss with fixed source and load impedances, usually 50 Ω. The actual insertion loss realized in a practical circuit may be different from 50 Ω. This aspect should be taken into consideration when specifying or using EMI filters (Reference 31).

The number of filters required for a given device and the amount of insertion loss required in each filter should be determined as early as possible in the design phase of a program. If the equipment is small and relatively simple, it is possible to establish filter requirements as part of the original circuit design. Usually a filter is inserted on each power input lead. Filters are mounted so that the filter body forms a part of the basic enclosure, with the input and output terminals of the filter on opposite sides of the enclosure (Reference 32).

If the equipment is large and complex, the filter requirements are generally established by circuit EMI evaluation in the breadboard stage. Interference specifications limit the amount of conducted EMI that may be present on the power input leads and the amount of radiated EMI that may be present at some fixed distance from the test sample. In the breadboard design stage, it is possible to make accurate measurements of conducted EMI on the power lines; this is compared to the specification limit. The amount by which the EMI present exceeds the level of the specified limit is the amount of filtering required.

Evaluation of the radiated EMI characteristics is somewhat less direct. Of course, the actual level of radiation can be measured. This level will be reduced by the shielding of the final enclosure.

Another method for evaluating EMI in the breadboard stage of development is based on a Fourier analysis of waveforms in the various circuits. Repetitive nonsinusoidal waveforms give rise to harmonic components at frequencies far above the fundamental repetition rate. Whether radiated or conducted, these components become a major factor in all broadband EMI problems, both radiated and conducted. A survey of the more powerful waveforms is made with an oscilloscope; the waveforms to be analyzed are photographed. The spectral distribution of harmonic components should be computed and plotted for each case. This set of graphs enables one to judge which circuits are most likely to be a major EMI contributor (Reference 33).

In the following paragraphs it is assumed that filtering requirements have been established.

Once the requirement for insertion loss is known for a given filter, the next step is to design the filter circuit. If actual circuit impedances are known in the frequency range of interest, then these values should be used for calculating circuit values. If they are not known, then an arbitrary impedance, usually 50 Ω, is assumed and a filter circuit is designed. This circuit is installed in the breadboard model and the reduction in conducted EMI is measured at several frequencies above the nominal cutoff frequency of the filter. The amount that this insertion loss differs from the calculated (50-Ω) insertion loss is noted. A final circuit is then designed, still based on the 50-Ω impedance but correcting for discrepancy in the insertion loss determined above. This procedure will yield the correct filter design.

Certain guidelines are helpful in deciding what type of filter circuit to apply in any given instance. If it is known that the filter will connect to relatively low impedances in both directions, then a circuit containing more series filter elements is indicated (a T-circuit, for instance). Conversely, a high-impedance system calls for a π-filter. If the filter is connected between two severely mismatched impedances, then an asymmetric filter circuit such as two L-section elements can be used. The series element faces the low-impedance side of the system.

At frequencies well above cutoff, a properly designed filter provides transmission loss that increases at 20 dB per decade per reactive element. This is sometimes expressed as 6 dB per octave per reactive element (Reference 34).

The classical π-type filter circuit, consisting of an inductor and two capacitors, can be used for any low-pass filtering in communications circuits or for suppressing electromagnetic interference in power lines.

Equations and charts for the design of π-type filter circuits are readily available in standard reference books; however, these were developed primarily for applications in which current and voltage levels are relatively low. When such data are used to design a π-network for use as an EMI filter, the current and voltage levels are usually so large as to require elements of impractical sizes.

For example, a Butterworth π-filter with a cutoff frequency of 10 kHz calls for a capacitance of 0.3 μF and an inductance of 1.6 mH. The design of this inductor presents no problem if it is required to carry only 10 or 20 mA. If it must carry 20 A without saturating or overheating, it may become too bulky to be practical. The desired transmission loss at frequencies above cutoff may be obtained with elements of practical size by using a different set of values for both series and parallel filter elements. The component values for such modified circuits are not easily derived from the design equations usually given in handbooks, and it is difficult to determine the transmission loss of the modified filter as a function of frequency (usually specified in terms of a matched 50-Ω system).

5.1 LOW-PASS FILTER DESIGN CHARTS

With the charts presented herein, practical low-pass filters can be designed to reject RFI above the upper cutoff frequency, and their transmission loss can be determined rapidly. Two related equations are the basis for the nomographs and charts (Reference 35). First, consider a simple circuit consisting of only a source and load, both of which have impedances of equal value as shown in Figure 2.34a. The output voltage for this circuit is simply

$$V_1 = \frac{V_{in}}{2}. \tag{2.15}$$

Next, consider the circuit with a π-type filter inserted between the source and load impedances, shown as Figure 2.34b. Writing the loop equations for this circuit and solving for the load voltage, the output voltage is

$$V_2 = \frac{V_{in}}{2 - 2\omega^2 LC + j[(\omega L/R) + 2R\omega C - \omega^3 RC^2 L]}. \tag{2.16}$$

Substituting $V_{in} = 2V_1$ from (2.15) into (2.16) and determining the square of the absolute magnitude of the voltage ratio leads to

$$\left|\frac{V_1}{V_2}\right|^2 = 1 + \omega^2\left(\frac{L^2}{4R^2} + C^2R^2 - LC\right) + \omega^4\left(\frac{L^2C^2}{2} - LR^2C^3\right) + \frac{\omega^6 R^2 L^2 C^4}{4}. \tag{2.17}$$

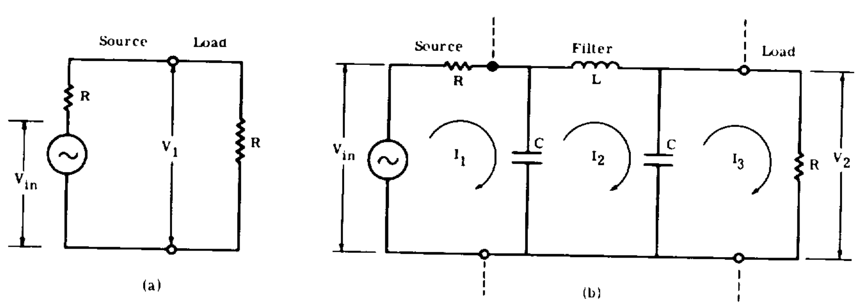


Figure 2.34.–Source and load connected directly and with filter interposed.

Filter insertion loss in dB is defined as

$$\text{I.L.} = 10 \log_{10} \left| \frac{V_1}{V_2} \right|^2 \tag{2.18}$$

To simplify (2.17), let the damping ratio be represented by d:

$$d = \frac{L}{2CR^2} . \tag{2.19}$$

As ω becomes very large, the ω^6 term dominates (2.17). A cutoff frequency, $f_0 = \omega_0/2\pi$, can be defined from this term, where

$$\omega_0^6 = \frac{4}{R^2L^2C^4} \tag{2.20}$$

so that

$$\omega_0 = \left[\frac{2}{RLC^2}\right]^{1/3} .$$

If the expressions for d and ω_0 are substituted into (2.17) and appropriately rearranged, the insertion loss in dB is

$$\text{I.L.} = 10 \log_{10} \left[1 + \frac{\omega^2}{\omega_0^2}\left(\frac{1}{d^{1/3}} - d^{2/3}\right)^2 - 2\frac{\omega^4}{\omega_0^4}\left(\frac{1}{d^{1/3}} - d^{2/3}\right) + \frac{\omega^6}{\omega_0^6} \right] . \tag{2.21}$$

This expression can be simplified further by letting

$$D = \frac{1}{d^{1/3}} - d^{2/3} \tag{2.22}$$

and by normalizing the frequency of interest with respect to cutoff frequency f_0:

$$F = \frac{f}{f_0} = \frac{\omega}{\omega_0} .$$

The insertion loss in dB for a π-type filter circuit can then be expressed as

$$\text{I.L.} = 10 \log_{10}[1 + F^2D^2 - 2F^4D + F^6]\,. \tag{2.23}$$

The insertion loss response curves (Figures 2.35 and 2.36) are based on this equation. The response curves are numbered from 0 to −6 for Figure 2.35

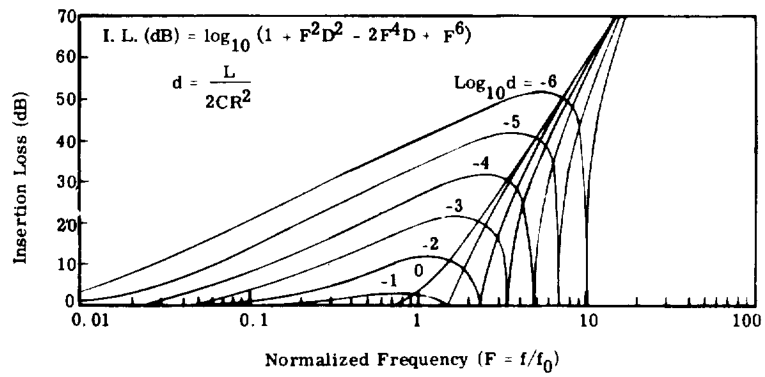


Figure 2.35.–Curves for insertion loss for an under-damped, low-pass, three-element, π-type filter. (After J. C. Shifman, IEEE Trans. EMC, 1965)

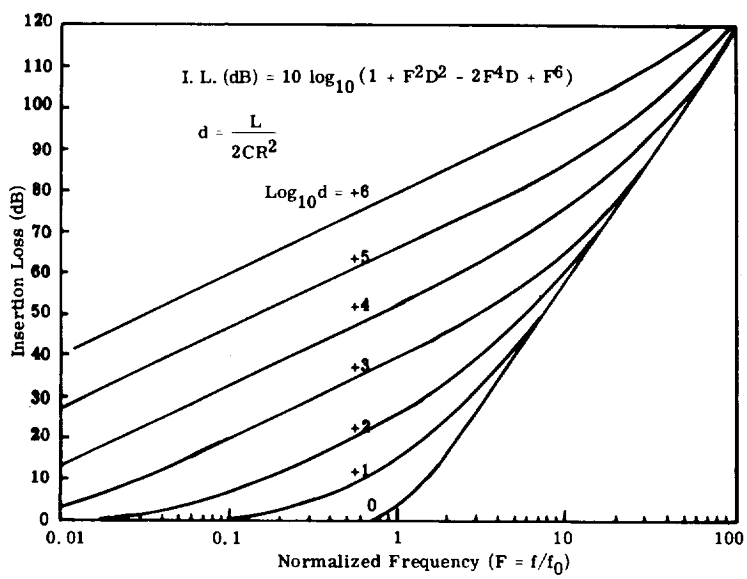


Figure 2.36.–Curves for insertion loss for an over-damped, low-pass, three-element, π-type filter. (After J. C. Shifman, IEEE Trans. EMC, 1965) and from 0 to +6 Figure 2.36. These numbers correspond to the logarithm of the damping ratio, $d = L/2CR^2$. Thus, for a value of $d = 0.01$, the -2 curve of Figure 2.35 represents the transmission response of the filter.

5.2 INSERTION LOSS CALCULATIONS

As an example of the use of these curves, consider the case in which it is required to find the insertion loss at 10 kHz for a three-element filter having a cutoff frequency of 50 kHz and a damping ratio, $\log_{10} d = -4$. The normalized frequency is $10/50 = 0.5$. By entering the graph at the abscissa value 5 and projecting upward on Figure 2.35 to curve -4, the value of the insertion loss is read from the left-hand scale as 20 dB.

As another example, let it be required to find the insertion loss at 40 kHz for a three-element filter having a cutoff frequency of 80 kHz and a damping ratio, $\log_{10} d = +3$. The procedure is the same as that described, but since the damping ratio is positive, Figure 2.36 must be used. For the normalized frequency of $40/80 = 0.5$, the insertion loss is found to be 34 dB.

The nomograph of Figure 2.37 is based on (2.19) and (2.20) and represents the relationship between inductance L, capacitance C, cutoff frequency f_0, and response curves of Figures 2.35 and 2.36. Any straight line intersects the L, C, and f_0 scales at values that satisfy $\omega_0 = (2/RLC^2)^{1/3}$, where $R = 50\ \Omega$. Each sloping guideline in the nomograph is labeled with the logarithm of the value of d associated with the slope.

The data for Figures 2.35, 2.36, and 2.37 are valid only for $R = 50\ \Omega$. For some other resistance, $R' = KR$, the required values of inductance and capacitance are $L' = KL$ and $C' = C/K$, where $K = R'/R$ is the scale factor and the primed letters designate the required circuit values after scaling. Impedance scaling has no effect on the shape of the response curve, as the same values of attenuation occur at the same frequencies as before (Reference 36).

An example of how to use Figure 2.37 to determine the insertion loss of a symmetrical π-type low-pass filter circuit in a matched 50-Ω system is represented by the dotted line in the nomograph. For this example, L is 50 mH, C is 0.1 μF, and f_0 is 7 kHz. The projected response curve is obtained by noting which of the slanted lines marked, Damping Ratio, d, is most nearly parallel to the straight line connecting L, C, and f_0. In this case, the +2 line is used. Insertion loss as a function of frequency for π-circuits having damping ratios lying between -6 and +6 is shown in Figures 2.35 and 2.36 where the frequency scale is normalized to $f_0 = 1$. To get the actual frequency values, the frequency scale must be multipled by the cutoff frequency determined from Figure 2.37.

The insertion loss for the above π-type circuit is therefore determined by the curve labeled +2 in Figure 2.36. For this case, f_0 is 7 kHz and the normalized frequency scale in the figure must be multiplied by 7×10^3 to obtain the actual frequency values.

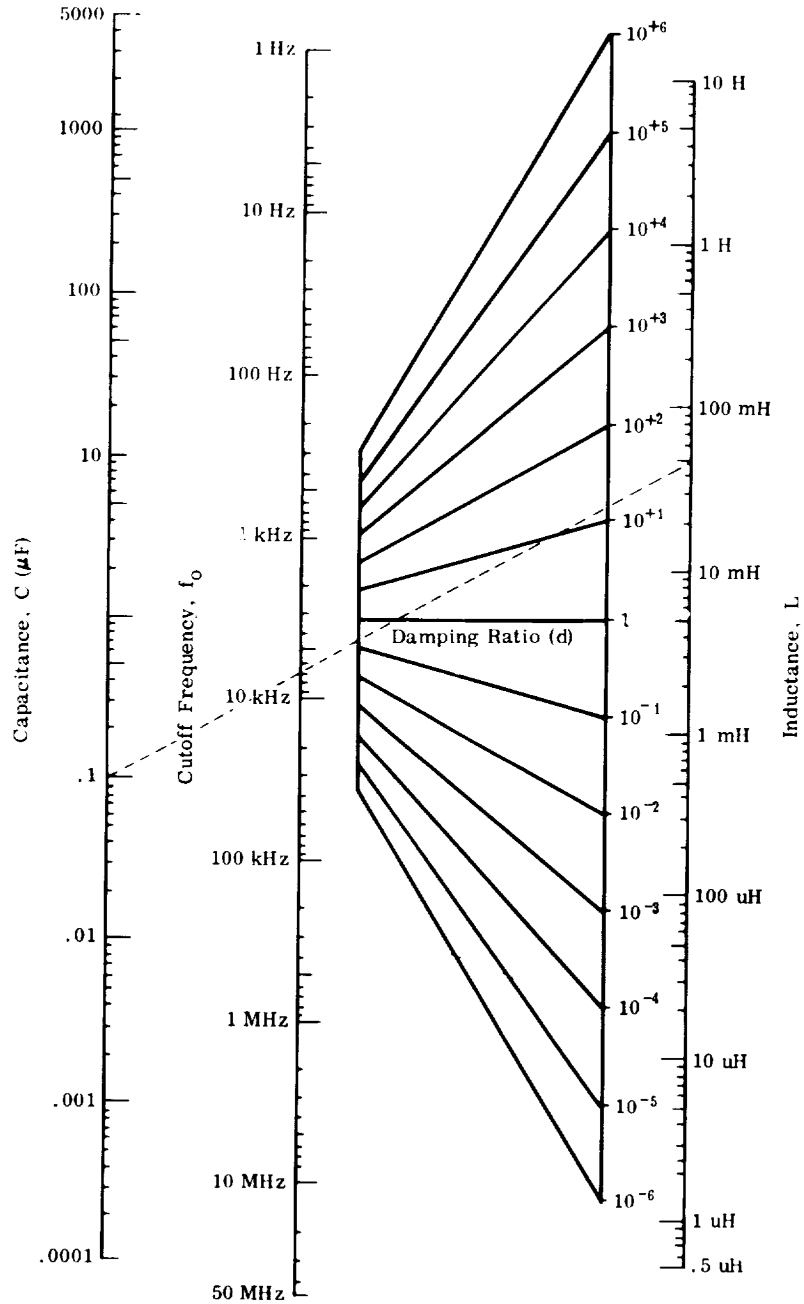


Figure 2.37.–Nomograph for computing constants of a π-type filter. (After J. C. Shifman, IEEE Trans. EMC, 1965)

5.3 THE Π-CIRCUIT DESIGN

Suppose it is necessary to design a π-type filter circuit that has 35 dB of insertion loss at 50 kHz. Assume that a zero damping ratio is the most desirable; then, from Figure 2.35 or Figure 2.36, at 35 dB read $F = 3.8$. From this determine f_0 by

$$F = \frac{f}{f_0} = \frac{50 \times 10^3}{f_0},$$

$$3.8 f_0 = 50 \times 10^3,$$

and

$$f_0 = 13.2 \text{ kHz}.$$

Place a straight edge on the nomograph (Figure 2.37) so that it lies parallel to the line of unity filter damping ratio and passes through 13.2 kHz on the f_0 scale. Read $L = 1.1$ mH and $C = 0.20$ μF, which are the component values for the π-circuit. If there are no restrictions on the component values or their physical size when determined as indicated, then the assumption of unity damping ratio in Figure 2.37 is valid.

When establishing requirements for filter insertion loss it must be kept in mind that different filters of the same design will display somewhat different interference characteristics due to component tolerance variations. For this reason, it is important to allow a safety margin in insertion loss requirements. A common practice is to allow at least 6 dB for this margin in the stop band.

5.4 SELECTION OF INDUCTORS

Filter inductors are usually wound on toroidal cores of powdered iron, molybdenum permalloy, or ferrite material. The choice of materials is determined by the operating frequency and current rating. The powdered iron cores can be used for all dc applications and for most 60-Hz power frequency applications. For high-current 60-Hz and 400-Hz applications, molybdenum permalloy cores should be used. For extremely low current (<0.1 A) applications, the ferrite materials should be considered. The size of the core is determined by the required inductance and current rating. The product of the number of turns and the peak current must be limited to a value that will not drive the core more than 50% into the magnetic saturation region.

Windings should be placed on the coil so that input and output turns are separated as much as possible. The resistance in the windings plus losses in the core cause all the heating that will occur in the filter. This heating should be taken into consideration when rating the filter for operation at a specified ambient temperature. An empirical relationship has been developed which indicates the approximate temperature rise of the containing case as

$$\text{Temperature rise in C}^\circ = \frac{P}{0.006A} \qquad (2.24)$$

where

P = power dissipated in watts

and

A = total surface area of filter case in square inches.

This expression is based upon the typical heat dissipation characteristics of tinned steel cans.

5.5 SELECTION OF CAPACITORS

Capacitor selection is determined by the voltage, temperature, and frequency ranges in which the capacitor must operate. Most EMI power line filters are rated for certain standard voltages. For 28-Vdc applications, capacitors with a 100-wVdc rating are quite adequate. Metallized mylar units offer the most compact capacitor with good reliability and low dissipation factor. Lead length should be kept short to improve the high-frequency performance.

If a large amount of capacitance is required in a small space, tantalum capacitors may be considered. However, this type of capacitor is sensitive to overvoltages and can be damaged by reverse polarity. The dissipation factor is considerably higher than for Mylar units and the high-frequency response is worse.

For 120-Vac applications, the capacitor should be rated as a 400-wVdc unit, suitable for ac use. A Mylar and foil unit or a paper-Mylar and foil unit is recommended. The dissipation factor is low and high-frequency performance is good. For 240-Vac applications, an oil-impregnated paper and foil unit is recommended.

If good capacitor performance is to be expected above 50 MHz, it becomes necessary to make use of designs using feedthrough capacitors which eliminate the lead inductance problem.

5.6 FILTER ENCLOSURES

An EMI filter should be enclosed in a steel case with a conductive outer finish. The terminals should be arranged so that the input and output leads are physically separated by the greatest possible spacing. The internal design should be such as to provide maximum isolation of individual filter circuits. If crosstalk is critical, a solid metal barrier between circuits should be used. The mounting plate of feedthrough capacitors should form a sealed barrier so that interference cannot "couple past" the feedthrough. Filter circuits carrying high current should be isolated from low-level circuits.

5.7 FILTER INSTALLATION

When an EMI filter is installed, care must be taken to ensure that its performance is not limited by poor mounting techniques. Ideally, the filter should be mounted as part of a natural barrier, such as the main circuit chassis or enclosure. This tends to prevent interference coupling from the input to the output. In no case should the filter input and output leads be bundled together.

The mounting surface for the filter should be a clean conductive area. An anodized surface on an aluminum chassis has poor conductivity; an iridite surface has much better conductivity. If the filter is mounted in a cutout in the chassis, the use of EMI gasket material is recommended to ensure a good bond. The mounting bond must not deteriorate with time; therefore, bare ferrous surfaces should not be used.

The leads on the output of the filter should not be routed through a region containing interference fields. If this is unavoidable, then the leads should be shielded.

5.8 FILTERS FOR RECEIVERS

Filtering may be required to reduce both reception and transmission of interference along conductors. The following guidelines are recommended:

(1) Tunable rejection filters may be useful at the RF inputs to reject off-channel interference.

(2) RF bandpass filters should be used where necessary to suppress spurious radiation (Reference 37).

(3) A low-pass filter between the output of the RF amplifier and the input of the mixer will reduce spurious reponses. A low-pass filter is used before the RF amplifier in field intensity receivers to reduce spurious response above the tuning range of the receiver.

(4) Low-pass filtering should be used on all RF filament leads.

(5) A harmonic rejection filter used between the output of the local oscillator and the mixer will reduce spurious responses.

(6) A tunable, active-notch rejection filter located at the input to the first IF amplifier is useful in controlling adjacent channel interference.

(7) An active audio-pulse cancellation filter may be useful in the audio stage to eliminate a single frequency, such as a heterodyne beat.

(8) Low-pass power line filters on the power input lines are recommended to reduce susceptibility to conducted and local radiated interference (Reference 38).

5.9 FILTERS FOR TRANSMITTERS

Filters for the transmitter are usually designed to reduce the output of RF energy at any frequency except the operating frequency. The following guidelines are recommended:

(1) The network coupling the final amplifier to the antenna terminals should be designed so that it incorporates the characteristics of a low-pass filter and suppresses the harmonics of the operating frequency (Reference 39).

(2) Interstage coupling should make use of filter techniques to reduce harmonic feedthrough.

(3) Low-pass filters should be used where signal and control lines penetrate the transmitter cabinet.

(4) Power lines should be filtered to prevent rectifier noise from being coupled from the transmitter into the power lines.

(5) Bypass filters should be used in the filament and plate power supply circuits to prevent harmonics from coupling into critical circuits.

(6) If the frequency source is a frequency synthesizer instead of an oscillator, coupling to it should be through a band-pass filter instead of a low-pass filter.

(7) In the frequency range above 100 MHz, transmission-line-type filters, resonant cavities, and waveguides are more practical than lumped parameter filters.

