HANDBOOK
ON
RADIO FREQUENCY INTERFERENCE

VOLUME 1

FUNDAMENTALS OF ELECTROMAGNETIC INTERFERENCE

FREDERICK RESEARCH CORPORATION
2601 UNIVERSITY BOULEVARD, WEST WHEATON, MARYLAND

HANDBOOK

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Radio Frequency Interference

VOLUME I

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ELECTROMAGNETIC INTERFERENCE

By

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PREFACE

Radio frequency interference has, in recent years, become a threat of the first magnitude to the operation of our military and commercial communication and electronic systems. This threat has received recognition at all levels of government and industry. Numerous large-scale programs have been initiated to study the problem, to conduct measurements, and to devise methods to prevent degradation of systems operations due to radio frequency interference. Over a period of years, a considerable wealth of information on RFI has been built up and this Handbook has been prepared in order to bring the information together in one place.

The Handbook is addressed to system planners, designers, field engineers, and other technical people who must deal with compatibility problems in communication-electronic systems. RFI theory, prediction procedures, measurement techniques, instruments, specifications, and design are covered in the four Handbook volumes, of which this volume is the first of the series. Much of the material is based on the many years of experience of the Frederick Research Corporation and its individual personnel in the RFI field. We are happy to be able to share the results of this experience with others who are dealing with the same complex problems.

While it is obviously impossible to mention here all of the people who have made contributions in the RFI field, we have listed many of these people and their works in an Appendix of Volume IV, and reference credit has been given throughout the Handbook. In particular, I wish to thank the Institute of Radio Engineers and the authors for permission to use material from "Noise Levels in the American Sub-Arctic" by N. C. Gerson, Proc. IRE, August 1950; "Noise Investigations at VLF by the National Bureau of Standards" by W. Q. Crichlow, Proc. IRE, June 1957; "Amplitude Scintillation of Extraterrestrial Radio Waves at Ultra High Frequencies" by H. C. Ko, Proc. IRE, November 1958; "Performance of Some Radio Systems in the Presence of Thermal and Atmospheric Noise" by A. D. Watt, R. M. Coon, E. L. Maxwell, and R. W. Plush, Proc. IRE, December 1958; "Radio Propagation at Frequencies Above 30 Megacycles" by Kenneth Bullington, Proc. IRE, October 1947; and "Some Technical Aspects of Microwave Radiation Hazards," W. W. Mumford, Proc. IRE, February 1961. I wish to thank the Institute of Radio Engineers, the Harvard University Press, and the authors for permission to use material from "The Radio Spectrum of Solar Activity" by G. Swarup, A. Maxwell and A. R. Thompson, Proc. IRE, January 1958, and the book "The Radio Noise Spectrum" published by Harvard University Press. My thanks also to the Armour Research Foundation of the Illinois Institute of Technology, to the United States Army Signal Corps, and to the author for permission to use material from "Fresnel Region Patterns and Gain Corrections of Large Rectangular Antennas" by E. Jacobs, Proceedings of the Fifth Conference on Radio Interference Reduction and Electronic Compatibility, October 1959. Acknowledgment is also made, with thanks, to John Wiley and Sons, Inc., for permission to use material from the book, "Ultra High Frequency Propagation," by Reed and Russell, page 95, Curves of Reflection Coefficients and Phase Shifts Versus Angles of Incidence.

I wish to thank, specifically, Dr. Robert Latorre, Senior Research Engineer, for his original planning and compilation of the overall material; Mr. A. H. Sullivan, Director, Advanced Systems Development, for the direction and editing of the program, with the assistance of Dr. R. A. Doering, Harry Zink, John Wibbe, Leslie Mackrill, and others. Messrs. Lightner, Humbertson, Agee, Taylor, Felty, and Hopkins with the help of others skilled in graphic arts assisted in assembling and reviewing the material. The art work and illustrations were accomplished by Robert Frederick and his staff. Typing and proofing was "well done" by Mrs. Roma Taylor, Mrs. Marjorie Brown, Mrs. Mary Schmidt, and Mrs. Frances Zello.

We earnestly hope that this Handbook will prove useful to all who have responsibility for the successful operation of communication-electronic systems.

Carl L. Frederick, Sr.
Wheaton, Maryland
10 January 1962

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HANDBOOK

ON

RADIO FREQUENCY INTERFERENCE

VOLUME I	Fundamentals of Electromagnetic Interference
VOLUME II	Electromagnetic Interference Prediction and Measurement
VOLUME III	Methods of Electromagnetic Interference-Free Design and Interference Suppression
VOLUME IV	Utilization of the Electromagnetic Spectrum

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INTRODUCTION

The general recognition for many years that the radio frequency spectrum is becoming overcrowded has led to requirements for planning for effective spectrum utilization as a conservation measure. In 1952 the IRE and RTMA Joint Technical Advisory Committee, after a broad study of the problem, prepared a report on Radio Spectrum Conservation. This report stated in the preface that "it has become increasingly clear that the spectrum is a public domain which must be conserved as carefully as if it were farm land, forest preserves, water power or mineral wealth."

Since 1952, the spectrum conservation problem has been compounded by the large increases in transmitter power outputs, by the increasing sensitivity of receivers, and by the rapidly increasing numbers of communications-electronics equipments. Various government agencies have embarked on long-range programs for more effective spectrum utilization. In the Department of Defense, for example, these programs have culminated in the Military Collection Plan for Spectrum Signatures and in the establishment of the Electronic Compatibility Analysis Center. In the area of non-military use of the electromagnetic spectrum, the Federal Communications Commission for many years has operated a radio network to monitor the various radio emissions throughout the spectrum.

Central to all efforts to "save the spectrum" is a knowledge of the radio-frequency electrical characteristics of transmitters and receivers. Such knowledge can come only from a detailed measurement and assessment of the equipment characteristics. From this information, system operation predictions can be made and action can be taken to prevent radio frequency interference (RFI).

The complexity of the problems created by electromagnetic interference has increased to tremendous proportions in recent years. Initially, of course, problems began with the advent of radio communications systems and grew in proportion as those systems became more complex, as they were more widely used, and as the RFI problem was allowed to grow. With the passing of time, more and more electronic equipment has been installed in military and commercial aircraft and seagoing vessels. More recently, missiles and space probes, loaded with sensitive electronic equipments, have further amplified the problem of RFI that now appreciably affects the performance of many systems. Pressures of the times, especially with relation to the general world situation, have given impetus to spectacular and accelerated development of all kinds of electronic equipments. While consideration of the electromagnetic interference problem has not been neglected throughout modern day development of individual electrical and electronic systems, nevertheless the overall problem has increased, and will continue to do so unless intensive RFI preventive measures are taken by all designers and systems users.

Brought about by rapid technological advances, we have adopted a procedure known as "Preventive Maintenance," the sole objective of which is to keep components and systems operational. Preventive Maintenance is obtained through a system of education. Relative to the problem of RFI, we have long needed a procedure that might be called Preventive Design. In such a procedure, we would consider the interference problem as an integral part of the design and educate our engineers to design interference-free systems. We cannot continue to treat this problem as secondary in favor of other considerations. It is a prime problem, and without continuous and even drastic effort to eliminate it we can only look forward to complete chaos. Too much time has been lost already. The problem tends to bog down our progress. We already have a colossal job in determining and employing methods of RFI suppression for existing systems. It is mandatory that we not compound the problem further. From the most insignificant electrical component and assembly to the most involved prime electrical or electronic system, each item must be given critical attention to design out of it the tendency to admit or to emit RFI. The scourge of electromagnetic interference invades all fields including navigation, guidance, control, and communications. Because of it, navigation systems fail, missiles veer off course, and control systems do not control according to their functional design. Communications become unintelligible or fail altogether. Even firing squibs detonate prematurely, destroying expensive missiles and creating a hazard to life. The seriousness of the problem we have allowed to grow about us cannot be overemphasized, and it is the responsibility of each one of us concerned in the field of electronics design and usage to wage war against the problem with unrelenting effort. We cannot live with it. We must eliminate it. No RFI source can be ignored.

The RFI problem has been attacked from many different angles by the Armed Forces, aircraft and equipment designers, and manufacturers. Electronic engineers, both military and civilian, have put much effort and attention to this subject in recent years. As a result, many papers on the subject prepared by individuals and societies have appeared in technical journals. Several technical manuals have been prepared by each of the military services. All these have been helpful and have emphasized the pressing need for an up-to-date manual, comprehensive in its treatment of the various aspects of the problem.

The Frederick Research Corporation has for many years been actively concerned in the field of electromagnetic interference. In publishing these volumes, it is hoped as a result of sharing this experience to impart an overall knowledge of the many facets of the RFI problem, how to measure interference, how to suppress existing interference sources, how to design interference tendencies out of new designs, all leading to a single objective - eventual elimination of the problem.

It is too often true that measurements and tests for electromagnetic interference are made only under ideal conditions. Compliance with the applicable clause in the equipment specification while ignoring the many adverse conditions that will be encountered in actual operation is often not enough. This may wholly or partially nullify the validity of such tests - theoretical and practical. Too frequently specifications are loose in this regard and facts relative to the environment of operation are not clear. Manufacturers may conscientiously attempt to eliminate or forestall RFI in their own equipment, but in such a way that such precautions do not preclude interference or interaction with various equipments with which their own may be associated operationally. Successful peak performance requires team work all along the line including the original designer, the installation personnel, and the maintenance personnel. The very remedies instigated for certain specific interference problems have sometimes actually increased the interference and have caused new interference in other equipment. This often results from "fixes" being undertaken without a thorough understanding of the basic theory underlying the attempted fix.

It is with these problems in mind that these volumes are written. It is of paramount importance to impress every designer with the concept that designing for correct functional performance is only half the job. He must be sure that side effects (unwanted signals) are not generated or accepted, thus destroying the usefulness of the equipment in the environment in which it is used.

"Radio interference" originates in many sources but affects us and our equipment only insofar as it affects a "receiver." For the purposes of these volumes, "radio interference" is defined as any electrical disturbance which causes an undesirable response or a malfunctioning in communication-electronic equipments or systems. The definition of the word "receiver" is extended in these volumes to include all types of electrical and electronic equipment in which radio interference may cause an undesirable response or malfunctioning. Thus, not only an earphone, but a navigation instrument or a radar scope, may be called a receiver, and even a relay, which may be tripped by a spurious signal, will fall under the heading.

There are three kinds of electromagnetic interference:

Man-Made: This includes interference from sources such as electrical motors and other electronic systems that radiate spurious signals.

Inherent: This is interference generated within the receiver due to thermal agitation, shot effect, and similar causes.

Natural: This interference is caused by atmospheric electrical disturbances and precipitation static.

Although the system designer and the system operator have different functional responsibilities, they have a common interest in electromagnetic interference. To meet their needs as well as the needs of system planners, we have divided the material into four volumes:

Volume I: Fundamentals of Electromagnetic Interference

Volume II: Electromagnetic Interference Prediction and Measurement

Volume III: Methods of Electromagnetic Interference Suppression and Interference-Free Design

Volume IV: Utilization of the Electromagnetic Spectrum

The first volume discusses the fundamental aspects of electromagnetic interference as the basis for prediction, measurement, and design. Included are both theoretical and practical approaches to system compatibility and interference reduction. The succeeding volumes discuss these various technical areas in detail.

1. NATURE OF INTERFERENCE

In considering the basic effect of the interference, i.e., its actual nuisance value, the most important factors are its magnitude in relation to that of the wanted signal and its position in the frequency spectrum. The nuisance value of an interfering signal varies directly as its magnitude, provided that it exceeds a certain minimum threshold value. This will remain approximately true even if the magnitude is measured not at the output but at the source or anywhere in the path of transmission of the interference from the source to the receiver. The transmission system is usually sufficiently linear to make the signal strength at any one point approximately directly proportional to the strength of the same signal anywhere else in the system.

As far as the position of the interfering signal in the frequency spectrum is concerned, no such simple statement can be made. For the final effect, the interference must either contain frequencies within the normal output range of the receiver, or it must be capable of making one or more stages of the receiver inoperative. In the first case, the final effect would be audio noise for an ordinary radio receiver, or it might be visual hash on a radar scope. In a navigational instrument, the final effect might be a false indication of an indicating needle. In the second case, the final effect would be the complete lack of an audio or visual indication. In either case, in order to produce this final effect at the output of the receiver, the interference at the input must contain frequencies within the band to which the receiver is sensitive. It is important to remember that the band of frequencies to which the receiver is sensitive is much wider for interference than what is normally considered its "bandwidth." The attenuation of frequencies outside the normal transmission band is never infinite, and very often there is insufficient rejection of large interfering signals, even though their frequencies may be considerably removed from those the receiver is designed to accept. This is one way in which an interfering signal may affect a receiver, even though it has no frequencies within its nominal acceptance band.

There is another way in which an interfering signal, outside of the acceptance band of a receiver, may gain entrance. During transmission, electrical signals may undergo one or several frequency translations, i.e., they may combine with other signals in nonlinear elements to produce entirely new frequencies. For example, the fundamental frequency output of a medium frequency transmitter may not itself fall within a band that is accepted by a low frequency receiver. But one of its higher harmonics may "beat" with the output of another transmitter in a nonlinear element in such a way as to produce an interfering signal which does affect the receiver. Therefore, even though only a fairly narrow band of frequencies is effective at the input of the receiver because of its selectivity, frequencies in that band may be produced by entirely different frequencies at the source, and it becomes necessary to treat signals of all frequencies as having potential nuisance value.

An interfering signal is always associated with a time-varying electric or magnetic field. If the variation of the field is sinusoidal, the resultant signal is completely specified by three quantities: its amplitude, frequency, and phase. More often, the variation will not be sinusoidal. In this case, an infinite number of parameters will, in general, be required for a complete description. This is evident from the following considerations: The time variation of the field intensity may be expanded either as a Fourier series or as a Fourier integral, the series being used when the variation is periodic and the integral when it is not. In the first case, the Fourier series will, in general, contain an infinite number of terms whose magnitudes and phases must be specific for a complete description of the series. In the second case, a complete description of the Fourier integral requires the specification of the amplitude and phase functions at all frequencies. Even though an infinite number of parameters is required for a complete description of the signal, usually only a finite number of terms or a limited range of frequencies need be considered because in all cases of practical importance the amplitudes of the high frequency components become too small to affect the receiver.

The physical interpretation of the above paragraph is that an arbitrary signal may be considered to consist of an infinite number of sinusoidal signals, superimposed. These sinusoidal signals are either of finite amplitude and occupy discrete frequencies, or they have infinitesimal amplitudes and a continuous frequency distribution. Such an analysis, usually called a "Fourier analysis," has the great advantage that if one deals with a linear network, for which the principle of superposition applies, the response of the network to any arbitrary input can be determined on the basis of a knowledge of the response of the network to sinusoidal inputs at all frequencies. This is the main reason why the consideration of nonlinear networks leads to much greater difficulties. The response of a nonlinear network cannot be found from an analysis of its behavior under excitation from sinusoidal sources.

The interfering signal, considered as a varying field, current, or voltage, is determined not only by the process of its generation, but also by the impedance into which the signal generator sends the impulse. Since the form of the resulting response (which is the only quantity that can be observed) is usually of greater interest than the hypothetical form of the generator output in the absence of any impedance, it is important to investigate the effect of various kinds of impedance on the form of any given impulse. To carry through such an investigation for all possible, or even all practically important, forms of impulses, is a prohibitive undertaking. A considerable insight into the problem may be gained, however, by concentrating on just one representative wave form. The particular form chosen is that of a rectangular pulse of unit area and of short but finite duration, because this form is often fairly closely approximated by actual interference pulses. Furthermore, a signal of any shape can be closely approximated by a succession of such pulses.

Considerable work is saved by utilization of the principle of duality. This principle states that, for certain network pairs, identical relations exist between the voltages, currents, impedances, and admittances, provided that all voltages, currents, impedances, and admittances in one network are replaced by the currents, voltages, admittances, and impedances, respectively, in the other. Network pairs which obey such laws are called duals of each other. Any planar network (i. e., one which can be projected onto a plane without any connections crossing one another) has a dual. The dual may be obtained by putting all series elements of one, in parallel in the other, and all parallel elements of one, in series in the other, by leaving all resistances unchanged, by replacing all inductances with capacitances, and by replacing all capacitances with inductances. A similar principle of duality may be stated for fields instead of networks.

A distinction is sometimes made between network pairs having dual configuration and actual duals. In the first case, it is necessary only that the above-mentioned relations exist between types of elements without regard to their magnitude. For actual duals, there must also be a relationship between the magnitudes of corresponding elements, as shown in Figure 1-1, which gives an example of two networks that are actual duals of each other. Here k is an arbitrary real constant.

By utilizing the principle of duality, it is found that the response of one network to a voltage impulse is the same as the response of its dual to the corresponding current impulse. For example, the current that results from applying a specified voltage to a series combination of resistance and capacitance has exactly the same form as the voltage that appears across a parallel combination of resistance and inductance when

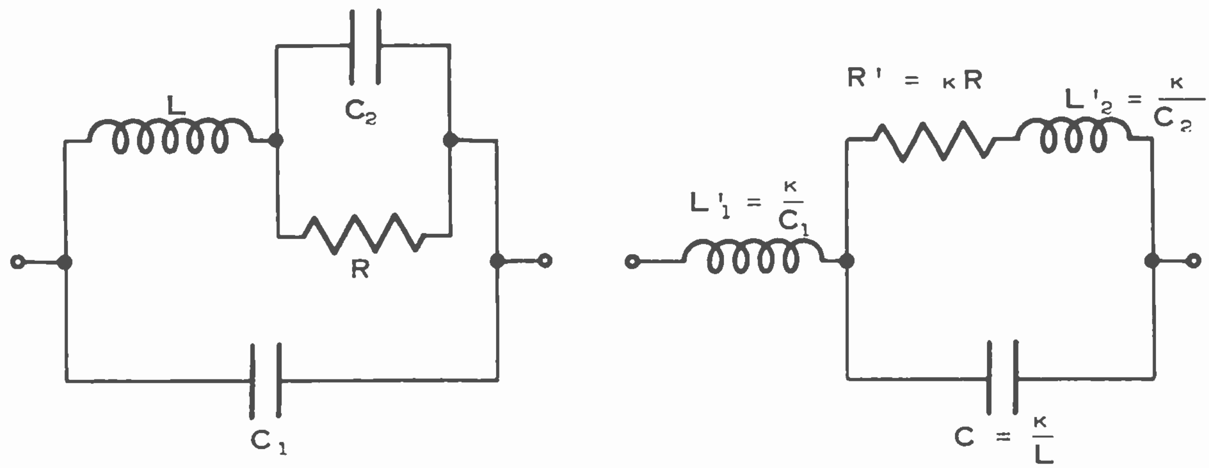


Figure 1-1. Dual Networks

a current of the same specified time variation is sent into it. In this way, each separate analysis immediately yields two significant results.

Figure 1-2 shows the responses of the simplest combinations of one, two, or three circuit elements to a short rectangular pulse. In each case the response shown is the current that will flow if a rectangular voltage pulse is applied across an impedance, Z, or the voltage across an admittance, Y, when a rectangular current pulse is flowing into Y. If the network Z consists of several branches in parallel, or the network Y has several elements in series, the result is simply a combination of the results shown, since the current in each branch or the voltage across each element may be found from Figure 1-2.

The actual interfering signal usually consists of a series of periodic or nonperiodic pulses similar to those shown in Figure 1-2. The responses will remain substantially as shown whenever the individual pulses occur so far apart that the energy of each pulse is practically dissipated before the beginning of the next one. (The cases without resistance, and hence without dissipation, need not be considered here since they cannot be realized in practice.) If, however, the pulses follow each other so closely that the initial conditions of each pulse are affected by the previous one, then the transient caused by one pulse is superimposed on the transient of the previous pulse, which did not have time to die out, and the resulting wave form may have very little resemblance to the response shown for a single pulse. In the case of a resonant circuit, for example, the output wave will deviate more and more

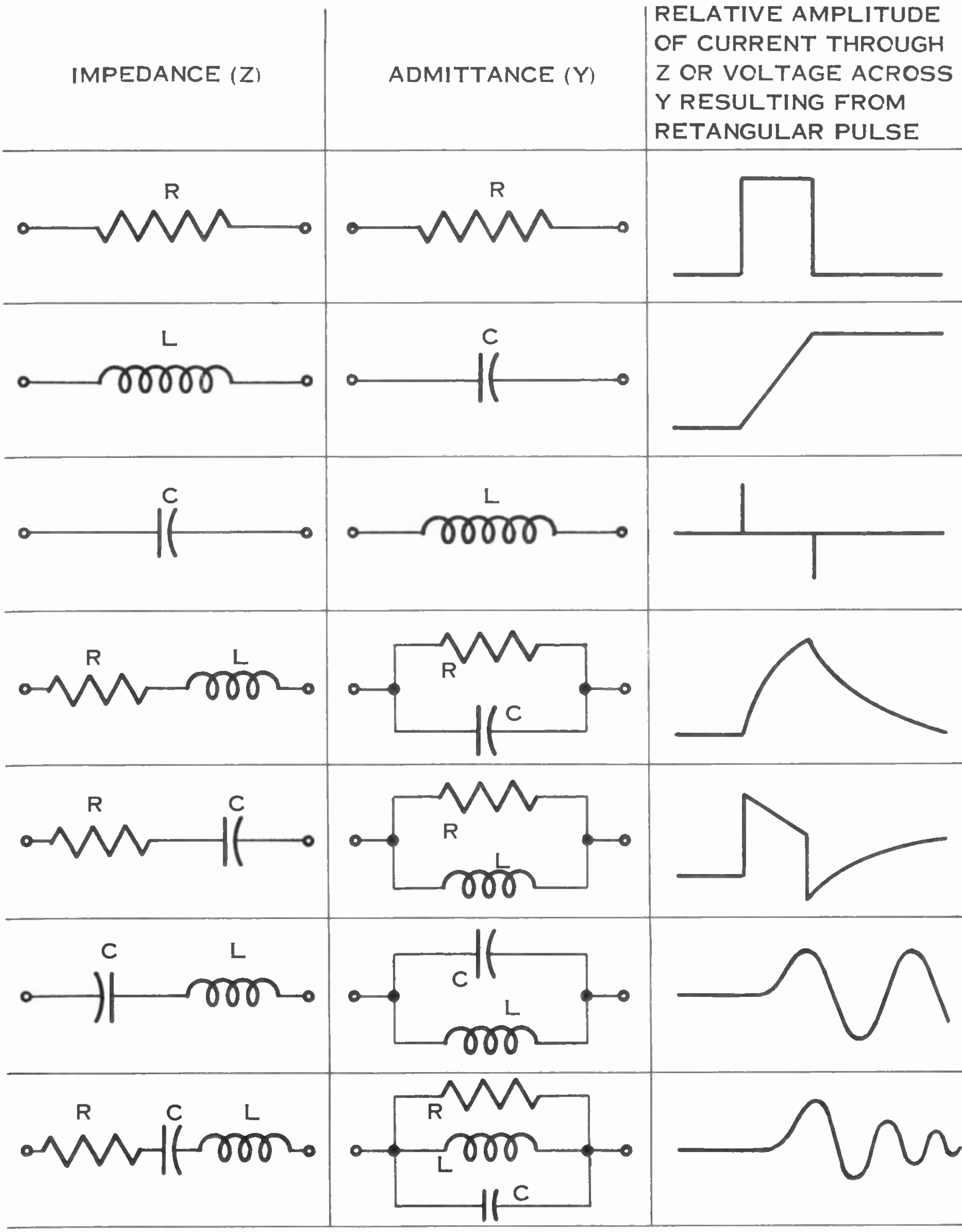


Figure 1-2. Response of Single Networks to Rectangular Pulse

from the damped sine wave shown in the last diagram of Figure 1-2 as the ratio of the resonant frequency to the pulse repetition rate becomes smaller.

The results of Figure 1-2 may be obtained in two ways: (a) determine what happens in the various elements as voltage (or current) is suddenly applied to them, or (b) consider the response of the network to sinusoidal excitations. In the first method, attention is focused on the way in which a capacitance stores charges, an inductance introduces inertia effects, and a resistance causes dissipation of energy. In the second method, a Fourier analysis of the rectangular pulse must be performed. This latter point of view shows that, as the frequency of the applied sinusoidal voltage increases, the current in a capacitive circuit will increase, the current in an inductive circuit will decrease, and the current in a resonant circuit will increase at first, reach a maximum at the resonant frequency, and then decrease. If the frequency spectrum of the exciting pulse is known, the frequency spectrum of the response may then be predicted, and the response itself may be determined.

For simple circuits, the first method is not only sufficient, but actually more enlightening, because it is easier to understand what actually goes on in the network. But for more complicated networks and general analyses, the second approach is indispensable. An analysis of this type proceeds as follows:

An arbitrary signal, which may bear no resemblance to the rectangular pulse used as an example before, is given as a function of time, $f = f(t)$. Also, there is given a network whose response to sinusoidal excitation is specified at all frequencies, in the form of a complex function $G = G(\omega)$ where ω is the angular frequency in radians per second. The task at hand is to determine the response of the network. The procedure is to find the Fourier transform of $f(t)$, designated as $F(\omega)$:

$$* \; F(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} f(t) \; e^{-j\omega t} \; dt \tag{1-1}$$

---

* The form $F'(\omega) = (1/2\pi) \int_{-\infty}^{\infty} f(t) \; e^{j\omega t} \, dt$ can also be used as a Fourier transform. It will be recognized that F' is simply the complex conjugate of F. The form used in the text is preferred in most of the literature.

and to obtain the Fourier transform of the response simply as the product of GF. If the response is indicated by h = h(t) and its Fourier transform by H = H(ω), then

$$H(\omega) = G(\omega) \; F(\omega) \tag{1-2}$$

and

$$h(t) = \int_{-\infty}^{\infty} H(\omega) \; e^{j\omega t} \, d\omega = \int_{-\infty}^{\infty} G(\omega) \, F(\omega) \, e^{j\omega t} \, d\omega \tag{1-3}$$

The square of the absolute value of F(ω), $|F(\omega)|^2$, is proportional to that portion of the energy of the signal f(t) that is associated with the angular frequency ω. Therefore, $|F(\omega)|^2$ will simply be called the energy distribution of f(t).

Listed below are certain relationships between the properties of the function f(t) and its frequency distribution F(ω), which are important for the analysis of interference.

a. Reciprocal Spreading

For a pulse of finite duration, the spread in frequency is roughly inversely proportional to its duration. Examples of this are shown in Figure 1-3.

b. Effect of Steep Wave Fronts

For any pulse, the sharper its rise or fall, the greater is that portion of its energy which is concentrated in the high frequency components. This effect leads to the conclusion that only those pulses which have sharp wave fronts will cause appreciable interference at high frequencies. An example of this is shown in Figure 1-4. The computations for these, and the following examples, are given in detail in Appendix I.

c. Effect of Duration of Pulse

For any pulse, the longer its duration, the greater is that portion of its energy which is concentrated in its low frequency components. An example of this is shown in Figure 1-4.

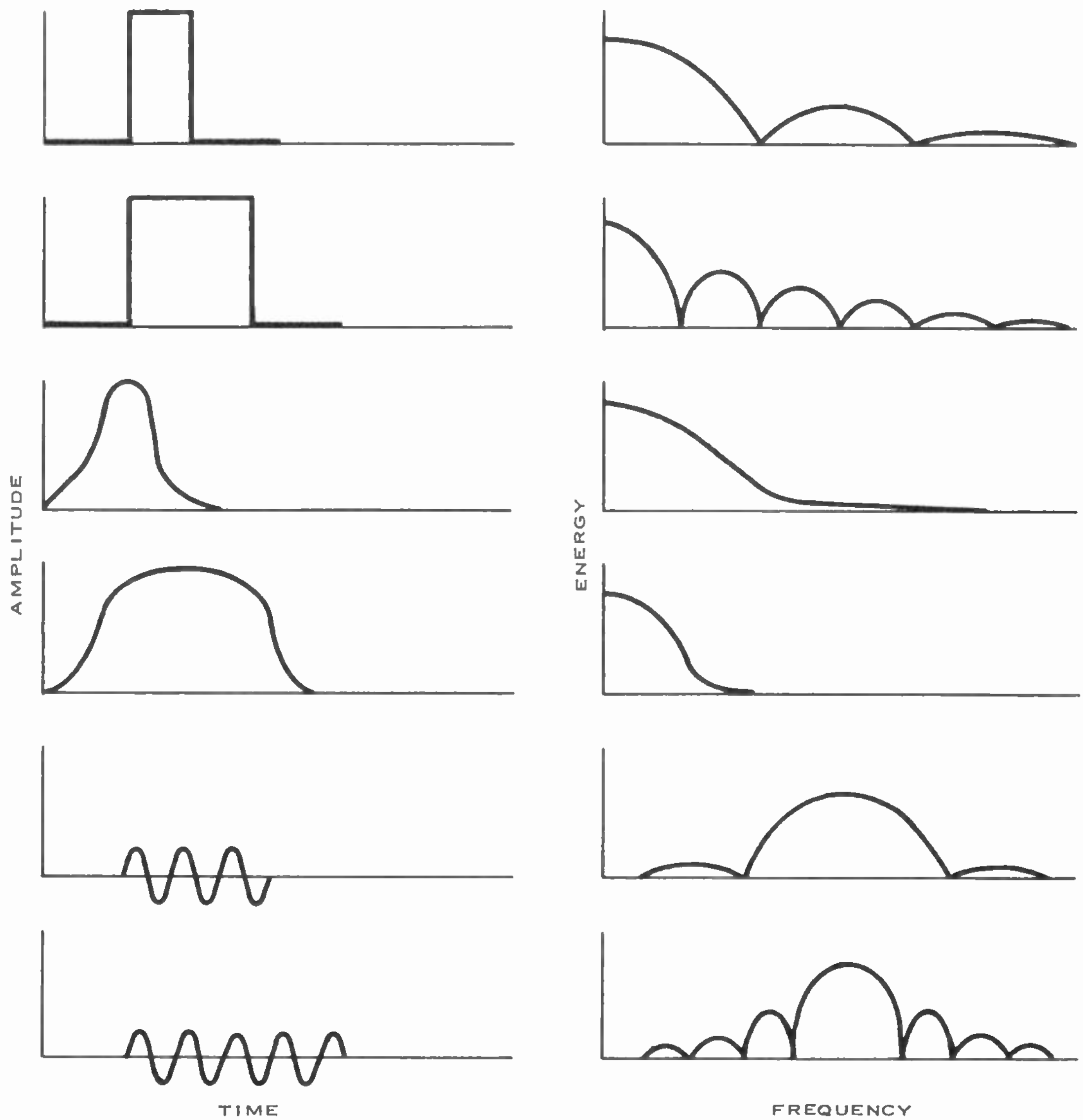


Figure 1-3. Energy Distributions Showing "Reciprocal Spreading"

These rules are qualitative rather than quantitative, since it may be impossible to assign exact values to the "duration" of a pulse which decays to zero exponentially, or to the "spread" of a function that goes to zero like a damped sine wave.

The method of Fourier analysis outlined above is extremely powerful and finds many practical applications. In this volume it will be used (1) to help formulate the impedance approach in Appendix II, (2) to predict the response of receivers to the most common types of

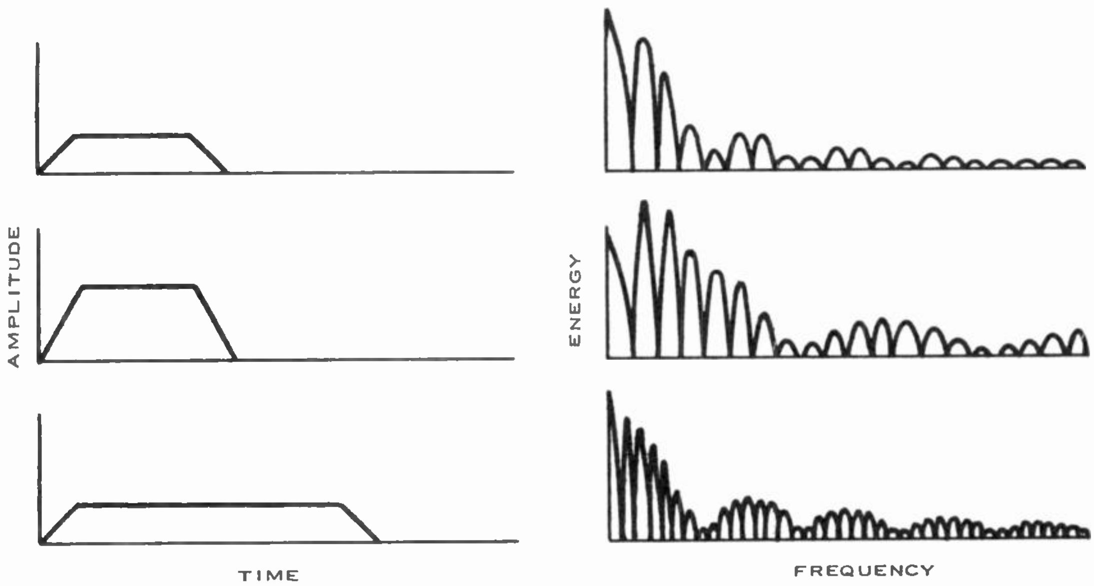


Figure 1-4. Energy Distributions Showing Effects of Steepness of Rise and Duration of Pulse

interference, and (3) to analyze the changes which the interfering signals may be expected to undergo during their transmission from the source to the receiver. For example, the Fourier analysis of the interference generated by a particular motor, may reveal a strong peak in the vicinity of, say, 2 megacycles per second, as shown in Figure 1-5. It will then be necessary to design a filter whose attenuation characteristic has somewhat the same shape as is shown in Figure 1-6. This could be accomplished by use of a series derived M-type filter giving an attenuation as shown in Figure 1-6, curve (a). A low pass filter, of the constant K-type, would probably not be sufficient to suppress the interference in the neighborhood of 2 megacycles, as shown in Figure 1-6, curve (b). Information relative to design of filters is discussed in Volume III.

2. NONLINEAR IMPEDANCES

Impedances whose magnitude depends upon the currents or voltages impressed are considered nonlinear. These impedances are in effect equivalent generators of electromotive forces, since they always give rise to harmonics of the conduction frequency. The situation is complicated further when the impedance is also a function of the time rate of change of current, as in nonlinear inductors such as swinging chokes. Many switching components such as diodes and vacuum tubes may be considered as nonlinear impedances.

In general, filtering and shielding are required to suppress the undesired frequency components generated by nonlinear impedances.

Impedances which vary in time cause varying currents, as will be discussed in the next paragraph. But impedances that vary with the current through them or with the voltage applied across them (so-called "nonlinear" impedances) act like equivalent generators of varying electromotive forces. This may be shown as follows. Consider the circuit of Figure 1-7. The generated voltage of the generator is $e_0 = E_0 \sin \omega t$, and the nonlinear impedance Z is assumed to be a function of the current i, $Z = f(i)$, which can be developed into a Taylor series:

$$Z = Z_0 + A_1 i + A_2 i^2 + \ldots \tag{1-4}$$

where the coefficients Z_0, A_1, A_2, etc., are assumed to be known. Then the current is

$$i = \frac{e_o}{Z} = \frac{E_0 \sin \omega t}{Z_0 + A_1 i + A_2 i^2 + \ldots} \tag{1-5}$$

This implicit equation for i may be solved explicitly by the method of successive approximations if the assumption is made that A_1 and A_2 are small and all higher terms are negligible. Denoting successive approximations to i by i_0, i_1, etc., one has:

$$i_0 = \frac{E_0}{Z_0} \sin \omega t \tag{1-6}$$

$$i_1 = \frac{E_0 \sin \omega t}{Z_0 + A_1\left(\frac{E_0}{Z_0}\right) \sin \omega t + A_2\left(\frac{E_0}{Z_0}\right)^2 \sin^2 \omega t} \tag{1-7}$$

$$\simeq \frac{E_0}{Z_0} \sin \omega t \left(1 - A_1 \frac{E_0 \sin \omega t}{Z_0^2} - A_2 \frac{E_0^2 \sin^2 \omega t}{Z_0^3}\right) \tag{1-8}$$

It is not necessary to go any further in order to see that the second term in Equation (1-4) gives rise to a term that varies as $\sin^2 \omega t = (1/2)(1 - \cos 2 \omega t)$, and that the third term in Equation (1-4) gives rise to a term that varies as $\sin^3 \omega t = (1/4)(3 \sin \omega t - \sin \omega t)$. Therefore, the current will contain a second and third harmonic that were not present in the original source. It is possible, in general, to replace any nonlinear impedance by a linear impedance in series with one or more generators whose frequencies are harmonics of the actual source frequency. Thus, the cir-

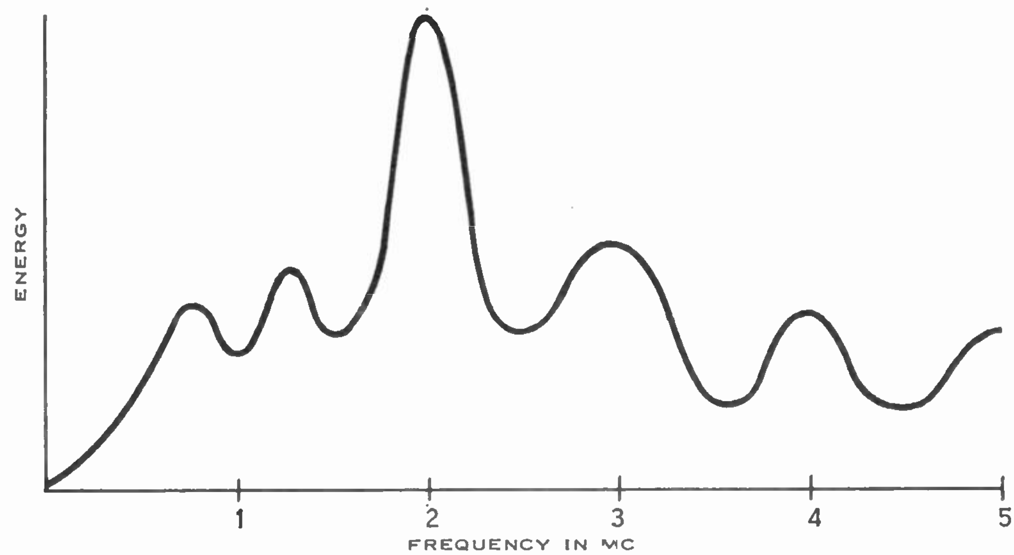


Figure 1-5. Typical Frequency Spectrum of Motor Interference

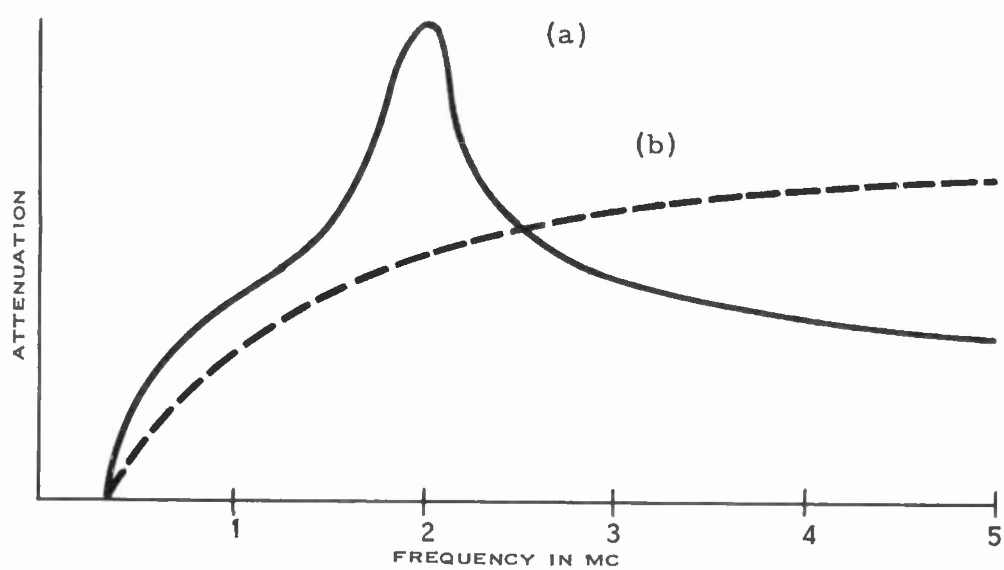


Figure 1-6. Attenuation of Filter for use with Motor of Figure 1-5

cuit of Figure 1-7 which contains a nonlinear impedance, is equivalent to the circuit shown in Figure 1-8, which contains only linear impedances, but has additional signal generators.

Still greater complications arise in cases where the impedance Z is a function of both the current i and its rate of change di/dt. This happens, for example, in a nonlinear inductance. While the mathematics involved are more complicated, the net results are essentially the same. It is still possible to replace the nonlinear impedance by a linear one in series with a number of harmonic generators.

Thus, it is seen that any nonlinear impedance is a possible source of interference.