5.10 FILTERS FOR OTHER EQUIPMENT

Servo systems, computers, command encoders, and time standards can be grouped together in the discussion of the applications of filters. Attention should be given to the following points:

(1) Power-line filters will reduce the probability of malfunctions caused by transients on the primary power line. They will also prevent rectifier noise from feeding back into the power lines.

(2) Input, signal, and control lines should be filtered to prevent the entrance of harmful interference.

(3) The output circuit should incorporate a low-pass or band-pass filter to suppress harmonics and spurious frequencies.

Commutator-type motors and generators are sources of serious interference. The switching transients resulting from commutation must be dealt with through the use of a number of design techniques. Modern motors usually incorporate shielded housings. Recommended interference suppression guidelines for rotating equipment are as follows:

(1) Each brush should be bypassed to the frame with a 0.05- to 0.25-μF capacitor. The voltage rating of the capacitor should be at least twice the operating voltage of the motor or generator. The capacitor leads should be kept short. If necessary, a feedthrough capacitor can be used in some applications.

(2) The armature terminal of a generator should be bypassed with a 0.05- to 0.25-μF capacitor whose voltage rating is at least double the output voltage rating of the machine. The leads must be kept short. A feedthrough capacitor offers improved high-frequency filtering.

(3) Portable hand tools, using commutators, should be filtered at the point at which the power line enters the case. Typically, the load current is 1 to 5 A and small π- or L-filters can be used. The case is grounded through the third wire in the power cable. This offers a common reference point for a balanced-type filter.

(4) The alternator should have capacitors installed at the brushes of slip rings and the exciter.

Filtering of teletype equipment is essential to reduce conducted interference. The following guidelines are recommended:

(1) Low-pass filtering should be employed at the output of the transmitter-distributor contacts.

(2) Low-pass filtering should be used at the input and output of signal lines.

(3) Low-pass power-line filters should be used at the power input to the equipment.

If a vending machine cabinet is an effective shield, the use of power-line filters may be sufficient to eliminate interference from this source. Generally, the cabinet shielding is poor, and it will be necessary to install filters as outlined below:

(1) All fluorescent lamps used on the machine should be filtered.

(2) All switching contacts associated with the coin switches and merchandise release solenoids should be filtered or bypassed with capacitors.

(3) All motor leads at the motor junction box should be filtered.

Filtering of motor vehicles can be accomplished by using the following guidelines:

(1) The battery terminal of the coil should be bypassed to ground by means of a low-inductance, 0.1-μF filter capacitor.

(2) The battery terminal of the generator should be bypassed to ground by means of a low-inductance, 0.1-μF filter capacitor.

(3) A low-pass filter should be inserted at the output of the regulator.

(4) The battery terminal of the regulator should be bypassed to ground by means of a low-inductance, 0.1-μF filter capacitor. The field terminal should not be bypassed.

(5) All gauges and sender units using make-and-break contacts should be bypassed with capacitors.

The voltage rating of capacitors should be consistent with voltages used in the system, with some allowance being made for transients. For automotive systems, this factor of safety is usually 4 and sometimes as high as 10.

6. WIRING AND CABLING

Methods for controlling EMI must include measures for reducing interference either caused by or aggravated by wiring and cabling. Wiring within an individual unit can cause problems by providing a coupling between EMI-producing components and nearby susceptible components or circuits. Cabling between units can cause coupling between leads within the cable and between cables. Wiring and cabling both provide entry points for harmful electromagnetic energy existing in the local environment. The design engineer should be cognizant of these various pitfalls, and utilize techniques for avoiding or minimizing them.

6.1 WIRING

The term "wiring" is defined as the interconnections among elements of one chassis or unit. Wiring is distinguished from cabling which is taken to mean conductors or groups of conductors connecting different units of a system. Wiring may include power leads, signal leads, printed circuit conductors, and control leads. Coupling between wires is more readily discussed by making a distinction between low-impedance (magnetic) coupling and high-impedance (electric) coupling (Reference 40).

1. Low-Impedance Coupling

Coupling between wires takes place by means of magnetic and electric fields. Low-impedance coupling is best visualized by thinking in terms of transmitting loops and receiving loops. Therefore, low-impedance EMI coupling is combated by decreasing the mutual inductance between these loops. It is assumed at the outset that everything possible has been done to reduce the magnitude of the interference current flowing in the transmitting loop.

The following methods should be considered as a means to reduce low-impedance coupling:

(1) The area enclosed by the transmitting loop may be reduced.

(2) The area enclosed by the receiving loop may be reduced.

(3) The transmitting and receiving loops may be oriented in mutually perpendicular planes.

(4) One loop may be broken and rejoined with a symmetrical cross connection to achieve cancellation of induced voltage.

(5) The impedance of the loads in either loop may be increased.

(6) The separation between loops may be increased.

(7) Shielding material may be placed between the transmitting and receiving loops.

Methods 1 and 2 are very effective solutions. If a current-carrying wire and its return wire are placed directly together, the effective area of the loop is made very small. However, this is not always easy to accomplish. In particular, it should be noted that use of a structure (chassis) as a return path may lead to large loop areas. Such loop areas may be minimized by running the hot wire directly against the structure and using the shortest possible length of loop.

Method 3 is often impractical since its effectiveness is critically dependent on perpendicular alignment.

Method 4 is a powerful technique for limiting interference coupling. The transposition of a wire and its return circuit is conventional practice in the construction of telephone lines. When carried over into electronic circuit wiring, it becomes the familiar twisted-pair wiring method.

Method 5 should be considered early in the design stage. If low-impedance sources are unavoidable, then efforts should be made to increase the impedance of nearby sensitive circuits.

Method 6 is probably the most obvious means of reducing coupling. Separating wires that carry interfering currents from those connected to susceptible circuits should be considered standard practice. A helpful factor is that induced voltage decreases in an exponential manner with increasing wire separation.

Method 7 is more difficult to implement than is at first apparent. There are two primary mechanisms by which shielding reduces coupling. A shield of high conductivity functions by developing a counter flux due to eddy currents excited within the shield. A shield of high permeability functions by providing a flux path of low reluctance for the interference field. The first shielding mechanism for reducing coupling becomes impractical at low frequencies because very thick materials are required for it to be effective. The usual shields on shielded wiring provide no significant protection against magnetic coupling at low frequencies. The second mechanism calls for the use of special materials such as Hypernik and Mumetal. To be most effective, a shield of high permeability must provide a closed flux path. Breaks in this path may themselves become severe sources of magnetic fields.

2. High-Impedance Coupling

Electric fields of relatively high value are produced in circuits that have high voltages and low currents. High-impedance coupling can be thought of as resulting from stray capacitance between leads. In general, high-impedance coupling is somewhat easier to eliminate than low-impedance coupling.

The following methods should be considered as means to reduce high-impedance coupling:

(1) Filters may be used at the source or at the susceptible circuit to reduce interference voltages.

(2) Coupling capacitance may be reduced by increasing spacing between wires.

(3) Shielded wires should be used.

(4) The input impedance of the susceptible circuit should be reduced.

(5) Balanced lines and balanced circuits should be used.

Method 1 offers a common solution; it is generally preferable to filter the source side if possible.

Method 2 is accomplished by separating wires known to be sources of interference and those recognized as pickups for susceptible circuits. Coupling can be reduced by perpendicular wire crossing.

Method 3 is very effective. The combination of filtering and shielding provides a means of reducing virtually all electric field coupling. This fortunate circumstance comes about because shielding effectiveness increases for electric fields at low frequencies where filtering becomes impractical.

Method 4 is recommended for consideration early in the design stage. However, the procedure recommended in method 4 conflicts with measures suggested for suppressing magnetic coupling. A compromise is usually made by designing for circuit impedances in the range of from 150 to 600 Ω.

Method 5 is effective for sensitive equipment when both wires are maintained at the same impedance to ground. Grounding, if required, is done at the center point of the source and load. The result is that coupled interference voltages appear equally on both leads and cancel at the input to the sensitive device. Wire shielding of the "twin-ax" type may be used to reduce the magnitude of the coupled interference. (The "twin-ax" cable is co-axial cable with two inner conductors somewhat off center.)

6.2 CABLING

In a general way, all of the interference reduction techniques outlined in the preceding section on wiring also apply to cabling. This section deals with problems more specifically associated with cables (References 41, 42, and 43).

1. Electric Coupling

Electric coupling between lines within a multiconductor cable can be severe in long cable runs. This is best minimized by isolating interference source leads in a cable separate from cables containing susceptible circuit leads. If this is not feasible, all interference source leads within a cable should be shielded. A further reduction in coupling can be achieved by separate shielding of susceptible circuits.

2. High-Frequency Coupling

In cables, one frequently faces the situation in which lead length is very long compared to wavelength. The distinction between high- and low-impedance leads then becomes less useful because standing waves may exist on wiring at high frequencies. A wire that exhibits a very high impedance or even an open circuit at one point becomes a very low impedance and is susceptible to magnetic fields at a quarter-wave distant. In this situation, it is necessary to treat the cable as a transmission line.

3. General Coupling Equations

The amount of coupling between wires in an actual installation is affected by stray parameters that do not yield to a simple mathematical analysis; however, an estimate of the coupling to be expected is of value to the designer and can be calculated by making certain simplifying approximations.

Figure 2.38 shows a circuit illustrating the basic coupling mechanism between two systems of parallel wires. The ratio of the noise voltage coupled into the input side of the susceptible circuit of the noise voltage (Reference 40) is given by

$$\frac{E_{2G}}{E_0} = \overbrace{\left(\frac{R_1}{R_1 + R_0} \times \frac{R_2}{X_C}\right)}^{\text{electric comp.}} + \overbrace{\left(\frac{X_M}{R_1 + R_0} \times \frac{R_{2G}}{R_{2G} + R_{2L}}\right)}^{\text{magnetic comp.}} = K_G f . \quad (2.25)$$

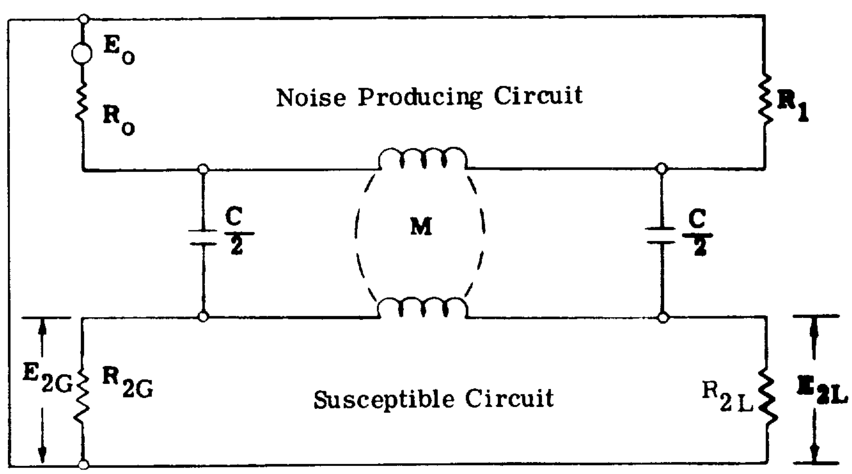


Figure 2.38.—Circuit representing electric and magnetic coupling between parallel wires.

The fraction of noise voltage appearing at the load side of the susceptible circuit is

$$\frac{E_{2L}}{E_0}=\overbrace{\left(\frac{R_1}{R_1+R_0}\times\frac{R_2}{X_C}\right)}^{\text{electric comp.}}-\overbrace{\left(\frac{X_M}{R_1+R_0}\times\frac{R_{2L}}{R_{2G}+R_{2L}}\right)}^{\text{magnetic comp.}}=K_L f\,, \quad (2.26)$$

where

E_0 = noise voltage in the interfering circuit,

E_{2G} = noise voltage coupled into the generator side of the susceptible circuit,

X_M = reactance component of the inductive coupling,

X_C = reactance component of the capacitive coupling,

E_{2L} = noise voltage coupled into the load side of the susceptible circuit,

K_G = coupling coefficient, generator side,

K_L = coupling coefficient, load side,

R_2 = $R_{2L}R_{2G}/(R_{2L}+R_{2G})$,

and

f = frequency of interfering signal.

In each of these equations, the first term in brackets represents the electric component and the second term represents the magnetic component of field coupling. Figure 2.38 identifies the resistive and reactive elements of the circuit. Note that the voltage coupled to the input end of the susceptible circuit, E_{2G}, is shown as the sum of the electric and magnetic components. At the opposite end of the suceptible circuit, the coupled voltage E_{2L} is shown as the difference between the electric and magnetic components. For the approximation made here, the two components of E_{2L} will indeed be in phase opposition. Rather than having to determine a difference voltage, the larger of the two components is assumed to be the predominant interference. It is common practice to assume that voltage transfers are proportional to frequency until unity transfer is reached and that the transfer is unity for all higher frequencies. Figure 2.39 shows the assumed voltage transfer ratio plotted as two dotted straight lines; one rises at a rate of 6 dB per octave, and the other has unit value at higher frequencies. Typical experimental data are

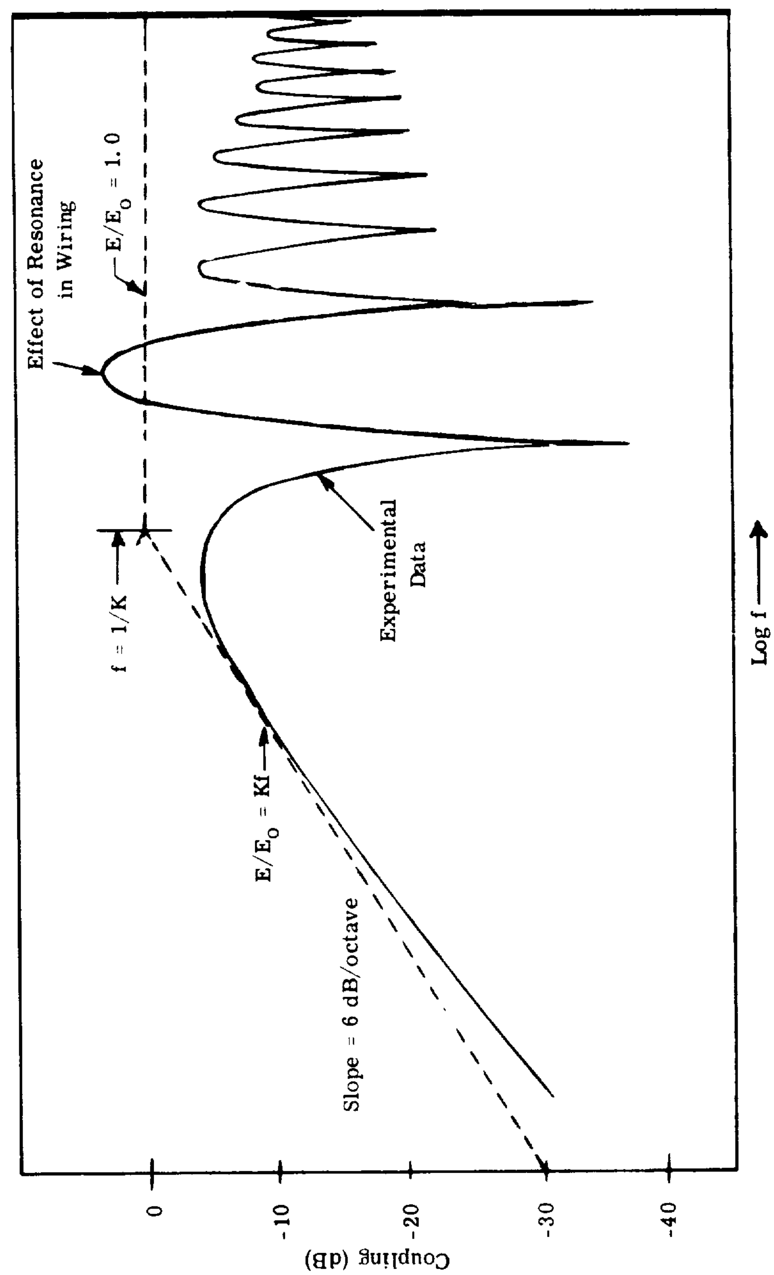


Figure 2.39.–Approximate degree of coupling produced by wiring in a signal circuit.

also plotted. The assumed curves show somewhat more coupled-in interference than is shown by the experimental data. Usually, this situation will be experienced in cases of light circuit loading. Resonances in the wiring can produce an inversion so that experimental data may rise above the approximation in the narrow frequency band of the resonance. For a considerable variation in loading above or below 300 Ω, $K_G = 2K_L$ is a very good approximation.

4. Graphical Calculation of Coupling

For certain simple situations it is possible to develop a graphical presentation of coupling effects. Such graphs are especially helpful in establishing cable and wire routing during chassis layout. Figure 2.40 shows the induced voltage E_i in a length of wire as the result of interference current I flowing in an adjacent parallel wire; the spacing D between wire centers is 1/2 in., and the wire height d above the ground plane is 2 in.

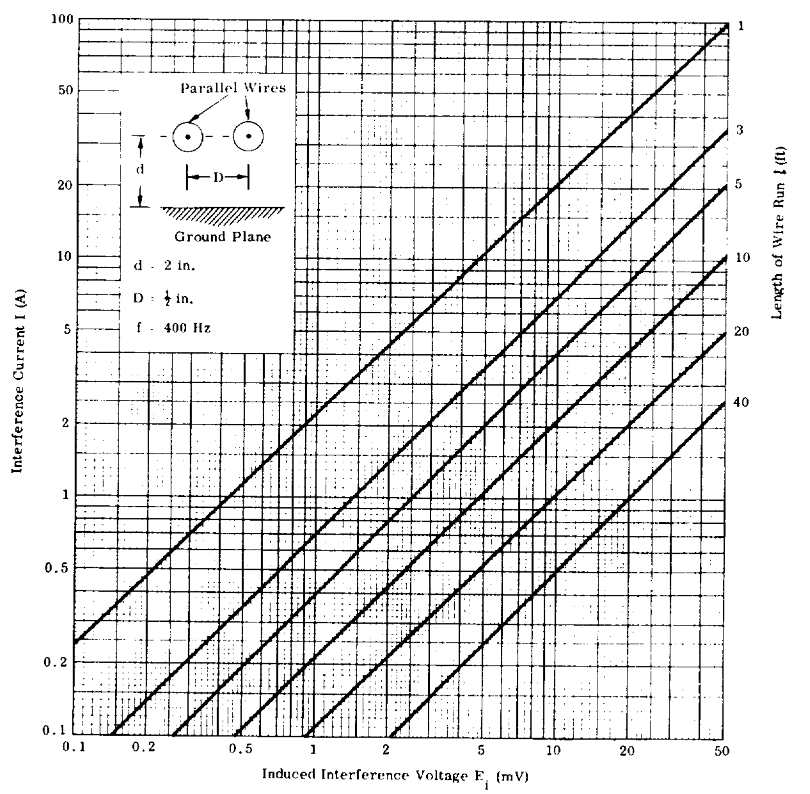


Figure 2.40—Voltage induced in a wire as a function of interference current I in a parallel wire. (After EMC Bulletin No. 8, E.I.A., March 1965)

The graphical data presented in Figure 2.40 may be generalized to yield the following empirical approximation, which gives the induced voltage E_i in millivolts:

$$E_i \approx 3.12 \times 10^{-4}\, lIfd/D, \tag{2.27}$$

where

l = length of wire run in feet,
I = interference current flowing in one wire, in amperes,
f = interference frequency in hertz, from 60 Hz to 10 kHz,
d = elevation height of wires above the ground plane, in inches, from 1 to 5 in.

and

D = distance between wire centers in inches, from 0.5 to 4 in.

6.3 GUIDELINES FOR WIRING AND CABLING OF EQUIPMENT

Wiring and cable routing should be planned in order to separate and isolate sensitive (susceptibility prone) wires from possible interference carrying conductors. The preceding guidelines are applicable to both receivers and transmitters.

The guidelines given in the previous sections also apply to support equipment. In addition, attention should be given to the following items (References 44 and 45):

(1) Separation of the various circuits into compatible groups is extremely important where diverse functions are incorporated into a single package (Reference 42).

(2) Individual separate ground returns are highly desirable to prevent common mode coupling.

The use of shielded wire is recommended in vending machines. As an alternative, metallic conduit or cable troughs may be used.

Vehicles in which interference is to be completely suppressed should have the ignition systems completely wired with shielded high-tension wires. Double-braided shield wire with connector fittings is commercially available for this purpose. In addition, the primary wiring between the generator and the regulator should be shielded.

7. CALCULATION AIDS

Some of the most frequently required mathematical conversion parameters are presented here as Figures 2.41 through 2.48. As users of these guidelines become familiar with the use of a given set of instrumentation, they will discover that many of the tabulated instrument calibrations, antenna factors, cable loss curves, and so forth can be combined in convenient chart form. If such charts are developed in permanent form, they will save much time and tedious labor.

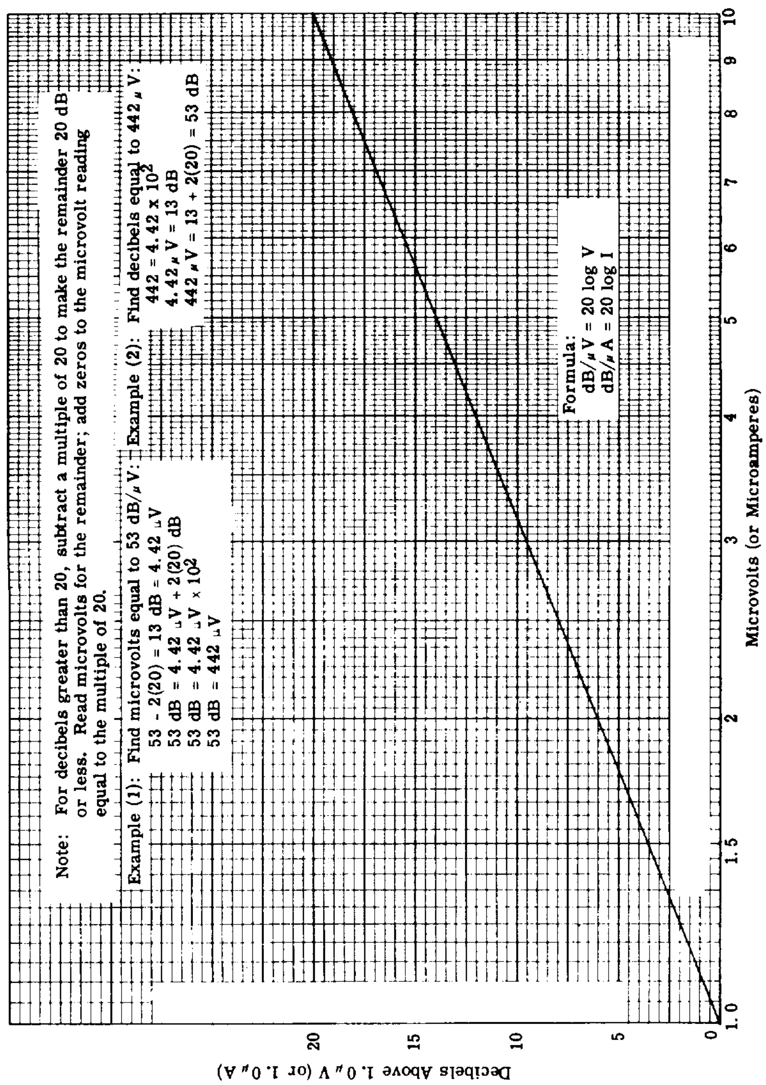


Figure 2.41.--Graph for converting dB above 1 μV (or 1 μA) to microvolts (or microamperes).