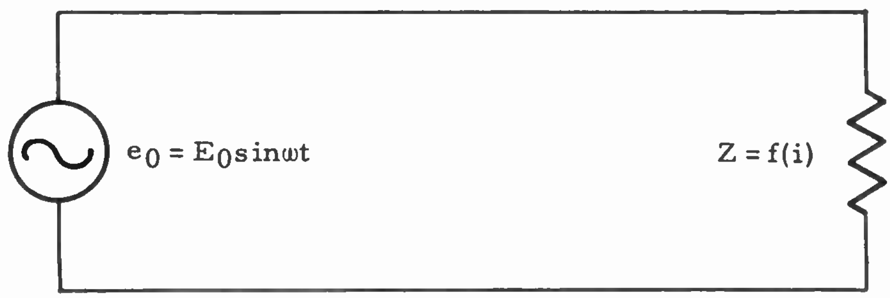


Figure 1-7. Circuit Containing a Nonlinear Impedance

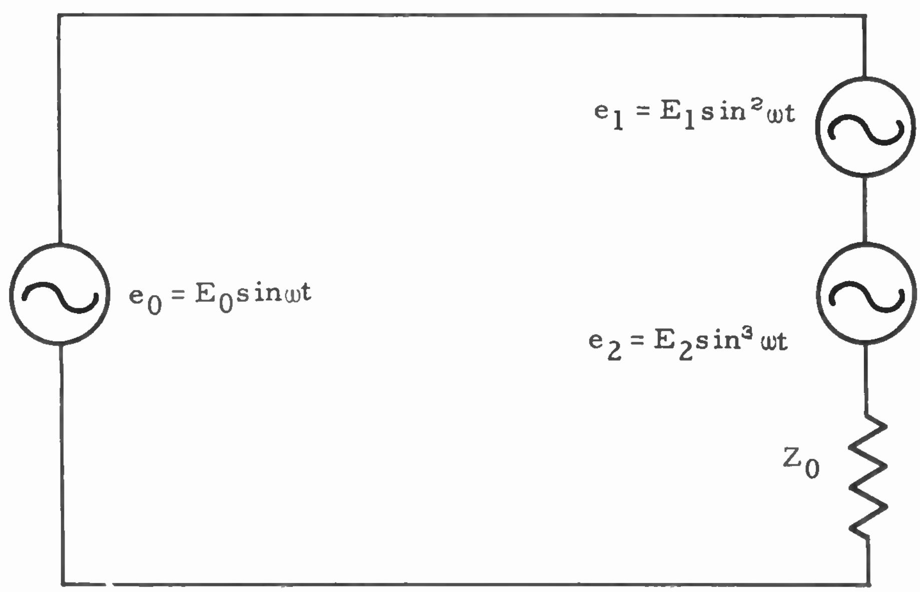


Figure 1-8. Circuit with Linear Impedance Equivalent to the Circuit of Figure 1-7

3. ELECTRICAL OSCILLATIONS IN STATIONARY CIRCUITS

Electromagnetic interference generated by the operation of surface based and airborne electronic equipment generally originates in stationary resonant circuits. These include local oscillators, transmitters, and modulators, which perform functions essential to the operation of both radio and radar equipments. In the case of aircraft, for instance, the stationary sources of interference, like the motional sources, are scattered throughout the aircraft in such fashion that each compartment and wing section may contain several pieces of electronic equipment. It is then possible that any offending electronic component may be mounted sufficiently close to susceptible components, or component wiring to introduce radio interference into these units and adversely affect their operation. The same circumstance can apply to surface based systems. Consequently, each electronic component must be designed so that electrical oscillations are confined to the unit itself and not permitted to enter other electronic components either directly or indirectly thus causing interference.

Oscillations may occur in electrical systems whenever the passive circuit elements are connected so that electrical energy may be exchanged periodically between them. Such oscillations or resonant conditions can be excited by any appropriate frequency component of the applied signal. Oscillatory action of this nature would be self-sustaining if the resonant circuit contained no resistance. However, since every practical circuit contains resistance, the supplied energy is dissipated gradually, and the oscillation is damped. Therefore, the peak values of the current or voltage oscillations will decay exponentially unless the dissipated energy is returned to the system. Circuits containing inductive and capacitive parameters, capable of storing energy, will resonate if properly excited. Spurious oscillations as well as the desired resonant condition may be set up in an electrical system by such action. In any case the sustained oscillations must be either self-excited or externally excited. Any amplifying device, such as a vacuum tube, is capable of generating and sustaining oscillations. Due to the amplifying characteristic of the tube, power available in the output is much greater than the required input power. This permits a certain percentage of the output to be fed back into the input in the same phase as the input energy, and thus supply its own excitation. Any parallel-resonant circuit is an example of external excitation when a constantly varying electromotive force is applied across its terminals. The same type of tuned circuit experiences a series of transient oscillations when subjected to a sudden pulse of energy.

Regeneration is a first requisite for sustained oscillation in any circuit. The amount of regeneration of in-phase feedback must be sufficient to overcome the resistance losses present in the input circuit. Practical oscillators must be self-starting as well as self-sustaining. The self-starting feature is provided by a random fluctuation in voltage due to such causes as, thermal agitation in the circuit and tube, fluctuations in tube current produced by shot effect, and contact differences of potential. These random voltages cover the entire frequency spectrum, but the tuned circuit selects that particular frequency to which it has a natural response and amplifies it, by the regenerative process, so that an oscillating condition is maintained. Undesirable or "parasitic" oscillations result in many circuits when the conditions for resonance are present. A feedback path for sustaining "parasitic" oscillations can be supplied by the grid-to-plate capacitance of an ordinary amplifying tube, the capacitance between adjacent circuit wires, or the distributed capacitance of transformer turns. Inductive coupling between transformers also becomes serious when there is relatively high gain between them. It may be necessary to screen one or both transformers with metal of low reluctance. Proper wiring layout will normally prevent coupling between adjacent wires in components; otherwise shielded leads will be required. Negative feedback may be employed to neutralize the effects of tube capacitance coupling. Design and interference suppression methods relative to parasitic oscillations, transient oscillations, arcing, ignition spark, and switching interference are discussed in detail in Volume III.

4. ARCING

When the electric field intensity in a dielectric between two conductors exceeds the breakdown strength of the dielectric, an arc occurs, resulting in very rapid and large variations of the impedance of the path between the conductors. The rapidity of impedance variation depends on the properties of the dielectric, as well as on the external circuit. In the case of a gas, it is the ionization and de-ionization time that is most important. Usually, the first is much shorter than the second; hence, the most rapid changes and the most severe interference take place during the actual breakdown of the gas. Arcing in solid or liquid dielectrics is not considered here, because its occurrence would be classified as a failure of the system and, therefore, would not be expected anywhere during normal operation. While not all gaseous discharges are arcs, it is not necessary to distinguish here between the various types of discharges. The word "arc" will be used for all types of electrical discharges discussed in these volumes.

Arcing occurs both by design and as an undesirable by-product. It occurs during switching processes. It may also occur when a potential difference is developed between two structural members that are almost, but not quite, in contact. If the distance between contact points is variable, due to shock or normal vibration in an aircraft, there is the additional effect of the varying capacitance, which introduces a varying impedance, even in the absence of an arc. Since structural members in aerospace vehicles are frequently used as ground return paths for power currents, and sometimes also for control and communication currents, it is clear that lack of good electrical contact at any point will be a serious source of interference.

Because arcing always produces large transients, special precautions must be taken when arcing occurs by design, as in the ignition system of an aircraft or vehicle, or the transmit-receive box of a radar set. These transients cause a large amount of radio interference unless they are prevented from reaching the receiver.

5. RANDOM NOISE

In statistical theory, the term "random noise" is used for disturbances that are completely without regularity in their detailed properties. These disturbances lend themselves well to treatment by statistical methods, and their theory has been extensively discussed in the literature. An example of radio interference which can be so treated is the vacuum tube "noise" due to thermal agitation and the shot effect. Into this group also belong atmospheric disturbances commonly known as "static." The characteristics common to these phenomena are their random amplitudes, random phases, and lack of periodicity. If a frequency analysis is performed on such disturbances, it is found that their energy is fairly uniformly spread throughout the frequency spectrum from zero up to a maximum, sometimes as high as 50 gigacycles (50,000 megacycles) per second, depending upon the source and drops to zero rather rapidly for frequencies above that maximum. A plot of energy versus frequency for such disturbances will, therefore, have a shape somewhat as shown in Figure 1-9. This "random noise" consists of an irregular sequence of pulses of arbitrary shape which bear no relation to each other. Most of the interference to be treated in these volumes do have some regularity. It is what in statistical theory would be classified as a "signal" rather than "random noise." The difference lies in that these interfering signals are usually periodic and show some regularity although their wave shapes and phases may be subject to statistical fluctuations. Like "random noise," their energy is often spread over a very wide frequency spectrum. Unlike "random noise," the distribution of their energy is

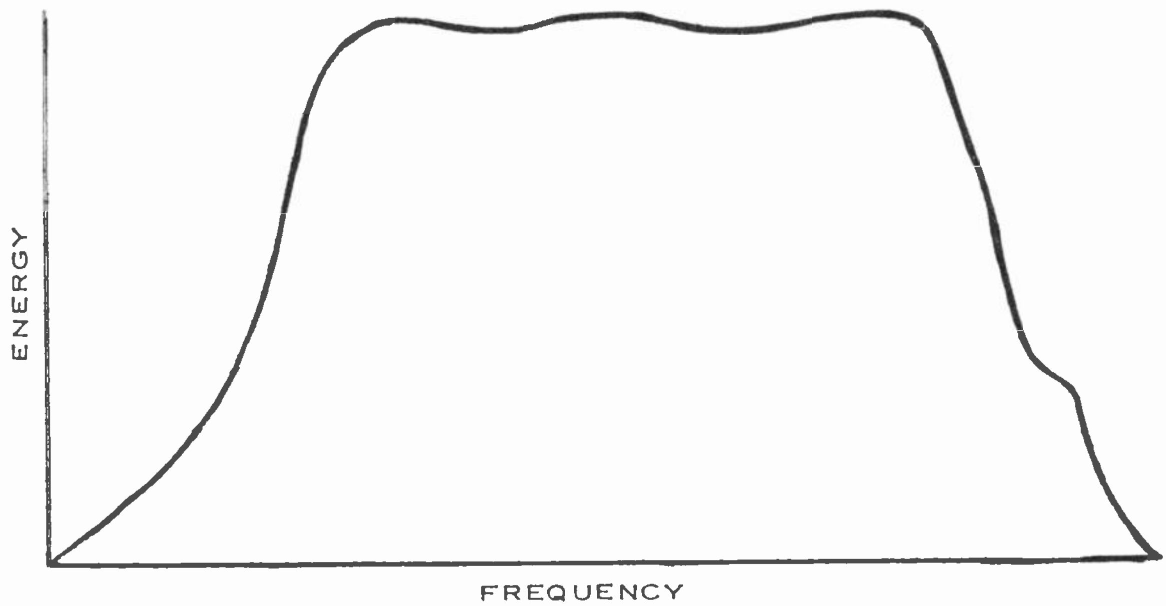


Figure 1-9. Typical Energy
Distribution of "Random Noise"

usually nonuniform, showing wide variations with definite maxima and minima as, for example, in Figure 1-10.

The differences in wave form between interfering signal and "random noise" are clearly brought out in Figure 1-11. Figure 1-11 shows an oscilloscopic trace of a typical interfering voltage generated at the brushes of a motor. The periodicity, as well as the random fluctuations, are clearly evident. Figure 1-12 is the trace of a truly random disturbance as obtained from a noise-generating diode exhibiting complete irregularity. Despite these differences, certain results of the statistical analysis of "random noise" find applications in the design of equipment for interference-free operation.

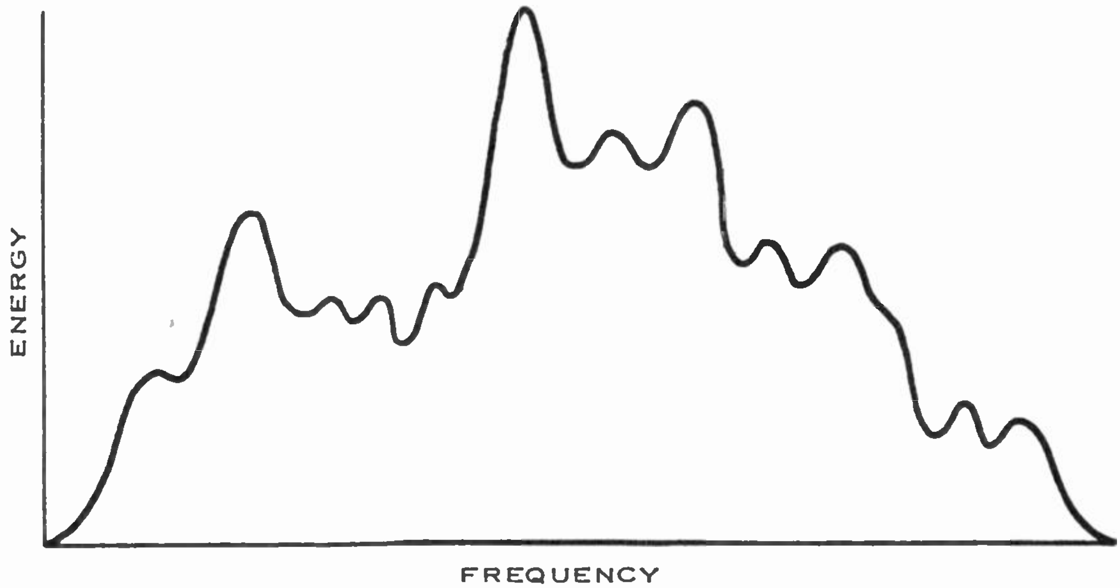


Figure 1-10. Typical Energy
Distribution of Radio Interference

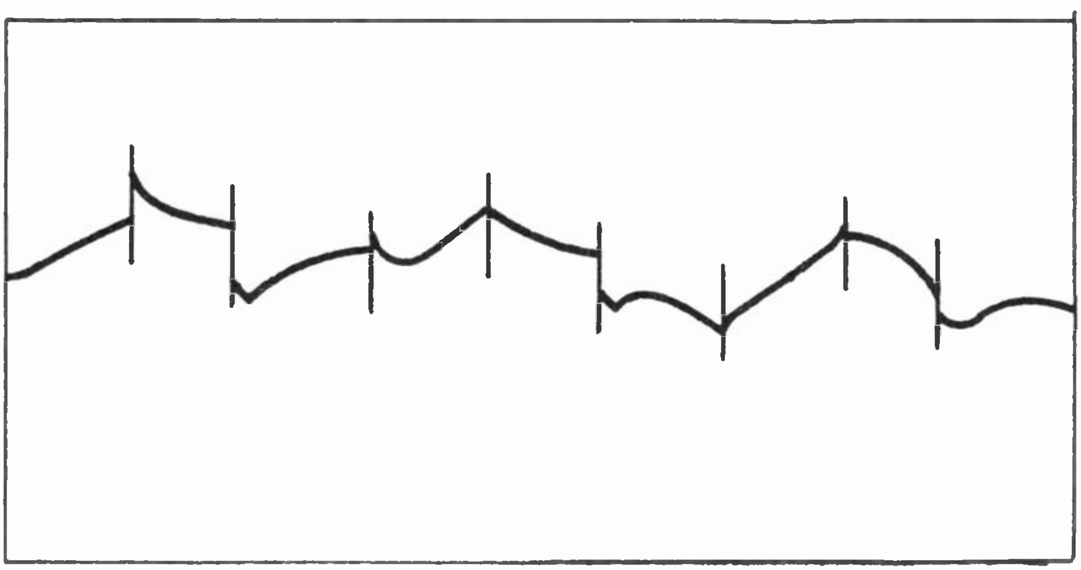


Figure 1-11. Oscilloscopic Trace
of Commutator Interference

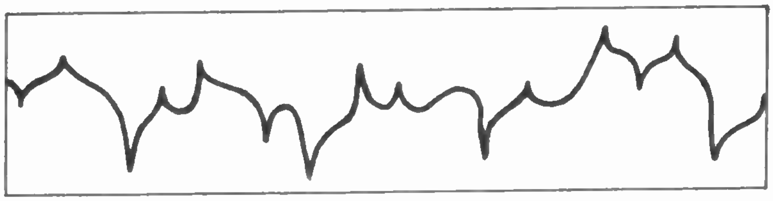


Figure 1-12. Oscilloscopic Trace of Typical "Random Noise"

6. THE GENERATION OF INTERFERING SIGNALS

We have said that interference is always associated with varying electric or magnetic fields. This is true, but the varying fields themselves are not the original source of the interference. What, then, causes varying electric or magnetic fields?

An electrostatic field exists wherever there are charges. If charges are stationary, the field remains constant and there can be no disturbance. Whenever the charges are moving, there is a magnetic field, but as long as the motion is uniform, the magnetic field is constant and again there can be no disturbance. It follows that before there can be any interference, there must be a non-uniform motion of charges, i. e., a varying current. And, indeed, as the various causes of interference in electrical equipment are investigated, one finds that in each case, a varying current is responsible. Once the current is considered, the charge may be forgotten, because a moving charge is a current, and one may base the discussion on one or the other, but not on both. It may be said, then, that the condition for no interference is

$$\frac{di}{dt} = 0 \qquad (1\text{-}9)$$

where i stands for current and d/dt for differentiation with respect to time. This expression is extremely simple, yet important to keep in mind as the ideal for which to strive. In some cases, a varying current is essential to the operation of the equipment, as, for example, in an alternator. In this case, Equation (1-9) cannot be satisfied. But even then Equation (1-9) remains valid, provided one lets i refer to that portion of the total current which remains after subtracting the desired current. Then the equation is not the condition for "no interference," but the condition for no more interference than is absolutely necessary for the proper functioning of the equipment.

To determine the actual origin of the interference, the causes of varying current must be located. The object of this procedure is not to make the obviously absurd attempt of finding the "origin of the origin of the origin," but rather to seek the points at which the application of corrective measures is most practical and effective. It is found that, in some cases, these points are reached when mechanical rather than electrical considerations are involved. In other cases, these points are reached when electronic means of current generation are involved.

What causes a varying current? A current may flow either in a conductor (conduction current), in a gaseous dielectric through which charged bodies or particles are moving (convection current), or in a dielectric void of free charges (displacement current). Displacement currents are negligible at all frequencies at which the considerations of ordinary circuit analysis are valid and need to be considered only in connection with radiation and other phenomena associated with high frequencies. Convection currents occur in electron tubes and in electric arcs and sparks. Also, any charged body in motion constitutes a convection current. Since an insulator will not carry an appreciable amount of net charge, the only bodies that need to be considered in this connection are conductors. While it is true that any motion of a charged conductor of finite dimensions will result in a varying field and, therefore, is a potential source of interference, it may be stated that such interference is very rare and in the case of aircraft not likely to be encountered. The currents which are important as sources of interference are the first mentioned convection currents, i.e., those consisting of moving electrons or ions, and the conduction currents.

These important currents may always be computed by an application of the basic equation $i = E/Z$, in which the current i is considered the effect produced by the electromotive force E against the opposition of the impedance Z. From this it may be concluded that variations of current may be caused either by a varying electromotive force or by a varying impedance. Therefore, there are two basic processes in which interference may originate: one is the generation of varying electromotive forces, the other the variation of impedances.

6.1 THE GENERATION OF VARYING ELECTROMOTIVE FORCES

A varying electromotive force may be generated in one of three major ways: generation by mechanical means in rotating machinery, generation by switching action in vacuum tube oscillators, and generation of equivalent voltages in nonlinear impedances. The generation of electromotive forces by chemical means in batteries may also be a source of interference, but such interference is extremely minute and not important for these volumes.

6.2 VARIATION OF IMPEDANCE

The variation of impedance to the flow of current is one of the most important causes of undesired current variations. Usually the variation is one of the resistance, but variations of inductance and capacitance may be troublesome also. The most extreme case of variable impedance is a switch, where the impedance changes from some finite value to infinity or vice versa. Less extreme cases occur, where the impedance changes more or less suddenly from one finite value to another; but from the discussion in Section 6, it is clear that very slow and gradual changes will not result in appreciable interference. On the other hand, very sudden changes, even though small, may cause a large number of undesired harmonics of appreciable amplitude. Important cases of trouble-causing impedance variations are discussed in Section 7 of this chapter. They include brushes, mechanical switches, commutators, electronic devices, and arcing.

7. MAN-MADE SOURCES OF INTERFERENCE

The origin of an interfering signal may frequently be associated with a particular component or device. These components may be either active or passive. The origin of the unwanted signal or signals may be attributed to a nonlinearity in the component. A component will be considered nonlinear if its electrical characteristics are dependent upon the level of the signal it is acting upon. The nonlinearity may be created to serve a useful purpose such as rectification, harmonic generation, Class B or C amplification, or clipping; or it may be used in a fairly linear region to approximate linear operation. Some of the components which exhibit these nonlinearities are diodes, transistors, gas tubes, vacuum tubes, varistors, and thermistors. In addition to these devices, there are also components which may have been damaged by exceeding their ratings. This might result in a leaky capacitor or a nonlinear resistor, etc.

The nonlinearities will frequently manifest themselves in a relationship between voltage and current, input voltage and output voltage or input current and output current. This relationship will demonstrate the creation of harmonic signals many of which may be unwanted and may be classed as interference. In addition, the nonlinearities provide a means for mixing action and additional interference signals are generated.

Virtually all components, devices, and systems using power, either directly or rectified, from a common 400-cycle distribution system, for instance, are both susceptible to interference and potential generators of interference through the power input connections. In the design of each component this problem should be recognized and minimized. This is particularly true for a complex weapon system where a number of components must simultaneously work together from a common 400-cycle supply without mutual interference.

Two general design rules should be emphasized. First, it is almost always best to suppress interference at the source before coupling to other components and systems can occur, rather than attempt to pretest each component. Second, susceptibility to interference should be minimized during design whenever possible. That is, component susceptibility to interference should be viewed with the same design philosophy as temperature variations.

7.1 MOTIONAL SOURCES

Much of the interference generated by the operation of electrical devices originates in commutator-type machines and arcing contacts. These include all motors, generators, vibrators, relays, and switches. Aircraft and missiles, for instance, have become a maze of such motional sources of interference. Every electrical component in the aircraft or missile may possibly be mounted sufficiently close to susceptible components or component wiring to introduce disturbing signals into these units and adversely affect their operation. The same applies to surface based systems. Consequently, each motional source must be designed so that inherent electrical transients are confined to the unit itself and not permitted to enter other components either directly or indirectly and cause radio interference.

7.1.1 ROTATING MACHINERY

Of all the sources of electromagnetic interference commonly encountered, rotating electrical machinery constitutes one of the largest groups. A good clean design is usually the least likely to cause interference. Peculiarly enough, considerations that seem secondary when interference prevention is neglected become primary when interference prevention is paramount. Such obvious points as symmetry in the windings, mechanical and electrical balance, accuracy of machined parts, and tolerances in the assembly acquire an entirely new importance when interference is considered from the beginning of the basic design.

In all rotating machinery, there is a relative motion between a set of conductors and a magnetic field. An electromotive force, E, is induced in the conductors, which may be computed according to the basic law,

$$E = B\,L\,v \tag{1-10}$$

in which B is the magnetic flux density, L the effective length of the conductor perpendicular to the field, and v the component of relative velocity perpendicular to L and B. The quantity B is varying in all cases encountered in generators, because it is not practical to have the motion take place in a constant field. Ideally, in an alternating current machine, the variation is such that the generated voltage is a pure sine wave. Ideally, in a direct current machine, the variation is such that the generated voltage, as measured at the terminals, is constant while the brushes slide on any one commutator segment, as well as when they slide to the next segment. Actually, deviations from the ideal are always present in both types of machines. The peripheral velocity, v, of the conductors will not be exactly constant, and the effective length, L, of the conductors will not always be exactly the same for all conductors. Thus, irregular variations occur in all factors on the right of the above equation, and, as a result, the generated electromotive force, E, will always contain undesired variations. It is the task of the machine designer to produce a machine in which these variations are as small as possible.

In a direct current machine, an additional difficulty arises. Even if the ideal is approached very closely and the output voltage appearing at the brushes is free from any "ripples," the voltages induced in the conductors must jump abruptly from one constant value to another each time the brushes pass from one commutator segment to the next, so that the current in the armature will show very rapid variations and therefore be very rich in harmonics, even though the external current may be free of them. This is one of the reasons why direct current machines are more troublesome from the point of view of radio interference than alternating current machines; the other is that the process of commutation itself practically always introduces harmonics into the external current. Obviously, the design engineer must develop rotating electrical machinery with the best possible mechanical and electrical characteristics in order to minimize the generation of interference.

7.1.1.1 Commutation

One of the most important examples of an interference-producing switching process is commutation in rotating electrical machines. The function of a commutator is to switch the output terminals from one set of terminals on the armature to another in such a way as to keep the current and generated voltage as constant as possible. The switching action of a commutator is explained with the aid of Figure 1-13. In each of the diagrams A, B, and C the individual armature conductors are represented by generators, the size of each generator indicating the magnitude of the voltage induced in it. In A, the generators (coils) G are about to undergo commutation. The voltage induced in them is very small, but one-half of the full line current is flowing through them in the direction indicated by the arrow. In B, the same generators (coils) have moved under the brushes and are now undergoing commutation. They are short-circuited and the voltage induced in them is ideally zero while the current is changing to a new value. In C, commutation in generators G has been completed. Now a small voltage in the opposite direction is induced in them, and the current in them is again one-half of the full line current, but the current now flows opposite to its previous direction. During commutation, the current must change from some positive value to an equal negative value. The variation may be as shown in one of the curves of Figure 1-14 which are called over-commutation, under-commutation, and linear-commutation, respectively.

This analysis makes clear that, even under ideal conditions, direct current machines with commutators must be a source of interference in three distinct ways: (1) the current in the coil undergoing commutation must change rather rapidly, (2) the voltage generated must vary as the voltage induced in each coil varies with its position in the magnetic field, and (3) the total armature impedance between brushes must change as some of its coils are short-circuited by the brushes. In addition, there are many opportunities for interference to be generated due to deviations from the ideal, such as the voltage induced in the coil undergoing commutation not being exactly zero, or arcing at brushes.

Since commutation is essentially a switching action, it is normally accompanied by interference-producing transients - called break transients. This commutation interference is apart from the brush interference, or surface contact transients, discussed in subsequent paragraphs. However, commutation is always achieved by brushes bearing on a commutator so that both contribute to the interference generated by commutator devices. This combined interference is commonly called "motor-hash" - a poor term to use in design practice, since no distinction is made between brush and commutator interference yet the suppression technique for each is different. Consideration of the requirements for good commutation may in some cases conflict with the choice of technique used to suppress brush interference.

Generally, good commutation receives first consideration. For this reason, machines requiring commutation are doubly troublesome from an interference design standpoint. Induction type motors, wherever

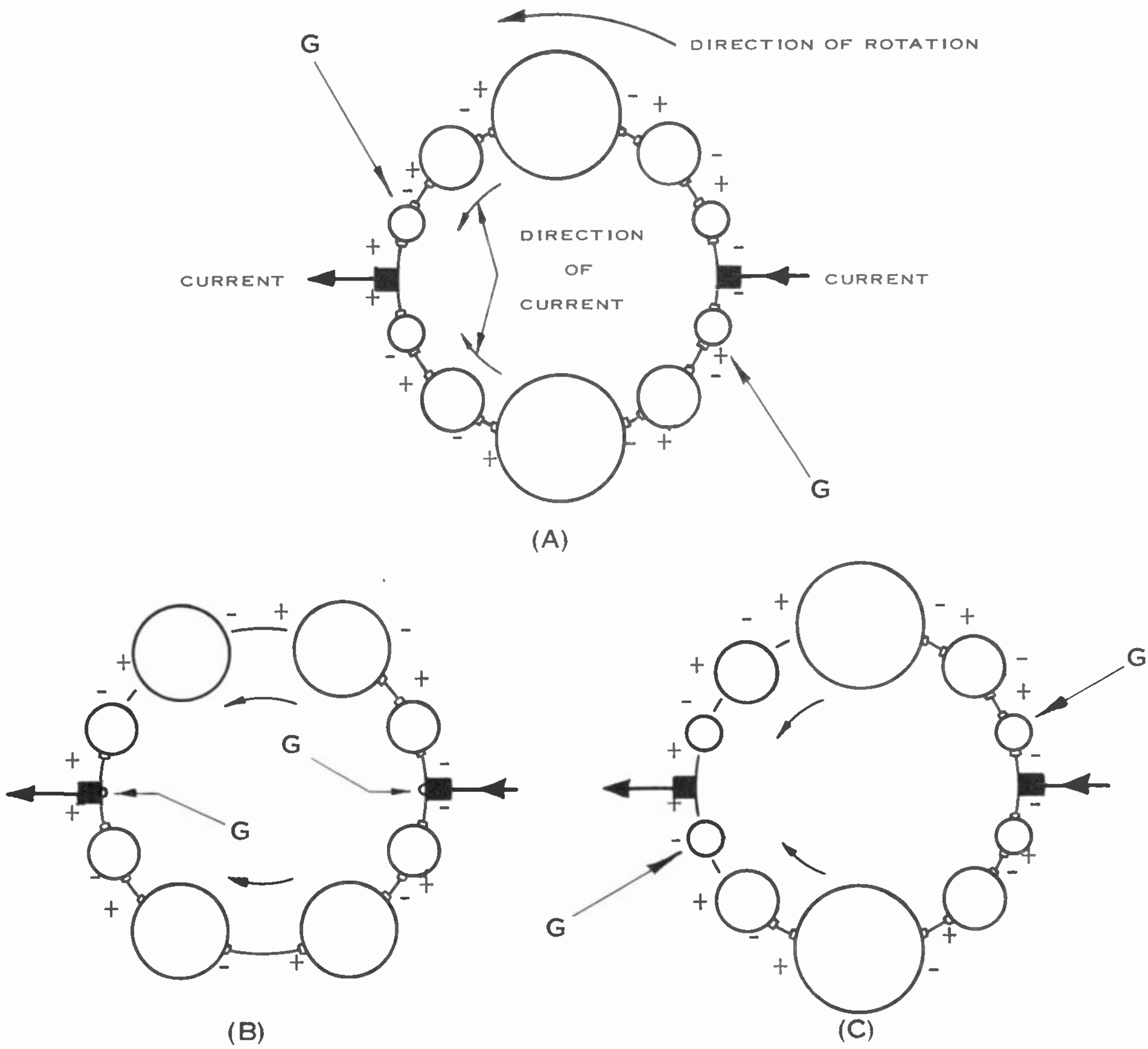


Figure 1-13. Illustration of Switching Action of a Commutator (Individual Armature Conductors Represented by Generators) (Magnitude of Induced Voltage Indicated by Size of Circles)

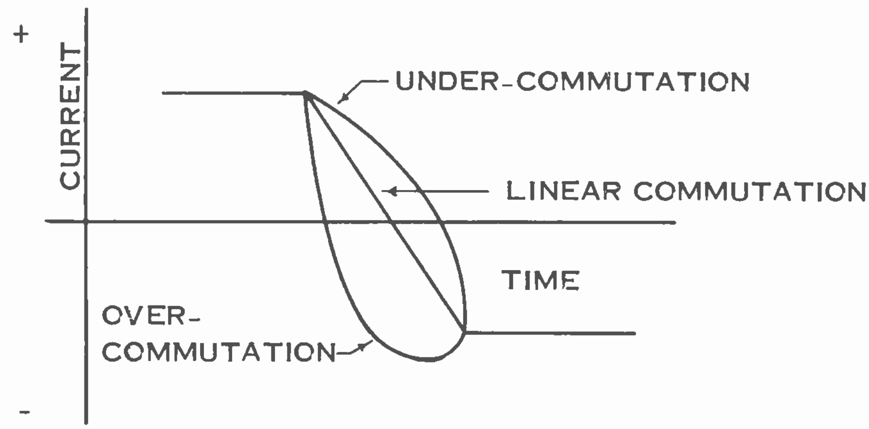


Figure 1-14. Variation of Current in Coil Undergoing Commutation

their use is possible, are by far preferable even at some sacrifice of cost, ease of wiring, or ease of control. Commutation interference, aside from brush interference, may be reduced by the use of interpoles, compensating windings, laminated brushes, and by careful machining techniques to insure clean symmetrical mechanical design. Capacitors and filters should be considered in the original design, as well as strict attention to adequate shielding. These design features and methods are discussed in detail in Volume III, together with discussions relative to different type motors, generators, alternators, inverters, and vibrators.

7.1.1.2 Brushes

Electrical contact between circuit components in relative motion is usually made by means of brushes that ride on slip rings or commutators. Due to the mechanical action of friction, both surfaces wear down gradually, although the brushes wear down faster than the harder metallic surface of the slip ring. The process of brush wear, which is gradual on a macroscopic scale, is actually very irregular and of random nature on a microscopic scale. Fairly large carbon particles are torn loose and either ejected or burned, so that the contact resistance, which depends both on the pressure and on the actual contact area, is subject to sudden random variations of considerable magnitude. In addition to this, vibrations may be set up in the mechanical system of the brushes, springs, and brush holders, which cause the pressure, and thus the resistance, to vary. In extreme cases, the variations of pressure may be so great that the brush bounces completely off the metal, thus causing a true switching action. These causes combine to make all brushes a serious source of interference.

In dc machines having good commutation, most of the interference is directly attributable to the sliding brush contact. This so-called brush interference may be reduced by careful design and choice of brushes and of the metal surface in contact with the brush face. Generated interference is affected by brush pressure. Increased brush pressure generally decreases the generation of interference at all frequencies. Increased brush pressure produces a more constant unit pressure over the entire brush face in contact with the metal rotating surface thus reducing contact resistance which produces what is known as surface contact transients. While the amount of brush vibration and chatter increases with the peripheral speed of the sliding surface in contact with the brush, nevertheless, the brush pressure selected for any application should at least be adequate to damp-out the vibration expected at the designed peripheral speed. It is true increased brush pressure increases wear rate, but the necessity of more frequent replacement should be considered a reasonable compromise for decreased interference.

Interference increases with increased current density for more heat is generated in the contact resistance and the temperature of the brush and commutator surfaces increases. This increase in temperature hastens the formation of an oxide film of considerable thickness on the sliding metal surface. Rapid variations in sliding contact resistance due to irregularities in the oxide film modulate the direct current and may give rise to radio interference. The magnitude of this interference is increased as the current density; hence, voltage drop across the brush face increases. Somewhat larger brush surface area should therefore be provided than is demanded only by consideration of the dissipation of heat and losses due to mechanical friction. However, if too low a current density is used, nonuniform grooves or threading develop on the rotating surface or commutator, and a high friction coefficient is apt to occur which will set the brushes into a noisy chatter thus producing interference.

When a copper commutator is in contact with a carbon or graphite brush, a layer of copper oxide, mixed with carbon particles from brush wear, forms on the commutator. The presence of this copper oxide film introduces unidirectional electrical properties (polarity effects) consistent with the well-known copperoxide rectification. The oxide layer displays a nonlinear resistance of higher value to a brush used as a cathode than to one used as an anode. The cathode brush passes current in discontinuous high current density surges. Approximately ten times as much radio interference may come from the cathode brush as from the anode brush. Surface treatment methods may be used such as with colloidal graphite material or by chromium plating the revolving metal surface to reduce generation of interference.

7.1.2 MECHANICAL SWITCHES

When a switch is opened or closed in a circuit, the impedance across the switch changes from essentially zero to infinity. This rapid impedance change, causing a transient in the voltage and current through the switch, creates a wideband interference signal. The transient effect is magnified if the switch cuts off current to an energy storage device such as an inductor or capacitor. The currents and voltages in the circuit must re-adjust themselves, and whenever the circuit contains inductances or capacitances, this adjustment cannot take place instantaneously, because the energy stored in the magnetic field of the inductance or electric field of the capacitance cannot change instantaneously. There is a brief time interval during which the re-adjustment takes place, and the changing currents and voltages during this interval are called transients. Since any transient may cause radio interference, any switch is a potential source of interference. This interference is conducted to all circuits connected to the switch, and also radiated from the circuits.

In addition to transients produced during the normal functioning of a switch, severe interference may be created due to arcing between the contactors of the switch, immediately before making or after breaking physical contact. If the voltage between contactors is high and the process of closing the switch comparatively slow, there may be a time when the points of contact are separated by a distance shorter than that required for breakdown of the insulation between them, and if the time required for ionization is shorter than that required to close the switch fully, arc-over will occur. A somewhat similar, but more frequent, phenomenon is arc-over between contacts during the process of separating them, providing that the circuit has enough inductance to induce the voltage necessary to maintain the arc. In either case, the resistance of the arc varies rapidly through a wide range of values, and, hence, varying currents, rich in interfering frequency components, are produced. Since an arc tends to maintain a current that is decreasing to zero, arcing actually decreases the severity of the transients by decreasing the rate of change of the current. But little is gained because the arcing itself produces severe interference. The result of normal switching transients, whether or not accompanied by arcing, is the familiar clicking sound so often heard in the earphones of radio receivers during switching processes.

It might be well here to point out that telephone exchanges and computer centers may be the source of radio interference resulting from relay and switching operations, and from the extensive use of pulse currents. This interference may be due to both radiation and conduction. It is particularly important to note that telephone conductors within a building may not be shielded and may act as antennas in radiating pulsating energy which has been fed into them from switching and relaying operations. The switching transients may also enter radio communications and electronics circuits at junction boxes which are used in common by the several circuits.

A capacitor connected across the contacts will reduce transient effects. If possible, filters may be placed in the switch leads. Leads to switches with heavy switching transients should be as short as possible and routed well away from wires to sensitive audio and radio frequency circuits.

Arcing during switching may be reduced by using spring-loaded rapid make-break switches. In addition, the "make-before-break" rotary switches for circuit changes will reduce transients.

One factor which cannot be neglected in high frequency interference suppression is the capacitance across contacts of a switch. In some cases high frequency signals will conduct through an "open" switch.

7.1.2.1 Relays

In addition to the problems associated with mechanical switching, transients in the relay coil will also produce interference. However, most relays are direct current operated, and the filtering problem is less acute. A diode connected across the relay coil such that the actuating voltage produces reverse bias on the diode will often reduce interference to an acceptable level.

Relay contacts are subject to bounce and chatter, giving rise to several switching transients. Interference from this source is reduced by connecting a capacitor across the contacts.

Either mechanical or electronic switching to actuate relays, is a prime generation of radio and audio frequency interference. The interference is transmitted and radiated from dc power supply lines to the relay.

7.1.2.2 Voltage Regulators

Voltage regulators are unique interference generators in that their function is the controlling of voltage levels. Regulator transients may produce interference over a large range of frequencies. There are several different types of voltage regulators which can be considered as sources of interference. Electronic type regulators may use V-R tubes or semiconductor devices. These devices create interference due to their nonlinear action which will be discussed more fully below. Relay type regulators will create interference and its nature will be identical to that of the switching action discussed in the previous paragraphs. Another type of regulator is a rheostatic type variable resistance consisting of stacks of graphite plates and silver button contacts and designated as a carbon stack voltage regulator. Generally, the resistance of the carbon stack is changed by shorting the individual graphite plates one at a time until the proper value of resistance remains in the circuit. This again is a switching action and interference will be created by transients in the line.

a. Reed Type Voltage Regulator

The reed type voltage regulator is one that has moving reed-like contacts such as those found in generator voltage regulators for charging a battery. The generator voltage regulator maintains the voltage output of the generator at a value that keeps the battery fully charged without overcharging it. The regulation is accomplished by a control unit that contains a reverse current relay, a current limiter relay, and another relay, called the voltage regulator.

The reverse current relay has a self-holding series coil that overcomes a spring tension and closes a set of contacts that short the generator output to the system and battery. This series coil holds the contacts closed as long as current flows from the generator. When the battery is charged the current flows toward the generator and the series coil cannot overcome the spring tension causing the contacts to open.

The current limiter relay opens its contacts when a predetermined current passes through its series winding. When these contacts are opened a resistance is added to the generator field dropping less voltage across the field thus decreasing the output voltage of the generator armature. When the current demanded of the generator exceeds the setting of the current limiter, the contacts continually open and close at a rate that keeps the average current at the maximum permissible value.

The voltage regulator relay is set to open at some voltage greater than that of the system. This voltage is 14.2 or 28.5 volts for 12 or 24 volt systems, respectively. The opening of the relay contacts inserts resistance into the field circuit of the generator again decreasing its output. The voltage setting can be adjusted by changing the spring tension of the contacts. When the relay opens the added resistance to the field of the generator causes less generator output and the voltage across the relay coil drops to a value that closes the contacts. This cycle is repeated by alternately inserting and removing resistance from the field at a rate that varies with the current demand. By these means the average field flux is controlled by supplying a constant voltage, regardless of the current demanded, as long as the generator rating is not exceeded.

Most of the interference generated in this type of voltage regulator is from the relays. The design engineer should use relays that have large silver contact areas to obtain lower current densities. Less variance in contact resistance is obtained when stronger reeds for greater contact pressure are used.

Interference is conducted through the system leads and radiated by the coil windings of the relays. In Volume III the routing of wires and cables, installation of components, shielding, bonding, and filtering will be discussed. Suppression of regulator circuits is accomplished by placing the unit in a shielding can and using coaxial cabling to the can. The regulator shield provides a ground termination for the shielded cables external and internal to the can. This unit should be installed away from areas where vibration is excessive.

Bypass capacitors at the generator, in the regulator or at the battery terminals can be used for some suppression but these capacitors lose their effectiveness when capacitances greater than 0.5 μfd are used. The series inductance inherent in conventional bypass capacitors decrease the effective coupling to ground of the offending interference. Also, large capacitors cause a time lag in the operation of the output voltage control.

The design of this voltage regulator is a formidable one because of the large currents the components must handle.

b. Carbon Stack Type Voltage Regulator

The carbon stack type voltage regulator is a rheostatic type variable resistance consisting of stacks of graphite plates. It is especially designed for automatic voltage control by regulating the main exciter field resistance of small and medium size generators. The graphite plates are pivoted at the center by metallic members and are separated from one another at the back of the stack by insulating spacers. At the front the plates have spring-mounted silver buttons that are normally separated, but can be closed by tilting the stack forward in sequence such that the silver buttons form a short circuit path when they are in contact. When the entire stack is tilted forward, the current path is through the buttons and the field has minimum resistance. The current path is through the center metallic member and all of the graphite plates when the stack is tilted back for maximum field resistance in the field of the exciter then depends on the number of plates tilted either backward or forward.

This type of voltage regulator has possible interference from the following areas: button pressure, button contact area, silver button surface, bonding between graphite plates and the pivot members and spacer, resistivity of graphite plates, and contact pressure between the metallic pivot member of the end insulating spacer and the graphite plate.

To the design engineer many of these problems are found when designing relays or brush commutators. Increasing the pressure between button contacts decreases the amount of interference generated because of the smaller variance in contact resistance. The upper limit of this pressure is the mechanical properties, or the amount of bending strain of the graphite plates.

Interference decreases when the current density of the button contact is decreased by using larger contact areas. The larger the area, though, the greater spring tension required for contact pressure and again a mechanical property of the plates must be considered.

Using silver for the button surfaces decreases the interference generated at the contact because of the conductive properties of the silver oxides. Normally, some metallic oxides cause a voltage drop across the contacts. Silver is more expensive but the life of the contact is lengthened.

Bonding between the graphite plates and the contacts, pivots, and spacers is another mechanical problem that must be considered. The detailed problems of pressure, current density, and surface treatment will be discussed in Volume III.

Generated interference is less for graphite plate materials of lower resistivity. Metallic impregnated graphite must be used to withstand the tensions in the plates.

7.2 TRANSMISSION LINES AND CONNECTORS

Interference created by transmission lines, connectors, brackets, insulators, etc., is dependent upon a large number of factors associated with the line design and construction. These will include conductor size, spacing, properties of the insulation, termination, physical position, and line hardware. There will be interference due to harmonics generated in nonlinearities. Interference may also be generated at trouble spots due to the sudden change of potential accompanying surges, sparks, or intermittent open circuits. Spark gap or corona formation results from ionization of the air created when a potential difference exists between adjacent surfaces. Frequent sources of interference on lines are primary cutouts, lightning arresters, broken bushings on transformers, loose connections including ground wires, defective insulators, imperfect contact between metal parts, loose tie wires, and loose and corroded hardware. The interference generated by these sources may be propagated by radiation from the line, conduction along the line, and conduction followed by radiation. The interference is normally broadband and may be either impulse or random in nature.

7.3 ELECTRONIC DEVICES

In many applications, a vacuum or gas-filled electron tube is used to produce switching action. These tubes are particularly useful as generators of nonsinusoidal waveforms and find wide applications as pulse generators, modulators, and oscillators. In these applications, the generation of harmonics is desired and is essential to the proper operation of the device. It is not surprising, then, that these devices also rank high as interference generators.

The basic circuit of an oscillator of the type discussed above (a so-called "relaxation oscillator") is shown in Figure 1-15. When the battery switch is closed, the condenser charges through a resistance until the voltage across the tube exceeds the breakdown voltage. When that point is reached, the condenser discharges very rapidly through the low impedance of the tube until the voltage across the tube drops to a value below that necessary to maintain the arc. After that, the tube impedance becomes infinite, and the entire cycle repeats. All electronic pulse generators or nonsinusoidal oscillators use this basic

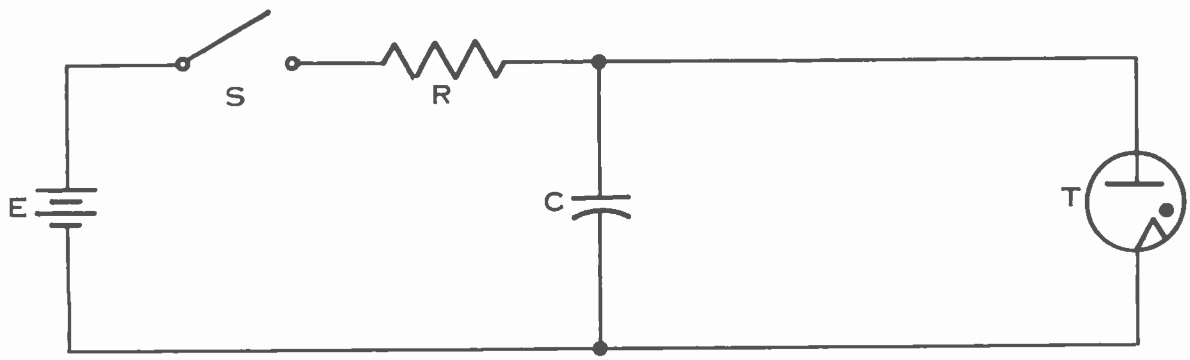


Figure 1-15. Simple Relaxation Oscillator

principle, although, in practice, the circuits will be more complicated. In particular, much more flexible control may be achieved by using a triode, and applying a control voltage to the grid, instead of relying entirely on the plate voltage to cause breakdown ("closing" of the switch) at the desired instant. If a triode is used, vacuum tubes may be substituted for gas tubes because a sufficiently large negative control voltage on the grid will "cut off" the plate current, thus opening the switch.

An undesired variation of impedance also takes place in vacuum tubes used as sinusoidal generators. Due to the nonlinearity of the tube characteristics, there will always be some harmonics generated in an electronic oscillator, together with the desired frequency. In fact, this effect is unavoidable, though it may be kept quite small, because an oscillator is essentially a nonlinear device. Were it linear, the amplitude of oscillation would be unstable, all values being equally possible. It is the nonlinearity of the tube that provides the stabilizing effect, by varying the impedance in such a manner as to counteract any small changes of amplitude which may occur. Thus, no vacuum tube oscillator produces a perfect sine wave, although in practice the harmonic generation can be kept extremely low. The larger the power output of the tube, the more difficult it becomes to keep the harmonics small. Therefore, it is usually the last stage of a transmitter, or similar electronic device, which is most likely to produce interference. However, the sensitivity of modern receivers is such that even very weak interfering signals may be very troublesome. Therefore, it becomes necessary to watch carefully all stages of transmitters and eliminate all harmonic generation, as much as possible.

7.3.1 DISCHARGE TUBES

Thyratrons and similar gaseous discharge tubes are often used as switches. These tubes have a very fast turn on time (1 microsecond or less) and can be built to handle large currents. However, these advantages for particular applications intensify the interference produced by these devices.

All of the methods discussed in previous sections apply to interference prevention. Any possible technique to reduce the time rate of change during switching, such as series resistors, or shunt capacitors, should be used. Thyratron circuits and connecting cables should be located well away from a susceptible receiver, in another compartment if possible. Also, deterioration of tubes produces increased interference, and worn tubes should be replaced.

7.3.2 DIODES AND RECTIFIERS

Semiconductor diodes and transistors are very frequently used in switching circuits, with advantages in size, weight, reliability, and power requirements over vacuum tubes. However, switching transients with these components produces particularly severe interference.

A widely used diode application is in a rectifier for a dc power supply. The diode conducts during one-half cycle of the sinusoidal input, at frequencies of 1, 2, 3, or 6 times the input voltage frequency. However, during the transition or recovery time immediately following the voltage shifts to reverse bias, a large reverse current pulse is generated. For a typical rectifier diode with an average forward current of 500 ma, a reverse current pulse of 1.75 amps peak and 40 μsec duration is generated. This pulse manifests itself as audio and radio frequency interference up to approximately 5 mcs.

This same form of interference generation takes place in most diode and transistor switching circuits. The size of the reverse current pulse is a function of diode recovery time. In design of diode circuits, particularly dc power supplies, the following design criterion should be followed:

a. Select a diode which will operate at lowest current density in proportion to maximum rated current, and with highest rated working and peak inverse voltage consistent with other design requirements.

b. Use lowest possible switching rate. This is contrary to ripple reduction and filtering requirements. However, the size of filters for both ripple and RFI will usually be smaller at the lower switching rate.

c. Select a diode with lowest possible recovery time to minimize the reverse current pulse.

Correction and filtering of the interference oscillations must be made right at the diodes. Actual harmonic filtering or transient current reduction is effective. In some cases, a small resistor in series with the diode is sufficient. Series inductors or shunt capacitors to the load may also be used. In transistor switching circuits conventional stage isolation of dc power supply at the offending circuit is used.

Unfortunately, semiconductor power supplies, with their attendant ripple reduction problems, are usually used in conjunction with transistor amplifiers in designs requiring minimum size and weight. Transistor amplifiers are particularly sensitive to ripple and interference in power supplies, since the dc voltage is directly coupled to transistor elements. In addition, power supplies must be of very low impedance to minimize interstage coupling. These factors magnify the interference reduction problem in the semiconductor circuits widely used in current electronic equipment.

In current aircraft design, a single transformer-rectifier unit (T-R) is used to supply all dc power requirements. Interference generated in the T-R unit may be unacceptable at frequencies up to 100 mcs. High frequency filtering at the T-R output is not extremely difficult to obtain, but low frequency receivers such as an Automatic Direction Finder are very susceptible to this interference. Additional pi filters at the input to the low frequency receivers may be necessary.

Audio circuits are also quite sensitive to low frequency interference noise on the powerline, especially transistor amplifiers. Additional filtering for these units may also be necessary.

7.3.3 NONSINUSOIDAL OSCILLATORS

Many electronic oscillators combine the interference production characteristics of both linear oscillators and switches. Examples are radars, pulse generators, and sawtooth generators. The interference spectrum produced by periodic nonlinear oscillators is characterized by a number of discreet components. Previously discussed suppression techniques are applicable.

The problem is particularly acute for high power pulses, such as in a radar system. Again, all interconnecting wiring should be held to a minimum, and equipment separated physically if possible. Filters at the input powerlines are usually required.

7.3.4 LINEAR OSCILLATORS

Linear electronic oscillators generating sinusoidal waveforms are prime sources of interference by virtue of the very function of their operation. These circuits are widely used in radio and radar equipment. Interference from this source can be prevented only by restricting fields and electromotive forces from electronic oscillators to the necessary circuits and transmission lines.

Proper shielding of circuits and transmission lines to prevent external coupling of oscillations is the first step in interference prevention. Equipment enclosures should be designed with bonding on access panels, and screening at all ventilation, meter, and other gaps in the enclosure. Whenever possible, all units of the equipment should be placed in a common enclosure to minimize the need for interconnection cable with high signal levels. Also, radio transmitter antennas should be located to take advantage of shielding provided by structures such as aircraft and missile skin. It is recommended that the transmission line to the antenna also be shielded.

All incoming power supply lines within the enclosure should be located well away from oscillator circuits. A line filter to remove oscillator components from the power line is often required, and this should be physically located at the point where the powerline enters the enclosure. Also, routing the powerline well away from all interconnecting cables is good practice.

7.3.5 ELECTRONIC VOLTAGE REGULATOR

A voltage regulator tube is a tube specially designed to keep the voltage at a practically constant value regardless of voltage regulation in the power supply or variation in load current.

A common type in wide use is the mercury vapor tube. The mercury vaporizes when the cathode is heated and ionizes when a certain plate voltage is applied. The tube is characterized by the fact that the voltages applied to the electrodes have little effect on the current flow once it reaches a certain minimum value. This minimum voltage difference between plate and cathode that causes electrons to move with sufficient velocity to produce ionization is called the ionization or firing voltage. Once the tube is ionized, the plate voltage can be increased or reduced somewhat without affecting the flow of electrons. Therefore, the voltage across the tube remains practically constant over a moderately wide range of current because the tube behaves like a low resistance that varies in value as a function of current.

During ionization or deionization, the effective impedance is nonlinear due to this function of current and the time rate of change of the current. This becomes a source of interference because of the harmonic generation.

Interference during operation of the tube is generated if the tube is very old or if it is subjected to severe vibration. Again the units containing the voltage regulator should be shock-mounted, installed where vibration is not excessive (away from engines), and accessible for the replacement of faulty tubes. Routing and shielding of the interconnecting cables and wiring, bonding of shields and covers, and filtering should be considered by the design engineer as discussed in Volume III.

7.4 TELEPHONE AND TELEGRAPH LINES, CIRCUITS, SWITCHES, AND DIALS

Telephone equipment such as rectifiers, commutator machines, switches, and dials may produce radio interference disturbances. Special filters have been designed for application to telephone dials when interference occurs with AM broadcast reception. Telegraph signals of the Morse code type involve on and off switching. The ground return circuits used for telegraph transmissions are less favorable from the interference standpoint than the two wire (metallic) circuits used for telephone transmissions. However, general experience has shown that neither telegraph nor telephone systems are sources of interference to radio communication when the accepted good practices for the respective systems are followed.

Telephone cables are very carefully designed and constructed to balance the capacitances between conductors, and between conductors and the sheath, to minimize cross talk. Telephone transmissions are kept at about the same level in all circuits in the cable. Signals of other types are sometimes transmitted in the same cable, and, when this is done, the current and voltage must be kept within the same limits as the telephone transmissions. To prevent interference when telephone high level signal levels must be carefully monitored, audio and keying signals are transmitted over circuits in a common cable. Under some conditions higher level signals can be attenuated before they enter the cable and then amplified to their normal level at the termination end.

7.5 RADIO AND TELEVISION RECEIVER OSCILLATORS

Radio and television receivers can cause interference with communications-electronic equipment in the same area. The high frequency energy generated by the oscillator may be transmitted to the antenna and radiated, or it may be radiated from the chassis. It, also, may be conducted through the power leads and radiated to other receivers.

Antenna radiation can be prevented by the use of a sufficient number of radio frequency stages, proper shielding, filtering, and bypassing, and the use of a compact layout for the receiver. Bypass capacitors and filters can also reduce conducted interference in power leads. Chassis radiation may occur particularly in the upper range of television frequencies where the size of the chassis may approach half-wave resonance and become an efficient radiator. This type of radiation can be reduced by appropriate shielding techniques.

Federal Communications Commission rules require receivers sold to the public to conform to standards which minimize unwanted radiation. New military equipment should be at least as free from interference. However, in some cases, when problems with existing equipment arise, it may be difficult to apply the necessary techniques and replacement of equipment in whole or in part may be necessary.

7.6 LIGHTING

7.6.1 INCANDESCENT LAMPS

Incandescent lamps are substantially interference-free except when being switched on and off. However, some lamps with old-type filaments act as a source of narrow band RFI between 20 and 100 mc. Lamps about to burn out may generate broadband interference in the lower frequency range. Determining the location of interference-producing lamps may be difficult and time-consuming.

7.6.2 FLUORESCENT LAMPS

Fluorescent lamps depend on conduction through ionized gas. On each reversal of the applied ac voltage, the tube changes rapidly from a good insulator to a good conductor, thus producing pulse-type interference with a repetition rate twice the power frequency. Interference may be conducted by and radiated from the power supply circuit. This interference may affect receivers in the LF and MF range, although higher frequency equipments are not ordinarily disturbed. Capacitors normally incorporated in the lamp "starter" reduce this effect, and line filters may be added if necessary. The effectiveness may be limited by the lack of a satisfactory ground connection for the filter. There is also direct RF radiation from the lamp itself which decreases rapidly with distance. This RF radiation is usually more pronounced in the higher frequency range, and the effect is somewhat reduced by application of capacitors in the circuit. Incandescent lamps have been preferred for locations in close proximity to sensitive electronic equipment, but the lesser heat dissipation makes fluorescent lamps attractive for confined areas. Special "interference-free" fluorescent lamps have been produced for this purpose, but good results are obtainable using conventional equipment with added interference suppression measures.

Neon signs and lights glow as a result of the flow of electrons in ionized gas. Interference may be radiated directly from the tubing or from the power supply circuit. Proper maintenance minimizes the interference but does not completely eliminate it. Sometimes interference is reduced by equalizing potentials at certain points along the tube by wrapping wires on the tube and interconnecting them. Filters can be applied in the supply circuit.

7.6.3 MERCURY AND SODIUM LAMPS

Mercury and sodium lamps have an RF interference radiation common to an arc discharge. Radiated energy is high due to the high voltage and current used. There will be a high level of radiation at both high and low frequencies, and lamps of this type should not be used unless sensitive receivers are adequately shielded or located at a distance. Suitable line filters should be used in all lamp installations.

7.7 WELDING EQUIPMENT

The three general classes of electrical welders in use today include the arc welder, stabilized arc welder, and resistance or spot welder. Broadband interference originates in all types. In addition, some narrow band interference originates in RF stabilized arc welders with an intensity much higher than that of the arc and resistance or spot welders. The Federal Communications Commission (FCC) has set aside the use of specific industrial frequencies for these equipments. However, harmonic and spurious outputs constitute a major problem.

7.7.1 ARC WELDERS

Electric arc welders are classified as spark-, glow-, and arc-discharge. In the spark-discharge type, a high voltage of approximately 1,000 volts is used with a low current. The glow-discharge type uses a lower voltage of about 100 volts with higher current, while the arc-discharge welder uses a very low voltage with a very high current. All three types of discharge have steep wavefronts with consequent broadband interference. There is a secondary interference resulting from the acceleration of ionized particles within the arc.

7.7.2 RESISTANCE WELDERS

Resistance welders are classified as spot, seam, butt, and projection welders. All are based on the principle of power loss in a resistance. In general, two pieces of metal are subjected to electrode pressure and a current impulse is passed through the contact. The current ranges upward from 5,000 amperes at 2 volts. This type of welding can be used only on sheets up to 1/8" thick due to electrode heating. Thicker sheets are welded by means of spaced impulses. This type of welding produces broadband high frequency radiations up into the microwave region. This type of welder is found on production lines in industrial areas and produces an audible signal in a radio receiver in accordance with the repetition rate.

7.7.3 RF STABILIZED WELDERS

This welder uses an inert-gas, shielded arc process. Radio frequency energy is superimposed on the welding current to provide for initiation of the arc and subsequent stabilization. The RF energy provides the high voltage necessary to initiate spark discharge. Although this welding system is efficient from a work standpoint, it generates strong radio-interference signals. The FCC has provided certain frequencies for operation as well as limits on spurious emissions. It is therefore necessary that filtered and bypassed power supplies be used, as well as extensive shields for cables and cabinets.

7.8 ENGINE-IGNITION-SYSTEM INTERFERENCE SOURCES

The ignition system is essentially an arc system and generates both conducted and radiated interference. The interference originates at breaker points, rotor gap, spark plugs, and ignition coil. This type of interference can be reduced by placing the capacitor and points in a grounded metal container. The peak current due to arc formation across the rotor gap can be reduced by resistors on both sides of the gap. A resistor of about 10,000 ohms in series with each spark plug will reduce HF radiation from this source. Resistor type spark plugs have resistors built in. Since some of the interference can be conducted back to the ignition switch and its complex of low-voltage wires, it is desirable to reduce it with a feed-through capacitor or a π-type filter in the line to the distributor. All ignition components and wiring should be enclosed in a properly grounded shielding harness.

7.9 ISM (INDUSTRIAL, SCIENTIFIC, MEDICAL) EQUIPMENT

Some heating and other operations used in industry, medical treatments, and even in commercial and domestic cooking are accomplished very efficiently with radio frequency energy, and in some cases cannot be accomplished by other means presently known. Two methods are employed to suppress RFI. First, shielding is used to prevent radiation. Second, ISM operation is restricted to the frequency bands permitted in the FCC rules. Since ISM equipments do not interfere with each other, only a very small percentage of the spectrum has been assigned for such usage. However, since different frequencies are needed for different operations, several bands have been assigned between 13 mc and 18,000 mc.

7.9.1 INDUSTRIAL RF HEATING EQUIPMENT

The two types of industrial RF heating equipment are induction heating and dielectric heating. Induction heaters usually are operated below 3 mc. The process is characterized by strong magnetic (low impedance) fields which decrease rapidly with distance. Elaborate shielding is not ordinarily necessary. Precautions are necessary to prevent radiation from the power line, and filters may be needed.

Dielectric heating usually operates at frequencies above 3 mc and is characterized by strong electric (high impedance) fields. A fairly high degree of shielding is necessary. Some processes can be carried on inside a small tightly sealed enclosure but others require fairly large electrodes in which case the only practical way to shield is to operate in a completely shielded room. All circuits entering the room must be filtered.

Radio frequency ("radar" or "microwave") cooking is accomplished in enclosed ovens, but the doors cannot be made tight enough for effective shielding at the frequencies used. Therefore, such units operate on the ISM bands at 2450 mc or 915 mc. Similar considerations apply to "microwave" industrial heating. Some dielectric materials which cannot be heated in conventional dielectric heaters can be heated effectively at 915 mc.

7.9.2 ELECTRO-MEDICAL EQUIPMENT

Various types of electro-medical equipment use large amounts of impulse power which can cause considerable conducted interference. Particularly troublesome from the radio-interference standpoint are diathermy and electro-surgical devices. In general, this type of equipment must be constructed and operated in accordance with FCC rules to prevent interference with communication-electronic equipment.

a. Diathermy Equipment

This type of equipment consists of induction and dielectric heaters used in therapy. Therapeutic effects result from the heat generated in the body tissues by the HF energy. Less interference may be encountered with the dielectric type than the induction type of diathermy equipment. In the dielectric type of equipment, the body is under treatment in an alternating electromagnetic field between the plates of a capacitor. The body acts as a lossy dielectric and the spacing between the plates of the capacitor can be reduced to confine the field. Most units are a combination of dielectric and induction types. To meet FCC standards most modern diathermy machines have crystal-controlled or stabilized oscillators, filtered power supplies, by-passed power lines, and containers which are well shielded.

b. Electro-Surgical Apparatus

This type of device utilizes a spark generated by an RF oscillator. The incision made by the surgical knife is cauterized by the spark to keep blood flow to a minimum. The best cauterizing action occurs with a succession of damped-sine-wave pulses, generally produced by a spark gap. To eliminate the interference from this type of device, complete shielding of the room is required. Shielding of the equipment container, filtering, and similar interference reduction methods should also be used.

7.10 HOUSEHOLD APPLIANCES AND BUSINESS MACHINES

Household appliances are of many types and may produce continuous or intermittent interference. Thermostats used on heating systems and appliances (such as electric irons and heating pads) may produce intermittent interference. The slow-acting type of thermostat may produce continuous interference. Vacuum cleaners, food mixers, laundry equipment, and other appliances using commutator motors, oil burners, and fluorescent lamps produce continuous interference. Interference produced by oil burners comes from the ignition system. Proper grounding of the furnace and shielding of the transformer minimize the interference. A balanced transformer with two electrodes produces less interference than one electrode and ground. Motor appliances can be suppressed in the same manner as other commutator motor devices. Universal motors with a split series field, half on each side of the armature, have lower levels of interference than other similar types of motors.

The motors in most electric razors usually break and make the circuit and therefore produce higher levels of interference than do ordinary commutator motors. Lack of shielding is also a factor. The electric razor is usually equipped with capacitors. Filters are difficult to apply because the razor has no accessible ground point.

Business machines, coin-operated vending machines, and similar devices may be sources of interference. When used in the vicinity of communication or sensitive electronic equipment, they may require suppression measures such as careful shielding, application of suppression to contacts, use of non-commutating type motors where feasible, and proper grounding. When equipment is metal enclosed so that the shielding is reasonably good, direct radiation is usually not a factor. The main consideration is interference conducted along the power supply circuits or radiated from these circuits.

7.11 ELECTRONIC COUNTERMEASURES

Electronic countermeasures (ECM) signals are a special class of interference since they are, in effect, "interference by intent." ECM may take many forms, and there are a variety of circuits and operational procedures which can provide a measure of protection. The engineer should obviously plan his systems with a basis of knowledge that ECM might be used against the systems. Much of the information on nature of this source of interference of this type will be found in security classified literature. Since ECM can partially or completely incapacitate a C-E system or equipment, the engineer should study possible ECM effects with great care and take necessary steps to make his systems as invulnerable as possible to this form of interference.

7.12 NUCLEAR EXPLOSIONS

The possibility of propagation blackout due to oscillating ionized plasmas in the upper atmosphere should receive consideration in system planning. Such conditions may result from high altitude nuclear explosions. Very little unclassified information is currently available on this subject, and the engineer should study the latest available information at the time that he is actually planning his installation.

7.13 NON-ELECTRIC EQUIPMENT

Any condition which results in the building up of a static charge followed by a sudden discharge can be a cause of interference. While "static" has been associated with interference, a "static charge" is not interfering when it remains static and becomes interfering only when it becomes dynamic. Static charges may be built up by friction or by induction. Discharge takes place when the potential exceeds the breakdown point. Rate of recurrence of breakdown is one of the factors to be considered. Effects on receivers are similar to those produced by the operation of switches and contacts.

Mechanical equipment, though apparently isolated, may be a source of interference to communications-electronic equipment. One example is the discharge of "static" from belt-driven machinery. The discharges take place through gears or shaft bearings. The lubricating film of the bearing may be broken down as the charge builds up or metal points may come in contact during rotation. In large motors and generators fairly large current may be induced in the armature shaft. These currents may produce sufficient potential to break down the lubricating film on the bearing. Sometimes induced currents are produced when gear trains operate in strong induction fields and discharges between gear teeth are produced. Usually the equipment itself is not a good antenna, but when the discharge is to a pipe or conduit coupled to electric wiring, the interference may affect a considerable area.

Bonding usually helps to minimize the interference such as thoroughly bonding together the motor, compressor, and other hardware of a refrigerator. Conducting lubricants have been tried experimentally but are not in general use. Grounding brushes can be applied to shafts. Metal parts should either be well bonded or well insulated. Proper grounding reduces the radiation of interference but multiple grounds may form a loop. Large discharge currents in such a loop would cause radiation.

7.14 ELECTRICAL CONTROLLER EQUIPMENT

Electrical controller equipment may be a variable precision device or a simple off-on switch. The variable control will cause slow variations of waveforms, producing very little interference. However, a control with make-break contacts will cause considerable interference.

7.14.1 GAP BREAKDOWN

As a switch is operated, gap breakdown occurs. This breakdown is dependent upon the type of load being supplied.

a. Resistive Loads

When a circuit with a resistive load is broken, the voltage across the switch will rise very quickly to the power supply voltage. This quick rise is due to charging of the small stray capacitance across the switch through the load resistance. During the initial break, the contacts move only a very short distance with a resulting high field, so that breakdown occurs and the voltage is reduced. When the voltage is reduced, the arc goes out. This cycle repeats itself several times during the initial phase of contact separation. Likewise, when the switch is being closed, as the moving contacts approach each other several arc cycles will occur before permanent contact is finally established.

b. Inductive Loads

When a switch in an inductive circuit is opened, a high electric field appears across the switch and breakdown occurs. The voltage is reduced and after the arc is extinguished, the voltage again rises and the cycle recurs. With the inductive load, the voltage across the switch can rise to many times that of the supply, resulting in considerably more switch arcing than in the resistive case. However, when a switch is closed in an inductive circuit, the inductive load resists any sudden change in current, and breakdown is resisted. As a consequence, arcing is less severe in this situation.

7.14.2 AUTOMATIC AND NON-AUTOMATIC SWITCHES

Manual switches are not considered to be a major source of interference due to infrequent operation. However, the frequent opening and closing of such devices as voltage regulators cause considerable conducted and radiated broadband interference at frequencies of operation varying from a few times a minute to several hundred times per second. Interference reduction techniques involve filters, shielding, and control of peak currents.

7.15 HIGH POWER RADARS

Of particular importance as a source of interference is the trend to higher output powers in radars. These powers are now in the multi-megawatt range and may go higher. Because of this high radar output power, the pulsed emission used in radars, and the high gain of radar antennas, there are high probabilities of interference at the fundamental frequency, at harmonic frequencies, and at non-harmonic spurious frequencies. This radar problem is considered so important that particular emphasis is placed on predicting radar interference and conducting spectrum signature tests to determine the actual interference radiation and susceptibility characteristics of high power radars.

Radar systems are discussed at greater length in Chapter III, Section 3.2 of this volume. Radar interference prediction and spectrum signature measurements are discussed in Volume II.

8. NATURAL SOURCES OF INTERFERENCE

Natural interference is defined as interference arising from a source other than man-made objects. Natural sources of interference include atmospheric disturbances, star activity, solar activity, and precipitation static. Atmospheric disturbances originate in the earth's atmosphere and are the most significant source of natural interference below 100 mc. Whereas star activity interference is measurable at frequencies up to 900 mc, solar activity is strong enough at 10,000 mc to be tracked by radar. Precipitation static is more pronounced in the LF and MF frequency ranges; however, VHF equipment may be affected under severe conditions. This section contains a detailed discussion of the phenomena, causes, and possible corrective measures associated with the various types of natural interference.

8.1 ATMOSPHERIC DISTURBANCES

Atmospherics that occur in the VLF range may be classified as follows:

a. Spherics which originate in and are propagated through the space between the earth and ionosphere.

b. Whistlers which originate in lightning discharges and are propagated through the ionosphere along dispersive paths.

c. VLF emissions produced by causes other than lightning, such as VLF hiss, discreet events of short duration and repeatable frequency time relation, and chorus consisting of overlapping VLF emissions.

The greatest source of terrestrial atmospheric disturbances is generated by thunderstorms, probably by the transformation of heat energy (released by condensing water vapor) into mechanical energy (manifested by winds and turbulence) and subsequently into electrical energy (revealed by the production and separation of charged particles). Two other possible sources of terrestrial atmospheric noise are: (1) storms of blowing particles that generate electric charges and potential gradients, and (2) random noise caused by high electrostatic fields between the ground and air immediately above it. On the average, the static produced by the latter two sources is normally much smaller than that caused by thunderstorm activity. However, when exceptionally large potential gradients exist, the noise produced by the resulting local corona discharge may exceed that due to distant thunderstorms. In general, the geographical source region of a very great portion of the static observed in middle and higher latitudes lies in the tropics or semitropics where the frequency of thunderstorms is the greatest.

The intensity of atmospherics depends upon location, season of the year, and time of day. The diurnal trend is absent or small during the winter season but increases as summer approaches. There is a rise and decline of noise intensity about the hours of sunset and sunrise, respectively. As the summer season approaches, the noise level increases at earlier hours in the day, until the maximum level is reached shortly after noon, local standard time. The daily cycle usually attains a minimum after sunrise and a maximum after sunset. Seasonally, the static intensity is higher in June than January. Some northern locations experience maximum level in February as a result of precipitation static caused by the increase in frequency and force of blizzards. The atmospheric noise level has been found to vary as a non-linear function of colatitude.

A Fourier analysis of a lightning impulse shows that it consists of a large number of components of different amplitudes and frequencies. Radio waves of different frequencies display different propagation characteristics; this applies equally to the different Fourier components of the impulses radiated by lightning flashes. Therefore, these different Fourier components travel via the ground, via the ionosphere, via the troposphere, or as an optical ray in exactly the same manner as other radio waves of corresponding frequencies. A generalized mathematical treatment of the problem based on Fourier analysis and statistics is not sufficient. The evaluations of the noise must also take into account the equipment which picks up the interference and the actual conditions under which the effect was produced.

8.2 STAR ACTIVITY

The intensity of radio stars, as received at ground level on meter wavelengths, is known to exhibit fluctuations. These fluctuations are analogous to the twinkling of optical stars on optical wavelengths and are often referred as "radio-star scintillation." Radio-star scintillation is believed to be caused by the irregularities in the earth's ionosphere. The irregularities have different radio-frequency refractive indices and thus present important effects on the propagation of radio waves.

Scintillation characteristics are markedly affected by the presense of auroras. During a period of auroral displays, scintillation has been observed at both lower and higher angles of elevation, characterized by a larger fluctuation index and a higher fluctuation rate.

8.3 SOLAR ACTIVITY

The radio emission from the sun is composed of a background thermal emission from the solar atmosphere, and bursts of radiation, sometimes very intense, which originate in localized activity areas on the sun. These bursts have complex characteristics which make classifications difficult at a single frequency. However, observations over a large continuous band of frequencies make it possible to base classification on the essential physical characteristics.

Early spectral observations established the fact that a great majority of the solar bursts belong to one of four types. The classification is natural in the sense that the various types can be interpreted in terms of different physical processes on the sun.

For the frequency band 100-580 mc, the assumption can be made that the solar radiation at a given frequency originates near a level in the solar atmosphere where the plasma frequency corresponds to that of radiation. *

A noise storm consists of a long series of short bursts continuing over hours or days. They are superimposed on a background of slowly varying enhanced radiation which has been described as "continuing" although it is possible that the background itself may be composed of a large number of overlapping bursts. Noise storms are normally spread over a large frequency band, but are rarely seen above 250 mc. On many occasions the bursts have bandwidths of a few megacycles and lifetimes extending from a fraction of a second to nearly one minute. At other times, bursts of bandwidths of nearly 30 mc and lifetimes of less than a second may predominate. Noise storms are caused by localized disturbances in the solar corona above active sun spot regions.

A slow noise burst appears as a narrow band of intense radiation which drifts gradually, and sometimes irregularly, towards lower frequencies. The spectra sometimes show the presence of harmonics, but are often so complex as to preclude such identification. The characteristic velocity of the solar disturbances which give rise to these slow bursts may be deduced from their rate of change of frequency. This velocity is in the order of 1500 km, and as it corresponds to that of the so-called auroral corpuscular streams, it has been suggested that the slow bursts are caused by the passage of the streams through the solar atmosphere. Alternatively, the bursts may be caused by acoustic shock waves resulting from explosions in the lower atmosphere. These waves, being propagated outward at some small multiple of the thermal velocity of the protons, would also be traveling at a velocity of the order of 1500 km.

Fast bursts, a very commonly occurring phenomenon, have durations of a few seconds and show exceedingly rapid drifts toward lower frequencies. The source velocity which would be deduced from the slope of the fast bursts is in the order of 50,000 km. It has, therefore, been suggested that the bursts are caused by the outward passage of solar cosmic rays. Alternatively, it has been proposed electron plasma shockwaves may provide the mechanism for the outward passage of the solar cosmic rays.

* The Radio Spectrum of Solar Activity; A. Maxwell, G. Swarup, and A. Thompson; IRE Proceedings, January 1958, and Radio Noise Spectrum, published by Harvard University Press, 1961

8.4 PRECIPITATION STATIC IN AEROSPACE VEHICLES

8.4.1 GENERAL

Precipitation static, as the name implies, is radio interference experienced when an aerospace vehicle flight path is through precipitation (wet snow or rain). Flight tests have served to tie down some of the principal causes of the interference, although there remain some aspects of it which could profitably be explored more fully. The interference is ordinarily intense in the low and medium frequency ranges, and often completely blocks some communication and navigation equipments. VHF equipment is not ordinarily affected, although some few reports of interference to VHF under severe conditions have been received.

Current knowledge of the causes of precipitation static indicates that high-speed jet type aircraft and guided missiles with their properly streamlined built-in dragless antennas should be relatively immune to such intereference when flying through wet snow or rain. Attainment of immunity is dependent upon development, specification, and engineering design based on a knowledge and proper consideration of the requirements of high corona thresholds. Continued development of superior methods of controlling aircraft-charging characteristics will gradually ease current design requirements.

8.4.2 CHARACTERISTICS OF PRECIPITATION STATIC

Precipitation static interference is characteristically broadband, with a continuous spectrum and produces a very loud "rushing" sound in the output of a receiver, similar to amplified "shot" noise. Since it can occur when the aircraft is "on instruments" and in need of radio communication and navigation facilities, the interference can present a serious flight hazard.

Flight investigations have established that disruptive discharges in air are responsible for practically all of the interference. These discharges produce steep-fronted impulses of relatively high peak-magnitude, and the interference may be severe when the average current transported by the discharges is quite small, of the order of microamperes. In general, such discharges arise as a result of charging effects produced by impact of airplane surfaces with precipitation, although they may also occur as a result of electric fields surrounding charged clouds. Impact charging may produce interference as a result of the charging of the whole aircraft or as a result of localized charging confined to the vicinity of the antenna, although continued improved antenna design and installation techniques has reduced the problem.

8.4.3 ATMOSPHERIC CONDITIONS FAVORABLE TO PRECIPITATION STATIC

Radio interference of an atmospheric nature was observed long before it became of interest in connection with aeronautics. Radio antennas were observed to collect static charges during snow and dust storms, causing severe interference with the reception of radio signals. Metal windmills on insulated towers and trucks on rubber tires accumulated high potentials when hit by particles of dry snow, sand, or dust. Aircraft, operated at much higher speeds through dry snow, sand, or dust, experience more rapid charging which results in more serious radio interference. The major portion of experimental work on precipitation static has been primarily concerned with problems of the charging of the aircraft as a whole, because effects of charge accumulation on the aircraft with external antennas. Increased use of integral antennas has placed more emphasis, in recent years, on problems of localized charging effects, although it is still necessary to give consideration to overall charging characteristics.

a. Dust, Sand, and Smoke

The impact of sand or dust particles against a metallic surface produces, by friction, a charge both on the dust particle and on the metal. If a positive charge is carried away by a sand blast, the metal becomes negatively charged. The same effect would result from the impact of any form of dry powder against a dry surface. The carbon particles which constitute smoke would therefore also be effective in producing a charge. Aircraft flying through the dust storms of deserts and drought-stricken areas have repeatedly become charged in this manner. The denser the smoke or dust cloud and the higher the relative velocity of the wind and the airplane moving through it, the greater the number of impacts per unit area per second, and therefore the more rapid the charging rate.

b. Ice Crystals, High Altitudes, Low Temperatures

Fine ice crystals, such as are encountered in driving snow at high latitudes or in the fine ice spicules which compose cirrus clouds, are an effective source of precipitation static. Such ice crystals occur only at low temperatures - ten degrees below zero Centigrade, or less. The height of cirrus clouds varies with the season and with latitude, but they do not usually form below 30,000 feet. Ordinary dry snow crystals are produced very readily in cold weather at intermediate altitudes. Flying through dry snow is one of the most common causes of severe precipitation static, a fact which has been demonstrated many times in flights to the Aleutians, and to Iceland and Greenland.

Aircraft flying through snow almost always become negatively charged, and discharges from charged aircraft, whether by corona or otherwise, consist of negative electricity. Laboratory tests in which dry snow has been driven at high velocity against insulated airplane surfaces have also resulted in the production of negative charges on the plane.

8.4.4 FRICTIONAL CHARGING BY IMPACT WITH WATER DROPS

While flight through dry snow or through the fine ice spicules of cirrus and alto-stratus clouds invariably results in precipitation static, flying through wet snow, glaze, or rain usually produces relatively little radio interference. For completeness, however, it is desirable to consider it briefly and to point out that there is no sharp line of demarcation between the various cloud formations, but, rather, a gradual shift from one form to the next. Furthermore, in the turbulence above frontal storms, dry snow, wet snow, glaze, and rain may all be present simultaneously at different levels.

a. Flight Through Wet Snow

Wet snow is encountered only in regions where the ambient temperature is near or slightly below the freezing point, and where the relative humidity is high. Frictional charging requires dry surfaces and low relative humidity. The impact between the plane and a snowflake produces energy, $E = mv^2/2$, where m is the mass of the snowflakes and v is the numerical sum of the components of their velocities in the direction of flight. The heat energy thus released melts more snow to increase the magnitude of the liquid film on the surfaces of the airplane. Snowflakes that remain solid after impact are carried away by the airstream together with excess raindrops, taking with them any charges previously left on the metal surfaces. Should the aircraft enter a region filled with wet snow with temperatures of its metallic surfaces below freezing, the snowflakes would form an ice coating over which the impinging snow would slide with only mild charging effects.

b. Flight Through Freezing Rain or Glaze

Where a frontal condition exists in which warm, moist air has been lifted above a colder air mass, snowflakes falling from higher levels are turned into fine supercooled raindrops, and then descend into the heavier and colder frontal regions below. An airplane, going through a misty and foggy region which separates the warm upper layers from the cold frontal air masses near the ground, becomes covered with moisture particles that are small and quite near the freezing point. Such moisture particles seldom are electrically charged to any great extent but easily become attached to all the metal surfaces of the airplane and produce an even more serious hazard by adding a heavy load of ice.

c. Flight Through Raindrops

Many laboratory experiments have been performed on raindrops falling at various velocities and under a large variety of electrical field conditions. It has been demonstrated that there must be an updraft of more than eight meters per second in a thunderstorm in order to support small hailstones and the larger sized raindrops. It has also been shown that raindrops falling with greater velocity than eight meters per second are torn to pieces by friction with the air through which they fall. Furthermore, these falling drops are charged, some positively, others negatively. The main body of a large drop retains a positive charge while the smaller droplets which have been torn away from it carry negative charges. It follows that the splashing effect of large raindrops against a metal surface moving at high speed must result in the breaking up of the raindrops encountered, the larger drops adhering to the metal structure and the smaller ones carrying negative charges away. Since the airplane is usually charged with negative electricity where precipitation static is present, the positive charges on the larger fragments of the raindrops tend to discharge the airplane and thus decrease the amount of radio interference due to precipitation static.