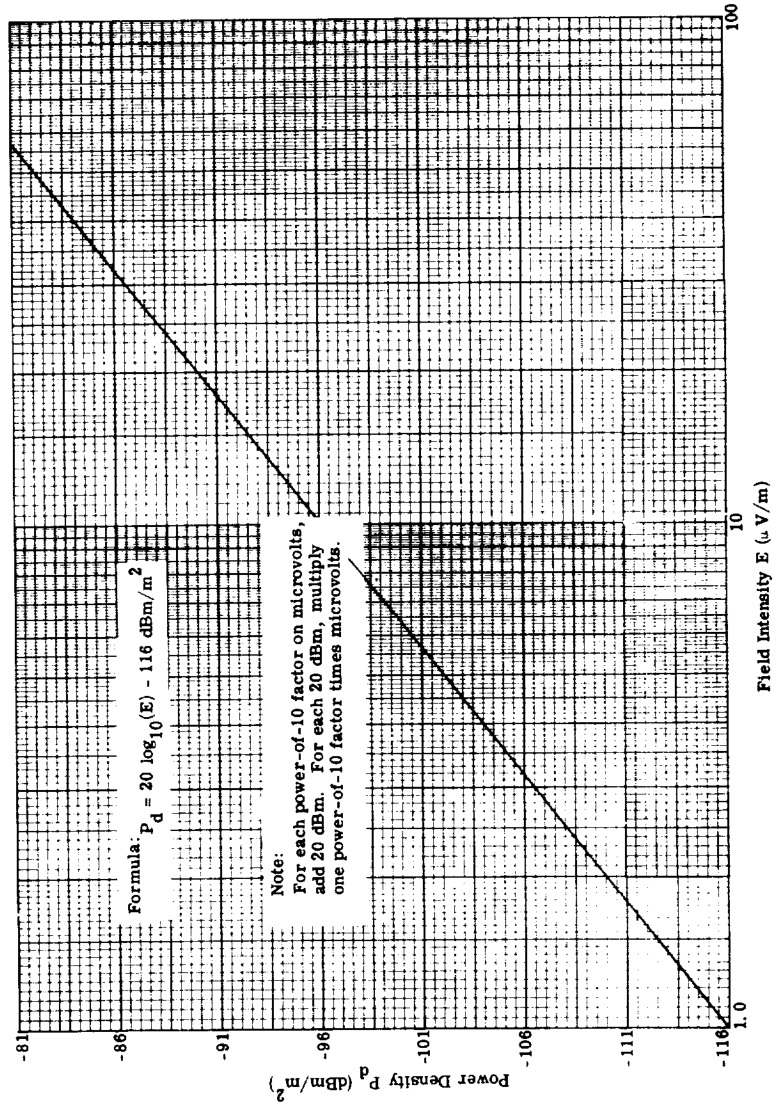


Figure 2.42.–Graph for converting field intensity in microvolts per meter to power density in dB above 1 mW/m^2 for an antenna matched in free space.

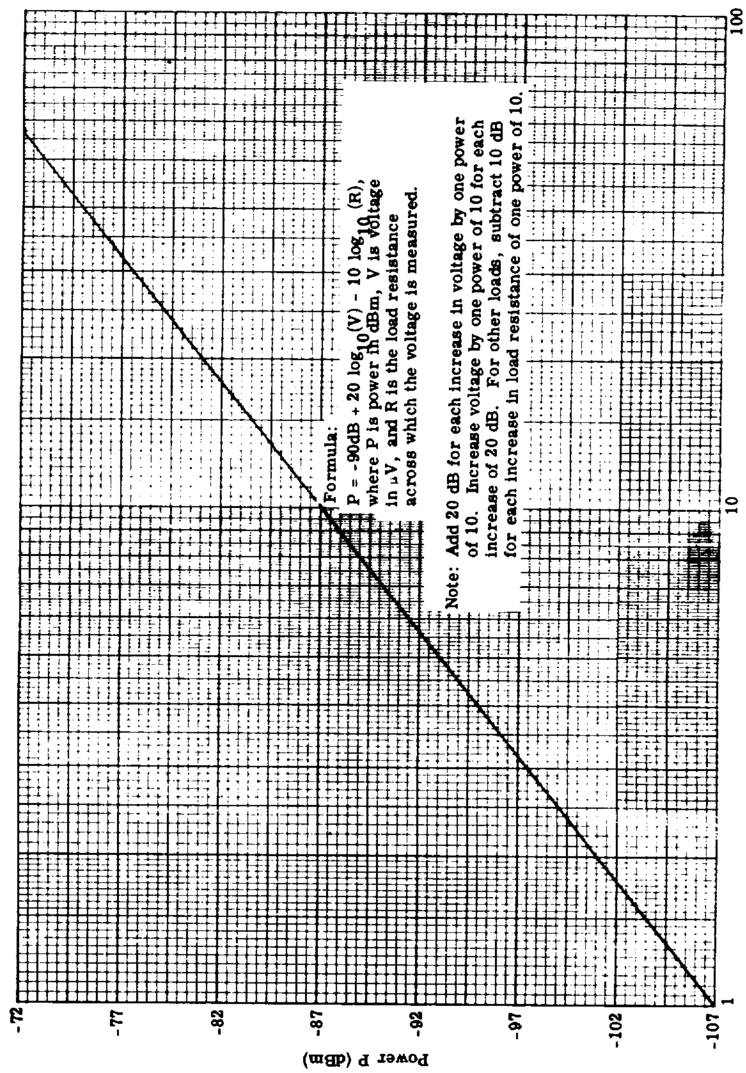


Figure 2.43.–Graph for converting voltage in microvolts to power in dB referred to 1 mW for a load impedance of 50 Ω.

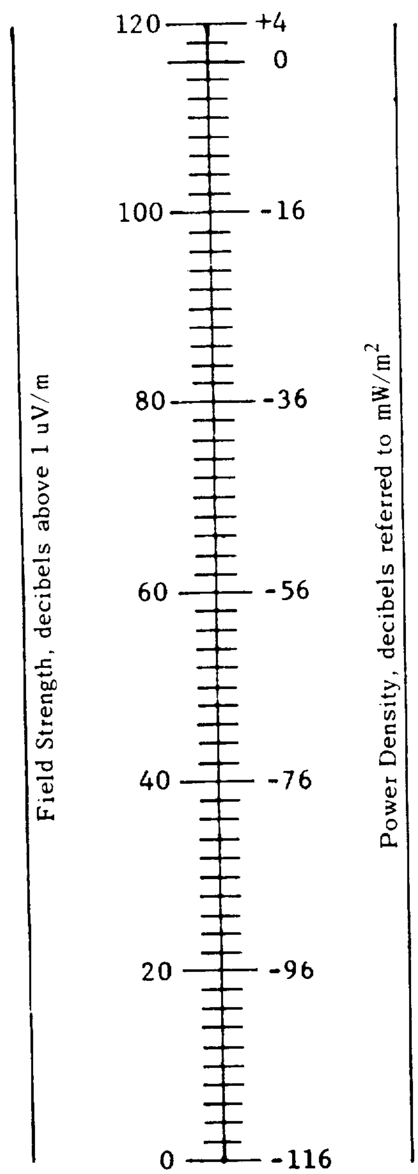


Figure 2.44.–Scales for converting dB above 1 μV/m to dB referred to 1 mW/m^2.

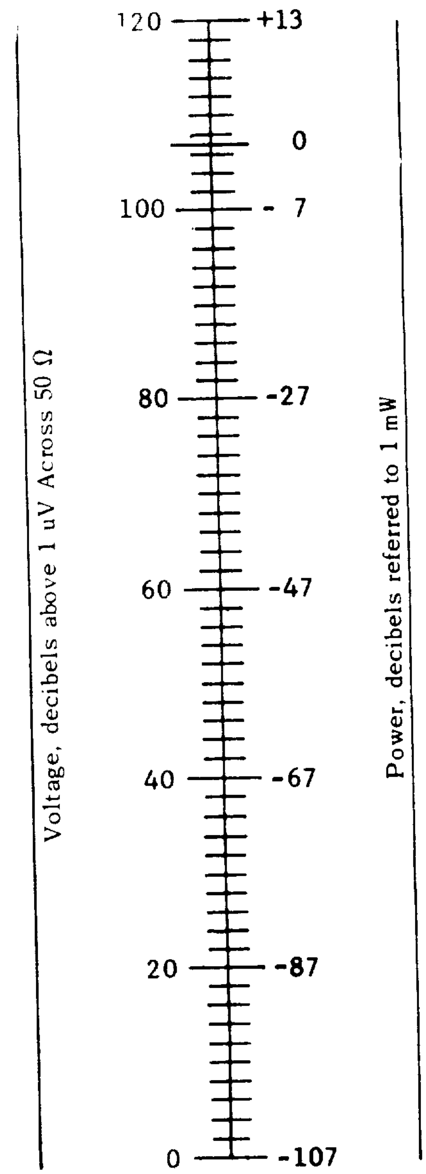


Figure 2.45.–Scales for converting dB above 1 μV to dB referred to 1 mV for a 50-Ω load.

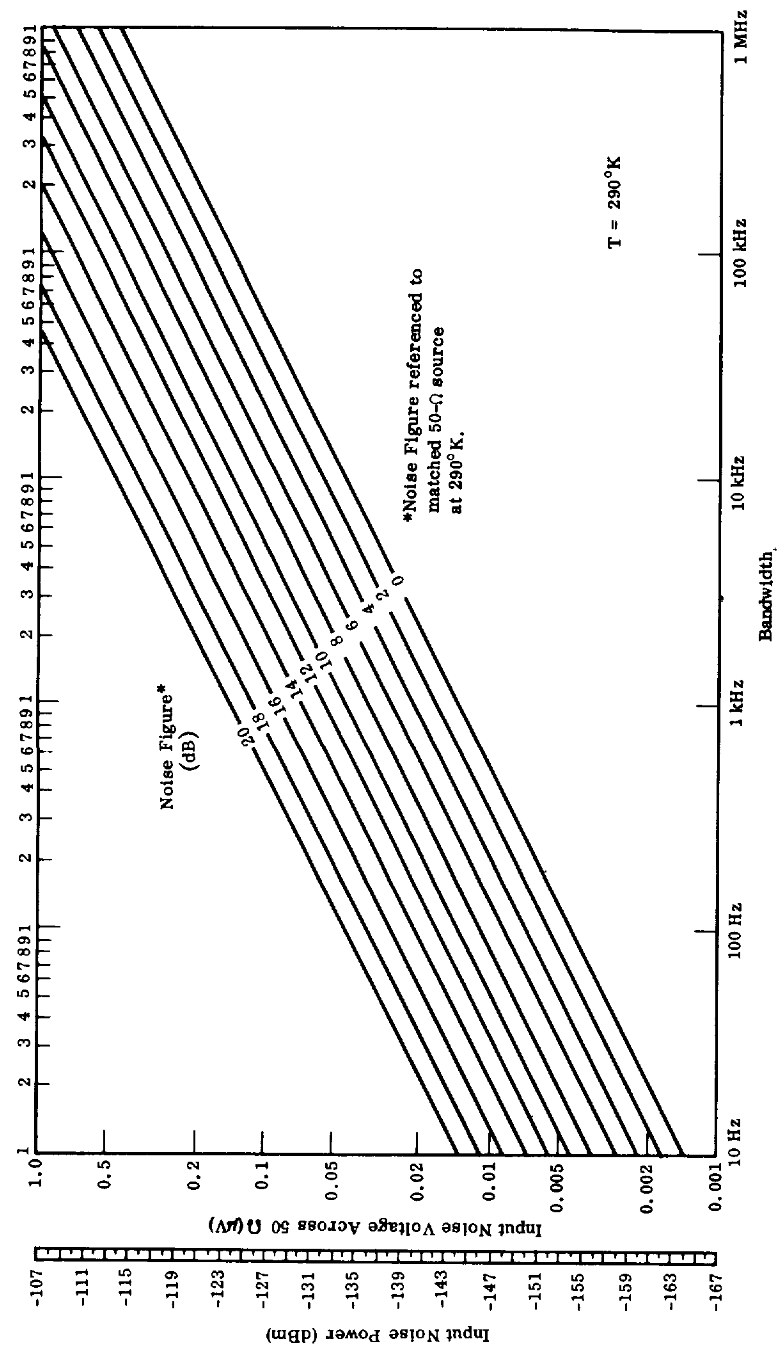


Figure 2.46.–Noise figure related to input noise power, noise voltage, and bandwidth.

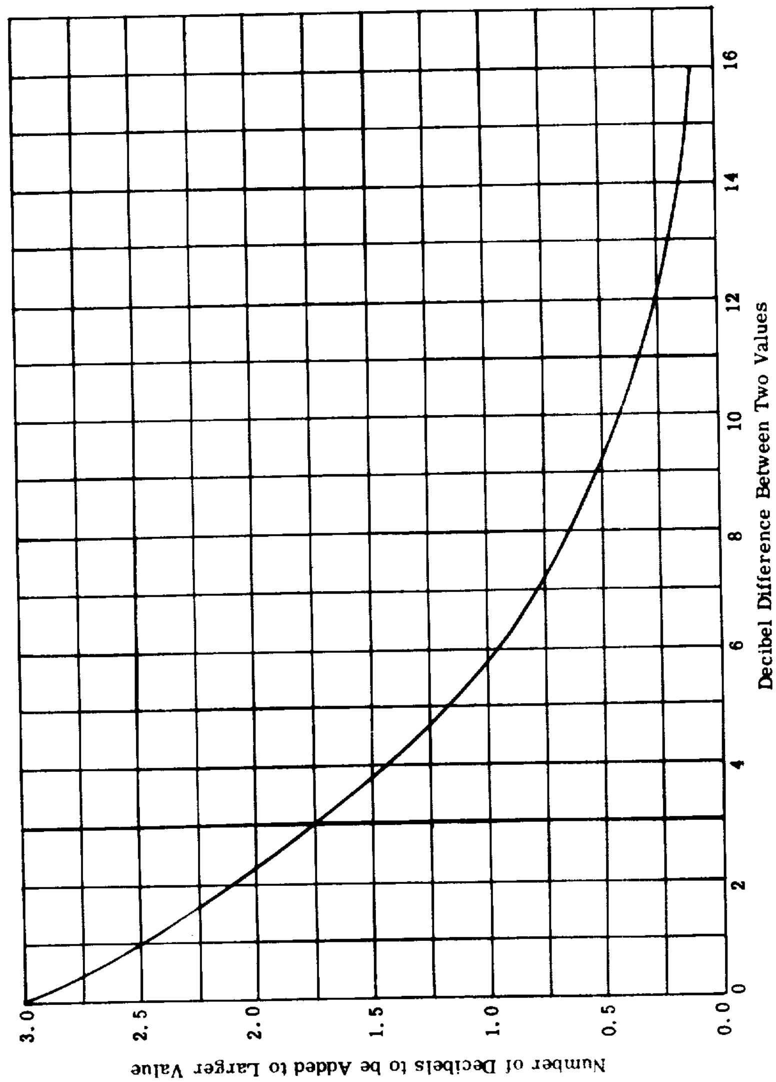


Figure 2.47.–Graph for adding noise or signals expressed in dB.

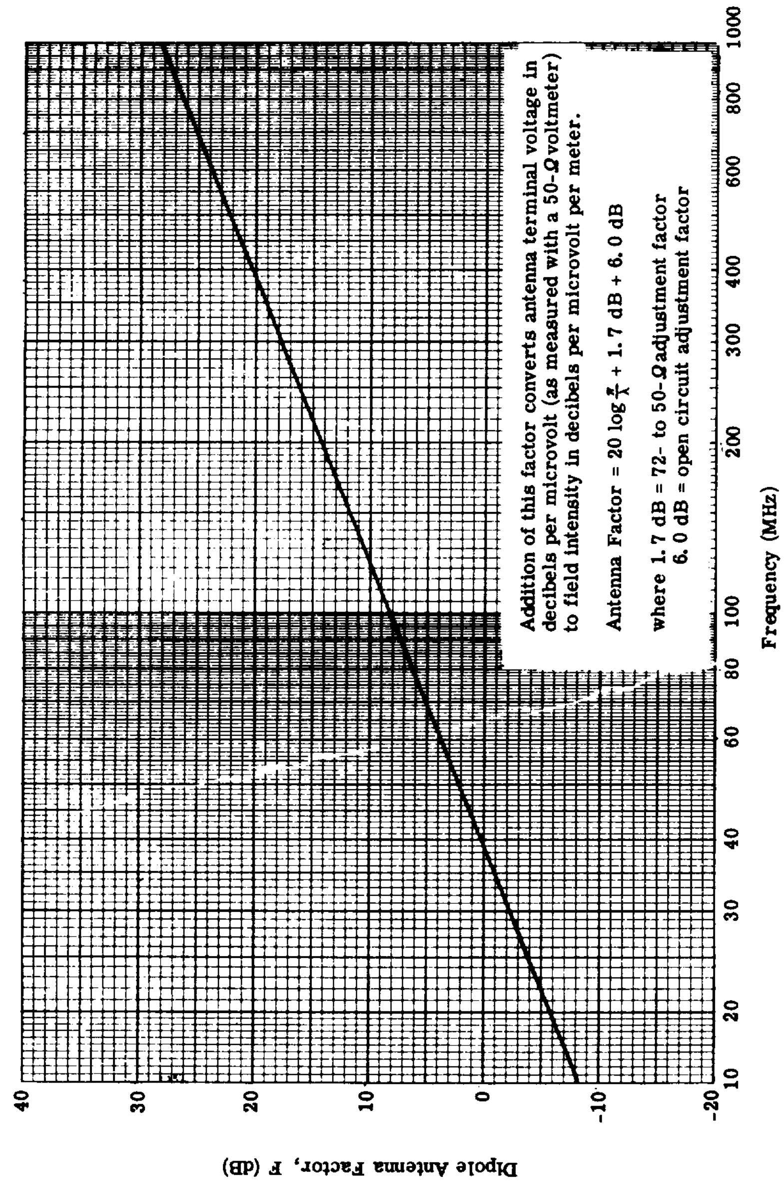


Figure 2.48.–Graph for converting the terminal voltage of a tuned dipole antenna to field intensity.

8. DEFINITIONS AND SYMBOLS

The following definitions are used or referenced in this section. (Reference 46). Following the definitions is a listing of symbols for physical quantities with their multiplying symbols. A standard method of decibel notation is explained.

(1) Ambient electromagnetic environment: The level of electromagnetic emission (conducted or radiated) indicated by a calibrated interference-measuring set with the equipment under test inoperative.

(2) Antenna factor: A multiplying factor applied to the voltage at the input terminals of the measuring instrument to yield electric field strength in V/m and magnetic field strength in A/m, for a given antenna and frequency.

(3) Antenna conducted emission, transmit condition: The undesired portion of signal spectrum appearing at the antenna terminal of a transmitter under full load transmit conditions.

(4) Antenna conducted emission: The undesired portion of signal spectrum present at the antenna terminal due to an operating receiver or a transmitter in standby condition.

(5) Broadband emission: Any signal emission having a spectral power distribution such that the impulse bandwidth of the measuring instruments is more than 1.2 times the repetition frequency of the signal when the impulse bandwidth of the receiver is also less than $1/t$, where t is the pulse width of the signal.

(For practical use in determining narrowband or broadband classification of emissions, detune the instrument by one impulse bandwidth or by one-half of the smallest frequency division on the dial, whichever is larger. If the peak reading is reduced by at least 3 dB, classify as narrowband. If the peak reading does not decrease at least 3 dB, classify as broadband. Note that this criterion makes classification as narrowband or broadband a function of the bandwidth of the measuring instrument.)

(6) Conducted EME: Electromagnetic emissions propagated along a power or signal conductor and measured by direct conductive or magnetic coupling to an appropriate probe.

(7) Component (equivalent to "equipment," as used in this handbook): The smallest recognized unit (black box) that generally is used in conjunction with other components to perform a particular function.

(8) Counterpoise: The ground (or reference) plane for an unbalanced antenna.

(9) Coupling coefficient (coupling factor used in the case of resistive, capacitive, self-inductive, and inductive coupling): The ratio of the impedance of the coupling to the square root of the product of the impedances of similar elements in the two circuit meshes. Unless otherwise specified, coefficient of coupling refers to inductive coupling, in which case it equals

$M/\sqrt{L_1 L_2}$, where M is the mutual inductance, L_1 is the total inductance of one mesh, and L_2 is the total inductance of the other mesh (all values in henries).

(10) Cross-coupling: The coupling of a signal from one circuit to another, where it becomes an undesired signal.

(11) Cross-modulation: Modulation of a desired signal by an undesired signal.

(12) Crosstalk: An undesired signal introduced by cross-coupling.

(13) Desensitization: The reduction in sensitivity of a receiver due to the presence of an undesired signal.

(14) Electrical equipment: Equipment that generates or utilizes electrical energy, but does not intentionally generate electromagnetic energy of a type that may be a potential source of interference. (Examples are electric motors, office machines, fluorescent lamps, and solenoids.)

(15) Electromagnetic emission (EME): Electromagnetic energy (desired or undesired) propagated from a source by radiation or conduction.

(16) Electromagnetic interference (EMI): A condition in which an electromagnetic emission produces an undesired response in a specific susceptible component or subsystem.

(17) Electromagnetic susceptibility (EMS): The tendency for an undesired response to be produced in a component by an electromagnetic emission.

(18) Ground plane: A metal sheet or plate used as a common reference point for circuit returns and electrical signal potentials.

(19) Impulse bandwidth (IBW): The ratio of the unmodulated sine wave in μV rms to the impulse amplitude in V/MHz required to produce an equal response in the circuit.

(20) Impulse interference: Interference causing nonoverlapping transient disturbances in a receiver.

(21) Intermodulation: Mixing of two or more signals in a nonlinear element to produce signals at frequencies equal to the sums and differences of integral multiples of the original signals.

(22) Malfunction: A change in the normal performance output that effectively prohibits the proper operation of a component.

(23) Microvolts per megahertz (μV/MHz): The broadband emission intensity in μV rms of an unmodulated sine wave applied to the input of the measuring circuit at its center frequency (which will result in a peak response in the circuit equal to that resulting from the broadband pulse being measured) divided by the effective impulse bandwidth in the MHz of the circuit.

(24) Narrowband emission: An emission that has its principle spectral energy lying at a single frequency within the passband of the measuring receiver in use. (See defintion No. 5.)

(25) Radiated EME: Electromagnetic fields in space, propagated by either induction or radiation.

(26) Receiver front-end rejection: The ratio of the power required at the receiver antenna terminal to produce a standard response or change in response to the power required to produce the standard response at the center frequency.

(27) Receiver spurious response: Any response of a receiver to a signal outside its intended reception bandwidth.

(28) Spurious emission: Any electromagnetic emission outside of the intended emission bandwidth.

(29) Standard response: The programmed or desired response produced in a test item by a specified input signal. [The standard response can be used to measure any deviation from normal performance which occurs during susceptibility testing (e.g., ratio of signal plus noise to noise on a receiver for a specified input signal).]

(30) Substitution measurements: A method of measuring signals which depends on the use of a calibrated signal generator with an output similar to that of an unknown signal. (The calibrated signal is substituted for the unknown signal to reproduce the instrument response. The error is then limited to the error of the calibrated signal source.)

(31) Test antenna: An antenna with specified characteristics designated for use under specified conditions in conducting tests of electromagnetic interference.