8.4.5 CHARGES INDUCED BY EXTERNAL FIELDS: THUNDERSTORMS

Precipitation static, quite independent from that resulting from frictional effects, although it may exist coincidentally, is also caused by the passage of the airplane through an electrostatic field such as that existing between two oppositely charged clouds. The greatest charge per unit area is built up around sharp edges or points such as antenna protrusions and propeller tips where such are present on the aircraft, and the edges of wings, ailerons and rudders. Maximum corona discharges, which are the source of radio interference, always occur from these sharply curved surfaces, whether produced by frictional impact or by induction from surrounding fields. Therefore, the aircraft designer avoids, wherever possible, sharp edges or points in order to minimize radio interference.

a. Effects of Altitude and Temperature

Charging by induction is limited to those synoptic conditions in which charged clouds produce an electrostatic field. Such fields may be vertical or horizontal, or a combination of the two. They are far from uniform and are constantly varying, especially under thunderstorm conditions.

It has been found experimentally that flight at the cirrus cloud level (about 30,000 feet) probably results in severe frictional charging but this altitude is usually free from electrostatic fields. Descent into the lower levels of nonturbulent clouds of the cirro-stratus, cirro-cumulus (about 25,000 feet), or alto-stratus type (about 12,000 feet) may relieve radio interference resulting from frictional charging as well as from charging by induction. This is particularly true of dense, fog-like clouds; since, if flight is through fog or vapor, there is no charging. However, depending somewhat on the height of the cloud above the flight level, discreet precipitation particles may be encountered and charging would still occur. The cirro-stratus level, depending on the latitude and the season, at 12,000 feet is approximately the upper limit for liquid raindrops, and at about this altitude they occur only together with freezing rain or snow at temperatures near 0°C (32°F). Above this approximate altitude, or freezing isotherm, snow is found which produces static interference due to friction. Below this altitude, icing conditions and induction charging in electrostatic fields between clouds, or between the clouds and earth, are encountered. In winter this freezing isotherm may be near the ground with frictional charging encountered from the ground level upward.

b. Effects of Turbulence in Thunderstorms

Turbulence is a prime characteristic of thunderstorms which in turn produce conditions for radio interference in aircraft as a result of induction and frictional effects. Where extensive local heating of air with high humidity results in a steep temperature gradient from the ground to the condensation level, a "heat thunderstorm" may develop. Adiabatic expansion of moisture-laden air in the ascending column causes precipitation with a resulting release of latent heat, which in turn warms the air and increases its velocity of uprush. By such means are produced the rapid convection currents necessary for the electrification of clouds. A "cold-front" storm results when cold, dry air underruns a mass of warm, moist air and lifts it to the condensation level, while a "warm-front" storm is produced when a mass of warm air is pushed up over a comparatively stationary mass of cold air. In each case, the conditions producing turbulence are present.

The typical storm in temperate regions is the cyclonic "low." The low-pressure areas which move slowly from northwest to southeast across the American continent are several hundred miles wide. The general direction of rotation in the Northern Hemisphere is counterclockwise. In such a general storm area, there is a warm front in the southeastern section and a cold front in the western and northwestern section. During the thunderstorm season, which extends from March to October in the temperate zone, the air in the warm front has originated in the Gulf of Mexico and adjacent land areas and has a high humidity. Its temperature is higher than the temperature of the surface air over which it flows. The cold front, on the other hand, consists of comparatively dry, cold air which has originated in the Canadian Rockies or in the plains toward the east. When such a mass of cold air rushes into the warmer surface air of the plains region, its approach is characterized by a long line of thunderstorms and corresponding turbulence. Such a "line-squall" may have a length of 400 to 500 miles and a height of 6 to 8 miles. In the tropics or near large bodies of water where the ascending air currents have an unusually large moisture content, the total height of the turbulent condition is usually less than 5 miles. In either case, it is obvious that a pilot will not be able to fly around the end of this storm area, rather than pass through its center, unless he happens to be near its periphery. It is also evident that he will seldom be able to fly over the top of such a line-squall without encountering precipitation static of the frictional type. Since the storm conditions are far from uniform along the squall line, he may be able to find a gap of minimum severity through which he may safely pass.

Nonfrontal thunderstorms are commonly called "heat thunderstorms" and may be encountered in both tropical and temperate latitudes. Heat thunderstorms usually reach their climax in late afternoon while line-squall storms may arrive at any time during the day or night. Heat thunderstorms are considerably less extensive than the other type and can ordinarily be circumvented without the plane getting seriously off course. Nonfrontal storms may, however, develop into general storm areas without the presence of any frontal phenomena. These constitute a special hazard to the pilot on cross-country flight under generally settled weather conditions.

c. Magnitude of Electrostatic Fields and Their Effects in Thunderstorm Areas

Laboratory tests show that the potential necessary to produce a spark between polished balls, 10 cm in diameter and 1 cm apart, at 76 cm of mercury pressure in dry air is about 30 kilovolts. For needle points, it is only 12 kilovolts, and is less at high altitudes where the pressure is lower. Field strengths of 400 volts per cm have been frequently encountered in test flights, and a field of 3400 volts per cm has been recorded just before an aircraft was struck by lightning. At 1000 volts per cm, a lightning stroke 3-miles long from cloud to cloud, or from cloud to ground, requires an initial difference of potential of about 500 million volts.

In frictional charging through snow or feeezing rain, or under dust-storm conditions, the same sign of charge, usually negative, is present over the entire plane and the density of charge at any given point on the aircraft remains essentially constant for comparatively long periods of flight. On the other hand, induced negative charges on the plane are found nearest the positively charged clouds, while the positive charges are found nearest the negatively charged clouds. In thunderstorm fields, both the sign and the potential gradient of the charges which cause induction vary rapidly with both time and the position of the plane with respect to the active portion of the storm. Resulting currents of an order of magnitude in the milliampere range surge back and forth from wing tip to wing tip and from nose to tail, all capable of producing extreme radio interference due to spark-over at poorly bonded joints.

The electrical capacity, in flight, of even a medium size airplane is quite small, about 780 μμf. Therefore, a current of 100 microamperes flowing into it would charge at the rate of 128 kilovolts per second. Sharp edges and points would therefore quickly reach the corona point when in the presence of strong electrostatic fields and tend to generate severe radio interference.

d. Lightning Discharges to Aircraft

Although the all-metal structure of aircraft afford protection to the crew against lightning discharges, a direct discharge to the aircraft may cause radio interference and damage to electronic equipment. Radio interference is at a maximum but usually of short duration due to the stroke itself. The after effects of the stroke may actually be more serious than the radio interference itself. The current in a lightning stroke may be as much as 100, 000 amperes. While the surface and framework of the plan can carry such a current without danger, even a small fraction of such a current, if it gains access to radio receivers, power supplies, and other electrical equipment, can easily fuse wires and put control apparatus out of operation.

The distribution of electrical charges in a thunder cloud is far from uniform, and authorities are not fully agreed on any general pattern. Whatever the distribution of electrical charges within the cloud, the potential difference between points at a distance equal to the wing span of an airplane undergoes rapid changes in sign and magnitude according to the part of the cloud in which the airplane is flying at any given moment. If it is in the region between maximum negative and maximum positive charge concentrations, the potential gradient is high. On the other hand, where the distribution of charges is of nearly the same concentration and of uniform sign, the maximum currents and potentials connected with a stroke through the plane cannot be very large. The probable minimum risk is taken by flying as high as possible above the active vortex of the storm or by flying to one side or the other of that vortex. When confronted with an extended line-squall storm with towering cumulus clouds that reach above the ice-spicule level, the pilot may be justified in flying under the cloud in the rain area between the cloud and the earth. This involves three hazards, the icing and induction-charging dangers, the possibility of intercepting a lightning stroke in the high potential gradient between the cloud and the earth, and the danger of encountering mountains or other obstructions.

8.5 METHODS FOR REDUCING INTERFERENCE FROM PRECIPITATION STATIC

Increased speed and higher altitude flying, the result of improved design and the development of powerful jet engines, both dictated by military and commercially economical necessity, has brought ever-increasing losses of precipitation-static type radio interference. The Military Services, aircraft manufacturers, and commercial flight operators have made persistent attempts to remove this hazard to aerial navigation. From numerous investigations, in flight and in the laboratory, there resulted a series of programs for "cleaning-up" airplane structure and design. These included the elimination of sharp-edged or pointed protuberances, the effect of various experimental coatings on the metal surface, and the bonding of separated parts of the aircraft structure. In addition, much research and development was done that made available dielectrically insulated antenna wire and fittings as well as devices for discharging static charges harmlessly from the aircraft while in flight.

Further details concerning the design of aircraft antennas and antenna fittings are presented in Volume III of this Handbook.

8.5.1 DESIGN OF AIRCRAFT STRUCTURES

From the standpoint of freedom from interference from precipitation static, the ideal shape for an airplane would be a smooth sphere. This being functionally impossible, there remained, as a means of a practical solution to the problem, reduction of the curvature of sharp-edged and pointed structures without interfering with their mechanical operation. This included elimination of protruding rivet-heads, projecting edges of sheet-metal, exposed surfaces of Pitot tubes and thermometers, all of which received attention. Since corona discharge begins at the point of sharpest curvature, it is necessary to reduce the curvature of all critical areas to the same value, so far as possible, in order to keep the potential of the whole aircraft uniformly high and to prevent breakdown below the operating voltage of static discharges.

The quantity of electricity, Q, on any metallic body is equal to the product of its electrical capacity, C, multiplied by the potential, V, applied to it, thus $Q = CV$. The electrical capacity of an insulated sphere in electrostatic units is equal to its radius, r, in centimeters; that is, $C = r$. Electrical charges distribute themselves equally over the entire surface so that $V = Q/r$, but if the sphere is surrounded by a medium of dielectric constant, k, then $V = Q/kr$. In the consideration of aircraft, k is approximately one for air.

The surface density of charge, s, equals the total charge divided by the area of the sphere; therefore,

$$s = \frac{Q}{4\pi r^2} = \frac{V r}{4\pi r^2} = \frac{V}{4\pi r} \qquad (1\text{-}11)$$

For any given potential, therefore, the charge density varies inversely as the curvature, 1/r. It approaches zero for a plane surface and becomes infinite at the tip of a sharp-pointed needle.

The atmosphere always contains considerable numbers of ions produced by ultra-violet light, cosmic rays, radio-active substances, engine exhausts, etc. Such ions are accelerated towards or away from intense fields, according to their signs, whether positive or negative, and may gain sufficiently high velocities to ionize more molecules by collision and thus initiate a corona discharge with its resulting static interference. Corona discharge generally begins at about 1/2 to 2/3 the potential required for a disruptive spark, although, depending on the geometry involved, it has been observed at 1/10 of this potential.

Laboratory experiements have shown that the breakdown potential between polished spheres, 1 inch in diameter, is approximately three times as great as for needle-points. One hundred kilovolts can bridge a six-inch gap between needle-points in dry air. Exact breakdown voltages have been shown to depend on materials used, temperature, and pressure.

The breakdown potential from a metal surface to the surrounding air depends on the density of the air, being a function of the mean free path of the molecules. For aircraft at high altitude the pressure decreases with the altitude while the temperature falls at the same time, increasing the air density. While these two factors affect the air density in opposite directions, the relations are not linear and the pressure effect predominated. At 5000 feet elevation, the breakdown potential is about eight-tenths (0.8), and, at 10,000 feet, sixty-seven hundredths (0.67) of that at sea level.

Because of the camber, or curvature, of the upper surface of an airplane wing, the air going over the top of the wing surface must travel farther, and hence have greater relative velocity, than the air passing the comparatively straight undersurface. According to the well-known theorem of Bernouilli, the greater the speed of the air over a surface, the less the pressure. Thus is produced the important lifting effect which keeps the plane in the air but incidentally increases the tendency to go into corona over these areas of low pressure. Because of the pitch and relatively high speed of the propeller tips on propeller driven aircraft, they are likewise surrounded by a region of reduced pressure as they plough through the ambient snow and ice particles. Consequently, the propeller tips tend to burst into corona almost as soon as other sharp metal points. Obviously, both of these Bernouilli effects are inherent and essential to the functioning of the plane. The change in pressure on the wing surface due to camber is not serious because of its large radius of curvature. In cases where propellers are involved, the remedy is to effectively bond them to the fuselage to keep their potential relative to the airframe as low as possible, and to keep their surfaces smooth and free from foreign matter.

9. INTERFERENCE IN INFRARED APPLICATIONS

Every object that has a temperature greater than absolute zero radiates electromagnetic energy, much of which lies in the infrared spectrum. While dealing with the subject of electromagnetic interference, its effects upon infrared systems must therefore be included. Infrared detection systems are all in some degree affected by "noise" interference, creating many problems for the scientist and engineer to solve. Present day concepts and their application apply mainly to military situations. The advance in technology concerning infrared scanning, detection, filter techniques, and display, dictates the advance in sophistication and capability of military infrared systems. There are many applications of infrared principles in present day systems and more will be forthcoming as the state-of-the-art improves. Those to be mentioned involve interference problems that tend in varying degree to encroach upon optimum operation of the particular application. One area in which infrared detection principles have important application is navigation and missile guidance. Navigational devices capable of determining precise location are extremely desirable. Such aids may lock-on and track the sun, the moon, certain stars and planets. Such precisely located celestial bodies relative to time and with relation to sensitive vertical determining means enables knowledge of accurate position on the earth. Other applications of infrared principles are in the fields of photography, mapping, and certain communications systems. There are other classified applications of infrared techniques that will not be discussed here. All applications suffer from the same "noise" problems, both internal and external.

9.1 INFRARED DETECTORS

Infrared detectors may be divided into two general classes: photon sensitive and thermal sensitive. The form of the radiant energy sensed by both types is the same; the distinction is in the mechanism of detection. When a quanta of energy is absorbed in a photon sensitive or photoconductive material, an electron is excited, causing a change in electrical characteristics. On the other hand, thermal detectors absorb radiation in the lattice structure, causing heating in the lattice to change the electrical or mechanical properties of the lattice. A detailed discussion on different types of infrared systems is contained in Volume IV, the subject of interference being mainly dealt with here.

9.2 INFRARED INTERFERENCE CONSIDERATION

Signal to interference ratios at the output of an infrared detector determine the absolute detectivity or sensitivity of the detector. The effectiveness of the detector is improved by increasing the signal or reducing the "noise." On the other hand, effectiveness is reduced by raising the "noise" level at the detector or reducing the signal in some manner.

Interference in the detector may be classified, in general, by the source. In any physical situation, some amount of background radiation will be present. The background radiation has a random fluctuation due to the random arrival of photons. This random variation sets the ultimate limit to the smallest detectable signal.

The desired signal also has a random variation in intensity. In present applications, the signal variation is much smaller than background variations. However, if signal or target fluctuations were increased, the effective interference level in the detector would also be increased.

Interference is also generated in the detector element itself by a variety of mechanisms. Although the goal of infrared technology is to reduce this internal interference below the level of the background interference, this ideal is not easily realized, and virtually all presently operational infrared detectors are limited by the internal "noise." The term Blip has been coined to describe the background limited photoconductor, where the predominant interference is background generated. Unwanted "noise" may also be generated in the amplifiers between the detector element and the display devices. Careful design will reduce interference from this source to a negligible level.

The interference at the output of a detector is probably best described by the mean-square fluctuation of unwanted signal about its mean value per unit bandwidth. Although the output of some types of detectors is not electrical (i.e., the Evaporograph), a voltage or current analog for both desired signal and unwanted "noise" may usually be found. To be general, then, voltage or current will be used as the independent variable in the following discussion of "noise."

It is assumed that the function x(t) satisfies the condition

$$\lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} |x(t)|^2 \, dt < \infty \qquad (1\text{-}12)$$

The power spectral density S(f) of the function x(t) is defined as the Fourier transform of the autocorrelation function, $\mathcal{R}(\tau)$:

$$S(f) = \int_{-\infty}^{\infty} \mathcal{R}(\tau) \exp(-j\omega\tau) d\tau \qquad \omega = 2\pi f \qquad (1\text{-}13)$$

This time autocorrelation function is given by

$$\mathcal{R}(\tau) = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} x(t+\tau) x^*(t) dt \qquad (1\text{-}14)$$

where $x^*(t)$ is the complex conjugate of x(t).

The spectral density gives the distribution in frequency of the power of a signal or a "noise." The integral of the spectral density is the average power of x(t):

$$\mathcal{R}(o) = \int_{-\infty}^{\infty} S(f) df = \lim_{T \to \infty} \frac{1}{2T} \int_{-\infty}^{\infty} |x(t)|^2 \, dt \qquad (1\text{-}15)$$

If x(t) is a stationary ergodic random process, the time average is equal to the ensemble average with probability one, and

$$\mathcal{R}(\tau) = E\left[x(t) x^*(t-\tau)\right] \qquad (1\text{-}16)$$

A problem which frequently arises in experimental work is the statistical estimation of the spectral density of a stationary random process when a finite time segment is known. A function of the independent variable is desired which converges in some sense to the true value of the spectral density. An often used and intuitively appealing function for spectral density estimation is the periodogram, defined below. However, unless the limiting process is carefully defined, the periodogram gives a very questionable estimate of spectral density.

For a random variable x(t), defined in the time interval $0 \le t \le T$, let

$$X_T(f) = \int_0^T x(t) \exp(j\omega t) dt \qquad \omega = 2\pi f \qquad (1\text{-}17)$$

Then,

$$S(f, T) = \frac{\left|X_T(f)\right|^2}{T} \tag{1-18}$$

is called the periodogram, giving the frequency decomposition of the power of x(t) in the defined time interval. The power spectral density may be estimated by

$$S(f) = \lim_{T \to \infty} E\left[S(f, T)\right] = \lim_{T \to \infty} \frac{E\left[\left|X_T(f)\right|^2\right]}{T} \tag{1-19}$$

The ensemble average must be carried out before the limit is taken. This relation will then hold for wide-sense stationary functions.

The various types of "noise" encountered in infrared detectors and the corresponding power spectral densities are considered below. The total spectral density in a detector is the sum of the spectrums of the "noise."

9.2.1 RADIATION INTERFERENCE

The random variation in the arrival of photons at the detector gives rise to radiation or photon "noise." This interference is present in all detectors, regardless of the mechanism of detection. In a very narrow band of radiation frequencies, Δf, the spectral density of the incident radiant power is

$$\Delta\left[W(f)\right] = \frac{hf\,\overline{P}\,\Delta f}{1 - \exp(-hf/kT_r)} \tag{1-20}$$

where:
- h = Planck's constant
- f = the radiation frequency
- $\overline{P}$ = the mean or average incident power
- k = the Boltzmann constant
- T_r = the radiator temperature

For a blackbody at a temperature T_r, this expression is integrated over the radiation frequency range to give the radiant spectral density:

$$W(f) = 4KT_r\overline{P} = 4Ak\sigma T_r^{5} \tag{1-21}$$

where: A = the effective area of the detector
σ = the Stefan-Boltzmann constant of radiation

It is assumed that the detector is immersed in the blackbody radiation field.

To obtain the power spectral density at the output of the detector, the radiant spectral density W(f) is multiplied by the square of the responsivity R(f):

$$S(f) = R^2(f)W(f) \tag{1-22}$$

Responsivity is defined as the ratio of input power to output voltage of the detector.

The "noise" spectral density has been defined for negative frequencies, so the total radiation "noise" power in a bandwidth B for a blackbody radiator is given by

$$P = \int_{-\infty}^{\infty} S(f)df = \int_{-f_o - B/2}^{-f_o + B/2} S(f)df + \int_{f_o - B/2}^{f_o + B/2} S(f)df \tag{1-23}$$

$$= 2\int_0^B 4Ak\sigma T_r^{5}df = 8ABk\sigma T_r^{5}$$

9.2.2 NYQUIST-JOHNSON INTERFERENCE

Nyquist-Johnson interference is a thermal agitation interference present at the output of all infrared detectors. This interference is due to the random movement of electrons in the detector material, and is commonly referred to as "resistor noise."

The spectral density of the "resistor noise" is given by

$$S(f) = 2kT_r R \frac{hf/kT_r}{\exp(hf/kT_r) - 1} \tag{1-24}$$

where R is the real part of the electrical impedance of the detector. In most cases, hf is small compared with kT_r, and the fractional term may be neglected.

9.2.3 GENERATION-RECOMBINATION INTERFERENCE

The random variation of the number and lifetime of thermally generated carriers causes a generation-recombination interference at the output of semiconductor detectors. An analytic expression for this interference is very difficult to obtain for any semiconductor configuration other than the most simple.

For intrinsic or very lightly doped semiconductor, the generation-recombination "noise" may be expressed as

$$S(f) = \frac{2\overline{I}^2 \tau}{n_0 \left[1 + (2\pi f\tau)^2\right]} \qquad (1\text{-}25)$$

where: $\overline{I}$ = the mean detector current
τ = the mean carrier lifetime
n_0 = the mean number of carriers in the semiconductor

9.2.4 MODULATION INTERFERENCE

The name "modulation noise" has been given to interference in semiconductors which is due to variations in conductivity from causes other than generation-recombination interference. The mechanism is not well understood, and the spectrum expression is derived from experimental measurements. The spectral density is approximately

$$S(f) = \frac{CI^2}{f\alpha} \qquad (1\text{-}26)$$

where: α = a constant of about unity
C = a constant of dependent on the material

In germanium, this interference has been observed over a frequency range from $2 \cdot 10^{-4}$ cps to $4 \cdot 10^6$ cps. This interference is sometimes referred to as 1/f interference, and is found in semiconductor detectors including bolometers.

9.2.5 CONTACT INTERFERENCE

Contact interference differs from modulation interference in that it arises at the terminal contacts of the semiconductor. The spectral density also varies approximately with the inverse of frequency

$$S(f) = CI^{\beta}/f^{\alpha} \qquad (1\text{-}27)$$

where: β = usually about two
α = approximately one

The interference in carbon button microphones is one example of contact "noise."

9.2.6 FLICKER INTERFERENCE

Flicker "noise" is a low frequency interference found in photoemissive detectors in excess of the shot "noise." The mechanism of generation is not well known, but is thought to be due to variations in the work function at the thermionic surface.

For temperature limited emission, the interference due to flicker "noise" may be written

$$S(f) = e\bar{I}^2 \frac{F}{fA} \qquad (1\text{-}28)$$

where: e = the absolute charge on an electron
$\bar{I}$ = the mean current
A = the emissive surface
F = a characteristic constant of the emissive surface

9.2.7 SHOT "NOISE"

The random emission of electrons in a photoemissive surface causes shot "noise." The usual expression for the spectral density is

$$S(f) = e\bar{I} \qquad (1\text{-}29)$$

for the temperature limited condition. Space charge effects will reduce the magnitude of the spectral density, and capacitance effects will set a high frequency limit to the bandwidth. For a filter with a bandpass B centered about f_o, the total "noise" power due to shot "noise" will be

$$P = \int_{-\infty}^{\infty} S(f)df = \int_{-f_0 - B/2}^{-f_0 + B/2} S(f)df + \int_{f_0 - B/2}^{f_0 + B/2} S(f)df \tag{1-30}$$

$$= 2 \int_{0}^{B} e\bar{I}\, df = 2Be\bar{I}$$

9.2.8 TEMPERATURE "NOISE"

In thermal detectors, "noise" arises from the random fluctuations in the temperature of the detection device. The density spectrum of the temperature variation is given by

$$W(f) = \frac{2kT^2 g(f)}{g^2 + (h + 2\pi fC)^2} \cdot \frac{hf/kT}{[\exp(hf/kT) - 1]} \tag{1-31}$$

where: C = the thermal capacity of the active element

$g(f)+jh(f)$ = the complex thermal conductance of the detector to the environment

$W(f)$ = the mean-square fluctuation of the temperature about its mean value per unit bandwidth

To obtain the output power spectrum, $S(f)$, the temperature spectrum must be multiplied by the square of the factor $Q(f)$ for conversion of temperature change to output voltage change:

$$S(f) = Q^2(f)W(f) \tag{1-32}$$

9.3 ATMOSPHERIC PROPERTIES INFLUENCING INFRARED TRANSMISSION

In most ground-based infrared applications, the transmission and attenuation characteristics of the atmosphere place the final limit on the usefulness of detection systems. In addition, the background radiation is determined in many cases by the absorption and emission characteristics of the atmosphere. These atmospheric characteristics are briefly reviewed below.

The three gases forming the largest portion of the atmosphere, nitrogen, oxygen, and argon, have no absorption bands in the infrared wavelengths. The molecules of two of these gases, N_2 and O_2, are homonuclear diatomic molecules, without electric dipole moments, and do not have infrared rotation-vibration bands. However, these nonabsorbing molecules will collide with molecules which do have absorption characteristics, producing pressure or Lorentz broadening of the absorption bands.

In the lower atmosphere, H_2O, and CO_2, are the principal absorbers of infrared radiation. The amount of water vapor in the atmosphere is highly variable, ranging from 1% to 0.001% by volume. Water vapor concentration is usually expressed in cm of precipitable water; that is, if the transmission path is replaced by a tube of uniform cross section, and the water vapor all condensed to one end of the tube, the depth of the water is a measure of the water vapor in the absorption path. At ground level, this measure may be calculated from the temperature and humidity.

Carbon dioxide is more uniformly distributed throughout the atmosphere, with an average concentration of 0.033%. Near the ground, however, this amount may vary as much as 50%, depending upon the CO_2 recently discharged into the air mass from various sources. The carbon dioxide concentration is customarily expressed in atm·cm; that is, the centimeters of path length of CO_2 in the path length if it were present alone at one atmosphere pressure.

Ozone will also absorb infrared radiation, but is not usually found in any concentrations near the surface. Other minor absorbing constituents, CH_4, N_2O, and CO are fairly uniformly distributed throughout the atmosphere, but have a relatively small absorbing effect at surface levels.

From laboratory and open-air studies, the infrared region between 0.7 and 15 microns may be divided into eight windows, or transmission regions between the principal absorption bands. To a good approximation, the window transmission data may be expressed by the equation

$$T = -k \log w + T_0 \tag{1-33}$$

where: T = the average transmission of a window
w = the water vapor concentration of the transmission path
k & T_0 = empirical constants for each window

Table I gives the window wavelengths and values for the constants.

Table I. Window Regions in the Infrared

Window Wavelengths microns	k	T_o
0.72-0.92	15.1	106.3
0.92-1.1	16.5	106.3
1.1-1.4	17.1	96.3
1.4-1.9	13.1	81.0
1.9-2.7	13.1	72.5
2.7-4.3	12.5	72.3
4.3-5.9	21.2	51.2
5.9-14	complex	complex

As another approach, the total absorption of the bands may be determined rather than the transmission through the windows between bands. For low resolution studies, particularly at ground levels, only the absorption by water vapor and carbon dioxide need be considered.

It has been found that the form of the absorption growth curve has two regions: one for weak bands where the absorber concentrations are low, and a second form for high concentrations. Where both concentration and pressure are varied, the total absorption may be expressed by

$$\int_{f_1}^{f_2} A(f)df = cw^{1/2}P^k \quad \text{(weak band)} \qquad (1\text{-}34)$$

$$\int_{f_1}^{f_2} A(f)df = C + D \log w + K \log P \quad \text{(strong band)} \qquad (1\text{-}35)$$

where f_1 and f_2 are the absorption band limits, w is the concentration of the absorber, P is the total pressure, and c, C, D, k, and K are empirically determined constants. For water vapor and carbon dioxide, the values of the coefficients are tabulated in the "Handbook of Geophysics for Air Force Designers."

The absorption regions of a gas, caused by quantum transitions in the molecule, might be expected to be monochromatic. Instead, collisions between the molecules of the absorbing gas and other gases cause a broadening of the absorption regions, thus changing the absorption coefficient of the absorption "line."

9.3.1 SCATTERING

In addition to the absorption of infrared radiation, particles in the atmosphere cause an attenuation of radiation due to scattering. The molecules of the atmosphere itself cause Rayleigh scattering, the mechanism producing blue sky and red sunsets. For all practical purposes, this scattering may be neglected for wavelengths beyond about 2 microns.

Aerosol particles in the atmosphere, such as smoke, haze, dust, and fog, with particle diameters of the same order of magnitude as infrared wavelengths, produce Mie scattering. With certain restrictive assumptions, the attenuation due to Mie scattering may be computed, but these assumptions are usually invalid in practice. However, an approximate attenuation coefficient may be computed as a correction factor to the transmission coefficient.

9.3.2 BACKGROUND RADIATION FLUX

At wavelengths near the center of a strong absorption band, the background infrared flux is approximately equal to the blackbody radiation at the temperature of the atmosphere. However, in the window regions, the background is largely determined by the emission of the physical background. In this case, to a first approximation, thick clouds and most types of ground cover radiate as a blackbody in the infrared.

At a given point in the atmosphere, the total radiation flux is the sum of the upward and downward fluxes. The upward flux is the sum of (1) radiation from the surface, multiplied by the transmissivity from the surface, and (2) the radiation from each layer of the intervening atmosphere as multiplied by the transmission between that layer and the measuring point. Since any strong absorber of radiation is also an emitter, the water and carbon dioxide molecules emit radiation at the frequencies of absorption. This emission contributes to the total flux.

Similarily, the downward flux is the sum of (1) radiation entering the atmosphere modified by the transmission coefficient from the point of entry, and (2) the radiation from each layer of the intervening atmosphere multiplied by the transmission between that layer and the measuring point. The random variations in background flux, caused by the random arrival of photons at the detector, set the ultimate limit on detector sensitivity.

ANALYSIS OF INTERFERENCE IN COMMUNICATION-ELECTRONIC SYSTEMS

CHAPTER 2

1. PLANNING INTERFERENCE-FREE SYSTEMS

One of the major problems in any communications or electronics system is that of receiving or causing interference. Military operations, for example, are highly dependent upon the capabilities of C-E (communications-electronics) systems for providing information and communications. Because any degradation of the C-E facilities will result in degradation of combat capabilities, C-E interference must be kept to a minimum. As the altitudes, velocities, and mobility of military units and vehicles increase, so too are the C-E systems increasing in complexity, power output, and radio spectrum utilization. Implementation of adequate interference prevention measures has become vital to combat capability. In time, interference aspects may predominate over all other considerations in engineering C-E systems.

In the past few years, the pressure of increased C-E capability requirements has caused intensified investigation of radio interference phenomena, sources, and measurements. Although there are many possible sources and causes of interference, most of the interference problems can be solved. It would appear that it is not necessary to "live with interference" if solution of interference problems is started in the basic design of the equipment and further work on the problems continued during installation and operation. Engineering measures leading to minimization of interference can be applied in each stage of the design-siting-installation-operation-maintenance process.

From the C-E system planner's point of view, there are three vital phases in planning an interference-free C-E system:

a. Prediction of interference based on known data and measurements concerning (1) system and equipment performance, (2) ambient electromagnetic environment, and (3) propagation and frequency data.

b. Engineering of the C-E system for maximum compatibility with the electromagnetic environment based on data obtained in a. above.

c. After system installation, the reduction of interference not previously predicted and eliminated.

As the C-E engineer approaches his system planning problem, he soon discovers that there are many limitations to his solution including such matters as cost, acquisition of real estate of the appropriate type, frequency assignments, and compatibility of the various equipments of the system both with the equipments of the system itself, and with other equipments of outside systems. All of these matters require detailed consideration and, for the most part, each is related to the others. The C-E planner must, therefore, conduct his planning activity in such a manner as to obtain an optimum system, using all of the information that can be obtained on the equipment, the frequencies to be used, the sites, and the system capability requirements.

Obviously it is desirable to adapt a systematic approach to the systems planning problem. Such an approach has many advantages, not the least of which is the saving of time in getting the system on a reliable operational status at the earliest possible date.

In order to properly plan an interference-free system, it is necessary to analyze the system actions and reactions which are the causes or results of interference. In general, this analysis pertains to signal sources, signal transmission, and signal reception.

2. METHODS OF SIGNAL TRANSMISSION

As far as the source is concerned, an interfering signal is sent into an impedance, which, in turn, may react back on the source to help determine the character of the signal. Whether or not the signal eventually reaches the receiver has no effect on the source. On the other hand, as far as the receiver is concerned, an interfering signal is received from an active network with a certain internal impedance. According to Thevenin's theorem*, it is only this impedance and the

*Thevenin's theorem states that, as far as the load is concerned, any two-terminal network containing a number of linear impedances and generators may be replaced by a single equivalent generator in series with a single impedance. The voltage of the equivalent generator is equal to the open-circuit voltage of the network, and the single impedance is equal to the impedance looking into the network with all the generators replaced by their internal impedances.

open-circuit voltage (or the short-circuit current) which need to be considered as far as the effect on the receiver is concerned. This section, however, is concerned with the way in which the interfering signal is transmitted from the source to the receiver; therefore, attention will be focused on the intervening network itself rather than on its input and output impedances.

In talking about networks (using this word in its most general sense), two points of view may be taken: viz., either the field or the circuit point of view. The field concept is by far the more general and is always applicable. The circuit concept is a special case, and, where applicable, leads to great simplifications. The ordinary circuit equations can be derived from Maxwell's equations, which describe the properties of the field under certain simplifying assumptions, the most important of which are the neglect of retardation effects, displacement currents, and phase variations of the current in a series circuit. All of these assumptions are excellent approximations at low frequencies, but become poorer as the frequency increases, and break down completely as the wavelength becomes comparable in magnitude with the circuit dimensions. In some cases it is possible to account for high frequency effects by introducing artificial circuit parameters, such as the radiation resistance of an antenna, while still adhering to the circuit concept; but in general there will be a frequency above which the circuit concept must be abandoned. Certain low frequency phenomena, such as the coupling due to stray capacitances and inductances, could also be treated better from the field point of view. However, since most engineers are more familiar with circuits, the circuit concept will be preferred in these cases. Nevertheless, the field concept will not be overlooked; and whenever the circuit concept proves inadequate, the field concept will be utilized.

Two circuits are said to be coupled when currents or voltages in one produce corresponding currents or voltages in the other. According to circuit concepts, two circuits may be coupled either by a mutual impedance or by a mutual admittance. Simple examples of mutual impedances are shown in Figure 2-1. A mutual impedance exists between two circuits when the current flowing in circuit 1 produces a voltage in circuit 2. The magnitude of the mutual impedance is the ratio of the open-circuit voltage of circuit 2, with all other voltage sources removed, to the current in circuit 1. A mutual admittance exists between a point of circuit 1 and a point of circuit 2 when the voltage between point 1 and some reference point (ground) produces a current to, or from, point 2, with this point connected to the same reference point. The magnitude of the mutual admittance is the ratio of the resulting current at point 2, to the voltage at point 1.

The mutual elements may be resistances, inductances, capacitances, or any series or parallel combination of these elements. Not all possible combinations, however, are equally important in the transmission of interference. When the coupling is of the mutual impedance type, the elements that must be considered most frequently are mutual inductance and the mutual impedance of a common ground return, which may arise, for example, from inadequate bonding. A good example of this is shown in Figure 2-2. Except for those cases of poor design or installation, a mutual resistance, capacitance, or self-inductance, i.e.,

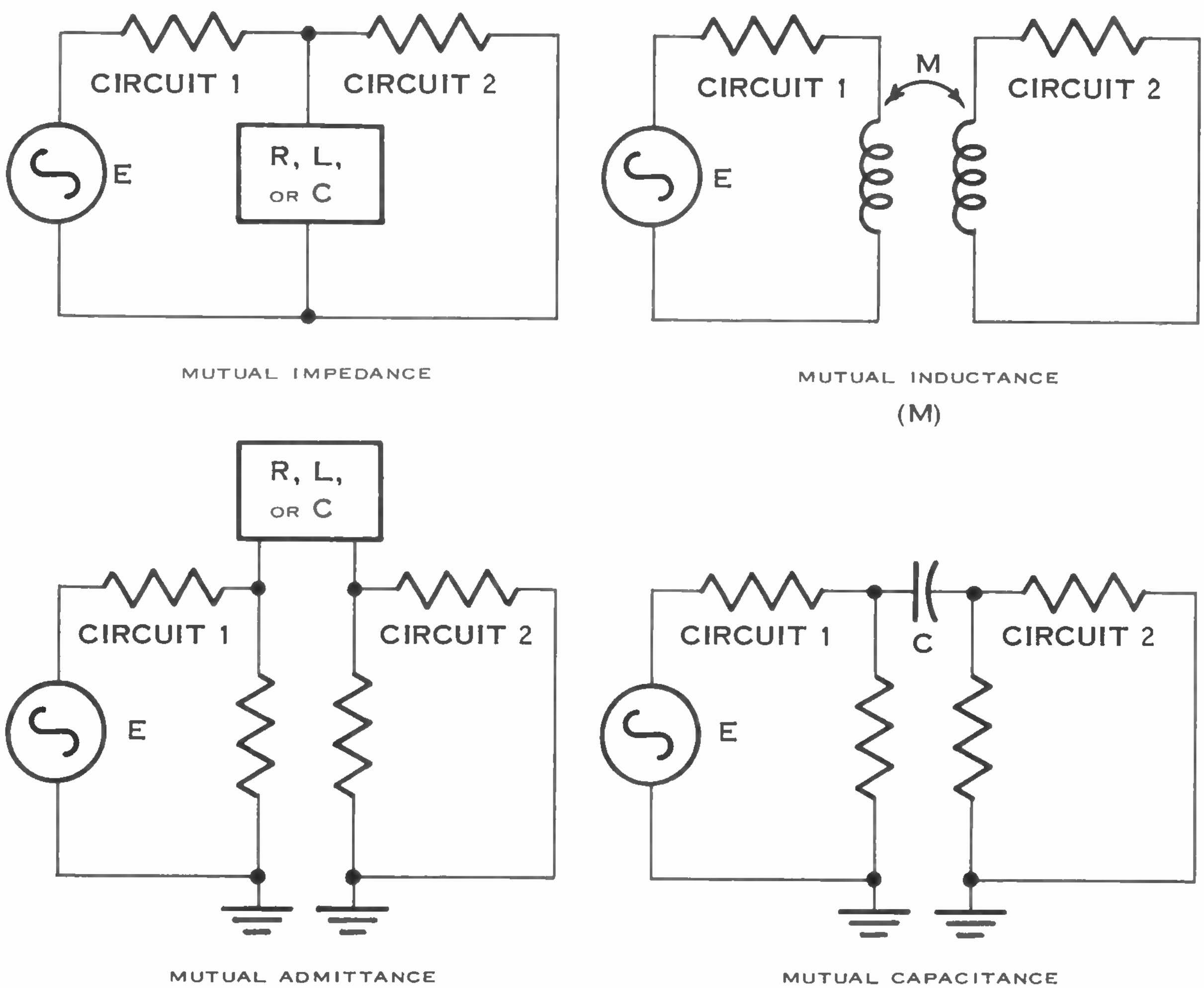


Figure 2-1. Coupled Circuits

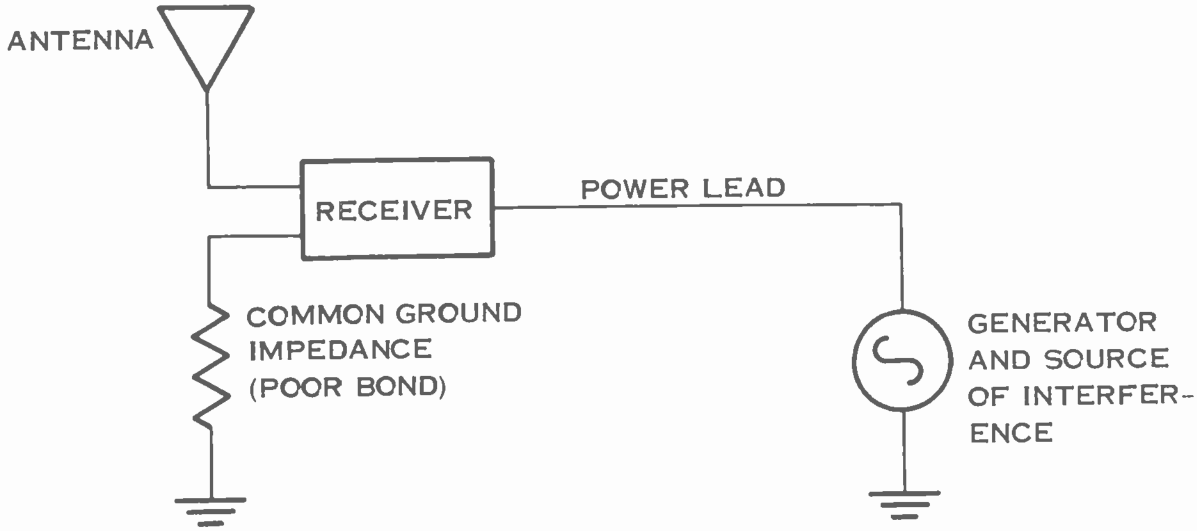


Figure 2-2. Mutual Impedance Coupling Through Common Ground Impedance

any such element that is common to both the circuit of the interference source and that of the receiver, will not normally occur in a practical system since it would serve no useful purpose. This leaves the mutual inductance to be considered, which may very well be present, and unfortunately, very often is. The presence of mutual inductance is nothing but a description, in circuit language, of the fact that the magnetic field set up by circuit 1 links with circuit 2, and therefore induces a voltage in it. But a magnetic field exists around any circuit carrying current, and its linking with other circuits is very difficult to avoid.