(32) Test item: Any separate and distinct component or subsystem which, if procured separately, would normally be subject to applicable tests as required by the specification. This term is used for brevity to refer to the item under test in this handbook, regardless of whether it is a component or a subsystem.

The following symbols shall be used to indicate the applicable physical quantity:

Physical Quantity	SI Unit	Symbol
Length	meter	m
Electric current	ampere	A
Electric field strength	volts per meter	V/m
Magnetic flux density	tesla	T
Magnetic field strength	amperes per meter	A/m
Voltage, potential difference, electromotive force	volt	V
Power	watt	W
Frequency	hertz	Hz
Electric resistance	ohm	Ω
Electric capacitance	farad	F
Inductance	henry	H

The following symbols shall be used to indicate the applicable multiplying prefix:

Prefix	Symbol
tera (10^{12})	T
giga (10^{9})	G
mega (10^{6})	M
kilo (10^{3})	k
milli (10^{-3})	m
micro (10^{-6})	μ
nano (10^{-9})	n
pico (10^{-12})	p

Decibel notations referred to one of the units obtained by applying a multiplying prefix to the physical quantity involved shall always be used as shown in the following example:

Quantity to be abbreviated and referenced:

Decibels Referencing Micro Volts Megahertz

Bandwidth factor

Physical quantity

Multiplying prefix

Above or below

Abbreviated as dB

The following abbreviations are interpreted as shown:

dBm*	=	decibels referred to one milliwatt
dBm/m^2	=	decibels referred to one milliwatt per square meter
dBm/m^2/MHz	=	decibels referred to one milliwatt per square meter per megahertz
dBμV	=	decibels referred to one microvolt
dBW	=	decibels referred to one watt

*Note that this is an exception to the rule, the W is understood; dBm is the generally accepted notation for decibels referred to one milliwatt.

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SECTION III

LIGHTNING PROTECTION PRACTICES APPLIED TO FIELD STATION INSTALLATIONS

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1. INTRODUCTION

Lightning, which has both plagued and awed man throughout history, is still effective in disrupting man's activities, including satellite tracking operations at NASA STADAN facilities, particularly those located in lightning-prone areas. This section discusses theoretical considerations of lightning phenomena and defines the magnitude of the magnetic, capacitive, and resistive components of lightning-induced voltages. As an example of the method discussed, a typical tracking station is analyzed from a lightning standpoint. Specific protective methods are discussed, including the implementation of high-surge-current voltage clippers for sensitive electrical and electronic circuits, grounding, counterpoises, and bonding. The safety of operating personnel is also treated as a prime consideration.

2. LIGHTNING STROKES TO STRUCTURES AND EARTH

One of the major factors to consider in determining the probability of lightning damage is the number of lightning strokes to earth in a given area and for a given period of time. Since precise quantitative data do not exist except at a few specifically instrumented structures, a secondary measure—the frequency of thunderstorms—is used.

For many years, Weather Bureau stations have recorded "thunderstorm days"—the number of days per year during which thunder is heard. This index is known as the "keraunic level."

It should be noted that, for several reasons, the information so collected is of limited value. First of all, no distinction is made between cloud-to-cloud discharges and cloud-to-ground strokes. Also, there is no allowance for the duration of a storm. A storm lasting an hour would be counted as heavily as one lasting several hours. Despite these limitations, the keraunic level is broadly useful and can be correlated at least partially with lightning strokes to earth-based objects. The United States Weather Bureau has compiled data on the keraunic level on a statistical basis and has published a map of the United States which shows isokeraunic lines; Figure 3.1 is an isokeraunic map of the United States (Reference 1). As shown, the incidence of lightning varies throughout the country, being lowest at the Canadian border and West Coast and highest over the Gulf Coast of Florida.

The expected frequency of strokes to ground, at a typical station such as that at Rosman, North Carolina, can be estimated by multiplying the keraunic level at the place of interest by a suitable empirical factor determined by analysis of data concerning lightning strokes to power transmission lines (Reference 2). This empirical factor has a value between 0.23 and 0.5 strokes

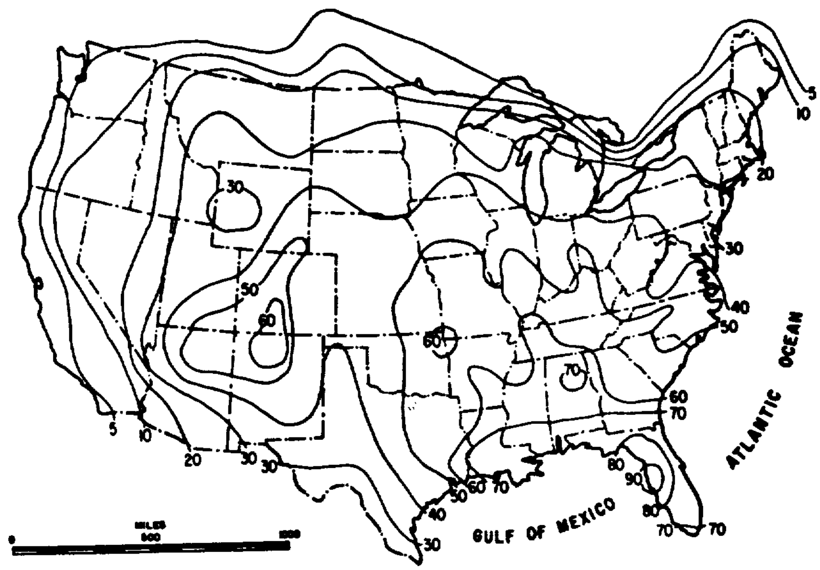


Figure 3.1.–Isokeraunic map of the United States showing the average number of days per year during which thunderstorms occur. (From U. S. Weather Bureau)

to ground per square mile. For example, at the Rosman site, where the keraunic level is 50, the expected stroke frequency would range from 11.5 to 25 (i.e., $50 \times 0.23 = 11.5$ to $50 \times 0.5 = 25$) strokes per square mile per year. For such numbers to be significant, it is necessary to determine the effective lightning-collecting area of the site.

In order to determine the effective lightning-collecting area, it is necessary to encircle the site with an imaginary geographical boundary line. This boundary line, defining the site lightning-collecting area, is located at a fixed distance x from site peripherals such as towers, cables, and so forth. Lightning strokes terminating inside the collecting area are assumed to be potentially damaging, whereas strokes terminating on or outside the boundary line do not affect the site.

Equation (3.2) is helpful in defining the fixed distance as

$$x = \left(\frac{\rho I}{2\pi E}\right)^{1/2}$$

The parameter E is the maximum safe voltage gradient (in volts per meter) that can exist along the surface of the ground and not cause damage to adjacent site structures or cabling. If this gradient is kept low enough, the magnitude of induced voltages and currents will be low; a heuristic value is $E = 10{,}000$ V/m.

Furthermore, if the current for an average lightning stroke is taken to be 40,000 A, and if soil resistivity is considered to be 6000 Ω-m, then $x = 200$ ft for a typical station. For a more conservative figure, the minimum safe distance should be increased to 500 ft.

Such a boundary line is shown drawn around the Rosman site plan (Figure 3.2). The area of the odd shape is computed to be 0.33 square miles. Using this number and the previously arrived at strokes-per-square-mile numbers, a stroke frequency from 4 to 8 strokes per year is calculated. These numbers compare favorably with the number of separate damage-producing storms recorded by site personnel.

Tall structures are frequent targets for direct lightning strokes. At the STADAN sites, the tallest structures are the collimation towers (between 100 and 200 ft high). From statistical data on strokes to tall structures, it can be predicted that, for a keraunic level of 50, a 100-ft-high structure would be struck 0.23 times and a 200-ft-high structure 0.70 times per year; these stroke rates are directly proportional to the keraunic level.

Since these numbers are significantly less than the expected 4 to 8 strokes in the general area of Rosman, the problem of lightning protection is

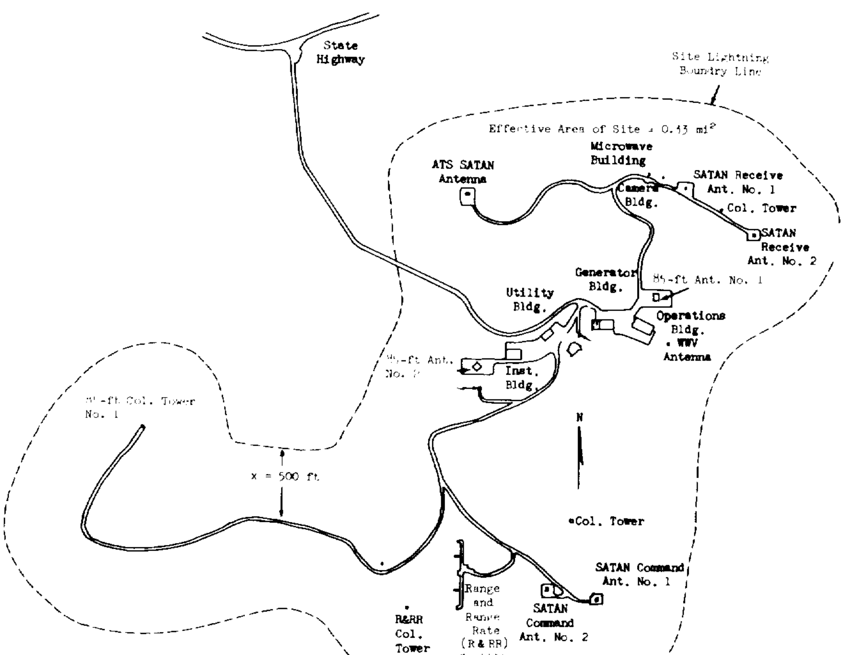


Figure 3.2.–STADAN site layout at Rosman, North Carolina, showing effective ground area of site for purposes of lightning protection.

not primarily one of protecting against strokes to the antenna towers. In this case, design changes and "fixes" to control lightning effects, although including those for the towers, must be directed primarily to deal with random strokes to ground that ultimately affect some part of the electric, electronic, data, or ground systems at the site.

To aid in making calculations of stroke frequency at other world-wide site locations, Table 3.1 (Reference 3) has been prepared.

Table 3.1.–Keraunic levels at world-wide STADAN sites.

Site Location	Keraunic Level
Fairbanks, Alaska	10
Carnarvon, Australia	1
Fort Myers, Florida	91
Greenbelt, Maryland	35
Johannesburg, South Africa	74
Kano, Nigeria	82
Kauai, Hawaii	6
Lima, Peru	15
Goldstone, California	5
Newfoundland	5
Orroral, Australia	9
Rosman, North Carolina	50
Santiago, Chile	3
Tananarive, Madagascar	132
Wallops Island, Virginia	40
Winkfield, England	14

3. THEORETICAL CONSIDERATIONS

3.1 LIGHTNING PHENOMENA

A lightning stroke consists of a rapid discharge of electricity. The current in the stroke generally rises to peak values of from 1000 to 100,000 A in time periods ranging from 1 to 10 μs. Some investigations have indicated that an average current changes at a rate of approximately 15,000 A/μs. The current generally decreases to one-half the initial value within approximately 50 μs. A typical time history of a stroke terminating on a very tall structure is shown in Figure 3.3. (This stroke terminated on the Empire State Building and is representative of many strokes to that building.)

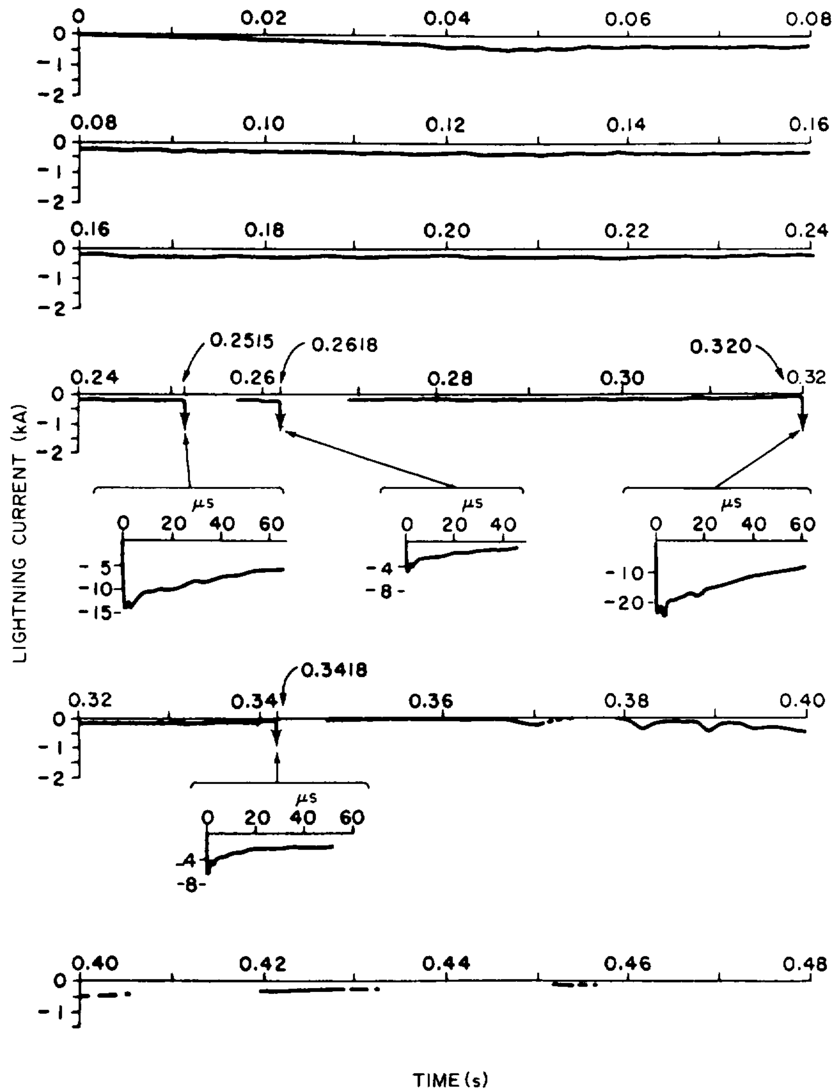


Figure 3.3.–Current peaks and continuing currents of a typical lightning stroke to the Empire State Building. Total duration, 0.47 s; total charge, -84 C. (After K. B. McEachron, Trans. AIEE, 1941)

Analysis of Figure 3.3 shows that the stroke begins in the upper left-hand corner with zero time and zero current (Reference 4). Elapsed time runs from left to right across the figure.

As can be seen, the initial stroke current reaches a negative value of about 250 A in 0.04 s and continues at that value until 0.2515 s has elapsed; then, a step negative current peak of about 15,000 A occurs. After several hundred microseconds, the stroke current falls again to 250 A; at 0.2618 s, a current pulse of 4000 A occurs. The phenomenon of a continuing low-amplitude current interspersed with high-amplitude pulses exists for the duration of the stroke. In this case the stroke continued for about 0.47 s and had four distinct current peaks.

The area under the current-time curve (measured in coulombs) largely determines the burning or thermal power of the stroke. This particular stroke discharged 84 C, of which 60 C were discharged by the continuing current before the first current peak. Most of the charge of the stroke is discharged by the continuous current and not the current peaks. Strokes to ground or low structures are not likely to exhibit the continuing low-amplitude current prior to the first current peak; consequently, these strokes normally have a relatively low charge. The time duration of a multiple-pulsed stroke may be as long as 1.5 s; however, most data indicate that the total duration of the average stroke is about 0.1 s. The time required for a single discharge varies widely, but 100 μs is not an unreasonable average.

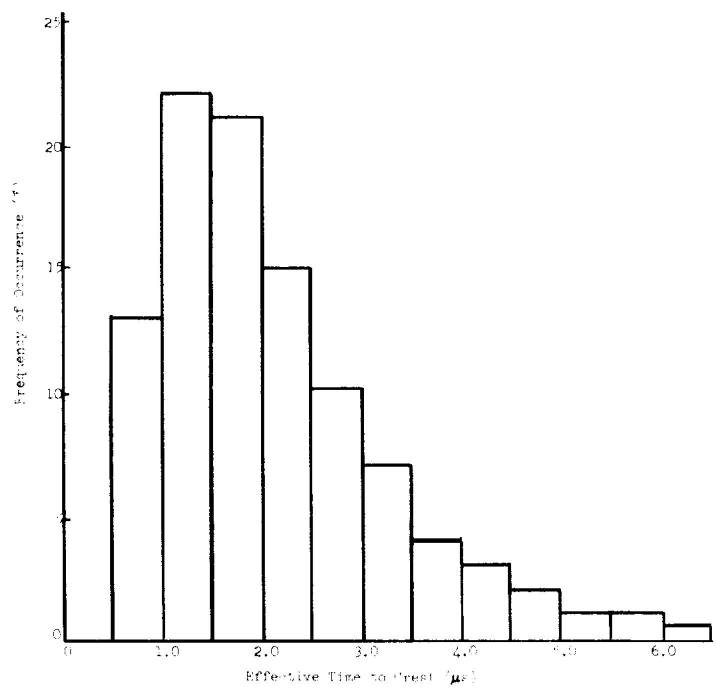


Figure 3.4.–Histogram of time to crest for lightning strokes. (After J. G. Anderson, Trans. AIEE, 1961)

The gamut of measured characteristics of lightning strokes is necessary to determine both damage possibilities and protective measures required.

Figure 3.4 shows a histogram of the average time for a lightning stroke to crest. One to two microseconds is the predominant time. The effective time to crest is an important parameter for determining the voltage induced in electrical circuits. The best of several sources of data (References 5, 6, and 7) is given.

Figure 3.5 shows the distribution of current amplitudes that can be expected from lightning strokes. The median is approximately 40,000 A.

Figure 3.6 (References 4 and 8) shows the effective rate of rise of lightning-stroke current peaks; the average value is 10,000 to 15,000 A/μs. This illustration shows that data of the kind shown in Figures 3.4 and 3.5 cannot be compared directly to yield a composite result. The tendency is for crest time to be delayed as current amplitude increases.

Figure 3.7 gives a measure of the charge in coulombs, a significant factor in determining the thermal burning action of the lightning stroke. As mentioned earlier, thermal action occurs largely in the long-duration portion of the stroke.

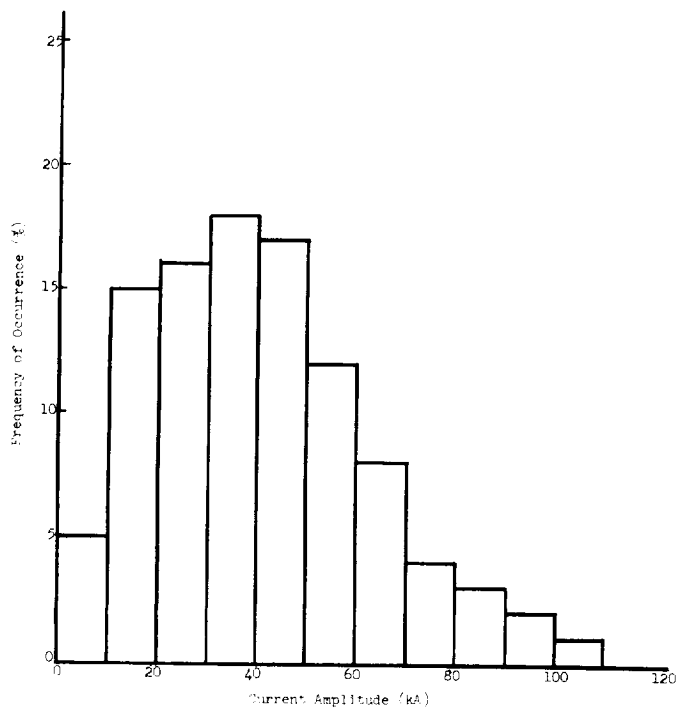


Figure 3.5.–Distribution of current amplitude for a lightning stroke. (After J. G. Anderson, Trans. AIEE, 1961)

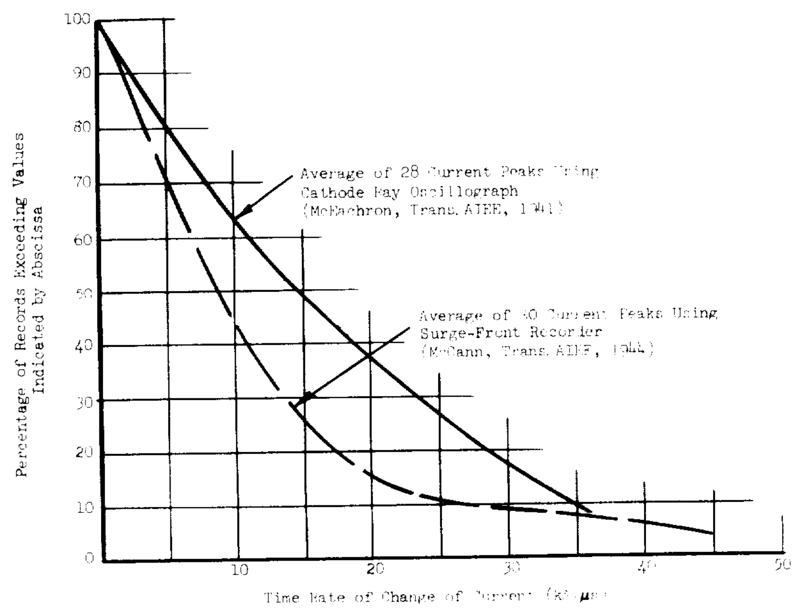


Figure 3.6.–Effective rate of rise of peak current in a lightning stroke.

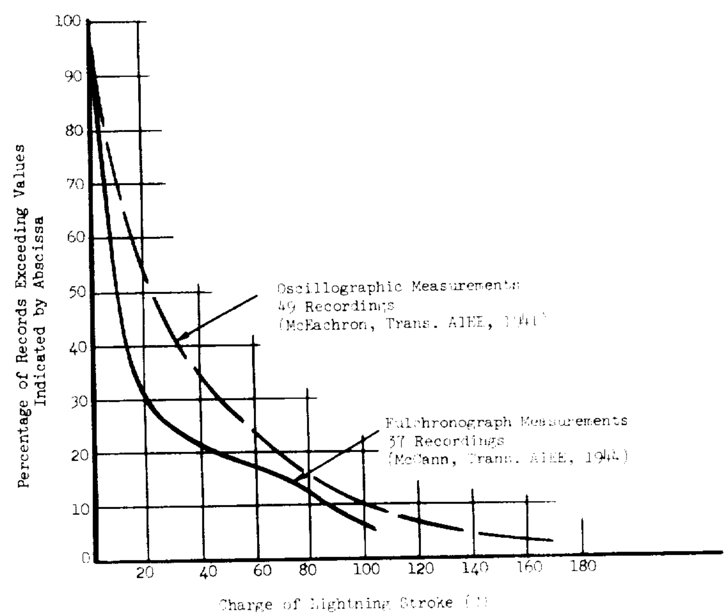


Figure 3.7.–Total charge in lightning strokes, measured at ground terminal point.

The importance of the preceding data can be summarized for practical applications as follows: First, current amplitude I determines the voltage amplitude E_R that will be developed across a resistance R in the path of the current flow. This is basically an application of Ohm's law, i.e., $E_R = IR$. Second, time to crest in conjunction with the current amplitude fixes the rate of rise of the current. This, in turn, affects the voltage developed across an inductance L, such as wires and structural members, in the path of the current flow (i.e., induced voltage $E_L = L\, di/dt$). It also determines the voltage that will be induced in circuit loops in the vicinity of the lightning circuit.

3.2 GROUNDS, COUNTERPOISES, AND BONDS

The requirements of a lightning protective system with respect to grounding, counterpoises, and bonding are based on the following premises:

(1) The lightning current must be conducted into the earth along a controlled path.