A somewhat similar situation exists when one considers the elements likely to become mutual admittances between two circuits. The ones most frequently encountered are mutual capacitances and the admittances offered by power or control cables connecting different pieces of equipment. Ideally, such cables should not carry any interfering currents, but when they do, the interfering signals are said to be transmitted by conduction. This method of transmission is important enough to be discussed separately in Paragraph 2. 3, not only because such cables may constitute a direct metallic connection between an interference source and a receiver, but also because cables leading to receivers from entirely interference-free equipment may have interfering voltages induced in them, which are then conducted to the receiver. Moreover, a cable which is connected neither to an interference source nor a receiver, may serve as an intermediate path for an interfering signal which uses other methods of transmission before entering and after leaving that cable. Except for this, the only element of mutual admittance that might be present without serving any useful purpose as such is that of mutual capacitance. The presence of mutual capacitance is nothing but a description, in circuit language, of the fact that an electric field exists in the vicinity of a conductor charged to some voltage, and that this electric field may induce a current in another circuit. Again, such interaction between two circuits is very difficult to avoid.

The effects of both mutual inductance, M, and mutual capacitance, C, increase with the frequency. This fact might seem contradictory if one considers that inductive reactance is $j\omega M$ and directly proportional to frequency, while capacitive reactance is $1/j\omega C$ and thus inversely proportional to frequency. The difference, however, lies in the fact that the mutual inductance is considered as a mutual impedance, i.e., the current in the influencing circuit produces a voltage in the influenced circuit, while the mutual capacitance is considered as a mutual admittance, inasmuch as it is the voltage in the influencing circuit that produces a current in the influenced circuit. In the first case, the voltage produced is proportional to the frequency because $V = IZ$ and $Z = j\omega M$. It is true that M itself is not constant, but increases with frequency, but this does not affect the basic argument. In the second case, the current produced is also proportional to the frequency because $I = VY$, where the admittance $Y = j\omega C$. The same conclusions would be reached on the basis of an analysis starting with Maxwell's equations. In the case of coupling through a magnetic field, the law of induction states that the voltage induced is proportional to the rate of change of current, and thus, for a sinusoidally varying current, is proportional to its frequency. In the case of coupling through an electric field, the resulting current is the time rate of change of the accumulated charge on the conductors. Since the amount of charge is proportional to the voltage, the resulting current varies as the rate of change of the voltage, and, for a sinusoidally varying voltage, varies directly as its frequency.

The interfering signal may also reach the receiver by radiation. This mode of transmission cannot be treated by circuit methods because the phenomena leading to radiation were expressly neglected in the transition from field to circuit theory. Hence, this must be treated as a separate and distinct method of transmission as is done in Chapter 4 of this volume.

2.1 MUTUAL INDUCTANCE

In considering mutual inductance, it is important to keep in mind that mutual inductance exists between two complete circuits. To talk about the mutual inductance between the parts of two circuits, i.e., between two wires, each of which is a part of a separate complete circuit, has no meaning. By its very definition, mutual inductance involves a current in circuit 1 and a voltage in circuit 2, both of which must be complete circuits. The voltage in circuit 2 is measured between two end points after that circuit has been broken, so that the sum of all voltages induced anywhere in circuit 2 is measured. In other words, the action of an induced voltage cannot be localized. No answer can be given to the question, "What is the mutual inductance between two long straight wires?", because a long straight wire is not a complete circuit. It is true that many expressions for mutual inductance are listed in various handbooks for many different shapes and configurations of pairs of open wires, but these expressions have meaning only if they are used to compute the individual contributions from the component parts of two circuits, which must then be added together to obtain the mutual inductance between the two complete circuits. For example, the mutual inductance between two rectangular loops may be obtained by computing 16 individual mutual inductances (one for the mutual inductance between each side of one loop and each side of the other) and adding all of them together. Each of the quantities may be computed by means of the expressions found in the handbooks.

In general, the determination of the mutual inductance between two circuits of arbitrary geometry is a very difficult task. A case of great practical importance for the purposes of this volume is that of two pairs of long straight wires all of which are parallel. A mathematical analysis of this configuration is carried out in Appendix III. The important results are given below.

The relative position of two pairs of long straight wires all of which are parallel may be specified in terms of the shortest distance, d, between the longitudinal axis of each pair, as shown in Figure 2-3, and the two angles, θ and φ, which the plane of each pair makes with the separation d. Mathematical analysis shows that, for values of d which are large as compared to the spacing between the wires of each pair, the mutual inductance varies inversely as the square of the quantity d. When d is small, the mutual inductance starts at a finite value and decreases at a much slower rate than would be given by the inverse square law. For every value of θ, there is one value of φ between 0 and

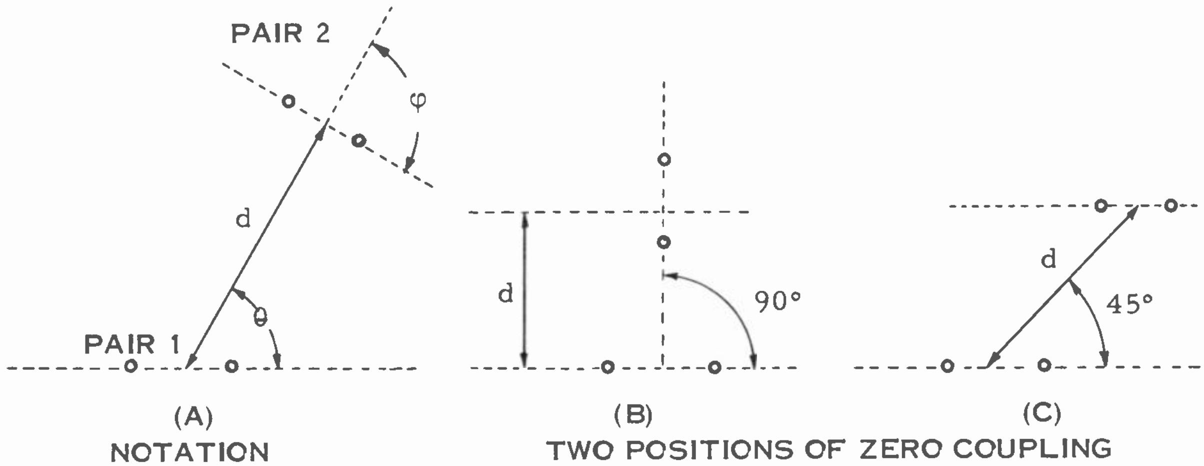


Figure 2-3. Mutual Inductance Between Two Pairs of Parallel Wires

180°, such that the mutual inductance is zero, and another value of φ, 90° removed from the first, such that the mutual inductance is a maximum. When d is large as compared to the spacing, the value for zero coupling is given by:

$$\varphi = 90° - \theta \tag{2-1}$$

This leads to the interesting result that, when d is in the plane of, or perpendicular to the plane of one pair, there is zero coupling when the planes of the two pairs are perpendicular to each other. But if d makes an angle of 45° with the plane of one, the coupling is zero if the planes of the two pairs are parallel. This is shown in Figures 2-3, (B) and (C). The situation is reversed if the coupling is to be a maximum.

This analysis was carried out for a rather special case. Two conclusions, however, can be expected to remain valid for other cases. First, for a pair of two-wire systems, the mutual inductance varies initially in some complicated way with the distance between the two circuits, but as the distance between them becomes large as compared to the dimensions of each, it will vary inversely as the square of the distance. This also applies to single-wire systems with ground returns because these may be reduced to two-wire systems by the method of images. This law, however, might be modified considerably by the presence of other conductors in the vicinity. Secondly, no matter how complicated the geometry of the circuits, for a given distance between centers, there will always be a position of circuit 2 relative to circuit 1 such that the coupling is zero. This last statement is obvious if one considers that the net magnetic flux must pass through circuit 2 in one of two directions, and that this direction can be reversed by reversing the position of circuit 2. In going from one position to the other, the flux changes continuously from maximum positive to maximum negative, and in doing so it must of necessity pass through zero for some position of circuit 2.

2.2 MUTUAL CAPACITANCE

While, according to the principle of duality, certain similarities exist between the effect of capacitive coupling and that of inductive coupling, there are also a number of significant differences. It is impossible to speak of mutual inductance except between closed circuits. On the other hand, the definition of mutual admittance introduces the concept of the mutual capacitance between just two points of two circuits. It is not very practical to use this concept because in all physical problems there will be capacitances between many points or regions of the two circuits. But for purposes of analysis, it does make sense, and it is much simpler, to use only one point at a time in each circuit. In magnetically coupled circuits, the induced voltage can be reversed by proper placement of one circuit with respect to the other. From this it could be concluded that there must be a position of zero coupling. On the other hand, in capacitively coupled circuits the direction of the current flowing to or from a point in one circuit depends only on the polarity of the voltage in the other. Therefore, it cannot be reversed simply by manipulation of the relative positions of the two circuits. From this it may be expected, and analysis shows it to be true, that capacitive coupling cannot be reduced to zero by rotation of one circuit about its center.

If the case of two long straight parallel wires is analyzed, it is found that the capacitance between these wires varies inversely as the logarithm of the distance between them. This would indicate that, for large distances of separation between two parallel wires, the capacitance between them decreases much more slowly with distance than the mutual inductance. This conclusion, however, cannot be generalized for other systems, because of the law of variation with distance for capacitance depends very much on the shape of the wires and their relative positions to other metallic objects in the vicinity. A general statement which can be made is that the capacitance between two points in two circuits decreases with distance somewhat more slowly than according to the inverse square law.

The definition of mutual admittance indicates that two circuits involved must have a common ground connection in order for a current to flow through the mutual element. It must not be concluded that the effects of mutual admittance may be prevented by eliminating all ground connections from one circuit. Practically, isolation is quite impossible without complete shielding (which would reduce the mutual admittance itself to zero) because even in the absence of a metallic connection to a common ground, there is always capacitance to some metallic object, which may be the airframe, providing a return path for the radio frequency current through the mutual element.

2.3 CONDUCTION

Whenever there is a direct metallic connection between two circuits, and in addition a return path, a conduction current may flow between these two circuits. The return path may be another metallic lead, or a mutual capacitance, or most frequently the metallic structure of the aircraft or missile acting as the ground return. The magnitude of the resulting current depends both on the potential difference between the points of exit and entry in the exciting circuit, and on the total loop impedance between the same points. It is important to remember to include the impedance of the return path as well as the impedance of the direct connection when computing this total loop impedance. If all impedances involved are linear, the current may be computed by a simple application of Ohm's law, $i = E/Z$, in accordance with the circuit concept.

The most common example of this is the transmission of interfering signals through the power and control leads, both out of the interference generators and into the receivers. The circuit shown in Figure 2-4 illustrates how an interfering signal may be transmitted from a motor into a receiver if both are connected to the same power supply.

The field concept is just as applicable to the case of conduction as to the cases of mutual inductance and capacitance discussed above. Strictly speaking, the energy associated with the interfering current is transmitted not through the metallic leads, but through the fields surrounding these leads, the leads acting only as guides. Yet it is precisely here that the circuit concept leads to the greatest simplifications and is, therefore, most frequently used. But in the discussion of radio interference it is often very useful to keep the field concept in mind, because it draws attention to the fact that there are always strong interfering

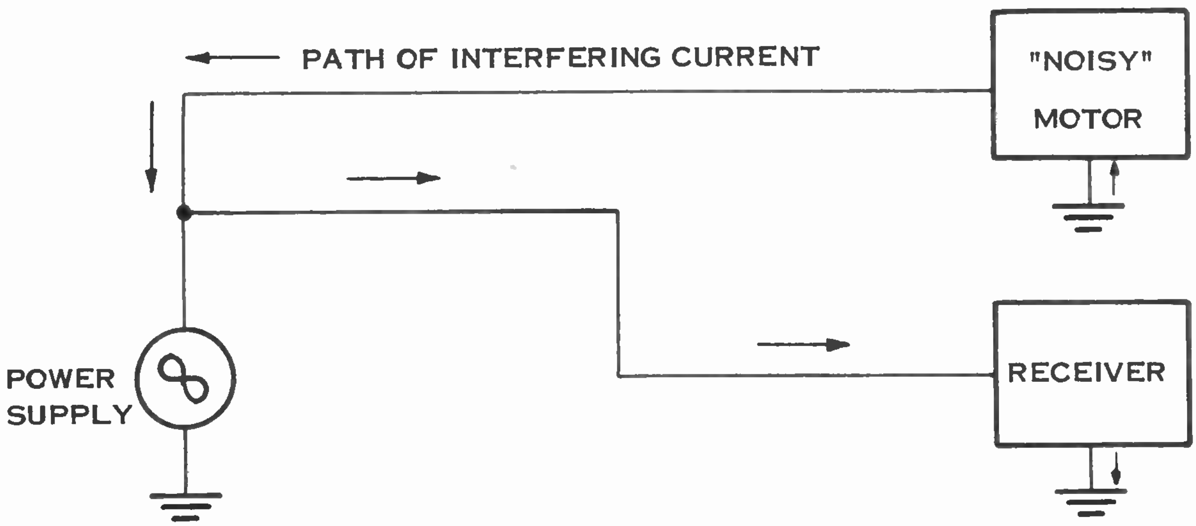


Figure 2-4. Transmission of an Interfering Current by Direct Conduction

fields in the vicinity of conductors carrying interfering currents. In fact, many troublesome interference problems are caused by bundling power or control cables leading to a receiver together with leads carrying interference currents, thus exposing them to strong interfering fields.

As an example of a case that should be treated by the application of the field concept, rather than the circuit concept, consider a cable connector with an unused pin. The pin does not actually connect two circuits, yet it may carry interfering conduction currents by having one end act as a receiving antenna and the other end as a transmitting antenna. Here the pin should be considered as a waveguide which transmits an electromagnetic wave from one side of the connector to the other.

3. INTERFERENCE MANIFESTATIONS IN RADIO RECEIVERS

The preceding has indicated some of the sources that can create extraneous signals in a communications system. These signals may assume many different forms such as broadband random interference, impulse interference, fundamental and harmonics, any other signal carried through the system lines or radiated by them, spurious signals created by transients, and extraneous signals created by mixer action. For these signals to manifest themselves as interference, they must be coupled in some manner into a system component and interfere with its operation.

One of the most common systems components in which interference becomes a problem is a radio receiver. The effect of the signals on the receiver is determined by the susceptibility of the receiver to the frequencies present. The signal may enter the receiver through the antenna, antenna lead-in, or antenna receptacle; through the power or control leads; through the output leads connecting the receiver to earphones, intercommunication systems, scopes, or other indicating devices; and through the casing enclosing the receiver, or any joints or openings in the casing. The effect on the receiver will be greatest if the interfering signal enters in the same way as the desired signal, because then it will immediately act on the most sensitive part of the receiver and may mix with the desired signal to such an extent that the separation of the two signals becomes practically impossible. For most receivers this most dangerous point of entry is through the antenna and its associated hardware. In others, such as a remotely controlled actuator, an interfering signal may enter through the control lead to produce a false actuating pulse.

If the interfering signal enters in a way different from that of the desired signal, the path of entry becomes just another link in the transmission system through which the interfering signal is moving. Eventually, it must enter the same path as the desired signal and mix with it. Otherwise it cannot reach the output stage of the receiver and cause undesired response or malfunctioning. Thus, if it enters through the power lead, it is transmitted into the receiver by conduction and may then enter the path of the desired signal, say, through inductive coupling of a power transformer with a radio frequency coil, or through capacitive coupling from the filament to the grid of an amplifier vacuum tube. If it enters through the headphone leads, it is again transmitted into the receiver by conduction and may then couple, within the receiver, with a sensitive circuit. If it enters through the casing or any openings in the casing, it is transmitted usually by capacitive coupling. Occasionally, it may be transmitted by inductive coupling as, for example, when an interfering field outside induces currents in the casing, which, in turn, produce an interfering magnetic field inside, or, in some cases, even by radiation. In all these cases, the problems encountered are truly problems of transmission as treated in this section. One-half of the proper design of receivers for interference-free operation is simply design for minimum transmission of interference into the receiver.

The remaining portion of this section will deal with the effect on the receiver of the interfering signals which have, in one way or another, entered the path of the desired signal in the receiver.

3.1 THE NUISANCE VALUE OF INTERFERING SIGNALS

The final effect of any interfering signal is to make it difficult or impossible for the receiver to function properly. The extent to which an interfering signal adversely affects the proper functioning of the receiver is called its nuisance value. Depending on the type of receiver, the effect may be complete inability to receive a message, or false indications of navigation instruments, or some such drastic "malfunctioning" as the premature explosion of a warhead triggered by a spurious control impulse. At any rate, in order for the interfering signal to have a nuisance value, either it must eventually contain one or more frequencies within the normal output range of the receiver, or it must be capable of preventing the receiver, or at least one of its stages, from functioning properly. In other words, the interfering signal must be capable of either producing effects similar to the desired signal or of preventing the desired signal from having its normal effects.

Normally, to produce these effects at the output of the receiver, the interfering signal must contain frequencies within the acceptance band of the receiver. But it must be remembered that the acceptance band of a receiver may be much larger for interfering signals, which may be of any magnitude, than for desired signals, which normally do not exceed a certain level. Most receivers have input circuits that behave somewhat like bandpass filters. Ideally the response to frequencies outside the passband is nil, but practically the attenuation in the attenuation bands is never infinite. Even though it will completely suppress all signals, outside the passband, that are of the same order of magnitude as the desired signal, it may still permit very strong interfering signals which lie entirely outside the passband to enter and to be transmitted past the first input stage. Once they have gone as far as that, they may produce one or more of the following three effects: (1) they may be strong enough to be transmitted directly to the output in spite of further attenuation in succeeding selective stages such as the intermediate frequency stages of superheterodyne receivers, (2) they may be strong enough to overload one or more stages thus making the receiver inoperative, and (3) they may combine with other signals in a nonlinear element in such a way as to produce a new frequency that is within the band of acceptance of the receiver. For example, they could combine in the mixer tube with a harmonic of the local oscillator in a superheterodyne receiver to produce a signal of the intermediate frequency. This last effect is sometimes incorrectly called "cross-modulation," a term that properly applies to the more complicated effect of a carrier being modulated by the modulating frequencies of another carrier nearby.

When the final output of the receiver is an aural or visual signal to be interpreted by a human operator, it is extremely difficult to assign any quantitative measure to the nuisance value of an interfering signal. In the final analysis, it is the operator who must decide what the real nuisance value of such a signal is, under the worst possible conditions. Certain general statements may, however, be made. If, in an audio receiver, the frequency spectrum of the interfering signal is fairly evenly distributed over the entire audio range, the result is usually considered to have a very "high nuisance value" by most operators. This is the case when the output consists of the crashes and background rumbles characteristic of atmospheric disturbances, or the pops and cracking sounds found in ignition interference, or the hash sounds usually accompanying interference from commutators. If, on the other hand, the interfering signal contains only one, or a very few frequencies, the resulting steady note of definite pitch may not be very bothersome at all to some operators unless it is extremely loud. For example, a 60 or 400 cycle power line hum of moderate strength, while unpleasant, may not impair the intelligibility of the desired signal to an appreciable extent. However, even this type of interference must be avoided because two or more "moderate" interfering signals may easily combine to produce a level of interference sufficient to drown out all desired signals. In general, the nuisance value of an interfering signal increases with the number of frequencies contained in it, but the effect of each frequency depends on the sensitivity of the individual operator. This sensitivity usually increases gradually for the lower frequencies, reaches one or more maxima somewhere between 700 and 3000 cycles per second, and then decreases to zero at about 15,000 cycles per second. Individual differences between operators are not quite so pronounced when the receiver is a radar scope; but here, too, the skill and experience of the operator play an important part in determining the nuisance value of an interfering signal during actual operations.

3.2 PULSE LENGTHENING

Any interfering signal that enters the receiver through the antenna or antenna lead-in must pass through the same stages of the receiver as the desired signal before it assumes its final form. During its passage through these stages, it may be modified both linearly and nonlinearly. It experiences nonlinear distortion whenever the output of a stage contains frequencies that were not present in the input. The appearance of intermediate frequencies in the mixer stage of a superheterodyne receiver is an important example of this. Linear distortion occurs whenever different frequencies experience different attenuations and different equivalent velocities of propagation. It is here that the difference between the desired and interfering signals is usually most important.

The input stages of a receiver are designed to pass the desired signals with a minimum of distortion of any kind. The desired signal usually contains frequencies within a definite range, say, a carrier and two sidebands in amplitude modulation, or a carrier and about 20 to 30 sidebands in frequency modulation. A well-designed receiver will pass a signal whose spectrum is confined entirely to its band of acceptance with little or no distortion. On the other hand, a signal whose spectrum extends considerably above and below its acceptance band may be modified beyond recognition by the same receiver. Most interfering signals extend over a frequency spectrum much larger than the acceptance band of the receiver. Therefore, the waveform of the interfering signals will be considerably modified in its passage through the receiver.

The most frequent and most important effect of this kind is the lengthening of interference pulses commonly referred to as "shock excitation" or "ringing." It occurs whenever a pulse of short duration enters a circuit whose bandwidth is narrower than the frequency spectrum of the pulse. In that case, the pulse at the output will be much longer than that at the input and considerably modified in shape. If the circuit has the characteristics of a resonant circuit, and if the pulses occur comparatively far apart, the output pulse will consist of a damped oscillation at the resonant frequency. The circuit has been excited by the "shock" of the pulse and, as a result, "rings" at its resonant frequency. But even if the characteristics of a resonant circuit are absent, the output pulse will be much longer than the input pulse. The only circuit that will pass a short pulse without distortion is a circuit without inductance or capacitance, or in other words, a circuit whose acceptance band is infinite.

An example of the lengthening of a short pulse in various stages of a superheterodyne receiver is shown in Figure 2-5. Part (A) of this figure shows an approximately rectangular pulse of 0.4 microsecond duration, which is applied to the input of the receiver. The next part (B) shows the resultant "ringing" of the radio frequency stage. (C) shows the further lengthening of the pulse in the intermediate frequency stage. Finally, (D) shows the pulse as it appears at the audio output after rectification. Note that the pulse is now almost 200 microseconds long, i.e., the initial duration has been multiplied by a factor of almost 500.

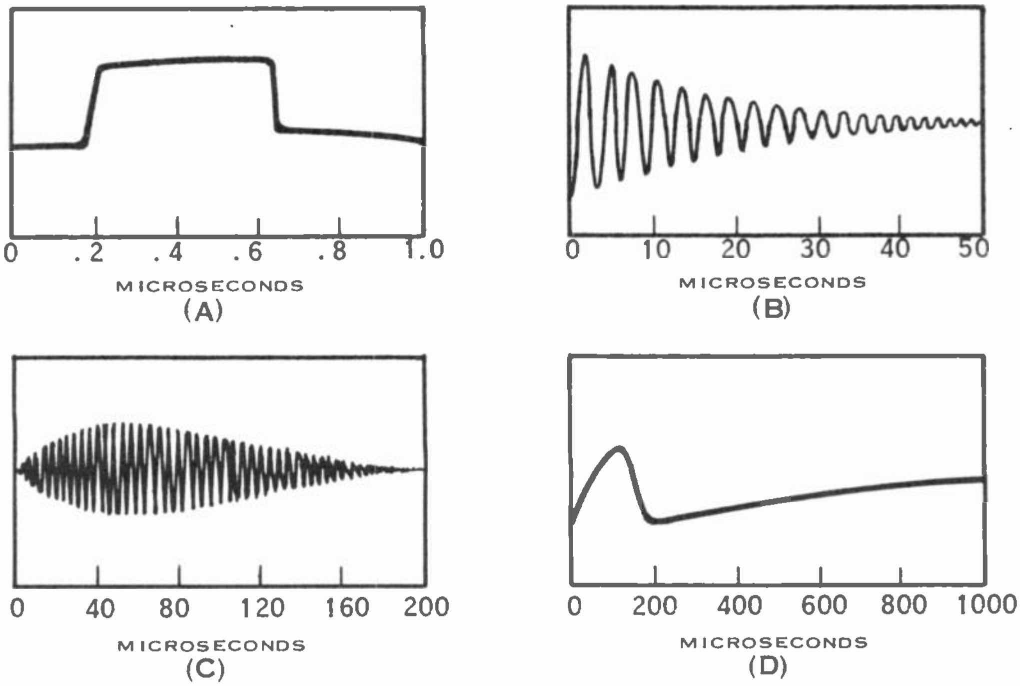


Figure 2-5. Pulse Lengthening in Various Stages of a Receiver

There is a definite relationship between the amount of pulse lengthening and the bandwidth of a stage. Roughly, the lengthening of a pulse is inversely proportional to the bandwidth of the stage through which it passes. This rule is valid both for unidirectional rectangular pulses and for high frequency sinusoidal pulses produced by amplitude modulation of a high frequency carrier with a rectangular modulation envelope.

3.3 DESENSITIZATION

If the amplitude of a signal is much larger than that for which a stage is designed, it may completely block that stage so that no useful signal can get through at all. Whether this limiting action is intentional or not makes little difference. The fact that the stage is inoperative for the duration of the high amplitude signal will always cause distortion in the desired signal. This effect is often produced intentionally because the resulting distortion may be less than that which would occur if the interfering signal were allowed to pass through unlimited.

The most common cause of overloading is the saturation of a vacuum tube. A tube is a linear device only for a certain range of applied voltages and currents. If an excessive signal is applied to the grid of a triode, the plate current may become temperature limited, which means that the output becomes independent of the input, and dependent only on the cathode temperature and surface condition. The same effect may occur in a diode that is called on to rectify an excessive signal.

3.4 EFFECTS OF MODULATION

Whenever radio frequencies are employed for the transmission of intelligence, use is made of some type of modulation. The original signal is not transmitted directly, but rather it is used to modulate a so-called carrier, whose frequency is much higher than that of the original signal. This is done both in order to increase the number of available communication channels, and because low frequencies (below about 100 kc) cannot be efficiently transmitted by radiation. The most common types of modulation are amplitude, frequency and phase, and pulse modulation.

The three types of modulation differ markedly in their abilities to transmit useful information in the presence of interfering signals. This ability consists of two different and entirely separate properties. One is the ability of the receiver to amplify the desired signal more than the interfering one. It is measured in terms of an interference-voltage reduction factor, defined as the ratio of the signal-to-interference ratio at the output to that at the input. The second is the ability to separate the desired signal from the interfering one, which is measured in terms of an improvement threshold, defined as the minimum signal-to-interference ratio necessary at the input to produce an intelligible signal at the output. The signal-to-interference ratio required at the output for intelligible reception depends on the type of signal, i.e., whether the intelligence being received is voice, code, or a control signal, as well as on the type of interference that is present. The improvement threshold may be thought of as the point at which the desired signal "takes over" so that it, rather than the interference, determines the main character of the output.

To explain the practical significance of these two properties, consider a receiver to which a combination of signal and interference is applied with ever-increasing signal-to-interference ratio. At first the interference is much larger than the desired signal and the latter is completely masked at the output. No useful information is received. As the desired signal increases in strength, a point will be reached at which the desired output signal becomes intelligible. The input signal-to-interference ratio at this point is the improvement threshold. At and beyond this point, the amount by which the ratio is reduced between the input and output is measured by the interference-voltage reduction factor.

The behavior of the three types of modulation with respect to these two properties is illustrated in Figure 2-6, in which the signal-to-interference ratio at the output is shown as a function of the signal-to-interference ratio at the input for some typical cases. The improvement threshold is the point of maximum slope on each curve, i.e., the point at which the desired signal "overtakes" the interference at the most rapid rate. The interference-voltage reduction factor in each case is given by the ratio of ordinate to abscissa. The values of this ratio of greatest interest to the designer are those for which the signal-to-interference ratio at the input is in the region of 10-20 db.

These characteristics of the various types of modulation are important for equipment designers to consider in the initial design stages especially when there is a choice as to what type modulation is to be used. Once the modulation is decided upon, the designer must make use of the best design techniques in order to insure that the system is interference-free within the limitations posed by the type of modulation.

Another consideration is important to the overall picture for interference-free operation of the complete system. It is the fact that all three types of modulation may be involved in the various equipments installed. Therefore, for interference-free operation of the complete system, techniques must be applied as required for satisfactory operation of the most interference susceptible equipments. Methods of electromagnetic interference reduction and electromagnetic interference-free design are discussed in Volumes III and IV respectively.

3.4.1 RECEIVER INTERMODULATION

Intermodulation may occur in a receiver system as a result of simultaneous reception of two or more signals if they, or any integral multiple, combine in such a manner as to produce either the frequency to which a receiver is tuned, or a spurious response frequency. Interference due to intermodulation may also occur as a result of simultaneous reception of two signals separated by the intermediate frequency.

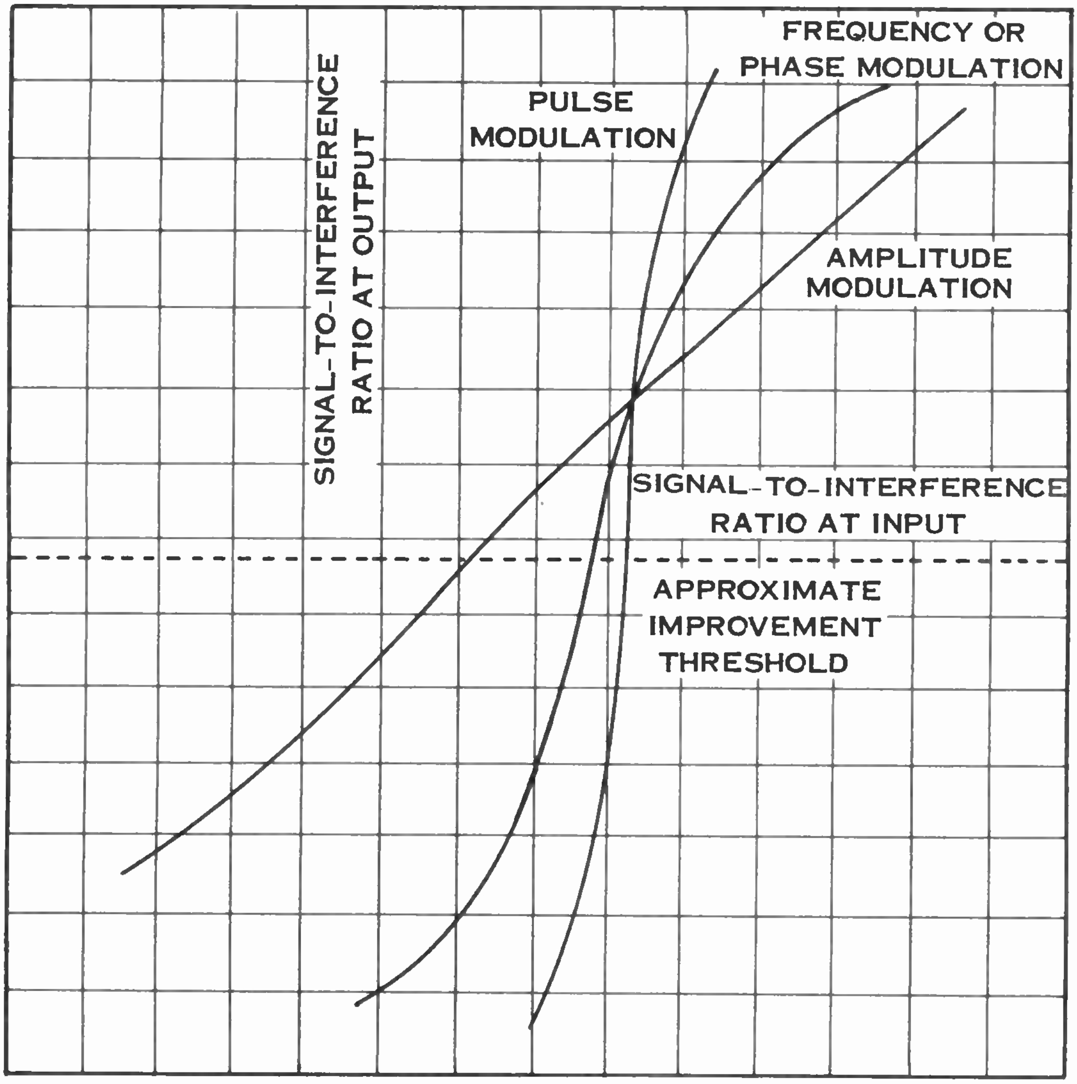


Figure 2-6. Relationship of Signal-to-Interference Ratio at Output to that at Input for Different Types of Modulation

In this case, the interference will occur even if the local oscillator signal is not present. Examples of the various frequency combinations which may lead to interference due to intermodulation in a receiver or antenna and input circuitry are shown in Figure 2-7.

Strong signals, although they may each be free of harmonic content, can generate harmonics and heterodynes in the input circuitry of a receiver. This is described as intermediate-frequency type interference. This form of interference will occur if a high-amplitude

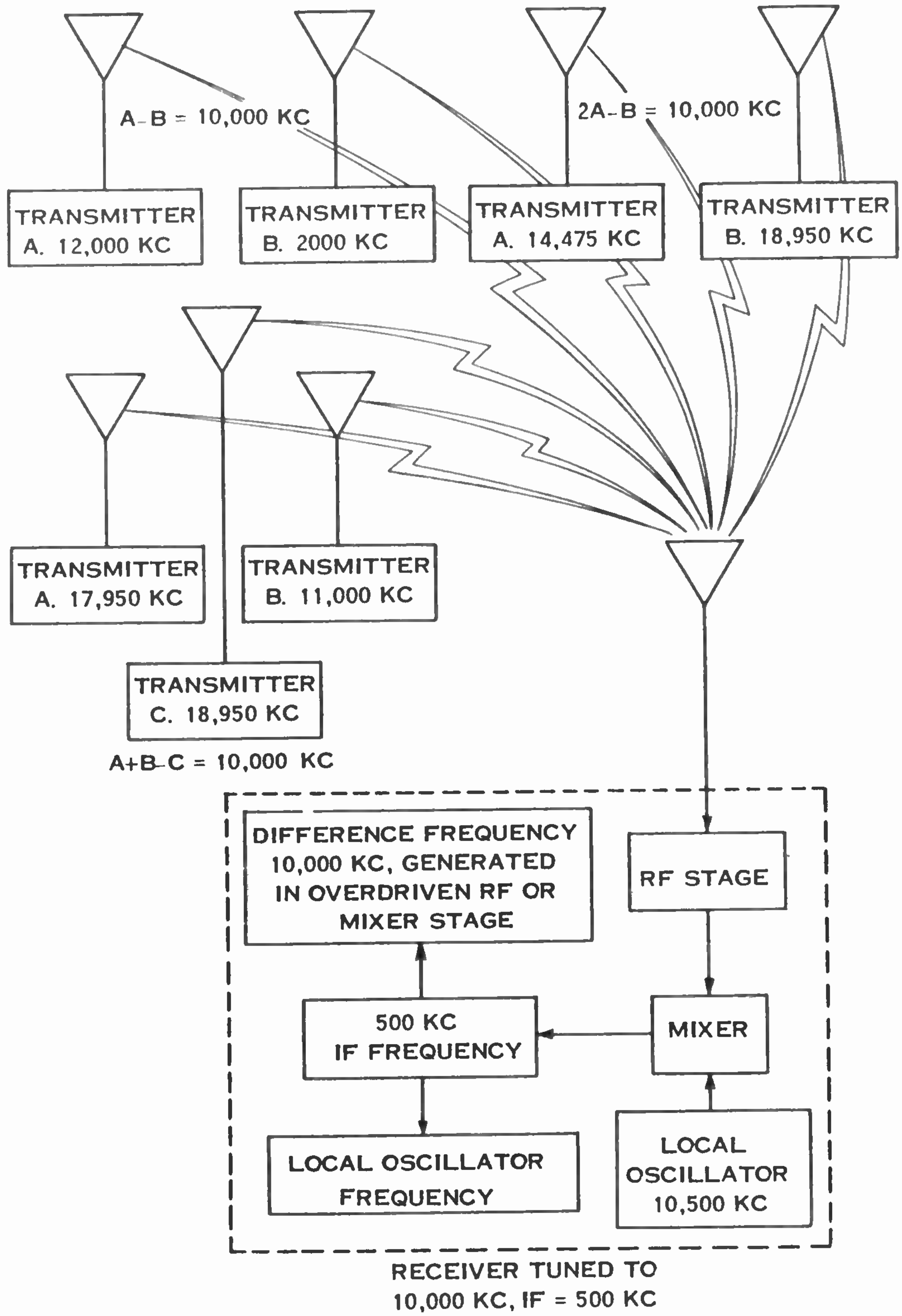


Figure 2-7. Receiver Intermodulation

signal at the intermediate frequency of the receiver reaches the mixer stage. This type of interference is characterized by its presence regardless of the frequency to which the receiver is tuned.

Interference due to intermodulation may be generated when the frequencies of two signals are separated by an amount equal to the frequency to which a receiver is tuned. If two such signals are passed through a nonlinear impedance in a receiver antenna or input circuitry, their sum and difference frequencies are generated by heterodyne action. These new frequencies are termed the A-B and the A+B intermodulation products. When these frequencies are of sufficient amplitude they will be detected in the same way as a desired signal if they are the same as the frequency to which the receiver is tuned, or the same as a spurious-response frequency.

Interference due to intermodulation may also result when the second harmonic of one signal and the fundamental of any other signal combine to produce a difference frequency equal either to the frequency to which a receiver is tuned or to a spurious-response frequency. This condition is termed the 2A-B intermodulation product. Any integral multiple of the fundamental frequencies such as 3A-2B, 3B-2A, 4A-3B, and so on to infinity may cause interference of this type when the frequency product is equal either to the frequency to which a receiver is tuned or a spurious-response frequency, and is of sufficient amplitude to be detected.

Interference caused by intermodulation may also occur when three signals combine to generate the frequency to which a receiver is tuned. The sum of two signal frequencies, when combined with a third frequency in a nonlinear impedance, will produce a difference frequency. If this difference frequency is equal to either the desired frequency or to a spurious-response frequency of the receiver, normal detection will take place and interference will occur. This frequency combination is termed the A+B-C intermodulation product.

3.4.2 CROSS MODULATION

Another type of intermodulation is cross-modulation type interference. Cross modulation is due to the modulation of the carrier that normally carries the intelligence by an undesired signal. It occurs in a receiver when the modulation of an undesired signal is impressed on a desired signal. The effects are apparent as the simultaneous reception of two signals, or if there is no modulation on the desired signal carrier the modulation of the undesired signal is detected. This process occurs in nonlinear elements of the receiver antenna input circuitry, the RF amplifier stage, and/or the mixer stage. These effects discussed above are called intermediate-frequency type interference. Several unusual factors are associated with this type of interference. First, the desired signal must be present for the modulation of the undesired to be detected; second, the undesired signal need not bear a harmonic relationship to either the desired signal, the local oscillator, or the intermediate frequency of the receiver; and third, cross modulation may occur with any frequency separation between the desired and the undesired signals if the undesired signal is of comparable or higher amplitude than the desired signal in the nonlinear element. In cross modulation, however, an undesired signal is usually near the frequency to which the receiver is tuned because of the increase in attenuation of signals outside the passband of a receiver as they are further removed from the tuned frequency.

3.4.3 HETERODYNING

When a sinusoidal wave is fed into a nonlinear impedance, detection occurs and harmonics are generated as a result. The output signal then contains the fundamental frequency as well as integral multiples or harmonics. If another sinusoidal wave is introduced into the same nonlinear impedance, its harmonics are also generated. In addition to the harmonics of the two fundamental frequencies, new frequency components are formed. These components, called heterodynes, are the sum and difference frequencies between the two fundamental signals and their harmonics.

One of these sinusoidal waves may be a desired input signal and the other an undesired input signal to a receiver. When these two signals reach a nonlinear impedance, new frequency components are introduced as outlined above. The total input current to the nonlinear impedance is the sum of the two sinusoidal input signal currents. If one input signal is removed or its amplitude is changed, it will have no effect on the amplitude of the other signal at the input. The total output current from the nonlinear device is the sum of the currents due to both input signals, all their harmonics, and all heterodynes. Therefore, a change in amplitude of one of the input signals will not only cause a change in its fundamental frequency output level and all harmonics or heterodynes of this fundamental frequency, but will also affect the output levels of the odd-order frequency components of the fundamental frequency of the other signal and its heterodynes as well. This effect, as has been explained before in the discussion on cross modulation, is termed cross modulation since it causes the amplitude of the output current at one frequency to be dependent upon the amplitude of a signal component at a different frequency.

The conditions outlined in the above paragraph may occur in an amplifier tube when the input signals are of sufficient amplitude to drive the tube into a nonlinear region of its grid-voltage plate-current characteristic curve, but not of sufficient amplitude to drive the grid positive. If the grid is driven positive by an undesired signal, grid current will flow and rectification will occur. The modulation components resulting from rectification in the grid circuit will in turn modulate a desired signal. In this case the modulation of an undesired signal will again be superimposed on a desired signal and will be detected along with the desired signal.