(2) The resistance of the current path should be as low as practical, and the path should be as short as possible to minimize inductance.

(3) In areas where electrical equipment is located, a uniform-potential ground plane should be established.

(4) In areas frequented by personnel, so-called "touch" and "step" voltages should be prevented. These are the voltages that can exist between two pieces of equipment which an operator could simultaneously touch; it is also the voltage difference that can exist between the point where a man is standing and equipment he could touch, or simply the voltage that could exist between a man's feet.

The first step in meeting these requirements is to provide an adequate path to conduct the lightning current into the earth.

1. Ground Rods

Lightning currents flowing through the resistance of the path between the stroke-termination point and the earth will produce voltage drops that may be dangerous to personnel and equipment. Since these currents can be quite large [40,000 A is an average value, with a possible maximum of 180,000 A, as was measured more recently (see Reference 9)], the resistance of the discharge path must be made as small as possible. For this reason, structures or equipment that may be struck by lightning or which form part of a discharge path for lightning current require a good electrical ground (a connection to earth by a low-resistance circuit). The resistance of this circuit is generally the sum of the resistances of the metallic structure or equipment and its joints and the ground resistance. Usually the ground resistance is the largest component.

To define ground resistance, the underground flow of lightning electric current must be understood. This current flows in the earth in three dimensions through a volume of earth that is, in general, not homogeneous. Therefore, a rigorous analysis of the distribution of currents is difficult. However, a quantitative analysis of the electric phenomena in the ground is possible if homogeneity of the ground is assumed. Such an analysis will allow numerical calculations to be made and will permit definite conclusions to be drawn. The available literature on ground resistance is extensive and needs little expansion. The purpose of the following discussion rather is to provide a capsule introduction to ground resistance (References 10 and 11).

To provide a better understanding of the phenomenon of current flow in the earth, a simple electrode in a homogeneous earth will be analyzed. (The meter-kilogram-second system (MKS) will be used because it yields results in practical electric units.)

The simplest electrode is a hemisphere of radius r embedded in the earth as shown in Figure 3.8. If a current I (in amperes) flows through this electrode and spreads out radially in the ground, the current density (in amperes per square meter) at a distance x (in meters) from the center of the hemisphere is

$$J = \frac{I}{2\pi x^2}. \tag{3.1}$$

According to Ohm's law, such a current density produces an electric-field strength given in volts per meter by

$$E = \rho J = \frac{\rho I}{2\pi x^2}, \tag{3.2}$$

where ρ is the resistivity of the earth in ohm-meters.

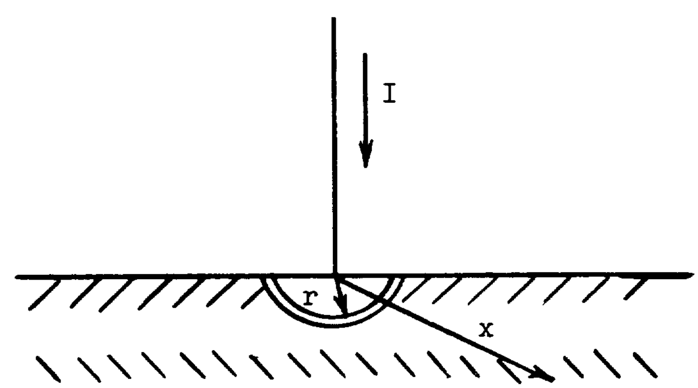


Figure 3.8.–Hemispherical electrode embedded in the earth.

The voltage, as the line integral of the field strength from the surface of the conducting hemisphere to any distance x, is

$$V = \int_r^x E\,dx = \frac{\rho I}{2\pi} \int_r^x \frac{dx}{x^2} = \frac{\rho I}{2\pi}\left[\frac{1}{r} - \frac{1}{x}\right], \tag{3.3}$$

where r is the radius of the conducting hemisphere in meters.

The total voltage between the hemisphere and a far-distant point with $x \approx \infty$ is

$$V = \frac{\rho I}{2\pi r}. \tag{3.4}$$

The total resistance experienced by the stream lines of current diverging from the hemisphere is

$$R = \frac{V}{I} = \frac{\rho}{2\pi r}. \tag{3.5}$$

Notice that the current I is not present in the last term in (3.5).

As an example, a hemisphere of radius $r = 1$ m embedded in soil of resistivity $\rho = 10\ \Omega$-m will have a ground resistance of

$$R = \frac{10\ \Omega\text{-m}}{(2\pi)(1\text{ m})} = 1.6\ \Omega.$$

This is the resistance encountered by current flowing through the entire earth surrounding the electrode. Most of this resistance is encountered in the immediate vicinity of the electrode. As can be shown from (3.3), 50 percent of the total voltage drop resulting from current flowing through this resistance occurs between the electrode and $x = 2r$; 90 percent occurs between the electrode and $x = 10r$. For the example calculated, these distances are 2 m and 10 m, respectively.

The general case described by (3.3) is plotted in Figure 3.9, where the potential at a point some distance x from the center of the electrode is given as a percentage of the electrode potential. This figure can also be used to determine the voltage at a point on the surface of the earth rather than at a point under the earth. For example, consider a hemispherical electrode of 1-m radius (r in Figure 3.10) carrying a current of 10,000 A. If the effective ground resistance of this electrode is 1.6 Ω, the electrode potential is 16,000 V. Now, take point a to be 2 m from the center of the electrode; it can be determined from Figure 3.9 that the potential existing at point a is 50 percent of the electrode potential; that is, point a is at 8000 V.

This example can be extended to show that a significant voltage difference can exist between two adjacent points on the surface of the earth. If point *b* in Figure 3.10 is 2.5 m from the center of the electrode, its potential will be 40 percent of the electrode potential, or 6400 V. Therefore, a voltage difference of 1600 V exists between points *a* and *b*, which are only one-half meter apart.

Thus far, the simple hemispherical electrode has been used to illustrate ground resistance, ground electrode potential, and the voltage gradient existing along the surface of the earth near a current-carrying ground electrode. However, the same phenomena will occur with the more commonly used rod electrode whose length is much greater than its diameter.

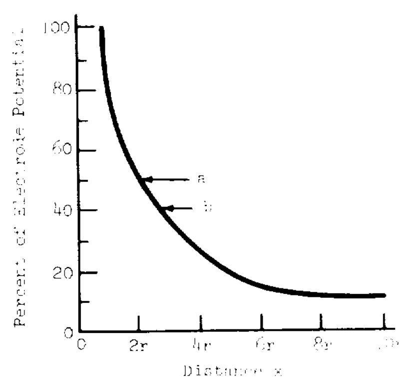


Figure 3.9.—Voltage gradient in the earth for a hemispherical ground electrode for a current *I*.

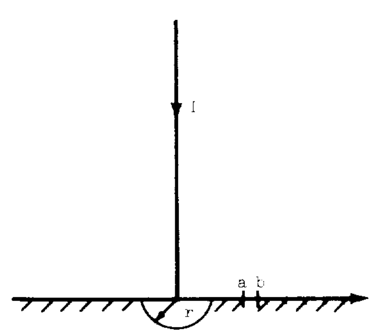


Figure 3.10.—Voltage along surface of the earth, produced by a current-carrying electrode.

Ground rods are commercially available in standard 8- to 12-ft lengths that may be joined by couplings for greater depth. Rod diameters are generally less than 1 in. For a rod whose submerged depth is much greater than its radius, the resistance of the rod (Reference 11) is given by

$$R = \frac{\rho}{2\pi l}\left[\ln\left(\frac{4l}{r}\right) - 1\right], \tag{3.6}$$

for $l \gg r$, where

ρ = ground resistivity in ohm-meters,

l = submerged length of rod in meters,

and

r = radius of rod in meters.

Within practical limits, the diameter of the rod is of much less significance than its length, so long as the logarithmic term, $4\,l/r$, remains unchanged. Since resistance does not decrease directly with the length of the rod, a point is reached at which further increase in length is accompanied by only a minor reduction in ground resistance. Table 3.2, from Sunde (Reference 12), shows the variation of resistance with rod length for various rod diameters; the data were calculated from (3.6) for an average ground resistivity of 100 Ω-m.

The practical approach is to determine the ground resistivity of the earth for a variety of soils and moisture content at a given site. Table 3.3, from Watt (Reference 13), gives representative values. Since ground electrical resistance varies directly with the ground resistivity, the data presented in Tables 3.2 and 3.3 can be used to determine the ground resistance of a typical ground rod.

The following relationship is helpful in determining an unknown ground resistance R_x, where

$$R_x = R_{\text{soil}} \frac{\rho_x}{\rho_{\text{soil}}}$$

$$= R_{\text{soil}} \frac{\rho_x}{100\ \Omega\text{-m}}.$$

The term R_{soil} is a resistance value obtained from Table 3.2, which is based on a soil resistivity of 100 Ω-m, and ρ_x is an earth resistivity value obtained from Table 3.3. For instance, a rod 1/2 in. in diameter and sunk 10 ft into marine sand (such as would exist at Cape Kennedy, Florida) would have a resistance ranging from 0.35 to 3.5 Ω. The range spread is predominantly affected by the moisture content of the sand.

Table 3.2.–Ground resistance (in ohms) for various lengths and diameters of ground rods.*

$2r$ = Diameter of Rod (in.)	l = Length of Rod (ft)						
	1	2	5	10	20	50	100
0.5	225	132	62	35	19.2	8.7	4.7
1	188	113	55	31	17.3	8.0	4.3
2	151	95	47	28.5	15.5	7.2	4.0
4	115	77	40	25	13.6	6.5	3.6
12	69	51	28.5	18.1	10.9	5.4	3.0
24	44	35	21	14.4	9.0	4.6	2.6

*Based on soil resistivity of $\rho_{soil} = 100\ \Omega$-m.

Table 3.3.–Representative values of earth resistivity (Watt et al., Proc. IEEE, 1963).

Type	Material	Approximate Resistivity, ρ_x
Soil	good	10 to 10^2
	average	10^2 to 10^3
	poor	10^3 to 10^4
Water	fresh	10^3 to 10^4
	sea	0.2 to 0.25
Sediments	marine sands and shales	1 to 10
	marine sandstones	1 to 10^2
	clay	10 to 10^2
	sandstone (wet)	10^2 to 10^4
	sandstone (dry)	10^4 to 10^7
	limestone	10^4 to 10^8
Igneous rock	granite	10^3 to 10^9
	basalt	10^5 to 10^9
Metamorphic	slate	10^3 to 10^5
	marble	10^3 to 10^8
	gneiss	10^3 to 10^7
	serpentine	10^3 to 10^7

When it is not possible to obtain the desired ground resistance from a single ground rod, several rods may be driven and connected in parallel. If the spacing between rods is large compared to the length of the individual rods, the resistance will be reduced in proportion to the number of rods. If the rods are close together, each rod will be in the electrical near-field of its neighbor and the overall resistance becomes

$$R = \frac{\rho}{2\pi l} \ln \left(\frac{2l}{A}\right), \tag{3.7}$$

where A represents the radius of an equivalent rod.

The expressions in Figure 3.11 show how the equivalent radius depends on the rod geometry. In each case r denotes the radius of the individual rods.

If the rods are moderately close to each other, the overall resistance will be more than if the same number of rods were spaced farther apart. For instance [from (3.7)], two 3/4-in. diameter, 10-ft rods in parallel and spaced 1 ft apart in a soil resistivity of 10 Ω-m will have a resistance of 2.5 Ω. The same two rods spaced 10 ft apart will have a resistance of 1.9 Ω. Lewis (Reference 14) gives additional information on ground rods in parallel as a function of spacing; also, see Ficchi's data given in Section II-4.5.

In closing this section on ground resistance, it is appropriate to include a short discussion on the measurement of ground resistance. The fundamental method of measuring ground resistance is shown in Figure 3.12. Current is circulated between the ground under test and an auxiliary ground. Preferably, this auxiliary ground should be located at a distance that is large compared to the dimensions of the ground under test, since it is not desirable to have interaction of the ground current distributions at the two electrodes. A voltage is then measured between the ground under test and a reference ground located somewhere between the two current-carrying electrodes. This reference ground should be so located that it is not in the electric field of either of the current-carrying electrodes. Assuming that the current density is negligible at the reference electrode, the resistance of the ground under test is $R = V/I$.

The measurement may be made with a voltmeter and ammeter, with the current being supplied by a transformer energized from the ac power lines. Alternatively, a bridge may be used for the measurement. Most often, however, ground resistance is measured with self-contained instruments such as commercially available ground-resistance testers.

When measuring the ground resistance of a large structure, it may be difficult to place the auxiliary current ground or the reference ground sufficiently far from the ground under test so that the electric fields of the electrodes do not interact. Curdts (Reference 15) has shown that, substantially, the correct ground resistance is measured if the reference ground is located at a point approximately 60 percent of the distance from the ground under test to the auxiliary ground. If the system is large, stray currents in the ground from outside power sources may introduce errors. These tend not to affect the

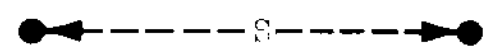
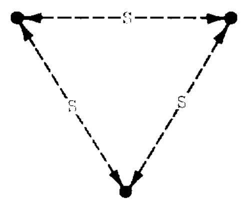
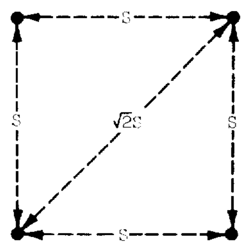
Geometry of Ground Rods	Radius of Equivalent Rod	Equation
	$A = \sqrt{rS}$	(3.8)
	$A = \sqrt[3]{rS^2}$	(3.9)
	$A = \sqrt[4]{\sqrt{2}\, rS^3}$	(3.10)

Figure 3.11.–Equivalent radius of multiple ground rods connected in parallel.

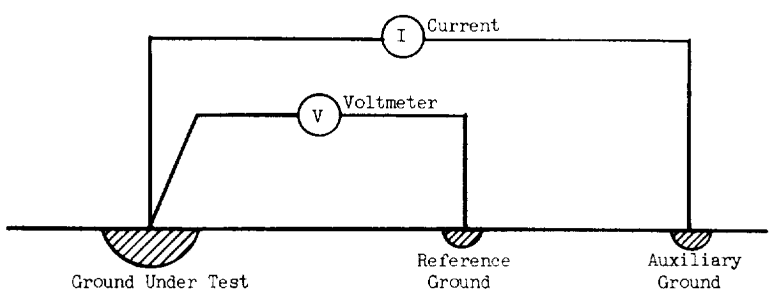


Figure 3.12.–Fundamental method for measuring ground resistance.

measurements when a self-contained ground tester is used, since the frequency of the current set up by the test set is generally different from the frequency of the interfering currents. Such interfering currents can also be balanced out in the measuring instrument. Duke and Smith (Reference 16) have described a 60-Hz test set, which contains such balancing circuits, for the measurement of low-impedance grounds.

2. Counterpoises

A counterpoise is a single, bare wire or a network of bare wires buried horizontally. The counterpoise functions to reduce the resistance of a ground electrode, interconnect ground electrodes, provide a convenient means for grounding equipment and circuits, reduce voltage gradients on the surface of the earth, and intercept lightning stroke currents that would otherwise terminate on equipment and circuits being protected.

From Sunde (para. 3.6, Reference 12), the resistance of a single, buried horizontal wire is

$$R = \frac{\rho}{\pi l}\left[\ln\left(\frac{2}{\sqrt{2rd}}\right) - 1\right], \tag{3.11}$$

for $l \gg d$, where

ρ = resistivity in ohm-meters,

l = length of wire in meters,

r = wire radius in meters,

and

d = burial depth in meters.

Table 3.4 shows how the resistance of a counterpoise varies with length of wire and burial depth.

Equation (3.11) is derived on the assumption that the potential is uniform over the entire length of the wire, which holds only if the wire has infinite conductivity. A very long wire will not be at the same potential along its entire length, and therefore (3.11) will be somewhat in error. However, from a practical viewpoint, the diameters and lengths of counterpoise wire normally used are such that the error introduced as a result of wire resistance may be neglected.

Table 3.4.– Variation of resistance (in ohms) with length of a horizontal wire for a ground at the surface and at a depth of 12 in.*

Burial Depth (in.)	Length of Wire (ft)						
	10	20	50	100	200	500	1000
0	80	45	19.4	10.4	5.6	2.4	1.29
12	47	27	12.8	7.1	3.9	1.75	0.45

*Earth resistivity = 100 Ω-m; wire gauge = No. 10 AWG.

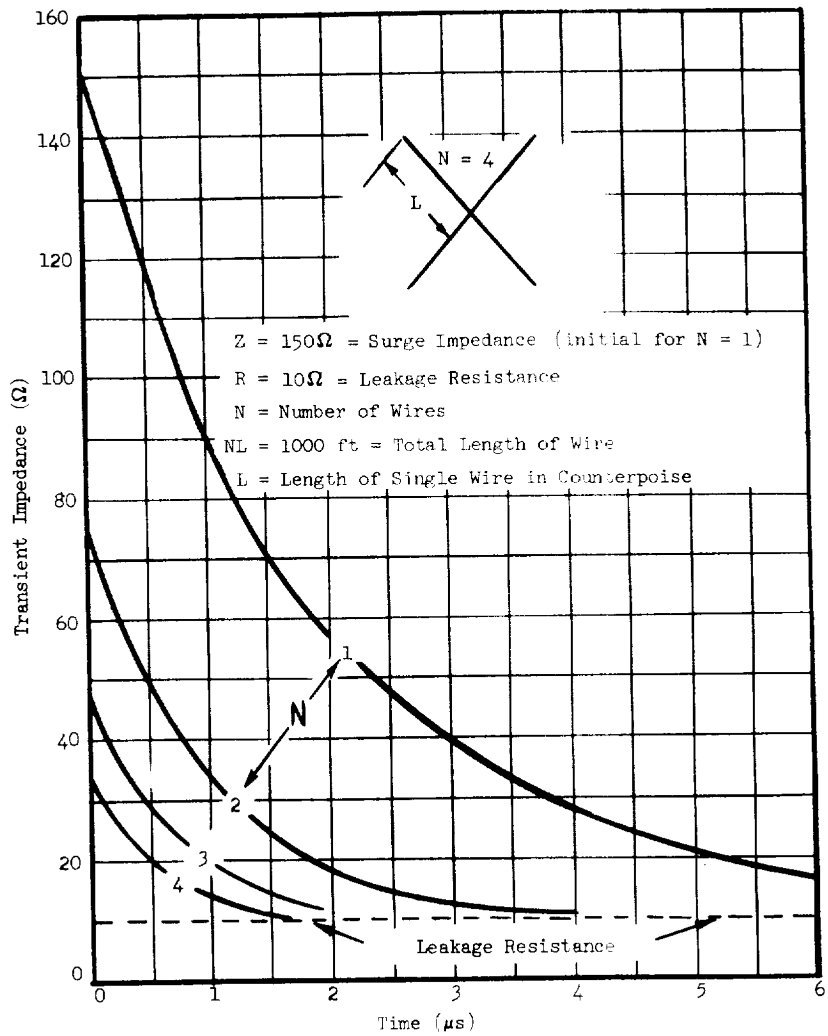


Figure 3.13.–Transient impedance of 1000 ft of counterpoise as a function of surge propagation time with the number of radial wires as a parameter. (After "Protection of Transmission Lines Against Lightning," by W. W. Lewis, 1950)

Another consideration in determining the effective resistance of the counterpoise is its transient response (i.e., the change in impedance along the counterpoise with respect to time) when a high current surge, such as a lightning stroke, is applied. Initially, the effective resistance is quite high, of the order of 150 Ω. The initial value of the resistance is defined as the surge impedance of the wire. As the surge propagates along the wire, the resistance decreases as the current spans more and more of the wire, thus making more effective contact with the earth.

The surge propagates in the earth at roughly one-third the speed of light. As an example, approximately 3 μs would be required for current to span a 1000-ft wire. For a given length of buried wire, the transient resistance will reduce to its steady-state resistance faster if the wire is arranged as several short radial wires than if it is laid as one long wire. Figure 3.13 (Reference 14) shows how the transient response of buried wires varies for several configurations of counterpoise wires. In the case of driven ground rods, the final resistance can be obtained quite quickly since the rods are fairly short. The final resistance of widely spaced ground rods cannot be attained until a surge current has reached the most distant rod. As a practical matter, the first 250 ft of buried conductor are the most effective for the grounding of lightning currents with counterpoise.

Ground resistance decreases with increasing current, at least until the current heats the soil moisture to the boiling point. The proportional reduction is less for grounds of low resistance than for grounds of high resistance. Figure 3.14 (Reference 14) shows typical measured values of ground resistance as a function of impulse current.

3. Bonding

Bonding can be defined as providing an electrical connection across mechanical joints between metallic structures, for example, the electrical connection of conduits and piping to structural metal, and the interconnection of reinforcing metals in cement structures (Reference 17). The bonding of these joints is necessary to provide a minimum resistance and direct path for lightning current from the point of stroke termination to the earth, to protect personnel from the shock hazard resulting from equipment internally power faulted, and to prevent the accumulation of static charge that would produce radio interference and sparking, thereby creating shock and explosion hazards.

3.3 VOLTAGES INDUCED BY LIGHTNING

The purpose of the following section is to describe the three principal lightning stroke phenomena that produce voltages in electrical circuits. These are as follows:

(1) The voltage induced in circuits that are magnetically coupled to conductors carrying lightning currents.

(2) The *IR*, or resistance voltage, resulting from lightning current flowing in the structure and ground resistance.

(3) The voltage induced in circuits that are capacitively coupled to the lightning stroke.

Lightning voltages in electrical circuits are generally some combination of the three components. For ease in understanding, the components will first be discussed separately.

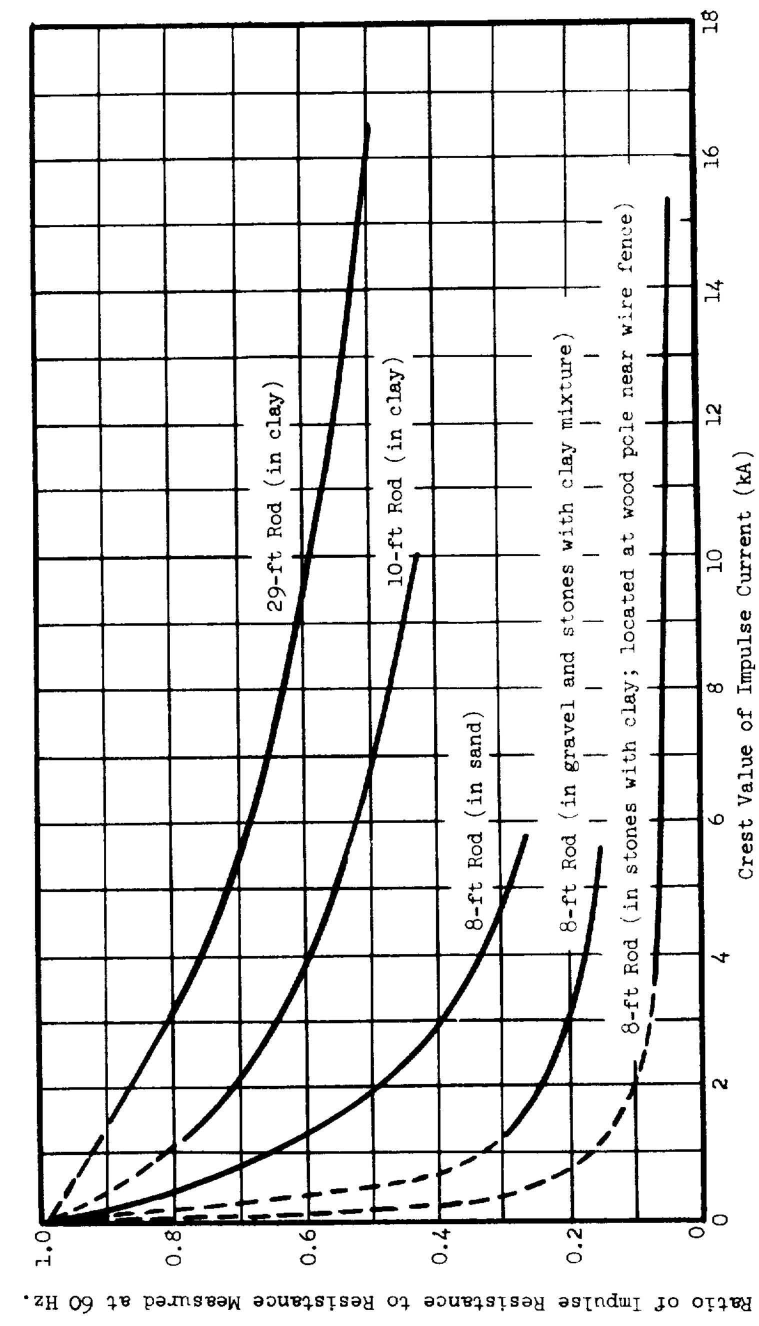


Figure 3.14.–Measured values of ground resistance as a function of impulse current for various ground rods and earth conditions. (After "Protection of Transmission Lines Against Lightning," by W. W. Lewis, 1950)

1. Inductive or Magnetic Component

The magnetically induced voltage is the most complex and, as previously stated, is the induced voltage in a circuit that is magnetically coupled to a current-carrying conductor. For instance, in Figure 3.15, the voltage is magnetically induced in loop A by the field produced by the changing lightning-stroke current flowing in conductor B.

As another example of induced voltage, consider Figure 3.16, where the current path forms part of the voltmeter loop; there is negligible current flowing through the voltmeter. This simple circuit is symbolic of lightning current flowing in a wire with a man touching the wire at two points, a and d. The voltage E is related to the rate of change of flux produced by the changing current dI/dt in the loop according to the following equation:

$$E = -N\frac{d\phi}{dt}, \tag{3.12}$$

where

N = number of turns

and

ϕ = flux in webers enclosed by the loop.

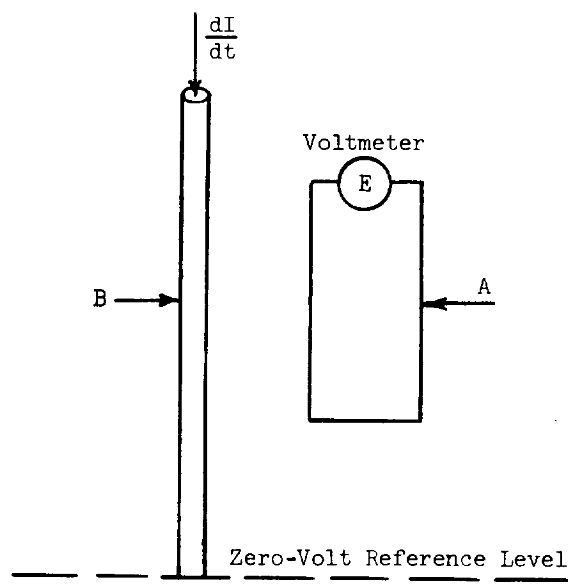


Figure 3.15.—Magnetically induced voltage E in a loop (A) by changing current in a conductor (B).

The dimensions of the loop and rate of change of current must be known to actually determine the voltage. First, the flux density in W/m^2 is determined from the relationship

$$\beta = \frac{\mu I}{2\pi r} \ , \tag{3.13}$$

where

μ = permeability of space, $4\pi \times 10^{-7}$ H/m,

I = current in conductor in amperes,

and

r = radial distance from conductor in meters.

Next, the flux linked by loop *abcda* is determined by taking the line integral around the loop to obtain

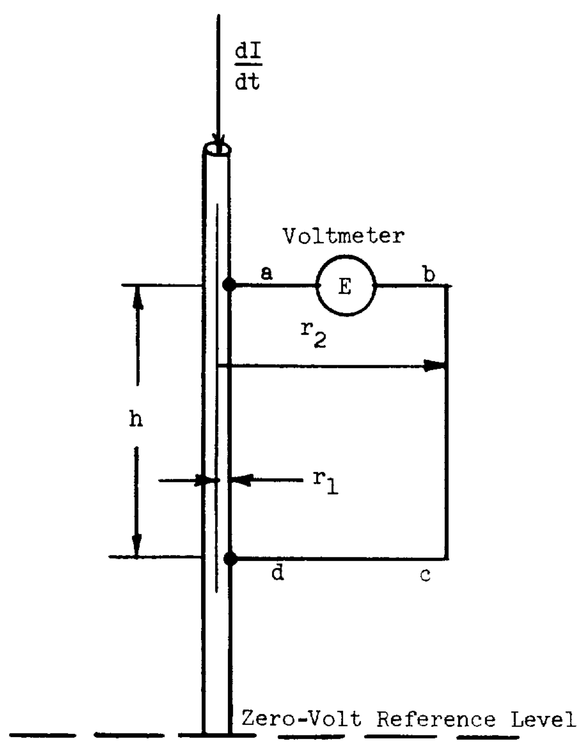


Figure 3.16.–Circuit for analyzing magnetically induced voltage in a voltmeter loop as the result of current flowing in an adjacent conductor.

$$\phi = \oint \beta dr = \frac{h\mu I}{2\pi}\int_{r_1}^{r_2} \frac{1}{r} dr ,$$

$$\phi = 20 \times 10^{-8} Ih \ln\left(\frac{r_2}{r_1}\right). \tag{3.14}$$

The time rate of change of flux is

$$\frac{d\phi}{dt} = 20 \times 10^{-8} h \frac{dI}{dt} \ln\left(\frac{r_2}{r_1}\right). \tag{3.15}$$

From (3.12) and (3.15) the voltage induced in the single-turn loop in Figure 3.16 is

$$E_{ba} = 20 \times 10^{-8} h \frac{dI}{dt} \ln\left(\frac{r_2}{r_1}\right) \tag{3.16}$$

where

h = loop height in meters

and

$\frac{dI}{dt}$ = rate of change of current in amperes per second.

For h in feet,

$$E_{ba} = 61 \times 10^{-9} h \frac{dI}{dt} \ln\left(\frac{r_2}{r_1}\right) \tag{3.17}$$

The equations for E_{ba} express the loop voltage in terms of loop dimensions and rate of change of the flux linked by the loop. From these equations it can be seen that the loop voltage is a function of the loop dimensions. For example, if one were attempting to measure the voltage between a and d, the value obtained would be dependent upon the dimensions of the loop and not simply the height, a to d. This concept of induced voltage is very important and, therefore, in an attempt to clarify it further, the same loop voltage will be described in terms of the familiar lumped inductance.

Consider the circuit in Figure 3.17 where

I = current in conductor ad in amperes,

L_{11} = self-inductance of the current-carrying conductor between a and d, in henries,

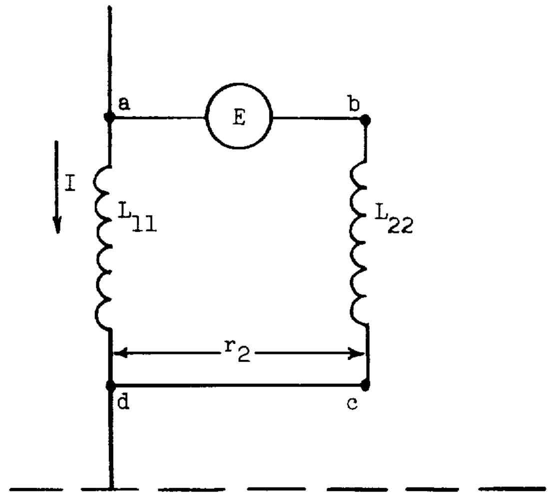


Figure 3.17.--Circuit for analyzing voltage induced in a flux-linked loop.

L_{22} = self-inductance of the meter lead, in henries

and

$L_{12} = L_{21}$ = mutual inductance between L_{11} and L_{22}, in henries.

Equating the voltages around loop *abcda* yields

$$E_{ba} = L_{11}\frac{dI}{dt} - L_{12}\frac{dI}{dt} = (L_{11} - L_{12})\frac{dI}{dt}. \tag{3.18}$$

If ϕ_1 is the magnetic flux produced by current I, then for a single turn,

$$L_{11} = \frac{\phi_1}{I}. \tag{3.19}$$

From (3.14), the flux in webers is

$$\phi_1 = \frac{h\mu I}{2\pi}\int_{r_1}^{R}\frac{1}{r}dr, \tag{3.20}$$

where

h = length in meters of the conductor being considered,

r_1 = radius in meters of the conductor,

and

R = radius to a distant point where flux density = 0.

Integrating (3.20) and combining with (3.19) yields

$$L_{11} = \frac{h\mu}{2\pi}(\ln R - \ln r_1)\,. \tag{3.21}$$

Similarly, for a single turn, the mutual inductance in series with L_{11} is

$$L_{12} = \frac{\phi_2}{I} \tag{3.22}$$

where the magnetic flux produced by the induced current is

$$\phi_2 = \frac{h\mu I}{2\pi}\int_{r_2}^{R} \frac{1}{r}dr\,, \tag{3.23}$$

where r_2 is the radial distance between the center of the current-carrying conductor and the meter lead. Integrating (3.23) and combining with (3.22) yields the mutual inductance:

$$L_{12} = \frac{h\mu}{2\pi}(\ln R - \ln r_2)\,. \tag{3.24}$$

Combining (3.18), (3.21), and (3.24) yields

$$E_{ba} = \frac{h\mu}{2\pi}\ln\left(\frac{r_2}{r_1}\right)\frac{dI}{dt}\,,$$

where

h = length of the current-carrying conductor in meters,

$\mu = 4\pi \times 10^{-7}$ H/m,

and

$\frac{dI}{dt}$ = rate of change of current in amperes per second.

Therefore,

$$E_{ba} = 20 \times 10^{-8} h \frac{dI}{dt} \ln\left(\frac{r_2}{r_1}\right). \qquad (3.25)$$

Note that (3.25) is identical to (3.16).

So far, a "resistanceless" conductor has been assumed. The addition of resistance to the circuit is illustrated in Figure 3.18. Equating voltages around loop *abcda* yields

$$E_{ba} = R_1 I + (L_{11} - L_{12})\frac{dI}{dt}, \qquad (3.26)$$

since current flows through R_1 but not R_2.

A comparison of (3.26) with (3.18) shows that, in the case of this loop, the total voltage drop across the resistance is simply added to the net drop of induced voltage.

Thus far, cases have been considered in which the current-carrying conductor formed part of the loop. In the more general case, the loop is electrically isolated from the current conductor, as shown in Figure 3.19. In this case, the drop in induced voltage is

$$E_{ba} = (L_{12} - L_{13})\frac{dI}{dt}, \qquad (3.27)$$

where L_{12} is the mutual inductance between L_{11} and L_{22} [see (3.22)] and L_{13} is the mutual inductance between L_{11} and L_{33}.

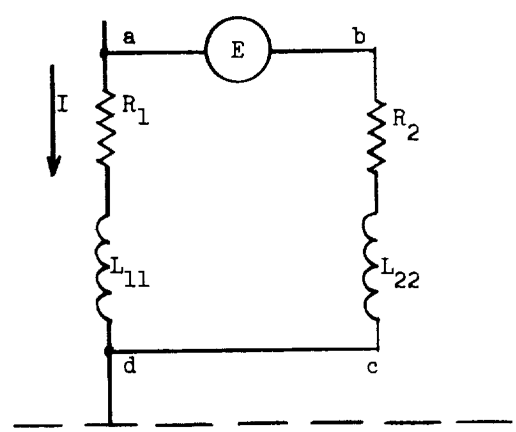


Figure 3.18.--Circuit for analyzing voltage induced in a flux-linked inductive loop containing resistance.

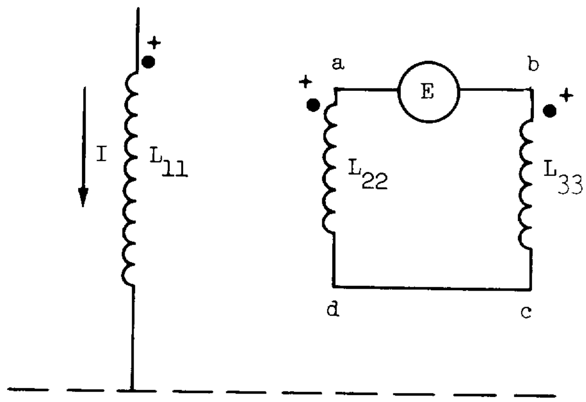


Figure 3.19.–Circuit for analyzing magnetically induced voltage in an isolated flux-linked inductive loop.

Equation (3.27) has exactly the same form as (3.18), and expressing (3.27) in terms of circuit dimensions instead of inductance yields (3.25), repeated here for convenience:

$$E_{ba} = 20 \times 10^{-8} h \frac{dI}{dt} \ln \left(\frac{r_2}{r_1} \right),$$

where

r_2 = radial distance between L_{11} and L_{33} in meters,

r_1 = radial distance between L_{11} and L_{22} in meters,

and

h = height of L_{22} and L_{33} in meters.

Equation (3.17) expresses E_{ba} for circuit dimensions in feet instead of meters:

$$E_{ba} = 61 \times 10^{-9} h \frac{dI}{dt} \ln \left(\frac{r_2}{r_1} \right).$$

Expressing (3.17) in a slightly different form yields

$$E_{ba} = 61 \times 10^{-9} h \frac{dI}{dt} \left[\ln \left(\frac{W}{r_1} + 1 \right) \right],$$

where $W = r_2 - r_1$.

As an example of the application of (3.17), calculations were made of the voltage induced in a loop inductively coupled to the current-carrying conductor for loops of different widths and distances from the flux-producing conductor. Figure 3.20 shows the results of such calculations for current changing at the rate of 1000 A/μs; the induced voltages are given per foot of loop height h for a single-turn loop. The figure graphically illustrates the effect of moving the loop with respect to the current conductor and varying the width of the loop, which changes the amount of linked flux.

As another example of a voltage induced in a loop near a conductor carrying a changing current, consider Figure 3.21.

Here, the voltmeter has been replaced by a man, and the loop is the path from the man's hand, down his body, across the earth, and up the conductor. If $h = 5$ ft, $r_1 = 0.01$ ft, and $r_2 = 2$ ft, from (3.17) the voltage across the man would be

$$E_{\text{man}} = (61 \times 5) \ln \left(\frac{2}{0.01}\right) \frac{dI}{dt} \times 10^{-9} .$$

For a current changing at the rate of 50,000 A/μs ($dI/dt = 5 \times 10^{10}$ A/s), the induced voltage is

$$E_{\text{man}} = 80{,}800 \text{ V} .$$

These examples illustrate the well-known principle that the magnetically induced voltage depends on the flux linked by the loop and that the flux linked by the loop depends upon the loop dimensions.

2. Resistive Component

The resistive component of voltage is best described by refering to Figure 3.22. The lightning stroke terminates at equipment A, and the current I flows into the earth through the ground resistance R_F, producing a voltage drop E between the cable connected to A and the zero-volt reference potential of the earth. Since there is no electric circuitry between A and the zero-volt reference, the magnitude of E does not pose a burnout problem. However, at the other end of the cable, which is shown with the sheath isolated, a dangerous voltage E' exists between the cable sheath and equipment B. It is important to recognize that when there is an open-circuit discontinuity in the sheath, negligible current flows to ground through R'_F; equipment B remains at the zero-volt reference potential and the voltage E' is across the discontinuity. The voltage E_1, between the circuit component and the metal housing of B, will be some fraction of E', depending on circuit impedances. Both E' and E_1 are the resistive components of voltage resulting from the lightning stroke terminating at A. The most direct way to minimize the effects of E' is to connect the sheath to B and allow the sheath to carry a portion of the stroke current.

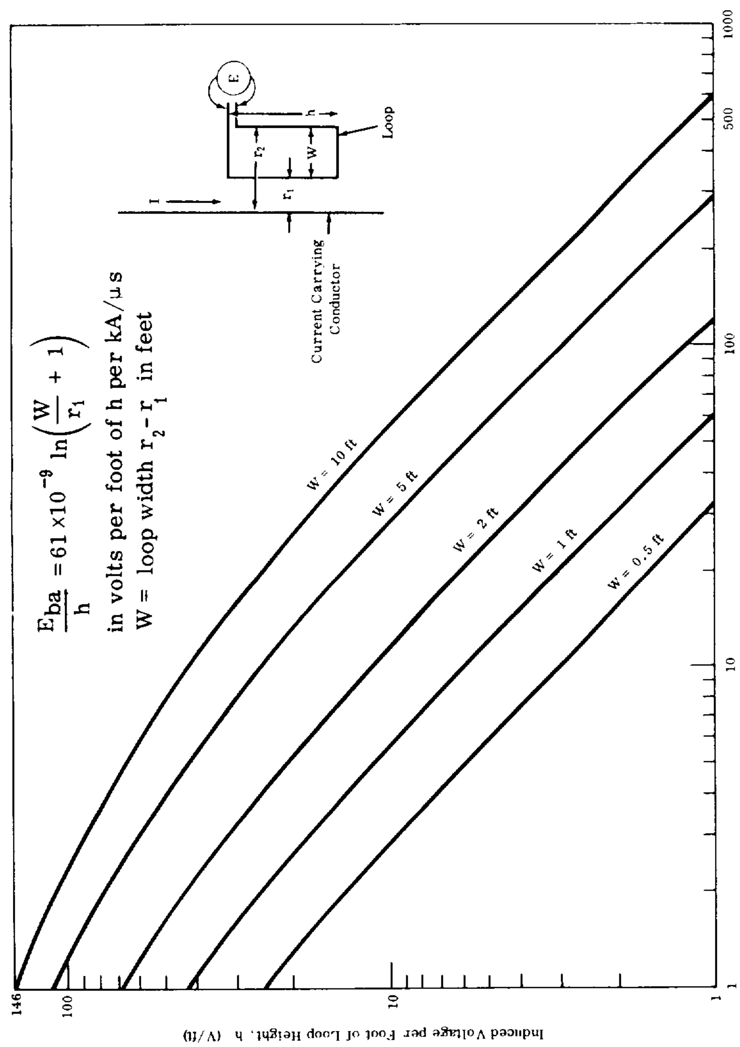


Figure 3.20.—Voltage induced in a single-turn loop for current changing at a rate of 1000 A/μs.

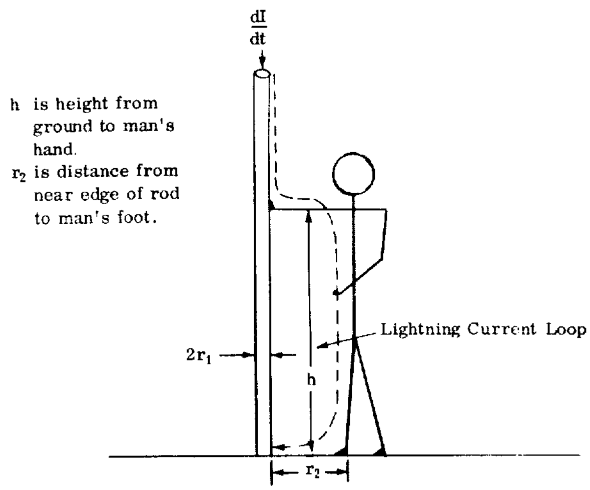


Figure 3.21.–Diagram of a lightning current loop formed by a man.

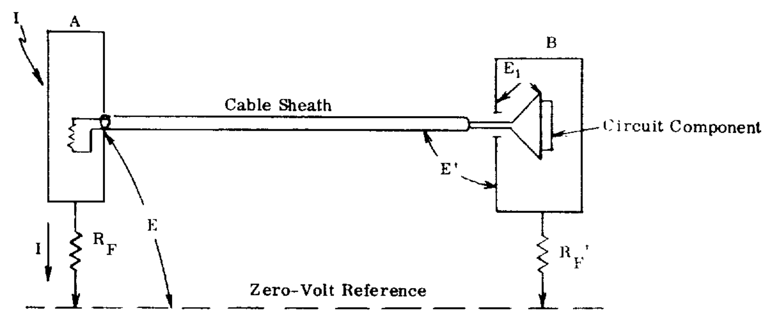


Figure 3.22.–Diagram showing resistive components E' and E_1 resulting from a lightning stroke.

3. Capacitive Component

Prior to a lightning stroke, an electric charge slowly accumulates on earth-based conducting objects in the vicinity of the electrified clouds. This increase in charge occurs slowly enough so that the potential of grounded conductors does not change appreciably with respect to ground, even when the conductors are grounded through a high impedance. When the lightning stroke terminates on a point such as *A* in Figure 3.23, the charge on the grounded objects suddenly becomes redistributed. The redistribution of charge manifests itself as a current flow through the grounding impedance of the conductors and produces a voltage across that impedance known as the capacitive voltage component of the stroke.

Referring to Figure 3.23, the voltage between the conducting object and ground can be expressed as

$$E = \frac{Q}{C} e^{-t/RC}, \tag{3.28}$$

where

Q = charge in coulombs,

$C = C_1 + C_2 + \ldots + C_n$, the total capacitance to ground in farads,

$R = R_f R_f' / (R_f + R_f')$, the effective resistance to ground in ohms,

and

t = elapsed time since stroke, in seconds.

Equation (3.28) shows that if the product RC is small, the exponential term will be large (for a time t of the order of 10 μs), thus making the value of a voltage capacitively induced on any well-grounded structure quite small for a typical stroke.

For example, if $RC = (1000\ \Omega)(1000 \times 10^{-12}\ \mathrm{F}) = 10^{-6}$ s, at the end of time $t = 10\ \mu$s, the value of E will be 1/22 V for $Q = 10^{-6}$ coulombs. For the larger value of $RC = (10^6\ \Omega)(10^{-9}\ \mathrm{F}) = 10^{-3}$s, the value of E will be 1000 V for a charge of the same magnitude.

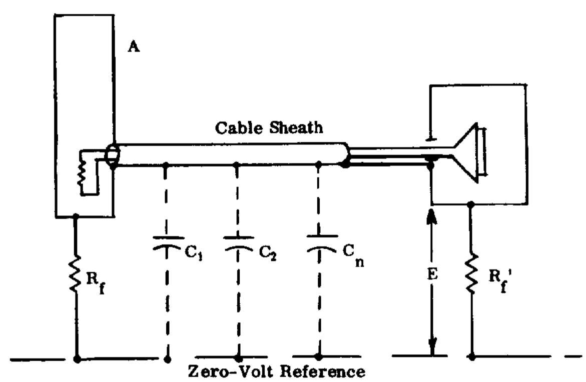


Figure 3.23.–Diagram showing capacitive component of lightning voltage induced in a circuit.

4. RECOMMENDATIONS FOR LIGHTNING PROTECTION

This section presents a discussion of the factors which should be considered in providing lightning protection for new installations, together with appropriate recommendations for implementing protective measures. This section is organized in four parts: protective grounding signal and data transmission circuits, power circuits, and surge protection devices for equipment.