3.4.4 TRANSMITTER INTERMODULATION

Interference due to intermodulation or cross modulation may be generated if two transmitter signals are mixed in a common nonlinear impedance. Before this can occur, electromagnetic energy from one transmitter must be coupled to another. Coupling may occur through radiation from one transmitting antenna to another and conduction by means of the normal transmission line; through conduction from one transmitter to another due to improper grounding or bonding; or through radiation from a transmitter antenna or improperly shielded, bonded, or grounded transmitter to another transmitter which is not shielded properly. After the signal from one transmitter reaches another by any of these paths, intermodulation or cross modulation may take place in its nonlinear elements such as loose or corroded antenna or ground connections, or stray coupling into nonlinear amplifier stages.

Several frequency combinations which may result in interference due to intermodulation are shown in Figure 2-8. Harmonics and heterodynes of integral multiples of the fundamental frequencies are generated when two transmitter signals are present in the same nonlinear element. The sum or difference frequency is known as the A+B and A-B intermodulation product. Interference due to intermodulation may also result from the second harmonic of one signal heterodyning with the fundamental frequency of another signal. The sum and difference frequencies are termed the 2A+B and 2A-B intermodulation products. This interference may occur in a third way if three signals mix in a nonlinear element. The sum frequency of two signals may produce a difference frequency with a third signal. This difference frequency is termed the A+B-C intermodulation product. The intermodulation pro-

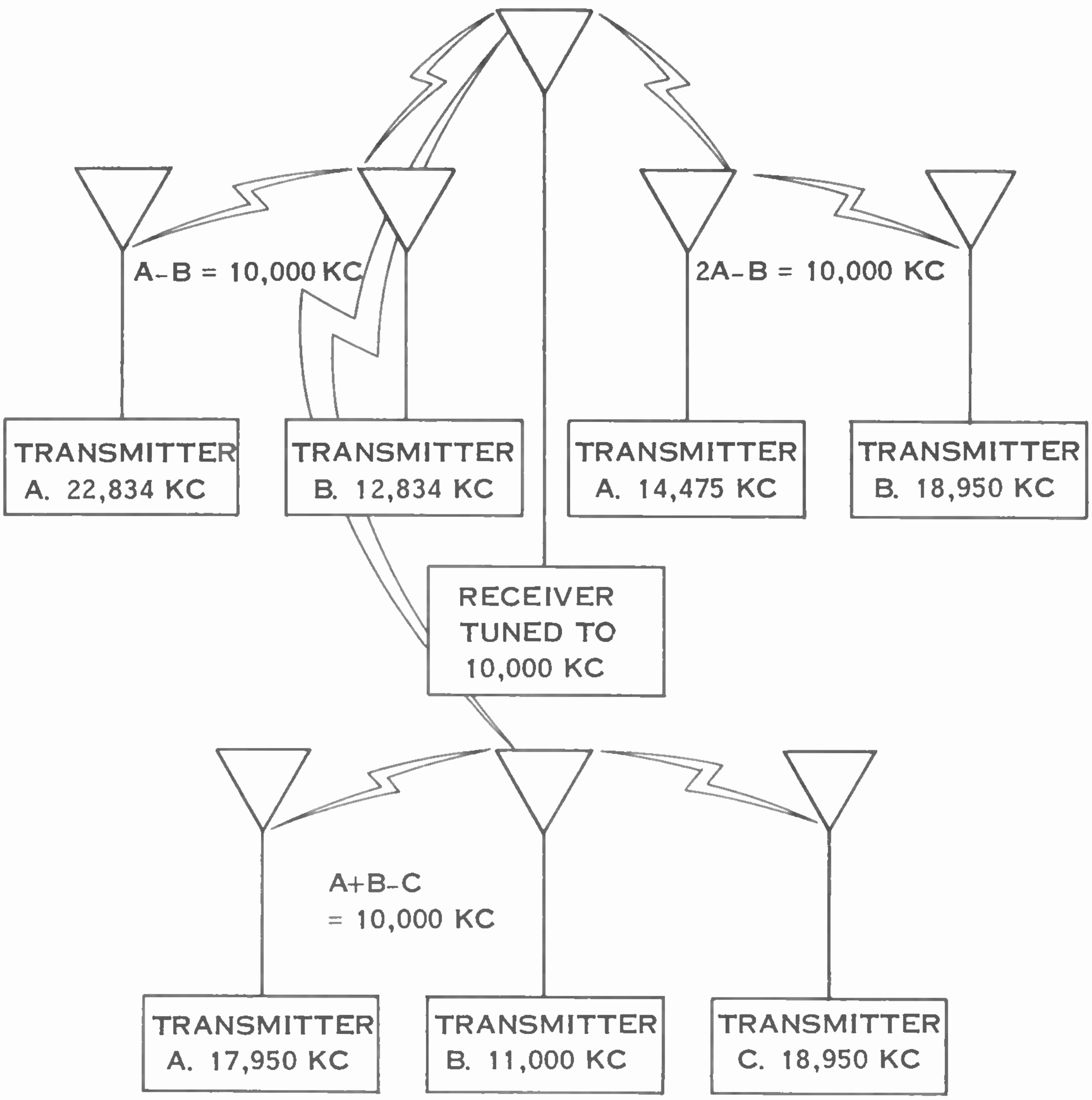


Figure 2-8. Transmitter Intermodulation

ducts which may be most troublesome are those which fall in an adjacent channel of the band in which equipment is operating. These are the 2A-B, 2B-A, 3A-2B, 3B-2A, and A+B-C. The higher order and sum frequency intermodulation products occur further from the fundamental frequencies and are usually not detected for this reason even though they are present.

In general, interference in a transmitter due to cross modulation and intermodulation has the same effects as in a receiver and occurs under the same circumstances, that is, two signals are mixed in a nonlinear element. For this reason the two are easily confused; however, the process by which they occur is distinctly different. In cross modulation, the modulation of one signal is impressed upon another signal whereas in the case of intermodulation, the interference is due to a signal occurring at a frequency other than either of the fundamental signals involved. When cross modulation occurs in a transmitter, the fundamental output signal then has the modulation of an undesired signal impressed upon it and therefore interference is contained in the radiated output signal of the transmitter.

4. CONSIDERATION OF AN ENTIRE AEROSPACE VEHICLE AS A SINGLE NETWORK

Many of present-day interference problems arise with respect to aircraft, missiles, and space vehicles. Special consideration is therefore given to an analysis of this subject in the following paragraphs. A similar analysis can be effectively applied to a ground or ship installation.

In order for any interference to become effective, it must, in some way or other, reach the receiver. Within the airframe there is rarely a single mode of transmission from the point of generation to the receiver, i.e., transmission is rarely entirely by radiation, or entirely by inductive coupling. In subsequent paragraphs, such single modes will be described and analyzed, but in this paragraph, emphasis is placed on the fact that an entire aerospace vehicle, including all its equipment and wiring, must be considered as one entity in any complete analysis of interference. Frequently, one particular mode of transmission will be more important than the others, but there is always great danger that, in over-emphasizing this one mode, important interactions with other modes will be completely overlooked. Often these interactions modify the original mode to such an extent as to make the isolated analysis of that mode alone completely invalid.

For purposes of analysis, then, the entire vehicle must be considered as a four terminal transmission network, characterized by four parameters: (1) a single complex input impedance at the terminals to which the interference source is connected, (2) a single complex output impedance at the terminals to which the receiver is connected, and (3) and (4) two transfer impedances or transfer constants. These last two will be different, in general, but will be equal in the special case of greatest practical importance, when the network contains only bilateral elements.

As has been previously pointed out, the character of a signal is determined not only by the process of its generation, but also by the internal impedance of the source, as well as the impedance into which the source is operating. This statement emphasizes that source, transmitting network or medium, and receiver cannot be treated independently, but must be considered as one unit. This idea is referred to for brevity as the "impedance concept" because, as far as the source is concerned, the entire vehicle and receiver can be replaced by a single complex impedance, and, as far as the receiver is concerned, the entire vehicle and source can be replaced by a generator in series with a single complex impedance.

The consideration of the aerospace vehicle as a four-terminal network is adequate for most applications, but lacks generality in two respects. In the first place, the concepts of networks and terminals become meaningless at frequencies for which the wavelength becomes comparable in magnitude with the dimensions of the equipment and wiring involved. The term "network" may be generalized so as to acquire meaning for this case also, but it will be more convenient to talk about "transmitting media" which are coupled to the source and the receiver and to each other through electric or magnetic fields, or through mutual impedances that cannot be localized at any terminals.

The other lack of generality lies in that this impedance concept does not apply readily in the presence of nonlinear impedances. If the principle of superposition does not apply, consideration of a four-terminal network is not sufficient, because of the possible interactions between the interfering signal under consideration and some other signal, which may be a desired or another interfering signal. Fortunately, nonlinear elements are very rare in the transmitting network, and occur mostly within the receiver itself. For the present, it is assumed that only linear elements need be considered, and, therefore, the consideration of four terminals is adequate.

The general theory of four-terminal transmission networks is extensively treated in several textbooks. There, it is shown that any linear, passive, four-terminal network is completely specified by four complex parameters, which, in general, are functions of frequency. Several sets of four may be used, but the two image impedances and the two transfer constants are the most convenient ones for the present purposes. The image impedances are those impedances which must terminate the network in order to have an impedance match (i.e., equal impedances looking into and out of the network, not the conjugate match required for maximum power transfer) at each pair of terminals. The transfer constants are one-half of the logarithm of the ratios of volt-amperes in to volt-amperes out when the network is terminated in its image impedances. For a network containing only bilateral elements, three, instead of four, independent parameters suffice to specify the network, since the two transfer constants become equal. If, in addition, the network is symmetrical, the image impedances are equal also, and only two independent parameters are left.

Consider, now, the circuit of Figure 2-9. The transmission network is linear and bilateral, but not necessarily symmetrical. It has image impedances, Z_{I_1} and Z_{I_2} and a transfer constant θ. A generator of voltage E and internal impedance Z_S, is connected to the input terminals, and a load Z_R is connected to the output terminals. It is assumed that the impedances are not matched on either side, so that $Z_S \neq Z_{I_1}$ and $Z_R \neq Z_{I_2}$. Then the currents in the generator and in the load impedance are given by:

$$I_1 = \frac{E}{Z_S + Z_{I_1}\left[\dfrac{1 + F_r e^{-2\theta}}{1 - F_r e^{-2\theta}}\right]} \tag{2-2}$$

where $F_r = (Z_R - Z_{I_2})/(Z_R + Z_{I_2})$, and

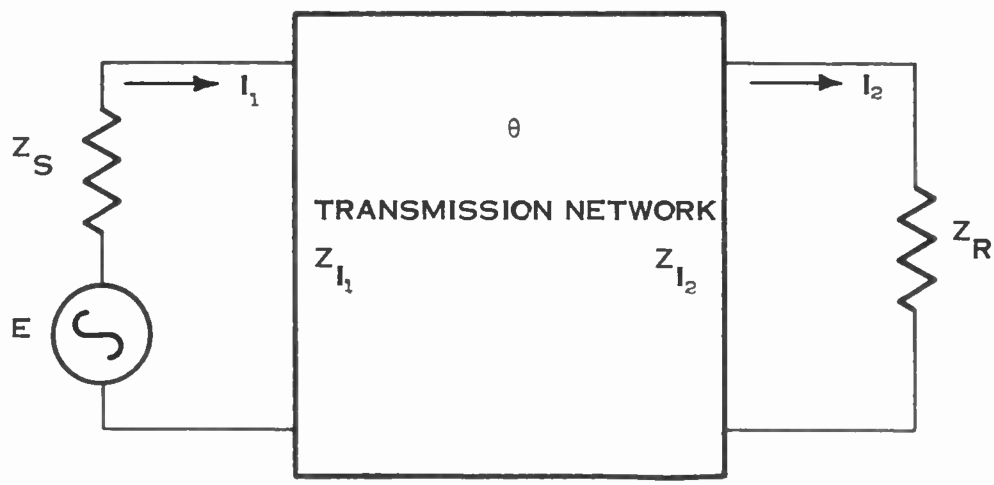


Figure 2-9. Generalized Four Terminal Transmission Network

$$I_2 = \frac{E}{2\,Z_S} \cdot \frac{2\,Z_S}{Z_{I_1} + Z_S} \cdot \frac{2Z_{I_2}}{Z_R + Z_{I_2}} \cdot e^{-\theta} \sqrt{\frac{Z_{I_1}}{Z_{I_2}}} \cdot \frac{1}{1 - \dfrac{(Z_{I_1} - Z_S)(Z_{I_2} - Z_R)}{(Z_{I_1} + Z_S)(Z_{I_2} + Z_R)}\, e^{-2\theta}} \qquad (2\text{-}3)$$

3

The expression for I_1 in Equation 2-5 means that the input current may be obtained by taking the ratio of the voltage E to the total impedance of the circuit. Hence, the second term in the denominator must be the impedance of the network when terminated in Z_R. The various factors in the expression for I_2 have very simple interpretations also. The first, $E/2Z_S$, is the current that would flow if the source were simply connected to a matching impedance Z_S. The second term, $2Z_S/(Z_{I_1} + Z_S)$, called the transmission factor, shows that this ideal current is modified in going from the impedance Z_S to the impedance Z_{I_1}. The third, $2Z_{I_2}/Z_R + Z_{I_2}$, is a similar transmission factor, indicating the modification due to the transition from Z_{I_2} to Z_R. The fourth, $e^{-\theta}$, gives the attenuation and phase shift due to a symmetrical network of transfer constant θ. The fifth, $\sqrt{Z_{I_1}/Z_{I_2}}$, indicates the transformer action of an asymmetrical network. The last, finally, is the so-called interaction factor, which arises from the multiple reflections in a network, connected into a circuit, where the impedances are not matched at either end.

If the transfer of power through the network is of primary interest, the two transmission factors and the interaction factor must be considered as introducing losses due to the lack of a good match at either end. But, if the quantity of interest is the output current or voltage, without any consideration of power, these factors may easily introduce gains rather than losses. For example, if the input impedance of the receiver is very high, as in the grid circuit of a pentode, the output current and power are very low, but the output voltage might become very large due to the transmission and interaction factors. This fact is of great importance to the transmission of interference, because it shows that it is by no means necessary to have an impedance match for maximum transmission of either voltage or current. On the contrary, a mismatch may actually introduce an increase in the interfering current or voltage at the output, or it may result in causing such large currents to flow from a power supply that the machine is damaged or circuits overloaded to the point where a fire hazard exists. Thus, the practice of inserting large capacitors in circuits to reduce interference without careful consideration of what effects impedance mismatches may cause, is to be avoided. The reflection gain or loss in decibels resulting from a mismatch of impedances as a junction in a transmission line is given in Figure II-6 in Appendix II.

Similarly, the "transformer factor," $\sqrt{Z_{I_1}/Z_{I_2}}$, which produces the same modification of current and voltage as a perfect transformer, introduces neither gain nor loss, as far as the power is concerned. But, if the quantity of interest is current or voltage, this factor may cause an unexpectedly large gain. These facts explain why it happens that an interference source sometimes produces large distrubances in a particular receiver, even though laboratory measurements show no excessive currents or voltages when the source is tested alone.

5. APPLICABILITY OF THE "IMPEDANCE CONCEPT" TO THE SOLUTION OF INTERFERENCE PROBLEMS IN AIRBORNE SYSTEMS

The problem of combating interference is a threefold one: to minimize its generation, its transmission, and its undesirable effects on the receiver. The impedance concept is useful in all three of these, but particularly in the second. The transmitting network, which is the aircraft with all its equipment and wiring, is not designed with any specific purpose. In fact, it is not designed as a network at all. It is a haphazard conglomeration of wires and pieces of equipment which happen to form network elements. It is not surprising, then, that often a shield or filter, designed on the basis of considering only a small portion of this total network, turns out to increase, rather than decrease, the interference at the receiver. In almost all cases of this kind, the failure can be traced to the lack of consideration of the complete picture. By suppressing one mode of transmission, another might be sufficiently favored to produce the undesired result. A device, designed to reduce the generation of interference, might greatly increase its transmission, thus minimizing or even cancelling the effect of the reduction.

The tremendous increase in the importance of the radio interference problem in recent years is due to the following reasons:

a. The large increase in the number of both receivers and potential interference sources in modern aircraft, missiles, and other surface based systems.

b. The necessity of having a large number of pieces of electronic equipment crowded together in close quarters such as in aircraft and missiles.

c. The increased opportunity of transmission of interfering signals due to the large amount of wiring required in modern airborne electronic systems.

d. The increased sensitivity of modern receivers.

e. The increased use of higher and higher frequencies for communication purposes.

All but the last two of these causes are such that a unified overall approach, such as is suggested by the impedance concept, seems not only advantageous, but absolutely essential. One can hardly hope to cope with a problem that arises from the combined action of all the elements of the airborne system by isolating just a small part of it and designing a solution for that isolated part. A further discussion of the use of the impedance concept will be found in Appendix II.

6. ANALYSIS PROCEDURE FOR A GROUND COMMUNICATION-ELECTRONIC SYSTEM

There are basically eight steps in the analysis procedure for setting up a ground C-E system. Efficient use of this procedure depends, to a large extent, upon obtaining technical data and information which is as complete as possible. Much of the information should be available in handbooks, equipment manuals, and other similar publications. Other information must be obtained by actual field survey and test. Some data, such as antenna patterns, may not be known. In such cases, it will be necessary to make measurements or to use approximations resulting from whatever information can be found, with interpretations based on good engineering judgment.

The analysis procedure consists of eight steps as follows:

a. Analysis and selection of equipment

b. Acquisition of technical data

c. Review of system characteristics

d. Preliminary analysis and prediction of interference possibilities within the system

e. Preliminary review and survey of possible sites

f. Final selection of sites

g. Analysis and selection of operating frequencies

h. Checkout for minimum interference after installation of equipments and antennas.

One of the most valuable sources of information is the data actually obtained by a field team visiting the proposed sites. The field team should make ground and air inspections of the sites which appear to be most desirable for C-E installations and should determine the actual physical location of existing installations of other systems and agencies. During this inspection, the survey team should also look for other new C-E installations which are under construction in the area. Existing installations which are inactive should be reviewed as possible future sources of interference.

An aerial survey will prove helpful in locating industrial areas and communications-electronics facilities in the vicinity of the proposed sites. A followup inspection on the ground will reveal additional information about the nature of the installations and possible sources of interference.

The proposed sites should be coordinated with other agencies to determine the probable location of future facilities in the area. If the proposed sites are located in the vicinity of heavily populated or industrial areas, local officials should be contacted for the purpose of determining future construction plans for the area. Contour maps, aerial photographs and aerial surveys should be studied. The site study and the field survey should give consideration to matters such as: accessibility of site, a.c. power availability, type of ground, obstructions, and size of area required.

The problems of mutual interference which might result to and from the proposed system and various other C-E equipments in the same general area and within range of the proposed C-E system, require that signal measurements and radio noise measurements be made at the proposed site to determine the extent of interference which might be expected. Possible interference sources include: radars; navigational aids; standard broadcast, television, and FM stations; microwave links as used by railroads, pipelines, electric power systems, telephone companies, and other organizations; communication equipments operated in the public service; and facilities under the control of such potential interference sources relative to site location, should be determined during the site survey.

The proximity of the receiving site to electrical installations of all types should be investigated. Typical of the kind of electrical noise which may originate in the area is that which is produced by rotating and commutating machinery, neon lights, flashing signs, fluorescent lighting, telephone switching systems, electric razors, arc welders, traffic signals, thermostatic controls, and other similar devices.

Interference poses a specific threat to weapons systems operations. In fact, today, when ICBMs must be intercepted, air vehicles and missiles of all types must be controlled, and large scale radio communications systems are fundamental to military operations, the C-E system may be the key to winning or losing major battles and wars. The exact amount of interference in the C-E systems must be known and action taken to alleviate the interference before military operational capability becomes degraded.

6.1 INTERFERENCE PREDICTION

In order to determine interference conditions adequately, which may affect a C-E system, it is desirable that a prediction of such interference be made before installation of the system. Due to the high power of transmitters and high receiver sensitivities in use today, some form of such prediction is almost mandatory in system analysis and evaluation.

Power output of modern military C-E transmitters may range up to 10 megawatts or higher (100 dbm and up). Antenna gains may be 20 to 40 db above an isotropic radiator. Receiver sensitivities range around -100 to -110 dbm. Ignoring, for the moment, the problem on the fundamental, a harmonic which fulfills the generally accepted requirement of being 60 db less than the fundamental would still have a power of 40 dbm or 10 watts. Such a power would, of course, have a serious effect on nearby receivers which are attempting to pick up weak signals on the harmonic frequency.

At the fundamental frequency, the entire problem is magnified because of high radiation power of the transmitter, high antenna gains, and high sensitivity of the receiver. As powers and sensitivities increase in the future, it will be absolutely necessary, in the design and planning stage, to devote considerable effort to the interference problem.

Interference prediction is becoming an exact science as sufficient data is being accumulated, through spectrum signature studies, on propagation, transmitter emission characteristics, receiver characteristics, and antenna patterns.

6.2 GENERAL PRINCIPLES OF INTERFERENCE PREDICTION

6.2.1 INTRODUCTION

The calculation of radio frequency interference probability, commonly called RFI prediction, involves in general a multiplicity of transmitters and receivers. The object of this operation is to determine if a certain class of receivers will be interfered with by any of the multiplicity of transmitters, or conversely if a certain class of the transmitters will interfere with any of a multiplicity of receivers. In general, these calculations are carried out only above 30 mc because the vagaries of ionospheric propagation make accurate predictions below this frequency very difficult. Lately, however, there has been increased pressure to attempt prediction below 30 mc. The application of the prediction process to particular RFI situations will be given a great deal of consideration in Volume II of this series. The present discussion will, therefore, discuss only the fundamentals of RFI prediction.

In its basic form the prediction of the severity of the interference which a given transmitter will generate in a particular receiver involves the determination of the signal-to-interference ratio (S/I) existing at that receiver. The multitude of factors which go into the calculations of signal-to-interference may be classified into three groups: (1) the parameters of the interfering transmitter at spurious as well as design frequencies, (2) receiver characteristics and susceptibility data, and (3) factors affecting the propagation path between the transmitter and receiver. Here it should be pointed out that the true value of interference is determined only after an acceptable S/I limit has been established for a particular receiver. A complete discussion of signal acceptability criteria appears in Volume II.

6.2.2 PROGRAM PHILOSOPHY OF AN RFI PREDICTION STUDY

The first step in carrying out a fairly large scale RFI prediction program is to determine the significant characteristics of the class of equipments that may be interfered with or that may cause interference. In other words, some RFI prediction investigations may involve the consideration of a certain class of receivers which may be interfered with by any transmitter that is approximately within the line-of-sight distance of the receiver, and in other cases the goal of the investigations will be to determine which of all possible receivers within the above-mentioned line-of-sight distance will be interfered with by a certain class of transmitter.

6.2.2.1 Receiver Susceptibility (Spectrum Signature)

If the class of equipments being investigated consists of receivers, then it is necessary to determine which transmitters will interfere with these receivers. To do so, it is necessary to know at which frequencies the receivers are sensitive to interference. In most cases, a receiver is sensitive to a large band of frequencies far in excess of its normal selectivity curve if enough power is applied to the input. In some cases, no response will be obtained even if sufficient power is applied to destroy the input circuitry, but with most receivers a continuous curve of response will be obtained over a very wide bandwidth. In these measurements special consideration should be given to recording the responses around the image frequency, the IF frequency, and response frequencies that may be caused by local oscillator harmonics as well as recording in detail the skirt shape of the designed pass band. This plot is called the susceptibility curve of the receiver. It is also technically known as the receiver's spectrum signature, but this term will not be used in the present discussion so that it may be reserved for use in connection with transmitters. A typical susceptibility curve is shown in Figure 2-10.

Since this information is not normally recorded by the receiver designer, it will be necessary to run special tests before the interference survey is begun. This information may be available if it were required that the receiver pass one of the military specifications on RFI. It is worth investigating this point before these tests are begun. Naturally, in a well-designed receiver this response will be 80 to 100 db down from the fundamental for most frequencies; however, this can still be significant enough to cause RFI if the interfering signal is strong enough.

6.2.2.2 Transmitter Spectrum Signatures

If the class of equipments being investigated consists of transmitters, then it is necessary to determine which of all the receivers within approximately the radio line-of-sight (RLOS) distance are likely to be interfered with by these transmitters. To do so, it is necessary to know at which frequencies the transmitter radiates and how much power is present at these frequencies. In general, a transmitter will radiate power not only at its design frequency but also at its harmonic frequencies, and in some cases at frequencies which are not harmonically related to the design frequency. All frequencies which are radiated, except the fundamental, are defined as spurious radiation. Frequencies which are not harmonically related to the fundamental are likely to exist around the fundamental and its harmonics, although in some

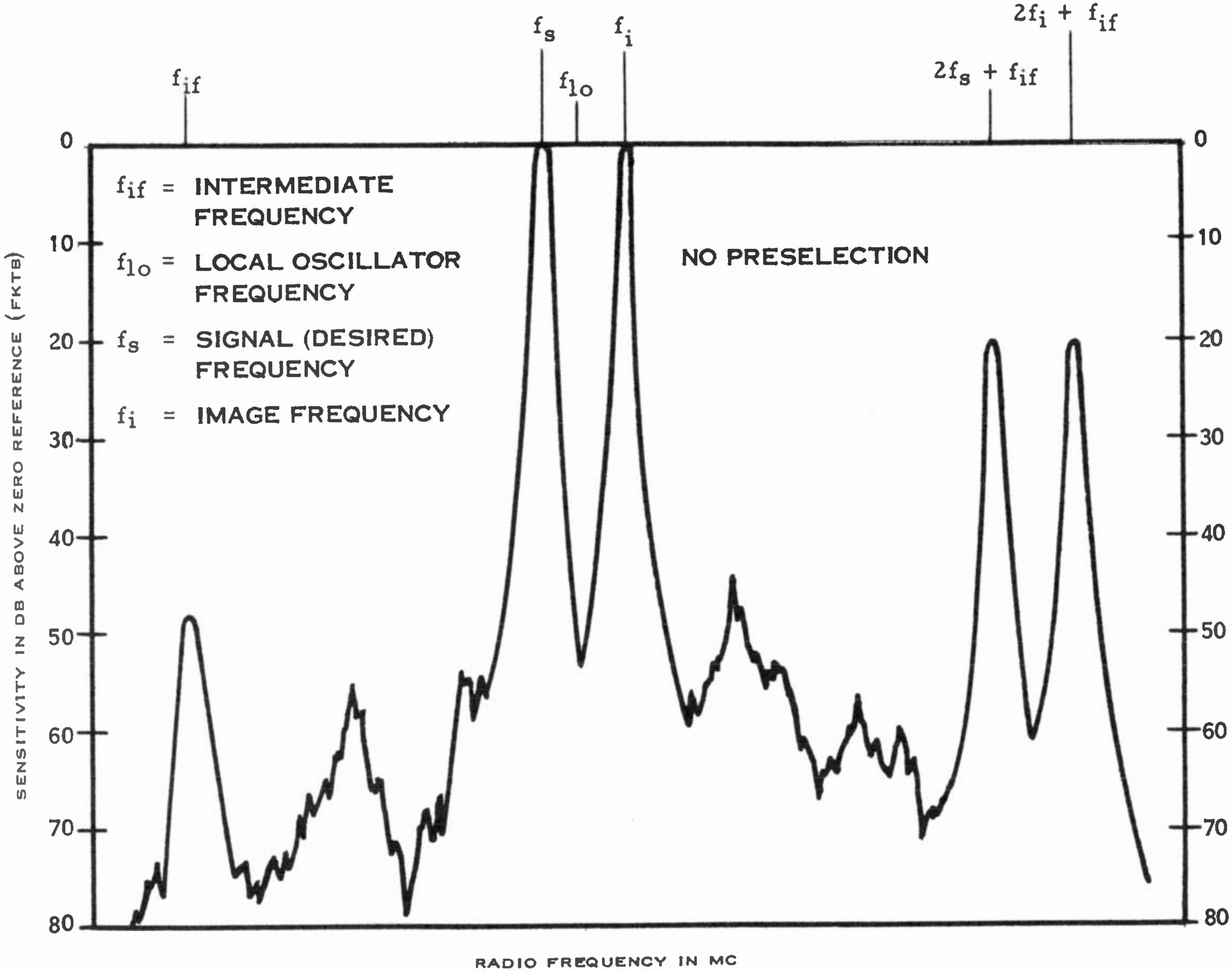


Figure 2-10. A Typical Superheterodyne Susceptibility Curve

cases they may exist in a continuous spectrum over a wide band of frequencies. As in the case for the receiver, it is necessary to measure the emissions from the transmitters at all frequencies having a significant power output. For very high-powered transmitters the spurious frequencies may be many db below the fundamental and yet still radiate significant amounts of power.

Because some of the spurious radiations may have frequencies below the transmitter design frequency, it is necessary to search the spectrum over a range which extends considerably below the design frequency and up to approximately ten times this frequency. The curve obtained from these measurements is defined as the spectrum signature of the transmitter. As in the case for receivers, very little information of this type is generally available. A large program is underway sponsored by the Department of Defense to determine the spectrum signatures of most radars, but much additional work still needs to be done to obtain similar data for high-powered transmitters and radars which are not part of the military establishment. A typical transmitter spectrum signature is shown in Figure 2-11.

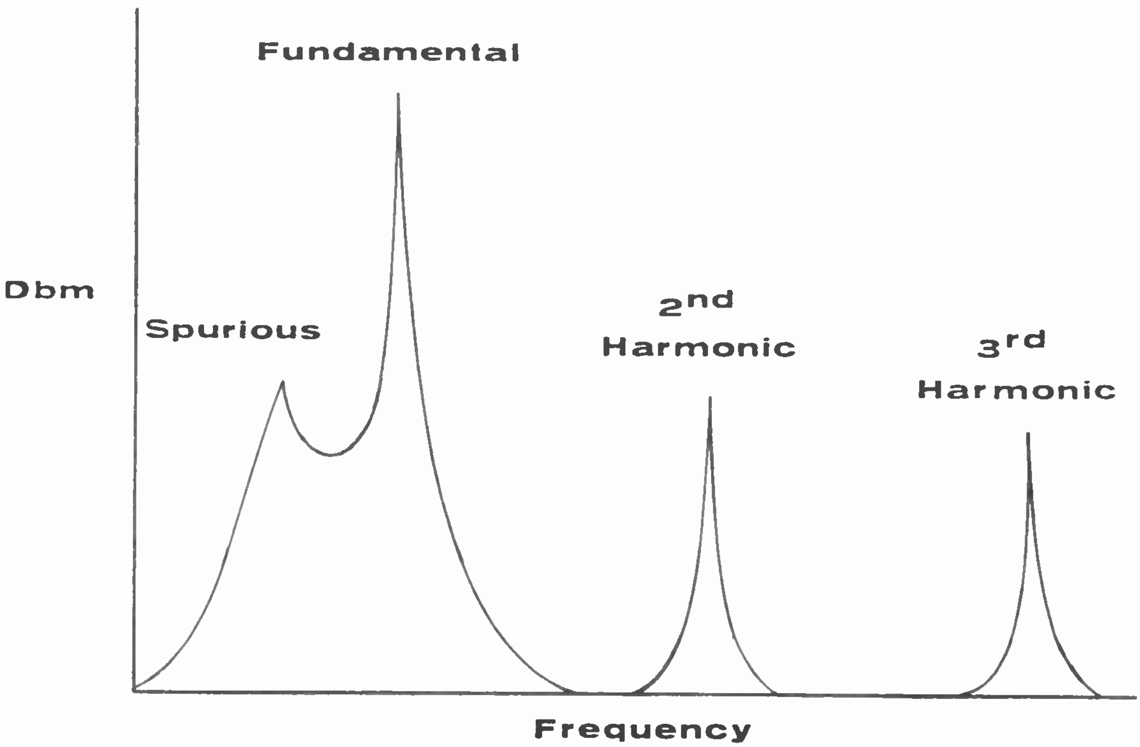


Figure 2-11. TX Spectrum

One problem which exists for transmitters but not for receivers is that while in the case of the receiver the susceptibility curve is almost entirely defined by passive circuits, in the case of the transmitter, when the output tube is changed the spectrum signature may change because it is determined not only by the passive circuits of the transmitter but also by the characteristics of the final output tube or circuit.

From the preceding discussion, it can be seen that to make a complete and thorough RFI prediction it is necessary to know the complete emission spectra of the transmitter being considered; that is, its spectrum signature and also the complete response of the particular receiver type being considered; that is, its susceptibility curve. It is obvious that in many cases this information will not be accurately known, and it will be necessary to make approximate calculations to determine the values of these parameters.

6.2.2.3 Determination of Interfering Equipments

After the characteristics of the transmitters and receivers under consideration are determined, the next step in a prediction program is to determine how many specific equipments must be considered. As stated above only calculations of the RFI threat to or from equipments of a given type, such as specific receivers, are usually made and it is necessary to determine which of the many other equipments, such as transmitters, are significantly located with respect to the receivers and are likely to cause interference. Only equipments in these categories require detailed consideration.

To make the following discussion specific, the above example will be continued wherein a specific class of receiver is being examined in relation to interfering transmitters. The first step in determining which transmitters must be considered is composed of two phases. The first phase is ascertaining which transmitter types have spectrum signatures which cause significant energy to be intercepted by the appropriate portions of the receiver susceptibility curve. Careful consideration will show that to solve this problem rigorously would involve the comparing of an extremely large number of spectrum signatures from different transmitters with the receiver susceptibility curve. The complete solution of this problem would assure that no possible interfering combination could be overlooked. Obviously, this cannot be done without the aid of a large amount of experimental data and a digital computer. While plans for this type of evaluation are underway, the complete implementation of them is not possible because very little spectrum signature data is available at present. The interim solution which must be used, especially if the services of a computing facility are not readily available, is to select only the transmitters which obviously have an interference potential. These include cochannel and adjacent channel transmitters and those whose harmonics fall within the design pass band of the receiver. Other equipments should also be considered if there is reason to believe that their characteristics will be such as to present an interference threat. From this it can be seen that judgment plays a very important part in this step and that at present proper selection is somewhat of an art.

6.2.2.4 Propagation Factors Effecting the Selection of Interfering Equipment

The second phase in determining which transmitting equipments must be considered in the interference program is finding how many of each type of transmitter that was included in phase one of this step lies approximately within radio line-of-sight of the receiver. In relatively flat country such as exists in the eastern and midwestern United States, the solution of this problem is more straightforward than it is for the western section which contains many mountains of greatly varying elevation as well as long flat valleys.

For relatively flat terrain, the area to be investigated in determining the number and position of the threatening transmitters can be calculated readily by the use of line-of-sight equations for a smooth earth. (Some allowance should be made for the fact that the transmitter and receiver may both be elevated and the line-of-sight should be extended accordingly.)

For mountainous terrain, the problem becomes considerably more complicated because of the possibility of extremely high receiver locations, shadowing by long high mountain ranges, knife edge diffraction over mountains and long distance propagation through valleys. For this situation a radio line-of-sight map is made using a terrain map and profiles are taken for the complete 360° sector around the antenna. A typical RLOS map is shown in Figure 2-12.

6.2.3 POPULATION CENSUS

After the transmitter types which must be considered are determined and the area around a specific receiver where the presence of such transmitters would be likely to cause RFI has been calculated, it is necessary to ascertain how many and what types of these transmitters are actually present. This determination is called a population census.

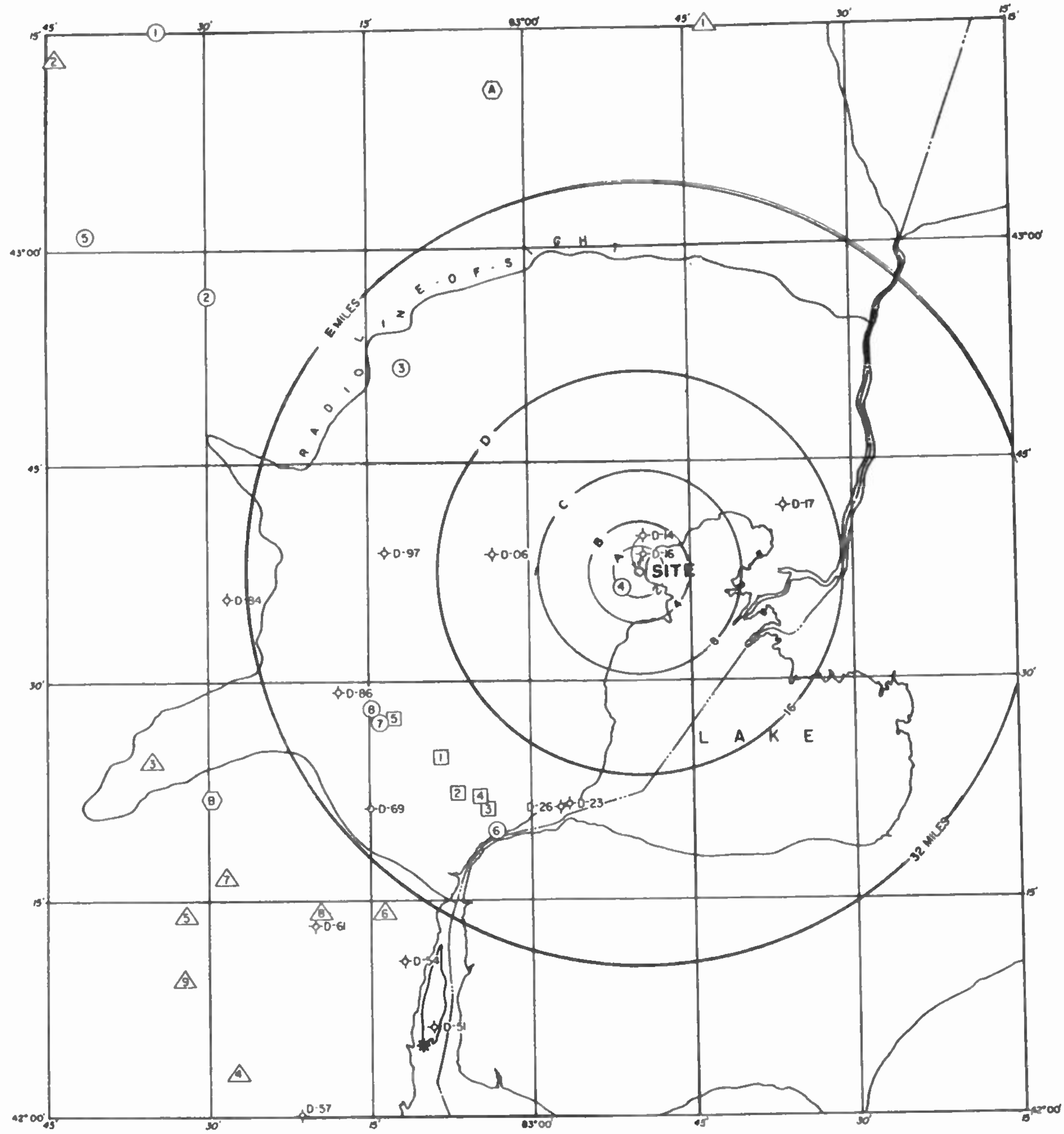


Figure 2-12. Radio Line-of-Sight Map

Naturally, it must be carried out around all the locations where the receiver types being considered are installed.

On first being contemplated, it would seem that a population survey would be one of the simpler aspects of the interference prediction problem, but in actual practice the opposite is often the case. The task of finding the exact location and number of all the potentially interfering transmitters without overlooking any and without including those which are no longer operating, can in certain areas become a very complex job. In general, there is no one authority in a given area which has complete information on the exact frequency of operation as well as the precise location and elevation of all the transmitters in question.

Another problem is that of obtaining information which is current. In many cases the latest available lists may be as much as a year old and, therefore, not necessarily up to date. The agency providing the data may make assurances that no additional equipments have been installed or that none of the listed equipments have been removed, but usually a few field checks will indicate that one cannot always place complete trust in these assurances.

At present some attempt is being made to have a single agency at one location have this information available for all areas of the United States. This is a very desirable goal but it will take some time before this kind of information can be assembled. One of the primary reasons for this is the mobility and changing operational plans of the military services which occupy a significant portion of the frequency bands above 30 mc. The military establishments are not limited to specific frequencies but only to bands of operation, and if a given installation decides to change frequency from one band edge to the other interference may suddenly appear at a frequency where all previous measurements and predictions based on the other operating frequency indicated no trouble should be experienced.

An additional problem in obtaining a complete and accurate census involves the matter of military security. If the prediction study is part of a military program and the equipments under investigation are part of the military establishment, then it is relatively easy to establish a "need to know" in order to obtain specific information concerning classified equipments and their location and operating frequency. If the prediction study is required for the operation of a commercial system, then obtaining a "need to know" clearance may be quite difficult and may require top level coordination between the commercial organization and the appropriate government agencies. No population census can be considered to be even remotely complete if it does not account for the presence of high-powered classified government equipment in the area. In the event that a "need to know" cannot be established, then the population survey cannot be completed by merely listing the known equipments in the area. Field measurements at the receiver location will be required to determine the nature of other unlisted signals which may be present at various times of the day.

Since, at present, information on military and civilian equipments is not completely tabulated at one location, it will be necessary to interview several agencies to obtain a complete listing of the pertinent equipments. In general, the area frequency coordinators for the various agencies are the appropriate personnel to be contacted. The United States is divided into geographical areas in which frequency assignment and control are handled by these coordinators. They are usually representatives of the military establishments, the FCC, FAA, and other cognizant agencies. They have the responsibility of coordinating the frequency assignments for the various equipments located within their area of interest. Major military installations such as the Atlantic Missile Range, the Pacific Missile Range, White Sands Proving Ground, Fort Huachuca, and others also have frequency coordinators. Their primary responsibility is to insure the compatibility of the range instrumentation, but in addition they have a considerable amount of information about other equipments in their area as well as those in the areas of other coordinators. At present then the initial step in obtaining a population census is to identify the appropriate area frequency coordinators and establish the necessary security clearances and "need to know" for the collection of the data.

In obtaining the census of equipments, it is important to include not only those in operation but those planned for future installation. This information may not necessarily be available from the area frequency coordinators, and it may be necessary to contact other installations to obtain this information. For the Air Force the GEEIA (Ground Electronic Equipment Installation Agency) is responsible for the installation of radars. Its headquarters is at the Rome Air Development Center (Rome, New York) but it also maintains three regional establishments, Eastern GEEIA at Rome, New York, Central GEEIA at Tinker Air Force Base, Oklahoma City, Oklahoma, and Western GEEIA at McClellan Air Force Base, Sacramento, California. These commands have information pertinent to radars installed, being installed, and planned for installation. This data will, in general, augment data furnished by area frequency coordinators. Similarly, Naval Districts and Army Districts will also serve to provide additional information as to equipments that are being considered for installation.

An additional perplexing problem which occurs in areas near important sea coast Naval bases concerns radars which are operated on Naval vessels. Frequently, these equipments are given cochannel assignments with land-based equipments. For proper operation these shipborne equipments should not be operated within line-of-sight distance of land; however, these equipments are rarely shut down at this point and the presence of a large Naval task force in a harbor may cause considerable interference which is not otherwise present. Obviously, it is essentially impossible to accurately account for the possible presence of these equipments when making an RFI prediction study. The most realistic rule to follow is to assume that any harbor will always be a source of potential interference and locate equipment to discriminate against radiations from that general area.

6.2.4 DEFINING ACCEPTABLE INTERFERENCE LEVELS

After the population census has been completed, it is necessary to determine what importance will be attached to a given interference level. When this is done, then various screening processes can be carried out to determine the cases where interference is likely and those where it is unlikely. The problem is usually resolved by laboratory measurements and calculations which determine the signal-to-interference ratio that can be tolerated before a receiver's performance is considered to be degraded.

Since the number of signal generators available for simulating interfering signals in the laboratory is usually very limited, the true field situation where many interfering signals of varying levels may be present at the receiver terminals cannot be adequately duplicated. In fact, rarely is more than one signal generator available to determine the effect of interference. Because of this, it is necessary to make theoretical assumptions about the weighting factors that should be given to various types of interfering signals at various frequencies and assumptions are also necessary concerning the manner in which the cumulative effect of these signals should be treated.

In the case of certain methods of transmission and specific types of information content, practical figures for tolerable interference levels have been developed over the years. In other cases the problem of interference is a new one either because the system is new or because it has previously been operating in an interference-free environment (for example, a radar in an isolated location), and it has not been necessary to consider the problem of acceptable interference levels.

Regardless of how detailed the determinations are or how many years of experience have gone into setting acceptable interference levels, there rarely exists a sharp dividing line where it can definitely be stated that any interference above this level is absolutely intolerable and any interference below this level is perfectly acceptable. In short, there always exists a gray area where the nuisance effect of the interference depends considerably on the final use of the information. As an example, it is usually stated that a high quality television line should have a signal-to-noise ratio of 30 db, but it is not likely that a signal-to-noise ratio of 27 db would be regarded as giving a degraded picture.

In general, it can be stated that voice channels are among the least critical as regards interference, especially if it is of the intermittent kind. Any system using sophisticated modulation techniques, such as pulse coded frequency modulation, is also very immune to interference if the power levels used are equivalent to those used for conventional links. In some cases the interference immunity of the sophisticated system is used as a justification for using less power, and in this case some immunity to interference is lost. As mentioned above, television is fairly critical to noise and interference. Even intermittent interference is highly annoying in this case. One of the most critical types of information links that will be encountered is a digital data channel. With the present state of the art, the redundancy in such a transmission is not great and any interference which could destroy a "bit" of information could cause disastrous results to the user.

The entire problem of setting acceptable interference levels will be discussed in considerably more detail in Volume II under the title "Signal Acceptability Criteria."

6.2.5 FIRST ORDER SCREENING

The first part of an RFI prediction consists of a preliminary (first order) sorting process which classifies the possible interference situations into one of three categories: (1) definite interference, (2) definite non-interference, and (3) gray area cases. This technique is made relatively simple and easy to apply by considering only those factors which grossly affect the calculation. In this way a large number of situations can be investigated quickly and the cases where detailed calculations are required are drastically reduced.

For the first order process the effect of the earth and objects surrounding and between the transmitting and receiving antennas are neglected. This implies that all propagation calculations are carried out on the basis of free-space transmission. This is normally expressed in so many db of loss between isotropic sources (or half-wave dipoles) in common use. See Figure 4-5, Chapter IV. In this calculation only the gross effects of the transmitter and receiver spectrum signatures are usually considered.

6.2.5.1 Off-Axis Pointing Considerations

Interference reductions due to the fact that either the transmitter or receiving antenna or both are not pointing directly towards each other are called off-axis pointing losses. Aside from the free-space propagation loss, these losses represent one of the major additional losses that are accounted for in the first order screening. Specific antenna horizontal polar plots should be used whenever they are available. Where high gain antennas (gain greater than 10 db) are employed and the data is not available, the pattern can, in general, be approximated by reference to available data on an antenna which is closest to the type under consideration. A discussion of the various antenna types and their patterns is given in Chapter IV. These typical patterns can then be used to calculate the off-axis loss to a close approximation for use in the first order process. These pattern approximations, of course, hold true only at the design frequency of the antenna and usually experimental data is required when spurious and other harmonic frequencies must be considered. Some idea of the patterns to be expected at harmonic frequencies are listed in the references given in Figure 4-4, Chapter IV.

Corrections for off-axis pointing in the vertical plane are usually neglected except for very unusual geometrics at this stage of the calculations.

When the effect of off-axis pointing losses appears to be important, it is usually helpful to make a site sketch showing a plan view of the antenna locations under consideration. A typical site sketch is shown in Figure 2-13.

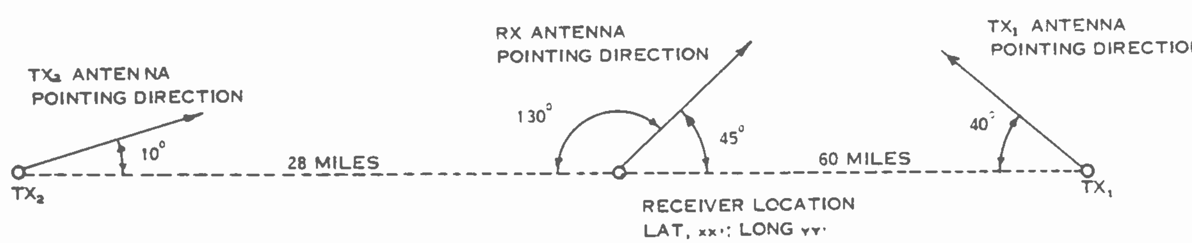


Figure 2-13. Site Sketch

6.2.6 SECOND ORDER SCREENING

The results of the first order screening will have eliminated a large number of the equipments from further consideration. Because the free space propagation calculations will give the highest possible RFI levels, then, if under these conditions it is found that no interference should result, it will be very unlikely that additional refinements will make interference probable. The equipments that remain to be considered at this point consist then only of those where gray area or positive interference situations were found.

The second order screening attempts to establish a definite division between positive interference cases and gray area cases. To do this, it becomes necessary to make more accurate predictions by taking more factors into account. At this point the effect of the earth must be accounted for since even for a smooth earth the RFI at a given receiver may be lower than predicted by the free space loss because the transmitted wave which reflects from the earth may totally or partially cancel out the direct wave from the transmitter. The effects of terrain features such as mountains, cities, forests, and other factors which influence the signal strength are also accounted for. The results of this order of screening will probably be that more RFI situations will be placed in the gray area; also it is quite likely that some situations which had previously been classified as positive RFI cases will be placed in the non-interference category.

6.2.7 THIRD ORDER SCREENING

The third order screening attempts to essentially eliminate the gray area by giving a fine grain structure to the prediction calculations. In this way most gray area equipments can be placed definitely in the class of either interfering or non-interfering situations. This level of investigation begins the detail process which accounts for the exact structure characteristics of the spectrum signature of the transmitters which have survived the previous screening and therefore need to be carefully considered. In general, the complete spectral output of each transmitter to be used should be known. This includes the fundamental, sideband, harmonic, and spurious output frequencies and signal levels. The antenna characteristics and feed system, coupling system and presence of possible non-linear elements in the final stage should also be known. When complete data on these factors are available, reliable computations of power densities can be made. At this point precise geometric considerations must be used to determine exactly the propagation path that will exist between the transmitting and receiving antennas. If possible, it is also desirable to account for the effect of the antenna pattern by using the actual measured pattern of the transmitter and receiver at the frequency for which the calculation will be made. In this final stage of the prediction process, it is necessary to make calculations not only for the fundamental frequencies involved, but for all possible combinations of frequencies that could cause interference to the receiver. For example, the second harmonic of a transmitter might not fall into the fundamental pass band of the receiver, but it might fall near the frequency of the image response. In this case, it would be necessary to account for the actual measured antenna patterns at this frequency. In general, this information will not be available and some assumptions will have to be made.

An example of the usefulness of the more complex propagation calculation which accounts in detail for the presence of the earth is shown in Figure 2-14 for a highly reflective ground condition. It can be

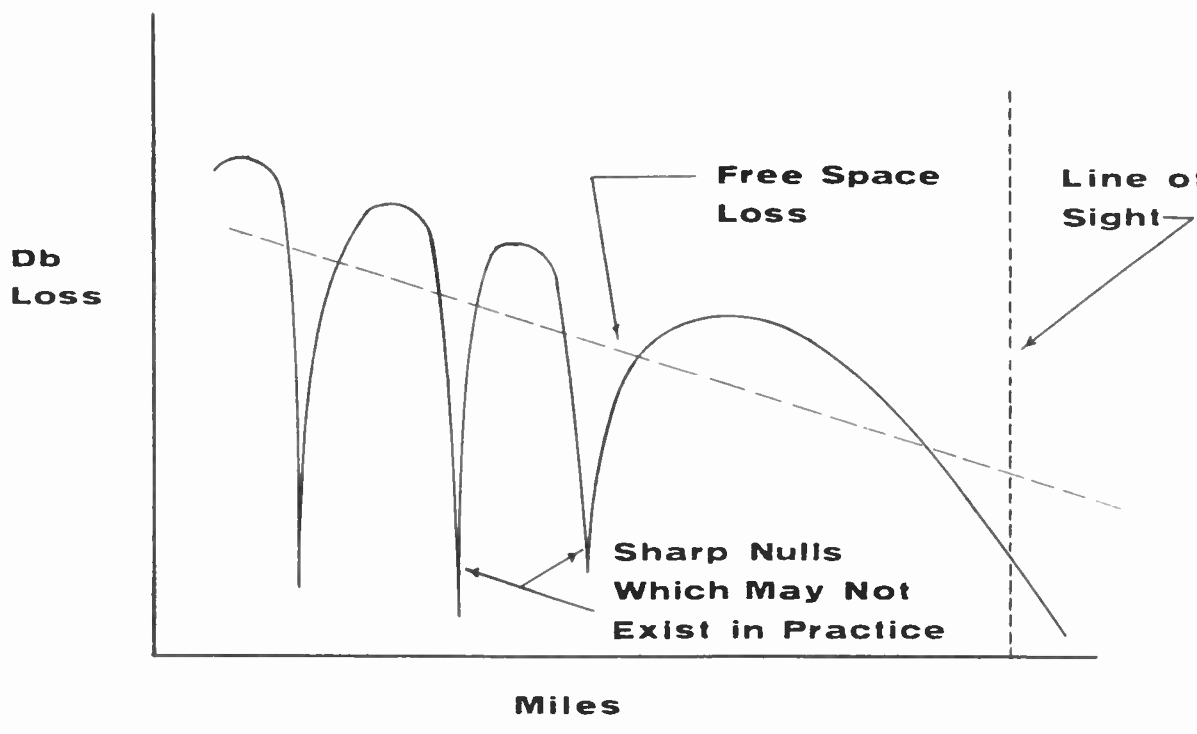


Figure 2-14. Propagation Loss Versus Distance Showing Sharp Nulls

seen that an equipment located in a null region would be relatively immune to RFI if the null actually existed. However, since atmospheric instabilities usually prevent the actual stable occurrence of a sharp null, it would be unwise to depend on such a prediction calculation for protection against RFI. Figure 2-15 shows a shallow null obtained from propagation over ground of different characteristics. The predicted field is about 10 db below free space. This type of null condition will be stable and can be depended upon to provide RFI protection greater than the free space calculations would have predicted. This illustrates that it is not always satisfactory to make the detailed propagation calculations only for the exact distance involved, but for several distances on each side of the equipment in question to determine the shape of the curve in this region.

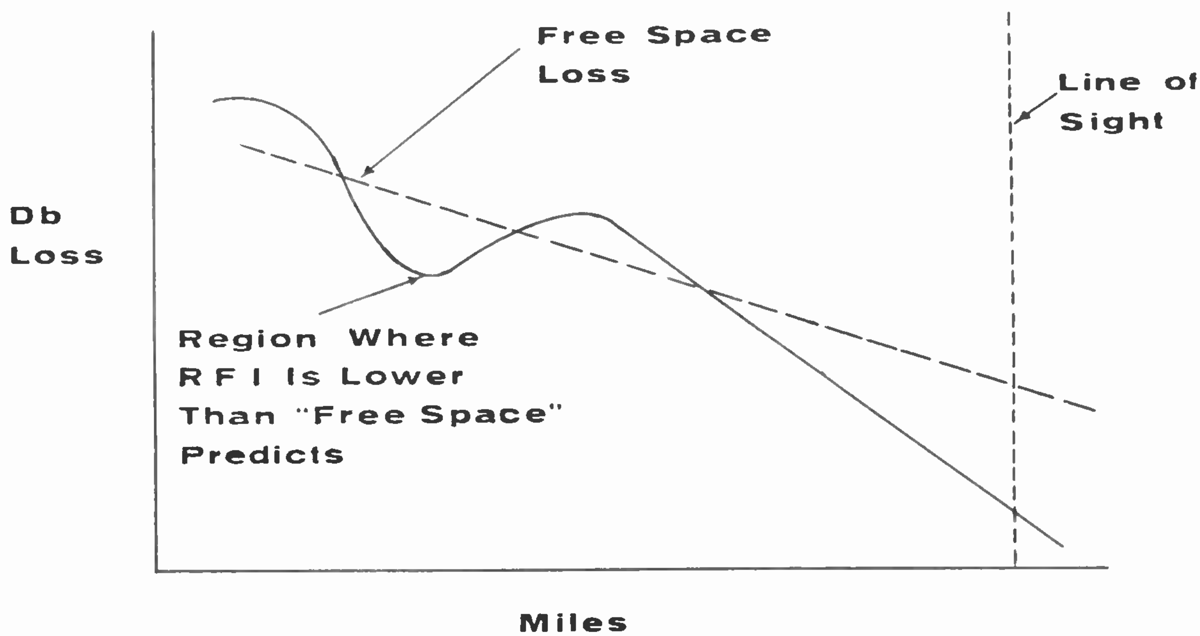


Figure 2-15. Propagation Loss Versus Distance Showing Shallow Nulls

6.2.8 DISCUSSION OF THE INTERFERENCE EQUATION

It was mentioned above that RFI prediction is basically the determination of the signal-to-interference ratio. A mathematical expression for this ratio will now be derived. This expression will not necessarily be in a form which is most useful for field calculations, but the derivation in this manner will serve to focus attention on the fundamental principles involved.

First consider an isotropic point source (in this case the interfering source) in free space radiating a total power P_T equally in all directions. To determine the power density at a distance R from the point source, the amount of power passing through a unit area on the surface of a sphere of radius R with its center located at the source is considered. Under these conditions the power density is:

$$P_D = \frac{P_T}{4\pi R^2} \quad \text{watts/meter}^2 \tag{2-4}$$

where $4\pi R^2$ = the surface area of the sphere

P_T = the total radiated power in watts

If the isotropic source is now replaced by an antenna with directive properties, the power density in the direction of maximum radiation is:

$$P_D = \frac{P_T}{4\pi R^2} G_T \quad \text{watts/meter}^2 \tag{2-5}$$

where G_T = the maximum power gain of the transmitting antenna

Note that the antenna gain is synonymous with directivity as applied here, since it has been assumed that there are no losses between the transmitter and the antenna output.

Now assume a receiving antenna with an effective aperture A_R located on the surface of the sphere in a manner such that it is illuminated by the main beam of the transmitter antenna. The power received is then

$$P_R = P_D A_R = \frac{P_T G_T A_R}{4\pi R^2} \quad \text{watts} \tag{2-6}$$

Since antenna gain and effective aperture are related under matched load conditions by:

$$A_R = \frac{G_R \lambda^2}{4\pi} \tag{2-7}$$

then the equation for the received power is more conveniently expressed as:

$$P_R = \frac{P_T G_T G_R \lambda^2}{(4\pi R)^2} \quad \text{watts} \tag{2-8}$$

where: λ = wavelength of the signal in meters

G_R = gain of the receiving antenna

The above equation, commonly known as the Friis transmission formula, enables the determination of the energy transfer from one antenna to another under optimum conditions. In general, these conditions will not exist in a practical situation and certain modifications to the equation must be introduced.

Since the total transmitted power may not lie entirely within the band pass of the receiver, as in the case of a receiver tuned to a spurious emission, the term P_T is replaced by P_{TR} where the latter includes only that portion of the transmitter spectrum which is contained within the receiver band pass. Further correction factors are introduced to account for possible misalignment of transmitting and receiving antenna polarizations, P, additional losses due to terrain conditions, A, and antenna line losses, L_T and L_R. Finally, the antenna gains are corrected for situations in which the main beam of either or both antennas is off-axis, either vertically or horizontally, with respect to a straight line connecting the two. The adjusted antenna gains are denoted G_{TA} and G_{RA}.

With the above considerations included, the equation for the receiver power, or in this case interference, becomes:

$$I = \frac{P_{TR} G_{TA} G_{RA} \lambda^2}{(4\pi R)^2 P A L_T L_R} \quad \text{watts} \tag{2-9}$$

It now remains to determine the second half of the S/I ratio; that is, the signal level. Here two possibilities exist: (1) the desired signal level at the receiver is known, as in the case of fixed point-to-point microwave communications, or (2) the signal level is not known, as in the case of a radar return.

In the first case the signal level will be known from system design criteria and the signal-to-interference ratio is easily computed once the interference level is known. In the second case some criteria must be established as to what signal level is to be considered in the computation of S/I. In general, this level is taken to be the minimum acceptable signal which may be defined as:

$$S_{Min.}\,(\text{dbm}) = N\,(\text{dbm}) + \frac{S}{N}\,(\text{db}) \qquad (2\text{-}10)$$

where: N = the internal noise level of the receiver (often referred to as the minimum discernible signal; that is, MDS)

S/N = the signal-to-noise ratio necessary for proper equipment operation

The interference to noise level then becomes:

$$I/N = \frac{P_T G_{TA} G_{RA}\,\lambda^2}{(4\pi R)^2\,P\,A\,L_T L_R} \cdot \frac{1}{N} \qquad (2\text{-}11)$$

or in logarithmic form:

$$I/N\,(\text{db}) = 10\log_{10} P_{TR} + G_{TA}\,(\text{db}) + 20\log_{10}\left(\frac{\lambda}{4\pi R}\right)$$
$$- P\,(\text{db}) - A\,(\text{db}) - L_T\,(\text{db}) - L_R\,(\text{db}) - N\,(\text{dbm}) \qquad (2\text{-}12)$$

Note that P_{TR} is now in milliwatts.

The term 20 log $(\lambda/4\pi R)$ is the free space transmission loss between isotropic sources and its value may be read directly from a nomograph for given values of frequency and separation. This nomograph will be discussed in detail in Volume II in the prediction section.

Using the above equation, the signal-to-interference ratio is found to be

$$S/I(\text{db}) = S/N(\text{db}) - I/N(\text{db}) \qquad (2\text{-}13)$$

Equations 2-12 and 2-13 provide the basis for the analysis of signal-to-interference conditions at a given receiver caused by a particular interfering source. It must be emphasized again that the results thereby obtained are in themselves insufficient in the comprehensive evaluation of interference situations. In any case, consideration must be given to such factors as the type of information being received, system reliability standards, and other user criteria.

COMPATIBILITY AND INTERFERENCE REDUCTION — CHAPTER 3

1. ENGINEERING FOR MINIMUM INTERFERENCE AS A BASIC PART OF SYSTEM PLANNING

The ideal method of dealing with interference is to design all equipment in such a manner that it does not generate or receive interference. This is not always possible, since in many cases a source of interference either is not known or cannot be controlled, or it may necessarily generate or receive interference while performing its normal function.

To avoid this situation, compatibility is a major consideration in the design of present and future equipment. The interference solution would be simple were it possible to assign a part of the spectrum exclusively to each user, as for example, a band of frequencies assigned to the Air Force, another to the Army, a third to civilian use, etc. Unfortunately, due to the physical nature of radio transmissions, certain frequencies are useful for certain purposes and others are not. This fact concentrates like devices into the same part of the radio spectrum regardless of the user.

In general, it can be said that a high degree of interference (possibly to the point of complete system inoperability) can be expected unless specific planning action is taken to prevent such interference. This planning must include appropriate frequency assignment, satisfactory geographical location, careful design of equipment, well-engineered installation, good grounding, bonding and shielding practices, and well-planned antenna location and orientation.

It is particularly important that the spectrum situation be carefully analyzed at the proposed location and that the area near the location be examined and screened for installations of other users which might be interfered with or cause interference. These two aspects of the planning activity should be carried out together since they are generally dependent on each other.

In general, cochannel and adjacent channel operation of C-E systems will result in some degree of mutual interference. Adequate frequency separations should be provided in the assignment of transmitter operating frequencies so that cochannel and adjacent channel operation is avoided. In cases where this cannot be accomplished, maximum use should be made of natural screening, isolation and separation of antennas, and directivity of antennas.

When planning an installation involving several pieces of electronic equipment, careful thought must be given to the location of the various units with respect to one another and to their surroundings. Much interference can be avoided at the outset by a suitable physical arrangement of the equipment to be used. Sensitive receivers should be installed as far as possible from units which may be interference generators. Any natural or artificial barriers which may be present should be utilized, where possible, to separate these units. All power, control, and other leads to the receivers should not be physically close to any interference generating equipment or interference carrying leads because of the inductive coupling which may exist between the leads.

Good bonding between allied electrical units can be very important in the reduction of interference generation, in minimizing the effect of interference once it is present, and in increasing the effect of suppression components such as filters, capacitors, and shields. Poor bonding between two points may allow large potential difference to build up between them. If these points become excessive, they may cause arc or spark discharges to occur.

By completely surrounding a source of radio interference with a metallic shield, the amount of interference escaping to the exterior of the enclosure can be greatly reduced. Shielding is useful not only as a means of preventing interference from leaving the neighborhood of the source, but also may be used to surround a receiver and keep interfering signals from reaching it.

Complete shielding of an interference source or a receiver is never possible because there will always be power and/or other connections entering the equipment. These leads provide paths of exit or entry for interfering signals. The reduction of interference transmission along these wires is accomplished with the aid of capacitors or filters. They may be used at the source to keep interference from being transmitted, or at the receiver to keep interference from entering. Other things being equal, it is usually better to have the suppression built in at the source.

Arcs occuring during switching and other processes should be suppressed as completely as possible because they are serious sources of radio interference and because they produce a rapid deterioration of the contacts. Where the existence of arcs is necessary, as for example, in ignition systems or welders, an attempt must be made to prevent the resulting interference from reaching receivers.

2. IMPROVEMENT OF EFFECTIVENESS BY USE OF HIGH POWER VERSUS INCREASE IN INTERFERENCE

Theoretically, the minimum power required for a radio signal is that required to produce at the receiver enough energy so that the signal will override the noise that is inherent in the circuits. Quite obviously this condition can only be obtained in a laboratory, as that is the only place where the presence of absolutely no external noise can be assured.

If a system is to be operated in a situation where interference is known to exist, a useful signal can be produced at the receiver output by the use of sufficient power output at the transmitter to override interference plus receiver noise. Such a course ignores the fact that additional interference to other systems may result from increased transmitter power output. In addition, while unlimited primary power may be available to some transmitters it may not be available for use with mobile and portable equipments. There is also a practical limit to the power which can be produced and radiated by communication-electronic equipment.

It is clear that the power to be used in a C-E system must take into account all of the above factors. In the design of such a system, the first consideration is the determination of the power required to deliver a usable signal. That is, the signal should have a sufficient margin of energy to insure that a reliable signal will always be maintained. The next consideration is in respect to the means that may be adopted to minimize the effect that the use of such power will have on the use of the spectrum by others.

In considering the first phase of the above study, such things as directive receiving antennas, filters and shielding come into consideration. These measures are designed to reduce and control the entry into the receiving systems of interference that cannot be avoided. They also increase the sensitivity of the receiver in the desired direction while reducing it in others. In the second phase, transmitter shielding and filters are used to prevent the radiation of unintended signals, but the use of directional transmitting antennas is of the most importance. Directional antennas will multiply the effect of radiated power many times, and, in addition, will considerably limit the geographical area in which the transmitter will generate signals at an interfering level.

To summarize, although interference may be overcome by the use of high power transmitter output in any single instance, the use of this means by large numbers of transmitters only compounds the interference problem. The use of unnecessarily high power for one transmitter may make impossible the utilization of many other systems which are of equal importance.

3. CHARACTERISTICS OF COMMONLY USED COMMUNICATION-ELECTRONIC SYSTEMS

To adequately consider compatibility and interference reduction in C-E systems, it is necessary that the characteristics of individual systems be considered. The discussion that follows in this section presents general information on radio, radar, and wire systems with particular emphasis on interference possibilities.

3.1 RADIO SYSTEMS

3.1.1 STANDARD AMPLITUDE MODULATED (AM) RADIO

Most radio systems for sound transmission in the lower frequency range use amplitude modulation (AM). Many systems in the VHF range also use this type of modulation. The envelope of the RF carrier is varied in proportion to the instantaneous amplitude of the signal. Thus, a sine wave signal produces a sine wave envelope. The modulated wave can be expressed as the sum of three frequency components, one at the carrier frequency, one at the carrier plus modulating frequency, and the third at the carrier minus the modulating frequency. The latter two are called sidebands. Since sound transmissions are complex waves, the signal is represented as a carrier with a similar spectrum on either side. The total width is twice the highest audio frequency to be transmitted. In the absence of transmitter harmonics and other unwanted frequencies in the transmission, there will be no interference outside the channel, except that receivers on an adjacent channel could be interfered with if they are not sufficiently selective. Under appropriate propagation conditions, the sky wave from a distant transmitter can be picked up by a receiver tuned to the same frequency and interference occurs. In general, AM receivers in the low frequency range do not cause interference. Regenerative receivers cause radiation but such receivers are no longer in common use. Local oscillators of superheterodyne receivers can radiate but this is not a problem at the lower frequencies.

3.1.2 KEYED CONTINUOUS WAVE MODULATED RADIO

Keyed continuous wave modulation is used for aural reception of Morse code and for similar transmissions. The signal is made audible either by means of a beat frequency oscillator in the receiver or by tone modulation of the carrier in addition to keying. Keying is equivalent modulation with short pulses and the signal transmitted can be represented as a carrier and two sidebands. The bandwidth necessary to transmit the pulse is inversely proportional to pulse duration. It requires a bandwidth at least 2 kc wide to transmit a one millisecond pulse. With this minimum bandwidth the pulse is somewhat rounded off but this effect does not detract from the usefulness of the code transmission. Since narrow band amplifiers are used before transmission, the only frequency components present in the signal are those within the channel, that is, the carrier frequency plus or minus half the channel width.

Code signals may be picked up on associated audio communication circuits such as the telephone lines under some conditions. For example, the rectifier used for the plate supply of a transmitter may produce harmonics in the power supply which produce noise in the audio circuits. When the rectifier is switched on and off during keying, the noise is also "coded." Harmonic filters on the rectifier remedy this condition.

3.1.3 FREQUENCY SHIFT KEYING

Frequency shift keying is used for automatic radio telegraph systems and teletypewriter systems. The amplitude of the carrier remains constant and the frequency is changed by a definite amount. The amount of shift depends on the rate of transmission. For 60 words per minute a frequency shift of plus or minus 50 cycles is sometimes used with a receiver bandwidth of about 170 cycles. Some teletypewriter systems use as much as plus or minus 425 cycles frequency shift. Under comparable conditions a system using frequency shift keying would work at lower signal-to-noise ratios than one using keyed continuous wave modulation but actually a higher signal-to-noise ratio may be necessary to obtain the desired rate of transmission. As in frequency modulation, frequency components are produced by the shifting which could cause interference but the spectrum is less extensive than it is in FM transmissions. Single sideband operation can be used with frequency shift keying in multichannel systems to conserve space in the frequency spectrum.

3.1.4 TELEVISION

Standard television transmissions contain an amplitude modulated (vestigial sideband) picture carrier, a frequency modulated sound carrier, and synchronizing pulses. In color television other carriers are needed for the color information, and the system was designed to prevent mutual interference. The usual picture carrier transmits brightness information and a pair of signals represent the color information. The two color signals are applied to a single color carrier in such a way that it is modulated both in amplitude and in phase. In order to obtain demodulation in the receiver at the proper time, a color phase signal is transmitted along with the synchronizing pulse.

There are several types of spurious radiations from television receivers that cause interference to other radio services. Included are: local-oscillator radiation; horizontal sweep circuit radiation; and IF amplifier radiation. Not only the fundamental, but various harmonics from these sources of interference are radiated and can cause interference to communications equipment. In each case, the interference leaves the television receiver via either the antenna, the power line, or by direct radiation from the receiver chassis or associated wiring and components.

An effective, properly designed, high-pass filter installed at the front end of the television receiver will prevent radiation by the antenna of signals from the IF amplifier and horizontal sweep circuit. Radiation of the local oscillator and its harmonics can be prevented by the Installation of band-pass filters in the RF and mixer circuits along with proper shielding of these stages or an additional number of RF stages. The power line can easily be filtered for all frequencies by using high-pass capacitors and RF chokes.

3.1.5 SINGLE SIDEBAND TRANSMISSIONS

In amplitude modulation all the information is contained in the sidebands, and since they are alike, only one of them is necessary. For transmission purposes the carrier and one of the sidebands can be suppressed and the frequency components in the other sideband transmitted. This in effect shifts the center frequency to another point. At the receiver a component which has the same frequency as the original carrier is superimposed to obtain proper demodulation. A modification known as vestigial sideband operation may be used for better transmission of low frequency audio components. One sideband is transmitted and the other is not completely suppressed. An advantage of the single sideband system is that it requires less space in the spectrum than standard AM (double sideband), and therefore more channels can be used without mutual interference. An AM signal can interfere with an SSB signal of considerably higher power if the carrier frequency of the AM transmission is at the center of the SSB transmission.

3.1.6 TELETYPEWRITER TRANSMISSIONS

The teletypewriter delivers a message in the form of a typewritten page in response to signals from a distant point. The signals may be transmitted by wire or by radio. For radio transmission either keyed CW or frequency shift keying may be used. Many installations use a five-unit start-stop permutation code. It is composed of five selecting intervals which may be either "marking" or "spacing." Sufficient combinations are available to transmit 32 characters.

When teletypewriters used with radio transmission are at some distance from either the radio transmitter or the receiver, tone keyers and tone demodulators are generally used so that the keying pulses can be transmitted and received over normal telephone lines.

Teletypewriters and printers contain relays and switches which are potential sources of interference. Suppression components are installed in most machines at the time of manufacture. The most common cause of interference is the absence of a good ground connection. A separate grounding point, from the one used for receivers, should be used for the interfering equipment. If the teletypewriter machines and receivers operate near each other, separate power circuits should be employed if available. If interference is due to defective filters in the teletype machine, they should be replaced only with those types specified by the manufacturer.

3.1.7 PULSED AND DIGITAL RADIO SYSTEMS

Pulsed systems generally operate in the higher frequency range such as the UHF range. In most of these systems the signal to be transmitted, considered as a curve of amplitude versus time, is sampled at regular time intervals. The interval is short enough so that there will be little or no loss of information. The transmission consists of a train of pulses which modulate the high frequency carrier. With pulse amplitude modulation the amplitude of the pulse is proportional to the signal amplitude at the point sampled. With pulse time modulation the interval between pulses is proportional to the signal. Another form known as pulse code modulation translates the signal amplitude into a code which is transmitted, decoded at the receiver, and the signal reproduced. Use of pulse code modulation offers one possibility for reducing the effects of interference in some cases because the interference has much less effect on the coded signal than on the signal itself. On the other hand, the transmission of pulses produces frequency components, which could interfere with other communication or electronic systems or equipment.

3.1.8 FREQUENCY MODULATION (FM) RADIO

Frequency modulation (FM) was originally developed as a means of overcoming the effects of atmospheric noise. Since noise is mainly amplitude variation, it has a greater effect on amplitude modulation than on other types of modulation. With FM the amplitude of the carrier remains constant and the frequency is shifted back and forth so that the deviation from the mean frequency is proportional to the amplitude of the signal being transmitted. There is another form of modulation known as phase modulation which is related to frequency modulation. The amplitude of the carrier remains constant and the phase angle changes. Limiters can be used in FM receivers to cut off noise peaks which exceed the carrier. Although FM systems were supposed to be "interference free," in practice lower power is used for FM systems than would be needed for AM systems so that the signal level is lower and therefore the relationship between signal and interference is about the same as in other systems.

FM transmitters are a common source of noise at very high frequencies. In addition to the carrier or center frequency, FM transmitters produce a continuous spectrum of sideband noise. The noise consists of random voltages with an infinite combination of amplitudes and frequencies. The amplitudes of these random voltages generally increase as the carrier frequency is approached, and decrease above and below this frequency to a point where they are negligibly small.

FM receivers can cause interference due to radiation from their local oscillators. This may be caused by either the fundamental frequency of the local oscillator, or by harmonics, particularly a second harmonic. Radiation of the local oscillator and its harmonics can be prevented by the installation of band-pass filters in the RF and mixer circuits along with proper shielding of these stages.

3.1.9 FACSIMILE

In a facsimile system a drawing, picture, or message is scanned and a signal is transmitted to a receiver which delivers a reproduction of the original. The signal can be transmitted by wire or by radio. Some types of facsimile receivers contain an amplifier and a grounded drum and stylus across which an electric potential is applied to obtain the graphic presentation. Sparking at the stylus may be a source of interference to other communications-electronic equipment. Resistance in series with the stylus reduces the interference and a small amount of shunt capacitance may help also. The equipment should be well shielded and the power supply should be filtered.

3.1.10 DATA-LINKS

The purpose of a data-link is to transmit data to a point where data is accumulated or processed. This point may be a center where data from a number of locations is brought together and processed. The data is put in the form of electric quantities which can be transmitted over wires or used to modulate the output of radio transmitters. Data in analog form may be converted to digital form for transmission. The transducers and associated equipment such as switches, motors, slip rings, commutators, and power supplies may contain sources of interference to other communications-electronic equipment. There may also be sources of interference in associated computers. When the sending and receiving locations are permanent, the radio link may use beamed radio transmission. This reduces the possibility of the signals being affected by external interference and also reduces the possibility of causing interference to other transmissions.

3.1.11 TELEMETRY

The purpose of telemetry is to reproduce the indication of a measuring instrument at some other location at any desired distance. The transmission may be by wire or partly by wire and partly by radio. There are many different kinds of telemetry. The simplest form is to convert the quantity indicated by the instrument into an electric current or voltage, usually alternating current or voltage. The amplitude of frequency is varied in porportion to the quantity being measured. The current or voltage can be used to modulate a radio transmitter by means of amplitude, frequency, pulse or pulse-code modulation. If pulse-type modulation is used, the current or voltage can be sampled at regular intervals and a signal transmitted. The receiver output is converted to a meter indication similar to the meter indication at the source. The signal may be transmitted from the source to the radio transmitter over existing telephone wires or cables in which case the voltages should not exceed 120 volts and the current should not exceed 0.1 amp. Harmonics should be suppressed so that audible noise is not produced in other telephone circuits in the same lead or cable.

For transmission over short distances, the measuring instrument may actuate the shaft of a transmitter such as a selsyn or autosyn which consists of a rotor and a stator. Electric currents are transmitted over wires to a similar device at the receiving end wherein an electromagnetic field is set up which causes the rotor to assume the same angular position as the transmitter. This equipment in itself does not cause interference.

3.2 RADAR SYSTEMS

Radars may cause interference not only to other radars but also to communications and other C-E equipments. Radar receivers are usually more susceptible to interference than the average communications receiver because of greater bandpass and poorer RF selectivity. Bandpass characteristics are necessarily wide for proper reception of pulsed signals and the design problems above 1000 mc make it difficult to provide a high degree of RF selectivity. The various causes of interference to receivers in the HF, VHF, and UHF ranges from nearby high-powered radar transmitters are covered in this section.

3.2.1 STANDARD PULSE MODULATION

Although the effect of interference from pulse modulation is similar in all cases, the method by which the interference is being produced must be determined before steps can be taken to reduce or eliminate it. In the following paragraphs, the interference is classified according to the method by which it is generated and some of the variable factors which may influence the degree of interference are discussed.

3.2.1.1 Interference From Pulsed Signal Rectification in the Second Detector

When an interfering radar signal reaches a detector, it will be rectified and the pulse repetition frequency will appear in the output of the receiver along with the desired signal. Normally, the interfering signal will be at a different RF than the receiver frequency; it may, therefore, be attenuated in the RF and IF circuits, and may not reach the second detector with sufficient amplitude to cause interference. If the receiver is not well shielded, it is possible that the interfering signal will be received directly at the detector. If the second detector is directly coupled to the video amplifier, a small amount of RF signal could develop enough voltage across the detector to bias the first video stage out of its operating range. This can be avoided by coupling the video amplifier through a blocking capacitor. The receiver will then continue to function normally until the IF amplifier is overloaded. The overload limit can be increased by design. The use of a gain control early in the IF amplifier is of value when strong RF signals are present. Grid gain control is satisfactory in this respect. To protect against either amplitude or frequency modulated RF, filters can be used which can be switched in or out as needed. For example, a filter might be used which shortens the time constant of the coupling element so that short pulses are passed with slight attenuation, whereas long pulses and very low modulation frequencies below, say 20 kc, are greatly attenuated when the filter is switched in.

3.2.1.2 Sidebands at the Receiver Frequency

Interference may be caused by pulsed signals because they have a large number of sidebands of appreciable amplitude extending far on either side of the carrier frequency. With a rectangular pulse, the frequency spectrum is a (sin x)/x function extending on either side of the carrier frequency throughout the RF spectrum. The amplitudes of these sidebands decrease with increasing frequency separation from the carrier and there are sharp nulls spaced at intervals as illustrated in Figure 3-1. The spacing between these nulls will be of the order of 2 kc or less for the pulse widths and repetition rates normally encountered in radar and IFF transmitters. For bandwidths normally used in communications receivers, these sidebands will form an essentially continuous spectrum. Since any interference to communications receivers would come from those sidebands below the output frequency of the pulsed transmitter, the upper sidebands can be disregarded. Most transmitters in the microwave region feed the antenna through a section of waveguide which has a cutoff frequency somewhat below the output frequency. For even short sections of waveguide, the attenuation at frequencies of 400 mc and lower is very high and the possibilities of interference are low. Interference should not normally result from sidebands generated in a pulsed transmitter above 1000 mc, if the antenna is fed through a waveguide.