4.1 GROUNDS FOR SAFETY PROTECTION

Protective grounding at fixed stations is intended to provide protection for operators, electrical and electronic equipment, and buildings and equipment housings. Some of the recommendations made to achieve effective protective grounding may be in conflict with desirable practices for grounding signal and data transmission circuits. However, these conflicts can usually be resolved in a satisfactory manner.

Common grounding connections between all metal parts in and around a building provide the best protection for operators and equipment. Parts to be grounded include all metal piping, conduits, structural members, and building outer skins, as well as equipment cabinets, racks, and enclosures.

Such a grounding arrangement should utilize a buried counterpoise around the periphery of buildings, with ground rods at the corners and spaced at intervals not exceeding 50 ft for large buildings. Such a grounding system is illustrated in Figure 3.24. This arrangement permits separate power and equipment grounding within the building. Should there be a requirement for a completely separate signal or instrumentation ground connected to earth at some point other than the building counterpoise, a protective device should be connected between it and equipment ground at each equipment cabinet, enclosure, or rack where the separate signal ground is used. Such a protective device that has proven to be extremely effective for this application is a high-surge-current, controlled avalanche selenium rectifier.

In general, the protective measures applied to the structural parts of all buildings should follow the recommendations given in Reference 18.

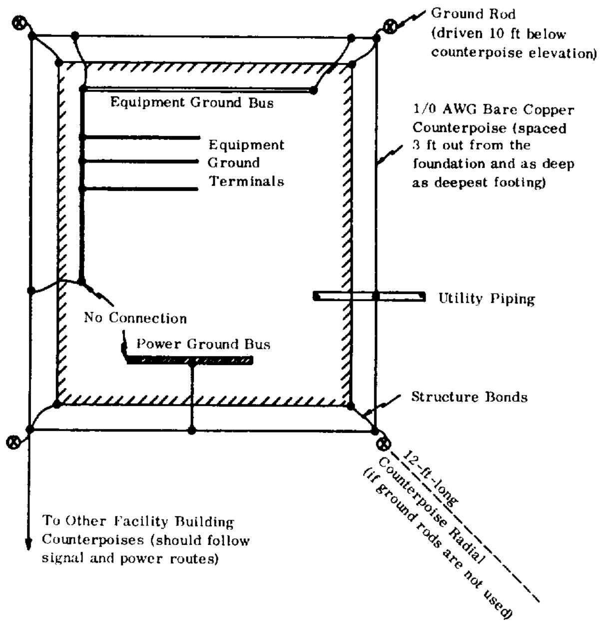


Figure 3.24.–Diagram of typical grounding for a building.

4.2 GROUNDS FOR SIGNAL AND DATA TRANSMISSION CIRCUITS

Signal and data transmission circuits provide paths through which lightning currents and voltages can enter sensitive equipment. Circuits totally contained within protected buildings and having no exposure to lightning currents can usually be designed without direct consideration of lightning effects. Circuits which are not totally contained within protected buildings or are connected to or are a part of circuits not within protected buildings must be designed to accommodate lightning effects.

The effects of lightning damage fall into two general categories. The first includes all types of burning and blasting mechanical damage that can occur when a lightning stroke contacts an object. The most direct way of minimizing mechanical damage is to provide a means of intercepting the stroke before it actually strikes. This interception can be accomplished by enclosing the circuits within metal conduit or by providing grounded guard wires such as are used on power transmission lines. The best protection for underground data circuits is provided by two bare guard wires buried 3 to 4 ft above and to either side of a cable trench. Data lines above ground should be protected by an overhead guard wire that is grounded at intervals not exceeding 250 ft and which is spaced at least 8 ft above the data circuit to be protected.

The second category of lightning damage involves the electrical properties of equipment circuits. Because electrical damage may arise from an immensely complicated situation, it is not wise to generalize on the proper protective measures to be applied. A complete analysis of the possible effects of lightning should always be made because an effective design from a lightning viewpoint should consider both the voltage tolerance levels of the equipment and the expected magnitudes of the lightning-induced voltages. The following sections outline lightning protective measures necessary to guarantee minimum induced voltages when common construction practices are utilized.

The use of rigid steel conduit to form a continuous grounded metal shield around signal and data transmission circuits results in a system that is immune to the effects of lightning. Such a system usually requires no other supplementary protective devices, since the use of a ferromagnetic conduit eliminates the need for special circuit grounding for the control of lightning.

For example, on circuits contained within 2-in. (electrical-trade size) rigid steel conduit, lightning-induced voltages will be of the order of 30 V between conductors and ground and 0.2 V between conductors; these results were obtained for an average lightning stroke directly to a 5000-ft-long conduit. Larger conduits and/or shorter lengths will result in lower voltages. The induced voltages are proportional to length and inversely proportional to the product of the square of the thickness of the conduit and its cross-sectional area.

The conductor-to-ground voltages are essentially those of the *IR* drop along the inside of the conduit, where *R* is a complex function of the conduit permeability, conductivity, cross section, and the rate-of-change of the lightning current. For the case of 2-in. rigid steel conduit and a simulated current pulse shaped like lightning, the effective resistance has been measured to be only 0.001 of the dc resistance (Reference 19).

The voltage between conductors is primarily due to the flux leakage at joints and bends and, hence, varies nearly linearly with length.

The use of non-ferromagnetic metal conduit (e.g., aluminum) will result in significantly higher voltages than ferromagnetic conduit (e.g., black iron pipe) even if the non-ferromagnetic metal conduits have a significantly lower dc resistance. For instance, the substitution of 2-in. aluminum pipe for 2-in. rigid steel conduit results in approximately a 42-fold increase in voltage, even though the dc resistance of the aluminum conduit is 0.06 that of steel. This apparent inconsistency is accounted for by the large permeability of steel relative to that of aluminum.

The type of construction utilizing above-ground cable trays with guard wires is frequently used in warm regions and provides the greatest flexibility for adding or deleting cables. This construction requires the following special bonds and grounds in addition to the guard wires described in Section 4.2.

(1) Grounding. The cable trays should be grounded to driven rods at intervals not exceeding 250 ft and to building ground systems at both ends.

The ground rods should be about 16 ft long. If soil conditions prohibit driving rods, equivalent counterpoise should be installed (see Section 3.2).

(2) Bonding. All trays should be bonded together so as to form a continuous electrical structure.

(3) Cables. All signal and data cables should have a continuous overall metal braid or shield (not spiral tape) that is bonded to the building ground system at each end.

Overhead cable trays with guard wires will not provide the protection that is obtained with steel conduit; some sensitive circuits may require special protective devices. Voltages between conductors in cable trays will be approximately an order of magnitude larger than for conductors in steel conduit. The conductor-to-ground voltages can be estimated utilizing the techniques given in Section 5.2.

The type of construction utilizing buried nonmetal conduits with guard wires does not provide the protection that can be obtained with cable trays or conduits. In addition to the guard wires described in Section 4.2, all signal and data cables should have a continuous overall metal braid or shield (not spiral tape) that is bonded to building ground systems at each end.

Voltages between conductors buried in nonmetal conduits with guard wires will be approximately 100 times the voltages between conductors in steel conduit; sensitive circuits connected to long runs will therefore require protective devices. The conductor-to-ground voltages can be estimated utilizing the techniques given in Section 5.2.

The type of construction utilizing buried conductors with guard wires is essentially equivalent to that described above, and the same restrictions and requirements apply.

4.3 GROUNDS FOR POWER CIRCUITS

Wiring and lightning protection for power circuits should in general follow commercial practices. Where sensitive equipment is used or where high reliability is desired, the following additional protective measures should be used.

The primary power should be distributed at as high a voltage as is economical. Lightning-induced voltages are primarily functions of equipment geometry and not system voltage. Therefore, the higher the system voltage, the lower will be the effects of lightning-induced voltage on a percentage basis. Also, the higher voltage systems have lower ratios of lightning arrester protective breakdown voltage to system operating voltage.

Lightning arresters should be applied at both ends of all primary distribution power lines regardless of whether these are buried or above ground.

Station-type arresters rather than distribution type arresters should be used.

Secondary arresters should be applied at all distribution panels fed from a transformer whose primary is served by a power line exposed to lightning.

4.4 SURGE PROTECTION DEVICES FOR CONTROL EQUIPMENT

In general, control circuits and communication systems are operated at voltage levels considerably below those used for electric power work; therefore, the protective devices for these circuits need not handle the very large amounts of energy encountered at higher voltage levels. Some of the devices to be described here will handle large current surges, but others are somewhat limited in their current-carrying capabilities.

According to their mode of operation, protective devices acting as overvoltage suppressors for control circuits and communication systems are classified as "crowbar" devices or switches (e.g., zener diodes), voltage clippers and nonlinear resistance elements (e.g., Thyrite), or energy storage elements such as a capacitor or inductor.

Crowbar devices or switches operate by abruptly conducting or breaking down electrically when the voltage between the electrical terminals reaches a specified value. This action results from the breakdown of a dielectric gas between two electrodes or from avalanche or other solid-state phenomena occurring across a semiconductor junction. In general, devices employing breakdown of a gas can control high-energy lightning surges if the tolerances on the voltage levels for which protection is required are rather loose. Semiconductor devices can provide very tight control over voltage levels, but they have limited thermal capacity to control high-surge energy.

The breakdown or avalanche effect may be triggered directly by the overvoltage, so that the device appears as a two-terminal device, or it may be triggered by an auxiliary electrode (gate) in a three-terminal device with the triggering signal supplied by an adjustable voltage-sensing circuit.

Many of these devices will continue to conduct after the initial triggering. In such cases, external means should be supplied to interrupt the power supplied to the device after the triggering surge has disappeared. Generally, a serious disturbance is introduced in the system voltage until the power actuating the device can be interrupted. Typical commercial devices that incorporate the interrupting means and which are currently available are listed in Table 3.5. This feature is limited to ac applications; for dc circuits, a definite voltage interruption is required to stop the current.

The initial conduction or breakdown action of protective devices is not instantaneous. As in the case of a spark gap, the time to breakdown is related to the rate at which the voltage is applied so that the voltage at which conduction is initiated increases when the rate of voltage rise is increased. The voltage-time characteristic of a typical crowbar device is shown in Figure 3.25.

The significance of the voltage-time characteristic is that there is a time lag during which the protective device has no effect on the overvoltage. For lightning surges having steep wave fronts, high voltages are left unsuppressed for time durations of the order of a fraction of a microsecond until the protective device turns on. In this respect, the behavior of the crowbar device is similar to that of the simple spark gap discussed earlier.

Spark gaps, gas tubes, and semiconductor switches have both advantages and disadvantages. Some of these are listed in Table 3.6; other features of these protective devices are enumerated in the following paragraphs.

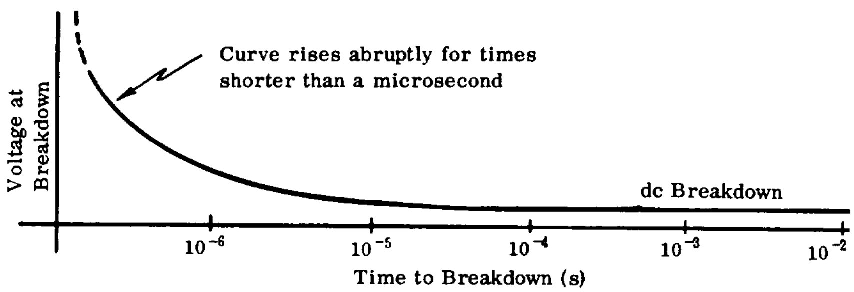


Figure 3.25.–Typical voltage-time characteristic for breakdown of a "crowbar" device.

Table 3.5.–Summary of characteristics of typical, commercially available, low-voltage surge suppressors.

Manufacturer	Model No.	Range of Minimum Breakdown Voltages*	Ratio of Max. to Min. Breakdown Voltage**	Peak Current (A)	Self-Interrupting
General Electric Company	730 B	250 to 6000	1.6	2000	No
AMARK Corp.	Cerbesis UA1	250	2.5	2000	No
EC & G Corp.	Fenotron	800	unknown	3000	No
Victoreen Inc.	VX-96	150 to 5000	unknown	unknown	No
Electronic Industries		750 to 50,000	unknown	1000 to 6000	No
Siemens Corp.	surge voltage protector	230 to 800	unknown	unknown	No
Western Electric	GA 51574, GA 51724	500 to 2600	1.6 or greater	6000	No
L. M. Ericsson Corp.		250 to 400	4	unknown	No
Amperex Corp.	Model 0369	200	7.5	unknown	No
	neon bulbs	70 to 120	4	unknown	No
General Electric Company	9 La 4C4	1000	2.5	10,000	Yes
Westinghouse Corp.	appliance protector	1000	4	5000	Yes
Dale Corp.	LA-9	1500 to 5000	6	300	Yes
General Electric Company†	Selenium rectifier, Thyrector types 6RS21SA2D2, 6RS15SA1D1, or 6KS32SA1D1	25 to 500	Zener type	300	Yes

*Minimum overvoltage at which the protector will operate. Some manufacturers provide gaps for various voltages.
**Maximum breakdown voltage is defined as the operating voltage at the 0.1-μs point.
†Operating temperature -70° to +160°F; also meets humidity requirements of MIL-Spec 202.

Table 3.6.–Advantages and disadvantages of miscellaneous types of lightning protective devices.

Device	Advantages	Disadvantages
Spark gaps	1. Simple and reliable. 2. Easily fabricated. 3. High energy handling capacity. 4. Very low voltage drop (arc drop) during conducting state. 5. Bilateral operation (same characteristics on either polarity). 6. Fast response time (start conducting in less than 1 μs if well designed). 7. Zero power consumption on standby. 8. Wide operating range. 9. Long life expectancy. 10. Low internal capacitance. 11. Require no auxiliary equipment, power supply, or maintenance. 12. Relatively unaffected by radiation.	1. Relatively high sparkover potential for their low-voltage ratings. 2. Simple gaps will not extinguish follow-on-current. 3. Seldom available in conveniently packaged assemblies; must be designed for each specific application.
Gas tubes	1. Low cost. 2. Small size (depending on bulb). 3. Low sparkover voltage--typically 60 to 100 V in firing times greater than 2 μs. 4. Can pass very high currents for short time. 5. Self-healing (usually).	1. Poor voltage-time characteristic. 2. Will continue to conduct if the driving voltage is above 60 to 100 V. 3. Possibly more sensitive to radiation than spark gaps in air at atmospheric pressure. 4. Will not absorb large amounts of energy.
Semiconductor devices used as crowbars	1. Good surge current ratings, although not as good as spark gaps. 2. Low voltage drop when conducting. 3. Suitable for use on low voltage dc circuits. 4. If properly applied, will interrupt follow-on current at the first zero following initiation of conduction.	1. Low thermal capacity to dissipate surge energy. 2. Must be triggered by an auxiliary circuit. 3. Will not interrupt follow-on current on dc circuits. 4. Limited in rate of buildup of current or rate of buildup of voltage that can be tolerated. 5. Expensive. 6. Not bilateral. (For protection on both polarities, two rectifiers and additional circuitry must be used. Bilateral devices are commercially available with two series back-to-back SCR's).

1. Spark Gaps

In operation, spark gaps tend to conduct with a low and reasonably constant voltage drop across the device. When the gap carries maximum current, the voltage across it is typically 10 to 20 V. With increasing current through the gap, the arc channel increases in area and the voltage drop across the arc remains about the same.

If spark gaps are used on power circuits, it is important to provide external means for extinguishing the arc by removing the applied voltage in some manner. This can be done by interrupting devices, such as circuit breakers or fuses, or by inserting resistance rapidly into the circuit by an additional element such as Thyrite or a gas-blast de-ionizer. By suitable design, spark gaps can be made self-extinguishing. Such self-extinguishing properties may make use of the magnetic blow-out principle or may employ other means for achieving the same effect.

2. Gas Tubes

Neon, argon, krypton, xenon, and other gases that ionize at low pressure are often employed as dielectrics in low voltage spark gaps. Such devices can be used as surge suppressors.

3. Semiconductor Devices

Crowbar devices are semiconductors, such as zener diodes and silicon-controlled rectifiers (SCR's), which conduct abruptly upon avalanche breakdown or upon triggering. Since the impedance of these devices reduces to a very low value when they conduct, it is often necessary to add a series impedance in which surge energy can be dissipated and which, at the same time, limits the magnitude of power-follow current. A nonlinear resistor (varistor) such as silicon-carbide is very effective for this purpose. This material is used in certain commercial surge arresters.

4. Voltage Clippers and Nonlinear Resistance Devices

Voltage clippers, as suggested by the name, limit the circuit voltage to a specified threshold value. Generally, voltage limiting is achieved by lowering the impedance of the limiter in proportion to the voltage rise. As a result, the corresponding current causes about the same (or a very slightly increased) voltage drop through the surge-voltage impedance.

The effectiveness of voltage limiters depends upon the ratio of their impedance under overvoltage conditions to the impedance of the surge source. The performance of these devices is best revealed by plotting current versus voltage. These characteristic curves exhibit either a knee or a curvature, as shown in Figure 3.26, in contrast to the straight line characteristic of a linear resistor. Under steady-state conditions, voltage clippers utilizing nonlinear resistance elements draw a small leakage current. When, for an initial steady-state voltage condition, an increasing amount of surge current flows through the device, the voltage rises gradually across it and across the

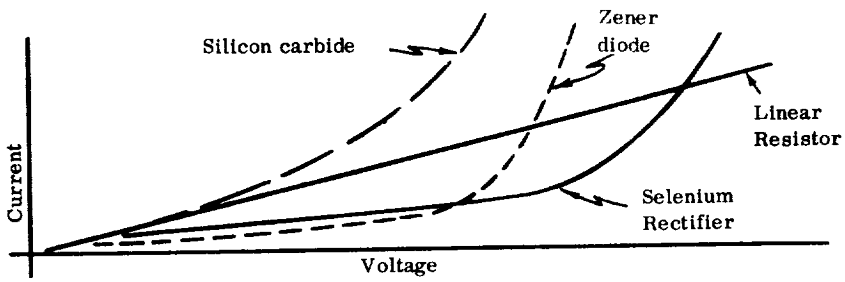


Figure 3.26.—Performance characteristics of typical voltage clippers compared to that of a linear resistor.

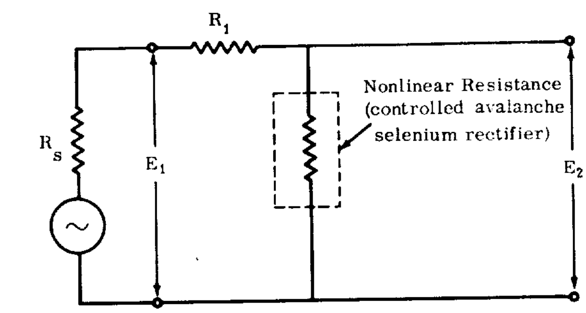


Figure 3.27.—Method of using a nonlinear resistor as a means of limiting the output voltage when the input voltage is subject to rapid rise.

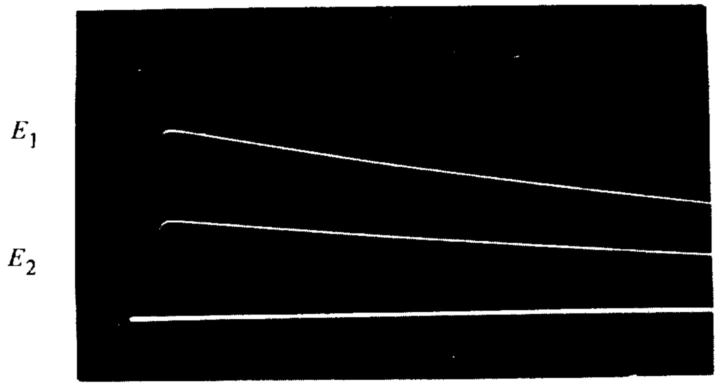
shunt-connected system to be protected, as shown by the characteristic curve. It is apparent that, under surge conditions, the voltage will increase a significant percentage above the steady-state level before reaching a limiting value.

Figures 3.27 and 3.28 demonstrate the action of a selenium rectifier in reducing surge voltage. For the case shown in Figure 3.27 the feed-through voltage, E_2, is reduced to a value less than 10 percent of the input voltage, E_1. For an applied surge voltage $E_1 = 600$ V, the feed-through voltage would be held to a value $E_2 = 0.097 \times 600 = 58.2$ V [see Figure 3.28(b)].

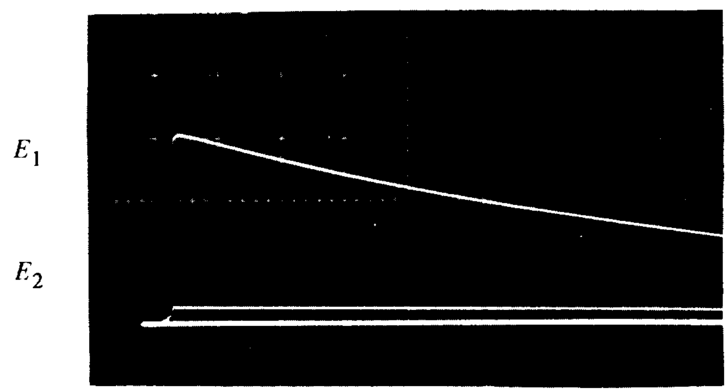
An example of a nonlinear protective device is Thyrite*, which is a nonlinear resistance element (Reference 21, pp. 12-5 through 12-7). The relationship between the current I, which flows through the device, and the applied voltage E across it is

$$E = (R_o I)^{1/n}, \tag{3.29}$$

*Registered General Electric Company trademark.



(a) Input voltage E_1 and output voltage E_2 without selenium rectifier. For conditions shown, $E_2 = 0.525\ E_1$, due to resistive voltage divider formed by source and rectifier impedances.



(b) Input voltage E_1 and output voltage E_2 with selenium rectifier. Rectifier maintains output voltage to a value $E_2 = 0.097\ E_1$.

Figure 3.28.–Effect of selenium rectifier in reducing surge voltage. For both oscillograms, sensitivity is 200 V/cm, and sweep rate is 5 μs/cm (Reference 20).

where $n > 1$ for Thyrite. Usually, n is approximately equal to 3.5, but n can equal 7 for higher resistivity Thyrite. For a common resistor, the constant R_o is the resistance in ohms and $n = 1$. Ordinarily, R_o is the initial resistance, in ohms, when voltage is zero (or very small). However, Thyrite has the property that the apparent resistance decreases as the applied voltage increases. Therefore, when the surge voltage from a lightning stroke appears across a Thyrite element in a circuit such as that indicated in Figure 3.27, it is clipped in magnitude because of the decreasing resistance of the Thyrite. The sum of R_s and R_1 should be very large compared with the maximum value of the nonlinear resistance of the Thyrite element. Typical Thyrite elements can handle large peak surge currents. The level to which an overvoltage is held is within the range of 10 to 20 percent above the peak value of the steady-state circuit voltage. A typical application is in 60-Hz primary power lines.

The rating of voltage limiters is directly related to their capability for storing or rapidly dissipating thermal energy, since clippers actually convert the surge energy into heat. On the other hand, as soon as the surge current decays and vanishes, voltage limiters recover their normal impedance so that a minimum of disturbance is introduced after the surge; their presence in the circuit does not interfere with normal circuit operations.

Typical devices in this category include silicon carbide (Thyrite) resistors, selenium rectifiers, and zener diodes.

5. Energy Storage Elements

Capacitors and ordinary resistors may be used individually or in combination across equipment electrical terminals to provide protective devices that function on the basis of stored energy.

Capacitors suppress overvoltage surges by temporarily storing surge energy which is gradually dissipated through the impedance of the equipment shunted by the capacitors; also, the inherent distributed capacitance of long cables feeding equipment can cause a progressive increase in the rise time of the leading edge of surge wave fronts reaching the equipment.

Linear resistors are often employed across highly inductive dc circuits. These dissipate the magnetic energy stored in inductive elements.

Combinations of capacitors and resistors across the equipment are also helpful. Surge energy is momentarily stored in the capacitors but is dissipated at a decay rate determined by the effective RC time constant. Such combinations are sometimes used to prevent excessive voltages at open switches accompanying surge reflection.

4.5 SAFETY OF PERSONNEL

The importance of personnel safety cannot be overemphasized. All operating and maintenance personnel, regardless of assignment, should be given instructions as to proper and safe actions during electrical storms. The following precautions should be taken.

(1) Maintenance or repair work on electric circuits should not be permitted during electric storm activity. If such work is mandatory, all wires should be temporarily grounded at the point of repairs and the workmen should wear approved insulated gloves.

(2) Personnel must be prohibited from working on towers and antenna structures during electric storm activity.

(3) Personnel should seek and remain in areas protected from lightning during electric storm activity. Examples of such areas are protected buildings, metal buildings, and hardtop automobiles. In the event of an emergency, shelter could be obtained under large dish antennas.

It is suggested that some means of monitoring electric field strength to establish the existence of dangerous electric storm conditions be obtained. There are several suppliers of this kind of equipment.

Commercially available sferics detectors monitor electric field strength and thereby provide a warning system that can be installed to predict the existence of dangerous electric storms in that area. (Reference 22, pp. 7-11, 7-12, and 7-13).

5. LIGHTNING PROTECTION PRACTICES APPLIED TO NASA STATION SITES

Ideally, from an economical standpoint, effective lightning protection practices should be implemented concurrently with new station construction. For example, ferromagnetic conduit for long underground cable runs, other metal structures, or counterpoises can be installed readily in a new site with minimum cost. On the other hand, it is costly to incorporate these protective features in existing installations.

However, the NASA Data Acquisition Facility (DAF) at Rosman, North Carolina, is a typical example of an existing field station that has been modified at minimum cost to achieve improvement in lightning protection of several orders of magnitude. The lightning protection practices applied at the Rosman site will be described as examples of desirable practices for lightning protection.

5.1 GROUND RESISTIVITY

As might be anticipated, the resistivity of the ground in the vicinity of a station site has a significant influence on the harmful effects of lightning. This is borne out by test data (Figure 3.29) obtained at two different station sites. Test data for the Rosman and Madagascar sites were taken in accordance with techniques outlined in Section 3.2.1. Data for the two installations differ not only in the magnitude of ground resistivity but also in the manner in which the resistivity varies with depth of ground rods. The parameter D in Figure 3.29 is the spacing between ground rods used in making the resistivity measurements. A spacing of D feet between adjacent rods, as shown in Figure 3.29, yields the average earth resistivity also to a depth D in the earth. Thus, D on the abscissa of Figure 3.29 also refers to the depth into the earth.

Since the Rosman station is located in mountainous terrain, lightning current injected into the earth's surface tends to remain fairly close to the surface, thereby creating good opportunity for the lightning current to find its way into the grounded conductors of buried cables in the vicinity. Primarily, this effect arises because ground resistivity increases with the depth to which ground rods are driven into the earth's surface, as shown by the test data of Figure 3.29. It has been determined that the primary cause of equipment burn-out at Rosman is the pickup of lightning currents on long interconnecting cables extending over a large area (see Figure 3.2).

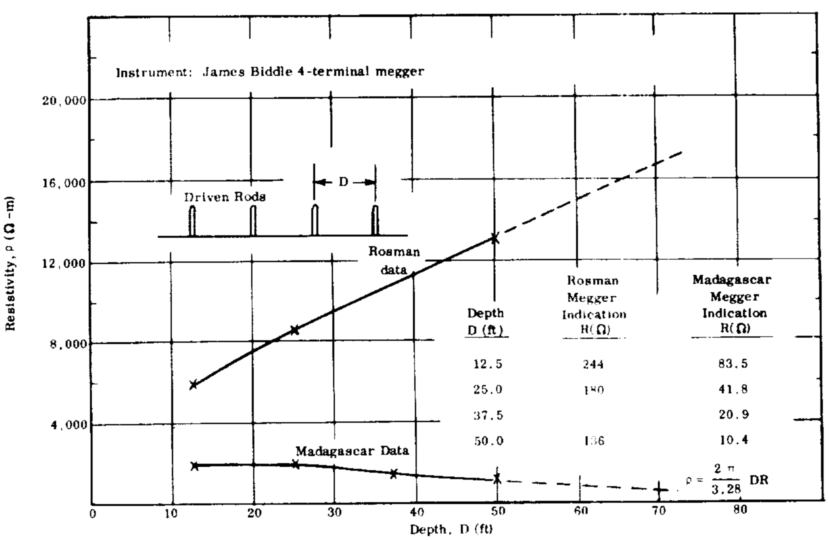


Figure 3.29. Measurements of ground resistivity for ground rods sunk to different depths at two DAF station locations.

On the other hand, the ground resistivity decreases with depth of ground rods for the Madagascar station at Tananarive, Republic of Malagasy. This is an ideal condition from the standpoint of best lightning protection. Therefore, it is to be expected that the Madagascar site should experience much less trouble during lightning activity than the Rosman site. This has, indeed, been the case according to both Madagascar and Rosman station reports.

5.2 LIGHTNING-INDUCED VOLTAGE IN LONG CABLES

Figure 3.30 represents a typical cabling problem wherein a lightning current injected into the system at Location 3 causes a current $I_{R_{g2}}$ to flow through the resistance R_{g2}. In fact, if R_{g2} is small compared to R_{g1}, and if the resistance of the ground conductor between points A and B is small, most of the injected lightning current will flow through R_{g2}. The voltage drop across R_{g2} will also appear at Location 2 and will be added to the magnetically induced voltage, V_{mag}, at that point.

If $V_{R_{g2}}$ is taken to be the voltage drop appearing between point A and ground as a result of the ground resistivity, and if V_{mag} is the voltage induced between points D and A as a result of flux changes in the loop $DCBA$ arising from the lightning stroke, the total voltage appearing between point D and ground at Location 2 is

$$V_{tot} = V_{mag} + V_{R_{g2}} = V_{mag} + R_{g2} I_{R_{g2}}, \tag{3.30}$$

where $I_{R_{g2}}$ is the current in the lightning stroke. For the purpose of simplifying computations, it is convenient to deal with voltages normalized for

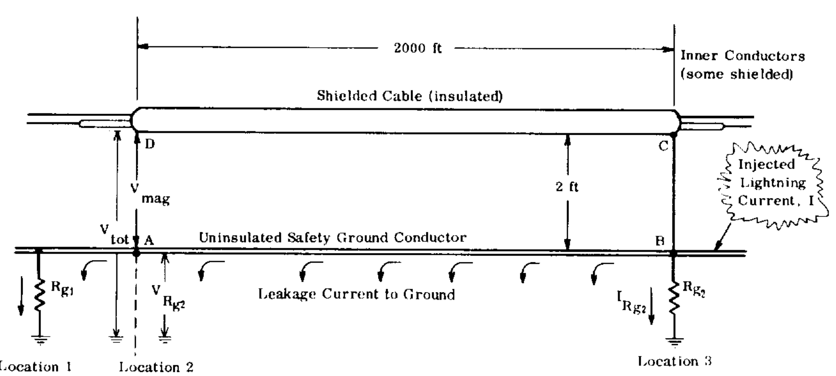


Figure 3.30.—Diagram for analyzing the voltage induced in a long cable as a result of a lightning stroke.

unit current rather than with absolute values of voltage. Typical values of normalized voltage are $V_{R_{g2}}/I = 1$ and $V_{mag}/I = 0.7$, for injected lightning current I. The lightning current may easily have a value $I = 1000$ A, and R_{g2} is typically 1 Ω. When these typical values are substituted into (3.30) the total voltage between points D and ground at Location 2 is found to be $V_{tot} = 1700$ V, for injected lightning current $I = 1000$ A.

5.3 LIGHTNING-INDUCED VOLTAGE BETWEEN EQUIPMENT RACKS

Figure 3.31 illustrates a typical configuration of station equipment cabinets wherein electrical sparking between cabinet racks was reported during lightning activity. As can be seen from Figure 3.31, the left-hand cabinet is bonded to the signal ground plate via a conducting path; a nonconducting loop formed by points $DEFABC$ is similar to the long cable loop $ABCD$ in Figure 3.30. A magnetically induced voltage V_{cab} can appear between points C and D if the cabinets are not electrically bonded together at these points. This voltage is the result of the magnetic fluxes passing over two paths from E to F and from points F to A.

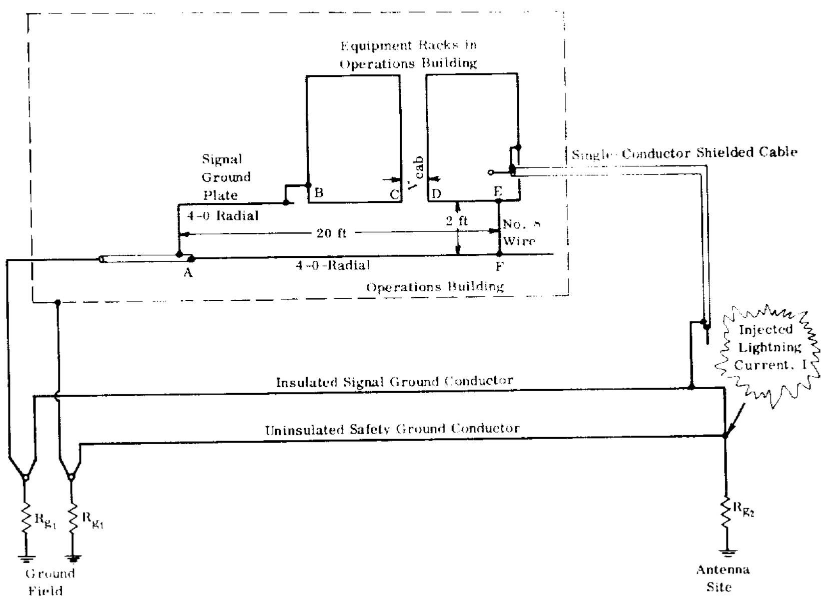


Figure 3.31. Diagram for analyzing the voltage induced between equipment racks as a result of lightning stroke.

The magnetically induced voltage V_1 results from lightning current flowing between points E and F, whereas V_2 is produced by the same current flowing between points F and A, shown in Figure 3.31. Voltages V_1 and V_2 are now computed for a typical example, and then combined in a root-mean-square manner to give a typical value for V_{cab}.

For the purpose of determining V_1, radius r_1 is set equal to 0.065 in. (the radius of a No. 8 wire conductor), radius r_2 equal to 240 in. (20 ft), and height h equal to 2 ft. From (3.17), the induced voltage resulting from the portion of the injected lightning current I flowing between points E and F is

$$\begin{aligned} V_1 &= 61 \times 10^{-9}\ 2 \ln\left(\frac{12 \times 20}{0.065}\right)\frac{dI}{dt} \\ &= (122 \times 10^{-9} \ln 3700)\frac{dI}{dt} \\ &= 10^{-6}\frac{dI}{dt}. \end{aligned}$$

Assuming dI/dt = 1000 A/μs yields

$$V_1 = 10^{-6}\left(\frac{10^3\ \mathrm{A}}{10^{-6}\ \mathrm{s}}\right) = 1000\ \mathrm{V}.$$

However, induced voltage V_2 also contributes to the magnitude of V_{cab}. In determining V_2, which is actually larger than V_1, the radius r_1 is set equal to 0.230 in. (the radius of a size 4-0 wire conductor), radius r_2 equal to 24 in. (2 ft), and height h equal to 20 ft. From (3.17), the induced voltage from the portion of the injected lightning current I flowing between points F and A is

$$\begin{aligned} V_2 &= 61 \times 10^{-9}\ 20 \ln\left(\frac{12 \times 2}{0.230}\right)\frac{dI}{dt} \\ &= (1220 \times 10^{-9} \ln 90.5)\frac{dI}{dt} \\ &= 5.65 \times 10^{-6}\frac{dI}{dt}. \end{aligned}$$

Again, assuming dI/dt = 1000 A/μs yields

$$V_2 = 5.65 \times 10^{-6}\left(\frac{10^3\ \mathrm{A}}{10^{-6}\ \mathrm{s}}\right) = 5670\ \mathrm{V}.$$

Since the magnetic fluxes set up by the current in these two paths are at right angles to each other, the voltages add vectorially, resulting in

$$V_{cab} = (V_1^2 + V_2^2)^{½} = (1000^2 + 5670^2)^{½}$$

$$= 5750 \text{ V}.$$

From this simple model, it is seen that the voltage between equipment cabinets may be appreciable. Hence, it is possible to develop a voltage sufficient to cause sparking between adjacent unbonded cabinets, thereby constituting a hazard to personnel.

5.4 INSTALLATION OF PROTECTIVE DEVICES

The judicious installation of selenium rectifier protective devices can significantly improve a station's vulnerability to lightning. Examples of practices employed at the Rosman station are illustrated in Figures 3.32 through 3.35, inclusive.

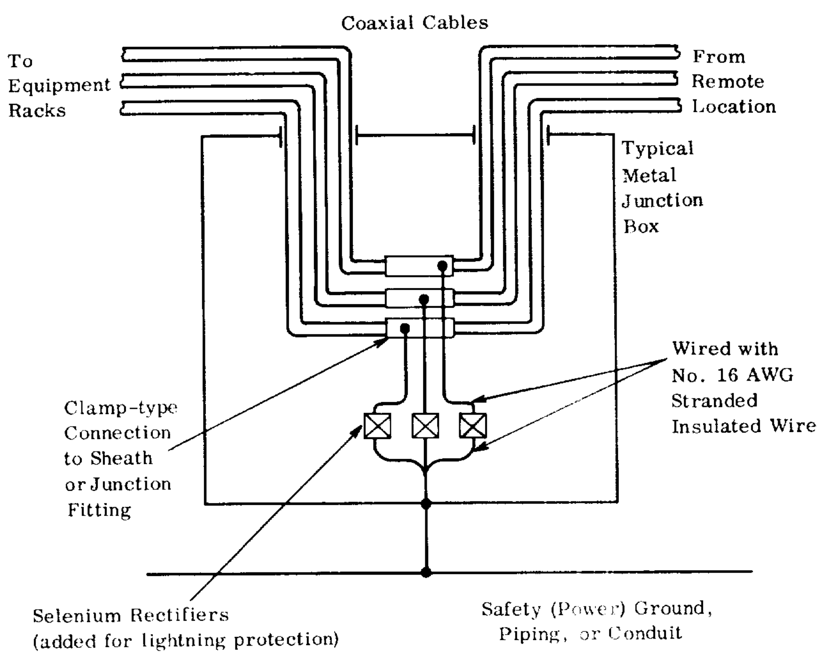


Figure 3.32.--Installation of selenium rectifiers for lightning protection of RF coaxial cable shields.

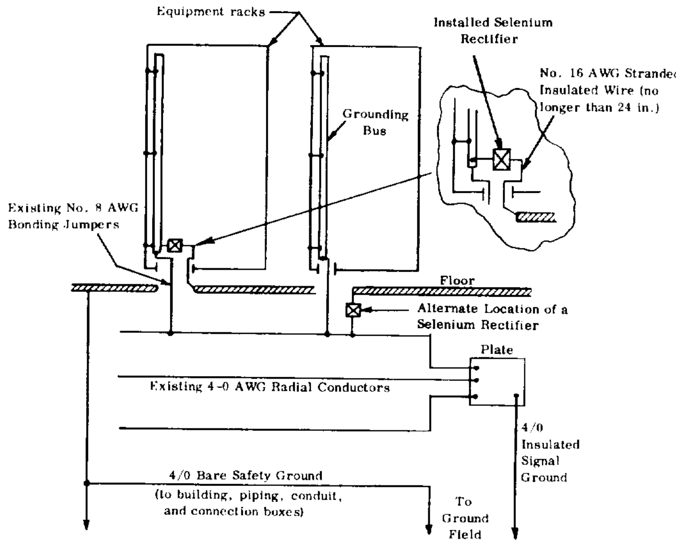


Figure 3.33.–Installation of selenium rectifiers for lightning protection of equipment racks.

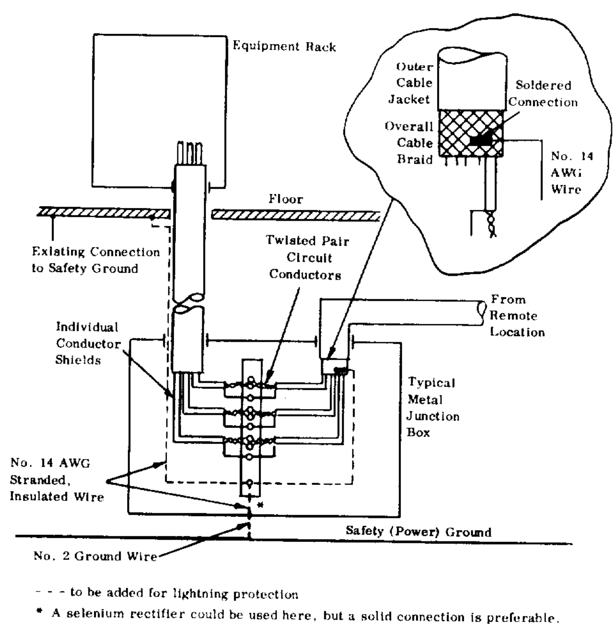


Figure 3.34.–Termination of shielded cables from remote locations for lightning protection.

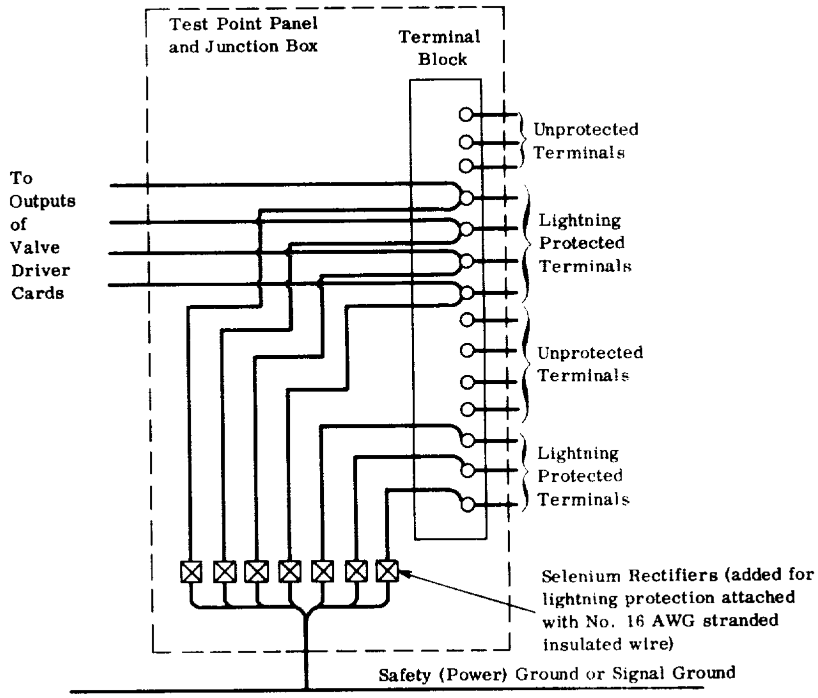


Figure 3.35.—Installation of selenium rectifiers on a test-point panel and junction box in a typical solid-state equipment control console.

The application of selenium rectifiers for protection of RF coaxial cables is illustrated in Figure 3.32. One method for protecting equipment racks against lightning-induced voltages is shown in Figure 3.33. In Figure 3.34, insertion of a selenium rectifier in a ground lead is used for lightning protection in terminating shielded cables from remote locations.

Figure 3.35 illustrates how a typical solid-state equipment can be protected from lightning surges by the installation of selenium rectifiers on critical voltage-sensitive circuit lines. The minimum breakdown voltage of each selenium rectifier should be selected to be slightly greater than the peak operating voltage for a given line.

6. LIGHTNING PROTECTION FOR NASA MINITRACK STATIONS

The 136-MHz Minitrack Interferometer system (Reference 23) has a unique lightning protection problem. The NASA installation at Fort Myers, Florida, continuously reported the burnout of coaxial switches in the antenna.

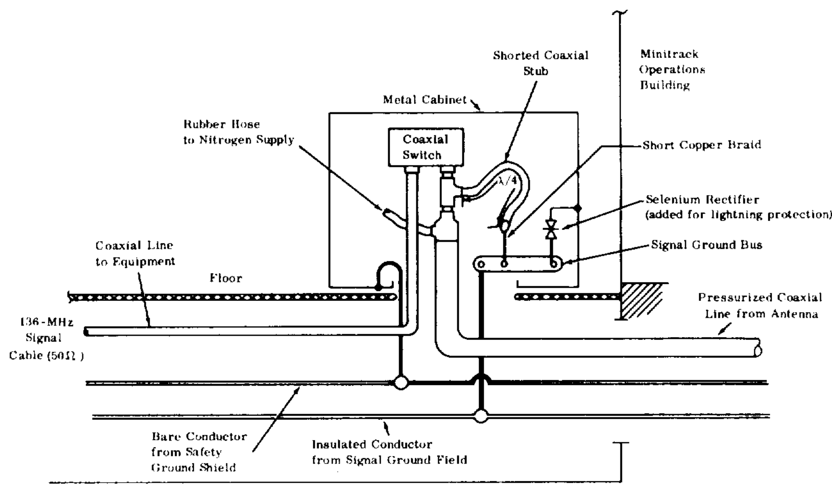


Figure 3.36.–Grounding details of a Minitrack antenna signal cable for lightning protection.

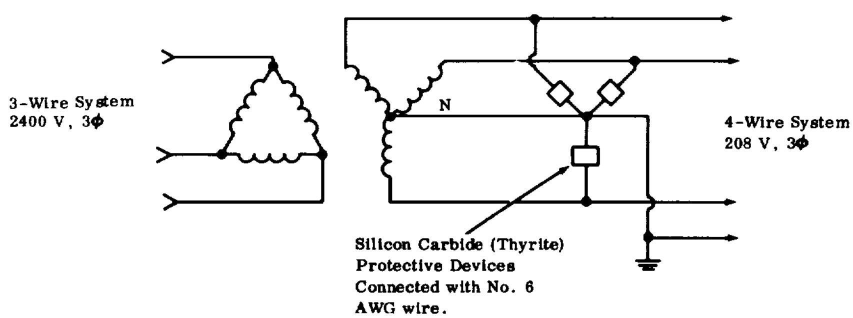


Figure 3.37.–Installation of silicon-carbide protective devices on secondaries of pad-mounted transformers for lightning protection.

An analysis (Reference 20) revealed that the coaxial switches could be protected by the use of a shorted coaxial stub, one-quarter-wavelength long, and a selenium rectifier protective device, as illustrated in Figure 3.36. The shorted, one-quarter-wavelength, coaxial stub provides a high parallel impedance to the 136-MHz signal in the 50-Ω coaxial cable; the coaxial stub provides a low-impedance path for the lightning current to flow through the selenium rectifier. Also, the selenium rectifier is effectively connected between the signal ground field line and the safety ground shield line, thereby providing additional lightning protection.

The primary power system was protected with silicon-carbide protective devices as shown in Figure 3.37.

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