3.2.1.3 External Rectification In Ground-Antenna System

External rectification of the pulsed signal can occur if poor connections (nonlinear elements) exist in either the antenna or ground system of the receiver. Although a certain amount of space attenuation will be present due to the spacing between the pulsed transmitter and the receiver, with a high-powered transmitter it is possible to have a sizable amount of power at the receiving antenna. If this signal is rectified, it becomes a low-frequency rectangular wave signal at the pulse repetition frequency. This signal will also contain harmonics with a (sin x)/x amplitude function extending from the repetition frequency through the high-frequency range, and possibly into the VHF and UHF ranges, with sufficient strength to cause interference in some cases. Even if the harmonics are not strong enough at the desired frequency to cause interference it is possible that lower order harmonics of greater amplitude will

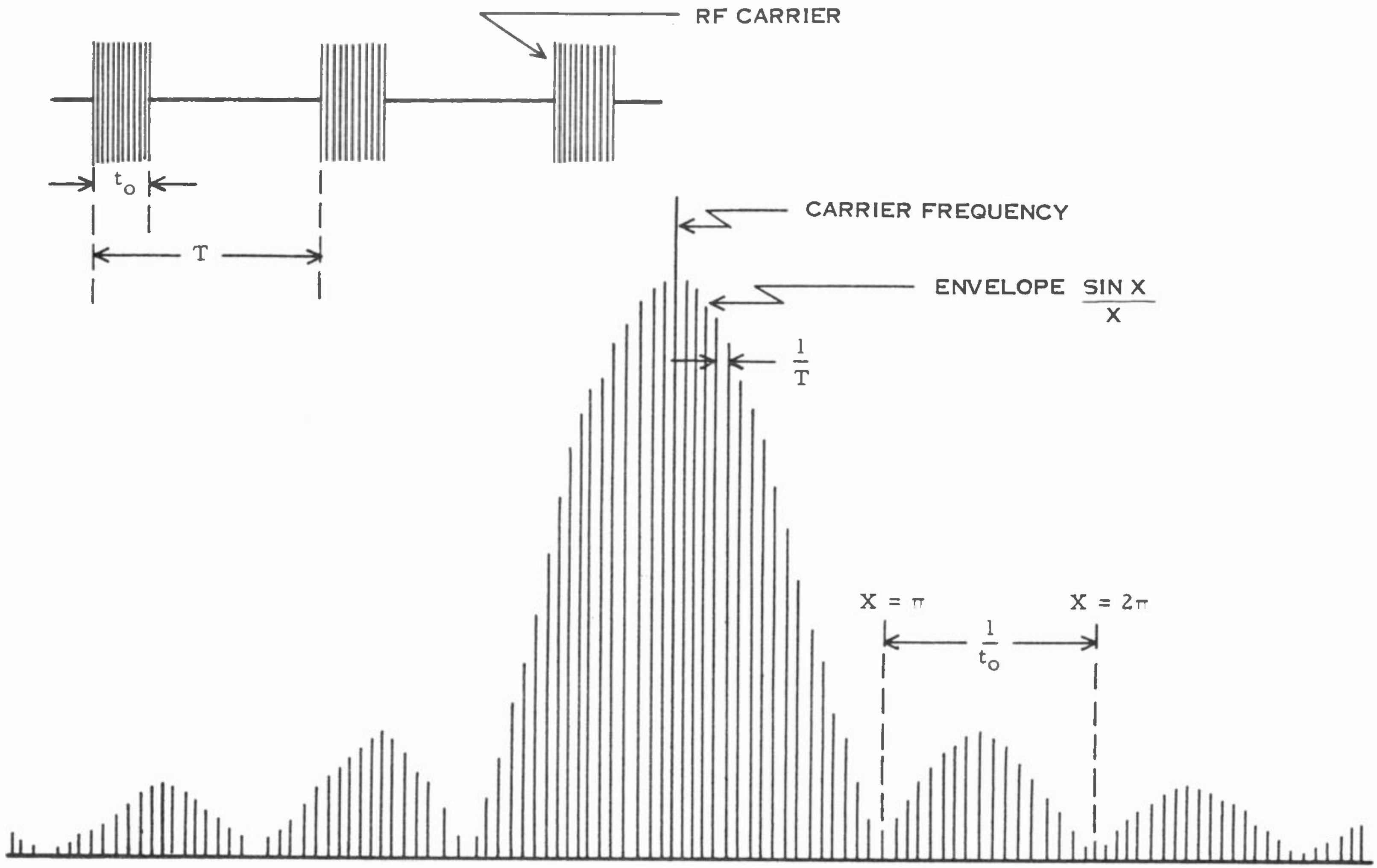


Figure 3-1. Radar Output Frequency Spectrum

pass through the RF stages of the receiver and be amplified by the IF system.

Another possibility of interference is cross modulation of the desired signal by harmonics of the rectified signal. Harmonics from an undesired transmitter may be generated by way of external nonlinear rectification and cause interference to a receiver. The extent of the interference is dependent upon the signal strength of the reradiated harmonic with respect to the signal strength from the desired transmitter. The strength of the reradiated harmonic is dependent upon the RF voltage applied, the efficiency of the rectifier, and the resonant frequency or frequencies of the system. External rectification in the antenna and ground systems can be prevented by providing well-soldered connections. In addition, insulation must be kept free of soot and rust, and copper wire should be kept free of corrosion.

3.2.1.4 Reradiation From External Detectors Not A Part of Antenna

Sources of rectification external to the receiver antenna system may be due to semiconducting joints in plumbing, electrical conduit, metal structural members in the building, equipment grounding straps, etc. It is possible for these devices to receive the pulse signal, rectify it, and reradiate harmonics of the pulse modulation.

Corrosion is a major cause of external rectification. The oxides found are generally good rectifiers and, if in contact with conductors, paths may form for conduction and radiation of unwanted signals. Corrosion conditions are accelerated in humid regions, particularly in coastal areas where salt spray is present. In industrial areas, corrosion may be increased by the presence of a slight amount of acid in the atmosphere.

Finding external rectifiers can sometimes become difficult, particularly when these devices are a part of the ambient building facilities. Suspected joints, connectors, and grounding straps can be physically moved and the interfering harmonic response can be watched at the receiver for any variations. Separating joints, cleaning connectors, and using shorting straps can also be effective for finding or for eliminating possible nonlinear sources. Direction finding using a grid dip meter and an auxiliary transmitter sometimes may be employed when rectification is occurring in hidden or in not easily accessible places.

RF radiation takes place most effectively from an antenna-like metallic body. Thus, when an external rectifier is connected to a metallic body, radiation effectiveness is considerably improved. Buildings should be recognized as complex antenna structures consisting of mazes of metal pipes, girders, wires, ducts, and fixtures.

3.2.1.5 Internal Rectification and Cross Modulation

Interference can occur from rectification of the pulsed signal at the grid of the first RF stage in the receiver. For this type of rectification to occur, the signal amplitude at the grid must be high enough to overcome the normal operating bias on the stage. When this bias voltage is exceeded, the grid and cathode of the tube act as a diode rectifier and produce an interfering signal. Since most receiving tubes used in RF amplifier service have a normal operating bias of one volt or more, interference will not be present unless the interfering signal is of large amplitude. In this case, the only method of interference elimination would seem to be a reduction in the amplitude of the signal at the grid of the first stage in the receiver to a point below the bias level. There would still remain some possibility of cross modulation, the severity of which would again be dependent upon the level of the interfering signal and the characteristics of the receiver.

3.2.1.6 Conducted Interference From Radars

Radar modulators are sources of interference which can be conducted through the power line. The audible effect is noise characterized by the pulse repetition rate. The modulator should be as close to the transmitter-receiver as possible. The case should be bonded to the structure. Primary power leads should be shielded from the transformer terminals to the point where they leave the modulator case, and there should be shields between the primary and the secondary of the power transformer and of the filament transformer. Rectifiers and thyratrons should be in separate shielded compartments inside the modulator case.

3.2.2 MONOPULSE RADAR

Monopulse radars are highly precise tracking radars which derive range and angle of arrival from a single target echo. This technique of angle detection is made possible through the utilization of electronic, rather than mechanical, beam scanning, and of more accurate methods of antenna angle sensing through phase measurement.

Monopulse radar antennas are arrays of accurately placed apertures capable of being fed a controllable distribution of signal intensity, resulting in the required scanning operation.

Angular measurement is characterized by either amplitude or phase sensing. Amplitude sensing utilizes two apertures with a common phase center and squinted diffraction patterns; phase sensing involves two apertures with separated phase centers and collimated diffraction patterns. In usual practice, the signal from both antennas is combined in a hybrid junction labyrinth from which elevation, azimuth, and quadrupolar components of the diffraction pattern are extracted. The echo signal is detected in an angle detector after appropriate amplification. Since monopulse radar is used for tracking purposes, the antenna pattern is normally a pencil beam, and therefore this type of radar does not normally present a serious interference problem.

3.2.3 DOPPLER RADARS

Doppler radars utilize the doppler effect which results when a wave is reflected from a moving target. The reflected wave differs in frequency from the outgoing wave by an amount which depends on the velocity of the target. The frequency difference is small but detectable. For some purposes, the bandwidth may be in the order of cycles and this has some advantage from the interference standpoint. Some doppler systems use pulse modulation. In one system the transmitter is turned on for half the cycle and off for half the cycle by means of a modulating square wave generator. The receiver input is subject to delay depending on the range and part of the received signal is cut off because the transmitter is off. The received signal is amplified by a superheterodyne receiver and then goes through a second detector. If the object is stationary, the second detector output is a train of pulses. If it moves, a periodic oscillation of the doppler frequency is superimposed on the pulses. A filter then attenuates the square wave repetition frequency, eliminating the signal due to stationary objects while leaving those due to moving objects undisturbed. The output is passed through a full wave rectifier. Then, the doppler sidebands and most of the noise are removed by another filter. The final output is a train of pulses; the trailing edge of which, relative to the modulating pulse, is a measure of the target range. This time or range may be displayed on the linear radial sweep of an oscilloscope by differentiating the voltage and using the result to intensity-modulate the beam. A wide band is not needed because only a few harmonics of the modulations square wave need be passed. The filter can have a fairly narrow band thereby reducing the noise.

3.2.4 FREQUENCY DIVERSITY RADARS

Frequency Diversity radars represent a class of modern heavy-duty equipment designed and constructed especially for permanent installation as part of the U. S. Air Defense System.

Frequency Diversity (FD) radars are characterized by very high powers and large aperture antennas, and this type of radar, therefore, warrants special consideration from an interference standpoint. The radiating systems may utilize unique design techniques, and the increased quantity and complexity of support equipment may require a much larger housing space. For example, the AN/FPS-35 radar is housed in a five-story building which contains the various components and units that make up the system. This is a very permanent and solidly built structure. The AN/FPS-28 uses an antenna structure in which the RF power is divided and subdivided into many individual waveguide feeds.

3.2.5 INTERROGATOR-RESPONDER SYSTEMS

Radar interrogator-responder sets transmit a signal which when received by a beacon or another interrogator-responder causes a return signal to be transmitted back to the first interrogator-responder. The received signal is a direct transmission, instead of a reflected wave, as in the case of a normal radar. Because of this, the energy content is higher and interference has less effect. With some systems such as the IFF, the return signal contains an identifying code.

3.3 WIRE SYSTEMS

Wire communication circuits and circuits associated with electronic equipment are subject to mutual induction unless precautions are taken. Early communication circuits consisted of one wire with the return through the earth. These ground return circuits are still used by telegraph systems subject to some limitations and by telephone systems for signaling. They are unsuitable for voice transmission because only a limited number of such circuits could be used without cross-talk and because ground return circuits are highly susceptible to interference. Cross-talk between two-wire circuits can be prevented by means of transpositions. Transpositions in different circuits must be properly arranged with respect to each other since otherwise the effect could be cancelled out. The continuous twist in a twisted pair of conductors has the effect of transpositions. Often the two conductors of one pair are used with the two conductors of another pair to obtain a third transmission circuit. Then the two pairs have to be transposed or twisted. Besides inductive coupling, capacitive coupling is a factor and may have to be considered especially when high voltage exists.

Frequency is an important factor. For voice transmission, only those frequencies between 200 and 4000 cycles are needed. A telephone system may be considered as functioning somewhat as a band-pass filter which passes that range. 60-cycle induction has little or no effect and therefore 60-cycle equipment used in communications-electronic equipment is considered as non-interfering when the wave shape is good. This does not apply when such equipment distorts the 60-cycle wave. 400 cycles is within the interference range and therefore 400 cycle equipment used in communications-electronic equipment needs special attention. Sometimes audio signals are used to modulate a carrier in the range of 30 to 300 kc or higher which is transmitted by wire. In many cases, regular power lines are used for such transmission for communication or other purposes such as control. The carrier may cause radiation of an interfering character. Since telephone lines with carrier are designed to transmit carrier efficiently with low power level, they are not expected to cause interference. Power lines, on the other hand, are not so designed, and require higher levels of carrier.

Interference problems have arisen in intercom systems such as used in offices and communication centers. These systems may include amplifiers, jack boxes, headphones, microphones, microphone switches and control panels. Coupling between the input circuit and the output circuit of the amplifier can cause audio oscillations which would interfere. In operating intercoms the signal is sometimes amplified at the source before being transmitted. This enables the signal to be transmitted in the presence of higher interference levels than could otherwise be tolerated. This method cannot be applied in the usual telephone systems because cross-talk limits the transmission level that can be used.

4. SYSTEM INTERFERENCE CROSS-EFFECTS

The table in Figure 3-2 shows various interference cross-effects which may be encountered in the operation of various types of C-E systems. The degree of interference will be dependent on the factors which have been previously discussed in this volume.

0 COULD BE INTERFERED WITH BY SOURCE IN VERTICAL COLUMN

X COULD CAUSE INTERFERENCE TO EQUIPMENT IN VERTICAL COLUMN

NOTE: THE TABULATION IS NOT INTENDED TO COVER ALL POSSIBLE SOURCES OF INTERFERENCE IN THESE EQUIPMENTS.

	STD. AM LF AND HF				KEYED CW LF AND HF				FREQ. SHIFT VHF				TV VHF AND UHF					FM VHF					RADAR				WIRE SYSTEM	
	TRANSMITTER SAME CHANNEL	TRANSMITTER OTHER CHANNEL	HARMONICS	RECEIVER	TRANSMITTER SAME CHANNEL	TRANSMITTER OTHER CHANNEL	HARMONICS	RECEIVER	TRANSMITTER SAME CHANNEL	TRANSMITTER OTHER CHANNEL	HARMONICS	RECEIVER	TRANSMITTER SAME CHANNEL	TRANSMITTER OTHER CHANNEL	HARMONICS	SIDEBANDS	RECEIVER	TRANSMITTER SAME CHANNEL	TRANSMITTER OTHER CHANNEL	HARMONICS	SIDEBANDS	RECEIVER	TRANSMITTER SAME CHANNEL	TRANSMITTER OTHER CHANNEL	SIDEBANDS	RECEIVER	AUDIO TRANSMISSION	SWITCHES AND DIALS
STANDARD AM, RECEIVER	0	0	0			0	0									0					0				0			
LOCAL OSCILLATOR				X				X																				
KEYED CW, RECEIVER		0	0		0	0	0									0					0							
LOCAL OSCILLATOR				X				X																				
FREQUENCY SHIFT, RECEIVER			0				0		0	0	0				0	0				0	0				0			
TELEVISION, RECEIVER			0				0				0		0	0						0				0	0			
LOCAL OSCILLATOR																	X					X						
SWEEP GENERATOR				X													X					X						
COLOR GENERATOR				X													X					X						
SINGLE SIDEBAND, RECEIVER	0	0	0			0	0									0					0				0			
LOCAL OSCILLATOR				X				X																				
TELETYPE, MODULATOR																											X	
SWITCHES				X													X											
FM, RECEIVER															0	0		0	0	0					0			
LOCAL OSCILLATOR																	X					X						
DATA-LINK (UHF)															0					0			0	0				
RADAR, RECEIVER																							0	0				
MODULATOR				X				X									X					X					X	
WIRE SYSTEM, AUDIO TRANSMISSION																											0	

Figure 3-2. Interference Cross Effects

1. EFFECTIVE RADIATED POWER

The criterion for judging the probability of RFI at a distant location caused by a particular transmitter-receiver-antenna complex is the effective radiated power which the complex radiates to the given location. Effective radiated power (ERP) at a given frequency is the product of the net antenna gain times the power, at that frequency, available for radiation. This power may originate from the transmitter fundamental or its harmonics, or other spurious frequencies, from the receiver local oscillator or its harmonics, or from other sources inside the complex which are not normally thought of as generating power which might be radiated. The term "radiation" is sometimes misused when applied to near field phenomena and this aspect is discussed in Section 3 of this chapter. As the term is used in this section, however, it applies only to true radiation to which the principles of antenna theory and propagation characteristics apply.

1.1 ANTENNA PATTERN AND GAIN CONSIDERATIONS

The ERP is a function of the antenna gain which, in turn, is directly related to the three dimensional antenna pattern. By making assumptions about the amplitude and phase distribution across the aperture of an antenna, a theoretical figure for the gain can be obtained without measuring the antenna pattern, see Section 1.1.3. However, in many cases where RFI is a problem, this will not be satisfactory, and it will be necessary to measure the "as installed" antenna pattern and calculate the gain from it.

1.1.1 A GRAPHICAL METHOD OF CALCULATING GAIN FROM THE MEASURED PATTERN

Although considerable information is available concerning the relationship between an antenna's pattern and its gain, a simple method for calculating gain when the pattern data is given is not readily found in the literature. The following is a straightforward method for calculating gain when pattern data is given in terms of conical patterns. (Even if pattern data is not available in this form, it can be synthesized by replotting the available data.) Such patterns consist of a series of plots of field strength, E, versus azimuth angle, ϕ, with ψ, the colatitude angle held constant, see Figure 4-1. As many of these as are necessary to include all the significant lobes are taken for values of ψ from +90° to

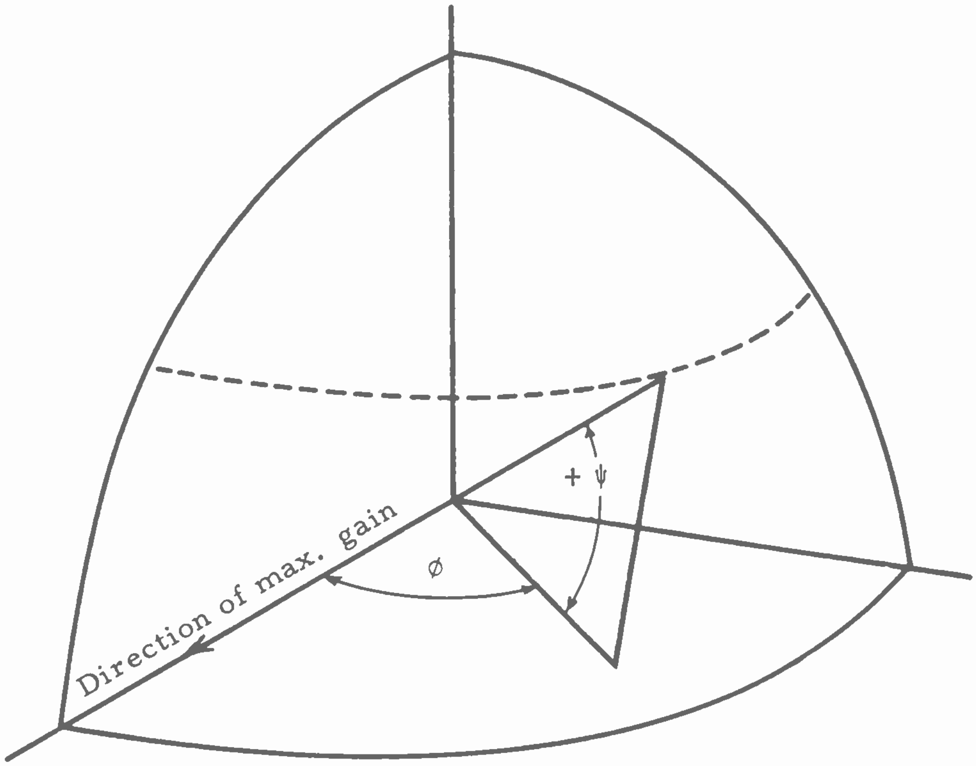


Figure 4-1. Coordinate System Showing Line Used for Measuring Conical Patterns

-90°. (Naturally when the antenna is mounted at or near ground level, it is not possible to take patterns for negative values of ψ. When the installation is mounted on a tower or a high prominence, however, radiation does exist for such values of ψ and therefore these patterns should be recorded. In some cases changing the orientation of the antenna will allow patterns to be taken for values of ψ which include -90°.)

A planimeter is a convenient device for measuring areas in the following procedure but counting squares on graph paper is almost as accurate. Each pattern of E versus ø is plotted in polar coordinates. The following steps are then executed:

1. Find the area, A, of each polar pattern in square cm.

2. Convert each area into "A units" where:

$$\text{A units} = \frac{\text{A in sq. cm.}}{r^2} \tag{4-1}$$

and r = the distance in cm from the origin to the point where $E(\phi, \psi)$ has its maximum value for all patterns

3. Multiply each area in A units by $\cos \psi$ where ψ is the colatitude angle for which the given pattern was taken.

4. Plot the function so obtained versus ψ from $\psi = +90°$ to $-90°$. Use a rectilinear plot with "A units" along the vertical axis and degrees along the horizontal axis.

5. Find the area, B, under this curve in square cm.

6. Choose a convenient length on the x-axis of this graph for a "B unit." (The example to follow will indicate the nature of a convenient length.)

7. The power gain is then given by

$$G(\text{power}) = \frac{2P^2M}{B_{cm^2}N} = \text{power gain with respect to an isotropic radiator at } E(\phi, \psi)_{max} \qquad (4\text{-}2)$$

where from the graph of step 4:

N = number of A units in one B unit
M = number of B units in 180 degrees
P = length of a B unit in cm
B_{cm^2} = value found in step 5

1.1.2 A SAMPLE CALCULATION

For an example, consider a pattern given by the equation:

$$E = \cos \psi \cos \phi \qquad E = 0 \text{ when } 90° < \phi < 270° \qquad (4\text{-}3)$$

This can be solved mathematically by double integration and the power gain is found to be G = 6. The above method is demonstrated by using this formula to calculate patterns which normally would be the measured pattern from field tests. The patterns are then plotted and the areas of each are measured. The table in Figure 4-2 shows the results of these measurements and calculations.

The curve of A $\cos \psi$ versus ψ is plotted in Figure 4-3.

ψ	A_{cm}^2* (Found from Plot of Calculated Pattern)	A Units $\left(\frac{A_{cm}^2}{r^2}\right)$	Cos ψ	A Cos ψ (A Units)
-90°	0.	0.	0.	0.
-80°	2.8	.0174	.174	.003
-70°	12.8	.0795	.342	.0274
-60°	29.0	.180	.500	.0895
-50°	48.6	.302	.643	.193
-40°	72.8	.451	.766	.344
-30°	92.8	.575	.866	.498
-20°	110.0	.681	.940	.637
-10°	122.0	.756	.985	.745
0°	124.6	.774	1.00	.774
10°	122.0	.756	.985	.745
20°	110.0	.681	.940	.637
30°	92.8	.575	.866	.498
40°	72.8	.451	.766	.344
50°	48.6	.302	.643	.193
60°	29.0	.180	.500	.0895
70°	12.8	.0795	.342	.0274
80°	2.8	.0174	.174	.003
90°	0.	0.	0.	0.

*For r = 12.7 cm.

Figure 4-2. Table for Sample Gain Calculation

For this curve B_{cm}^2 = 59.1 sq. cm. The polar graph paper had a full scale (E_{max}) radius of 5" = r = 12.7 cm; a B unit was chosen as 1 cm (because 1 cm = 10°) so P = 1. Horizontally, 1 cm = 10° so M = 18.

Vertically, the scale was chosen such that B unit = 0.1 A units, so N = 0.1. Therefore:

$$G = \frac{2 \times 1 \times 18}{59.1 \times 0.1} = 6.08 \qquad (4\text{-}4)$$

This checks within the expected limits for slide rule accuracy and graphical methods.

1.1.3 AN APPROXIMATE METHOD FOR CALCULATING THE GAIN OF AN APERTURE TYPE ANTENNA

Because many RFI problems involve aperture type antennas, such as paraboloids, horns, and lenses, a quick approximate method of calculating the on-axis, design frequency gain of such antennas is useful. The only data necessary is the size and shape of the aperture.[1,11*] The expression given below is a good approximation for antennas with gains above 15 db and is reasonably good below this figure.

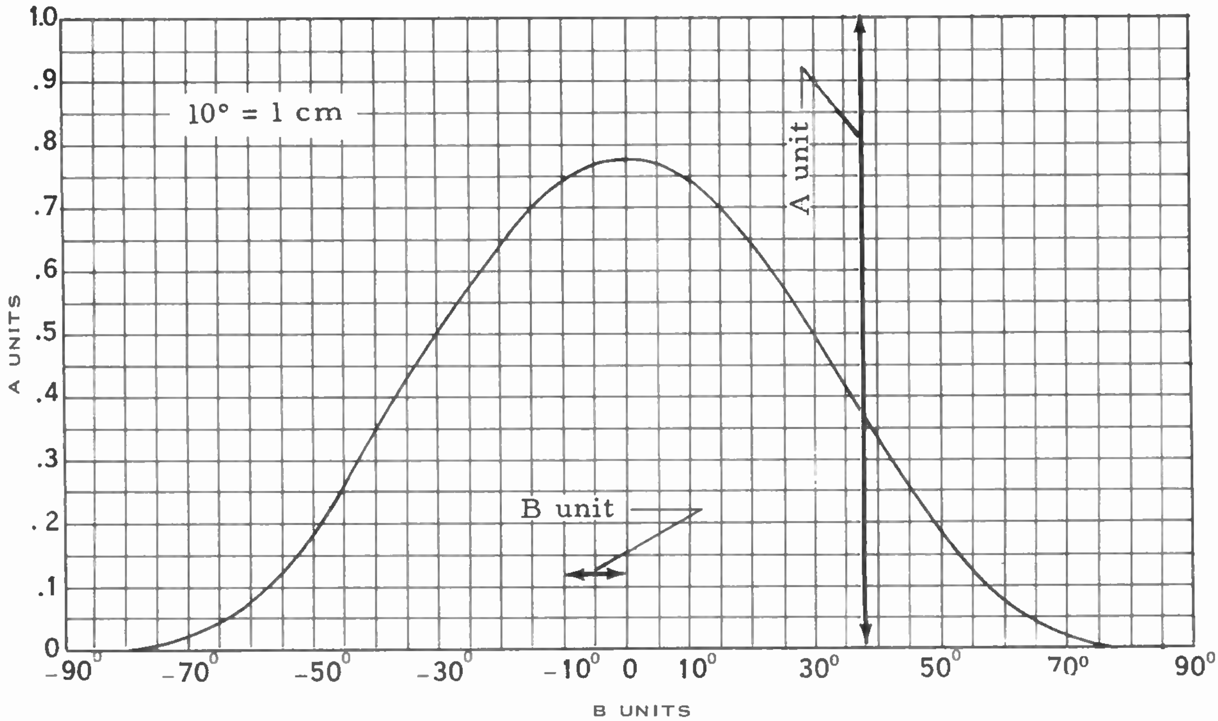


Figure 4-3. A Cos ψ vs ψ for Finding Gain in the Sample Calculation

*See end of chapter for references.

For high gain antennas, the power gain of an antenna with respect to an isotropic source is given very closely by

$$G = \frac{4\pi A_P F_H F_V}{\lambda^2} K \tag{4-5}$$

where: G = power gain (with respect to an isotropic source)
A_P = aperture physical area
λ = wavelength (same linear units as P)
F_H = correction factor depending on H direction aperture illumination
F_V = correction factor depending on V direction aperture illumination
K = efficiency, normally taken as 0.55

The correction factors, F_H and F_V, depending on illumination are shown below.

Type of Illumination	Correction Factor $F_{H,V}$
Uniform	1.000
Cosine	0.810
Cosine Squared	0.667
Cosine Cubed	0.575
Cosine Fourth Power	0.515

1.1.4 GAIN CONCEPTS AS APPLIED TO RFI

The above method gives the gain, neglecting any antenna or transmission line losses, with respect to an isotropic radiator. Since gains are usually given with respect to a $\lambda/2$ dipole which has a gain with respect to isotropic of 1.64, the gain obtained from the above calculation must be divided by this factor if a dipole is used as a reference. Gain in any direction other than the maximum $E(\phi, \psi)$ is found from the antenna patterns by the proportional relationship:

$$\frac{E^2 \max(\phi, \psi)}{E^2(\phi_1, \psi_1)} = \frac{G \max}{G_{\phi_1, \psi_1}} \tag{4-6}$$

Since RFI studies concern themselves with the radiation of spurious as well as fundamental frequencies, the gain of the antenna in the direction which may cause RFI must be known at each one of these spurious frequencies. The best way to determine this is by a direct gain measurement. If several directions are involved, it may be easier to record the patterns at the required frequencies and calculate the desired gains from them.

In case a direct measurement of either pattern or gain cannot be made, it may be necessary to attempt to obtain a rough estimate of the off-frequency pattern and assume a gain figure from this. Since most RFI measurements are concerned with orders of magnitude, a few db of error in this estimation will not be serious. If the spurious frequency is only 5 to 10 per cent off of the fundamental frequency, then it is fairly safe to assume the gain is the same as the gain at the fundamental. (Of course for narrowband antennas, the pattern gain will probably be the same but the effective gain may be considerably less because of impedance mismatch loss.)

If the spurious frequency is a harmonic or other frequency far removed from the design frequency, then estimating the antenna pattern may be quite difficult. For this purpose it is best to classify the antenna under consideration into one of four classes: (1) Resonant Antennas, such as dipole arrays or billboards; (2) Aperture Antennas, such as parabolic reflectors, horns, or lenses; (3) Traveling Wave Antennas, such as end-fire arrays, helices or dielectric rods; and (4) Frequency Independent Antennas, such as log periodic arrays or log spirals.

In some cases the off-frequency patterns may be assumed to a reasonable degree by reference to a good antenna book which will give the patterns of an antenna over a wide band of frequencies (See Figure 4-4[2, 3, 4]).

Dipole arrays show poor effective gain characteristics outside the design band because of the high mismatch loss and the close dependence of the pattern on the physical and electrical parameters[5] of the antenna. The horn and parabola are not as dependent on the physical parameters; instead, their radiation characteristics are dependent on the higher-order mode combinations that contribute to the pattern gain. The very erratic and unpredictable generation and combination of higher-order modes makes the prediction of gain and interference potential difficult for a given spurious frequency. Tests conducted in the past have indicated that the gain of a horn antenna remains essentially constant as the frequency is increased up to about the third harmonic.[6] Above the third harmonic there is a sharp drop in gain. The patterns of frequency independent antennas are periodic with frequency over bandwidths as high as 10 to 1 and are, therefore, relatively easy to predict. In general, the off-frequency characteristics of most antennas are not well known but the

A. Source: Antenna Engineering Handbook[2]

Page	Figure	Antenna Type
3-8	3-8	Center Driven Dipole
3-9	3-9	Center Driven Cylindrical Dipole
3-16	3-24	Asymmetrical Dipole
3-17	3-25	Asymmetrical Dipole
3-19	3-28	Sleeve Antenna
3-20	3-29	Sleeve Antenna
5-4	5-2	Two Antenna Array (point source)
5-12	5-9	Sixteen Element Array
5-16	5-12	Arrays of End Fire Couplets
7-4	7-5	Helical Antenna
11-5	11-11	Corner Reflector
11-6, 11-9	11-12, 11-21	Corner Reflector
18-19	18-19	Wire Trapezoidal Antenna
21-29	21-19	Broadband Curtain Array
22-10	22-9	Directional Antennas
24-30	24-22	UHF Corner Reflector
27-35	27-43	Annular Slot and $\lambda/4$ Stub
27-44	27-57	Aircraft Homing Array

B. Source: VHF Techniques[3]

Page	Figure	Antenna Type
255	10-31	Double Parabolic Spinner
250	10-26	Corner Spinner
148-150	6-13, 6-19	Horns
359-363	10-10, 10-13	Sectoral Horns

Figure 4-4. Table of References for Antenna Pattern Variations with Frequency and/or Length of Element

C. Source: Proc. IRE, Vol. 36, Pages 1101-1105[4]

Page	Figure	Antenna Type
1102	2	Large Collection of Horn Patterns
1103	4, 5	" "

Figure 4-4. Table of References for Antenna Pattern Variations with Frequency and/or Length of Element (Continued)

current spectrum signature contracts sponsored by the Department of Defense are expected to provide realistic methods for computing antenna gain at harmonic and other spurious frequencies.

1.1.5 ANTENNA SIDELOBE CONSIDERATIONS[2]

In many cases where RFI is a problem or where prediction of the possibility of RFI is desired, it is necessary to know the sidelobe characteristics of the antennas under consideration. Since the potentially interfering antennas may not be pointing directly towards each other, in most cases, the interfering signal may be transmitted and/or received by the sidelobes of one or both of the antennas. For very accurate studies, the sidelobe structure of all of the antennas under consideration must be measured at all of the frequencies where the calculations will be made. (The continuing spectrum signature programs of the DOD will add to the knowledge of specific types of antennas.)

When such information is not readily available, as will usually be the case, a rough estimate of the RFI situation can be made by using general information concerning antenna sidelobe magnitudes. The following discussion presents data on sidelobe levels of the more common antenna types at their design frequency.

The most common antenna that will be encountered is a parabolodial reflector illuminated by a point source feed. These antenna systems generally have sidelobes which range from 18 to 25 db down from the main beam. With the present state of the art, it is unlikely that sidelobe levels much lower than 35 db down will be encountered for angles in front of the reflector. However, the sidelobe levels may be down considerably greater than 40 db off the back of the reflector. When a fan beam is produced by employing a line feed to excite a paraboloid reflector, sidelobe levels may be about 27 db.

An array of discrete elements such as a mattress array of dipoles or a multi-element slot array may, if properly designed, have sidelobes as low as 30 db down from the main beam, even at angles close to the main beam. In actual cases, however, it is safer to assume a sidelobe level about 20 db down.

For rhombic antennas, the sidelobes in the forward looking direction will rarely be lower than 13 db down and sidelobes lower than 20 to 25 db will rarely be encountered in the back 180° sector. A very prominent lobe will usually occur exactly 180° away from the main beam. Its level may be only 10 db down from the main beam.

Arrays of helices are, in general, not very directive as compared to the previously discussed antennas, and their sidelobes are usually 10 to 15 db down from the main beam. It should be noted that since these antennas normally are circularly polarized, they will present an RFI problem greater than might normally be expected. This effect is shown in Figure 4-18 in Section 1.4.

Long linear arrays of waveguide fed slots when designed for optimum sidelobe suppression will have sidelobes of about 30 db down from the main beam.

Since a horn antenna can have many different shapes and sizes, no general rule can be given. If the physical parameters are known for a given horn, a rough estimate of the sidelobes can be gained from Source C^4 in Figure 4-4 which deals with the patterns of horn antennas. In general, a horn will have about the same sidelobe levels as a paraboloid of the same aperture with the likelihood that large aperture horns will have somewhat lower sidelobes than a paraboloid of the same aperture.

A hog horn antenna has the lowest sidelobes of any of the common type antennas. The hog horn is characterized by the use of a section of a paraboloid bounded by two parallel conducting plates. Stray radiation is minimized by extending the sides of the feed until they touch the paraboloid surface on three sides. On the fourth side, an opening is left for the reflected rays. Although this type of structure is useful in minimizing wide-angle sidelobes, it has no significant effect on the close-in sidelobes caused by the normal diffration effects. Sidelobe reduction is in the order of 13 db for the first minor lobe.

A corner reflector antenna falls in the same basic category as the helix, and the sidelobes will be of the same order of magnitude unless the corner configuration is constructed of extremely large plates of solid metal. In this case, the sidelobes may be substantially lower than the helix and may approach 30 db down.

In the periscope antenna system, a reflector is located in the Fresnel or near-Fraunhofer region of a conventional parabola antenna whose main beam is vertical. The reflector is located at a sufficient height to enable the final direction of radiation to be located above obstructions without using long lengths of waveguide. The radiation patterns of periscope antenna systems depends, in a rather complicated way, on the many system variables such as supporting tower reflections. However, for most systems encountered in practice, the first and second sidelobes of the periscope antenna with an elliptical reflector are about 18 and 25 db down, respectively; and, corresponding numbers for the rectangular reflector are about 14 and 18 db down.

The lens type of antenna provides high gain and a narrow main beam. The sidelobe level will be low in these antennas if the surfaces of the lenses are matched. As an example, the sidelobe level is approximately 20 db down for unmatched lenses, 34 db down for disk-matched, and 30 db down for corrugation matched lenses. The high sidelobe level of the unmatched lens is due to forward scattering of the reflected components by the walls of the feed horns. The sidelobe energy will be reduced considerably if absorbing material is provided along the narrow walls of the horn or if those walls are eliminated; however, then the reflected energy may be scattered in the rearward direction.

In a Cassegrain antenna, the main dish is a paraboloid and the auxiliary reflector is a hyperboloid. The first sidelobe in this type of antenna is about 19 db down and the second is about 27 db down.

Unidirectional log-periodic antennas are used for applications where the radiation pattern and input impedance of the primary radiator must be independent of frequency. The sidelobe level varies between 7 and 24 db for common types. The largest lobe nearly always occurs 180° away from the direction of the main beam.

1.2 FACTORS AFFECTING ERP IN THE PRACTICAL CASE

A number of variables determine the net antenna gain, such as transmission line mismatch losses, transmission line filtering if harmonics are being considered, circulating currents and insulator losses on the antenna, and radiation pattern distortions due to nearby structures. It must always be kept in mind that the installed antenna may have a far-field pattern substantially different from the calculated pattern or the pattern measured on a model range even if the effect of the earth is accounted for. Because of this, if truly valid RFI predictions are to be made, pattern measurements at the site are required.

The E. R. P. in the far field of an antenna is also a function of the transmitter power spectrum and the characteristics of the propagation path, including terrain effects, reflecting and absorbing objects and the time varying properties of the troposphere and ionosphere. Since E. R. P. is a function of frequency, its interference effects must be investigated for all likely radiated components.

Under conditions approaching free space, the far-field pattern of an antenna is, in general, invariant with time (there are exceptions but these are special cases), and, therefore, the gain and E. R. P. should be invariant. However, an antenna is rarely used under such conditions so that usually site effects resulting from the antenna's location cause a time variation of the far-field pattern. These effects are caused by trees, power lines, structures and other factors which influence the electromagnetic environment of the antenna (for example: the presence of moisture on these objects and the presence of leaves on trees causes a very substantial change in the effective antenna pattern in certain frequency ranges).

When one or both of the antennas are azimuth scanning types, an additional time variable is introduced. If only one of the antennas is scanning, the problem is rather simply handled by assuming worst case conditions such that calculations are made for the case with the scanning stopped and the antenna oriented for the maximum RFI. If both antennas are scanning, the problem becomes very complicated. Attempts to treat it by statistical analysis have not yielded realistic answers. If the rotations are correctly synchronized, it is obvious that little or no RFI will occur in most cases; however, if the rotations are nonsynchronized but near the same rate, there will be periods when the interference is very likely and periods when the interference is very unlikely. For this condition, probably a worst case analysis is the safest procedure; that is, both antennas oriented for maximum RFI. If the antenna scanning rates are significantly different from each other, then probably a statistical analysis based on the probable percentage of time that the antennas are oriented for maximum RFI is an acceptable method.

1.3 PROPAGATION THEORY

When the E. R. P. from an antenna is known, it is necessary to account for the propagation of this power from the source to the site of potential RFI. Since the literature on propagation theory is not too well organized, a summary of the pertinent factors will be presented so that reasonably accurate calculations can be made if necessary. (The following discussion pertains only to frequencies above 30 mc and therefore the effects of the ionosphere are neglected.)

1.3.1 FACTORS DETERMINING THE RECEIVED RFI

A considerable number of variables enter into the complete solution of a propagation problem. In order to preserve the simplicity of the presentation, these factors will be limited to those which contribute substantially toward a good approximation of the solution. It should be noted that regardless of the rigor of the most detailed propagation study, there are many unknowns which cannot be accounted for and, therefore, only the order of magnitude of the expected result is usually obtained.

The factors which must be considered when determining the characteristics of a particular propagation path are as follows:

a. Transmitter power (±dbm).
b. Line loss from transmitter to antenna (-db). See 1.6
c. Mismatch loss of transmitting antenna (-db). See 1.6
d. Gain of transmitting antenna referred to a $\lambda/2$ dipole (+db). See 1.1
e. Transmission loss between transmitting and receiving antennas when both are dipoles (-db). See this section
f. Gain of receiving antenna referred to a dipole (+db). See 1.1
g. Mismatch loss of receiving antenna (-db). See 1.6
h. Line loss from antenna to receiver (-db). See 1.6
i. Mismatch loss at receiver (-db). See 1.6
j. Loss due improper antenna polarization (-db). See 1.4

The bulk of this discussion will be concerned with calculations relating to item e, transmission loss between transmitting and receiving antennas when both are dipoles. The other factors are normally given as part of the design parameters of the system. Any that are not, can usually be assumed to a reasonable degree of accuracy.

The calculation of item e can be classified under two broad headings: (1) conditions which can be assumed to approximate free space transmission; and (2) conditions which require consideration of the effect of a reflecting surface, either sea water or some type of soil. The method used considers the entire system from transmitter output to receiver terminals as a "black-box" four-terminal network. The losses involved are tabulated in decibels and their summation gives a single figure which equals the insertion loss for the black-box network. Naturally, this loss is a function of the geometric relations between the transmitting and receiving antennas and the reflecting surface, if any is involved.

1.3.2 FREE SPACE PROPAGATION

In most propagation problems above 30 mc, a first approximation to a solution is the loss that would result if the antennas were located in free space with no reflecting surfaces present. This calculation is usually carried out by calculating the loss between two dipoles. The net gains of the actual antennas used are referred to the dipoles and the difference is accounted for in the final calculations of the transmission loss. For matched dipoles in free space oriented for maximum signal and of the same polarization, the ratio of the antenna system input power to the output power is given by

$$\frac{P_{out}}{P_{in}} = \left(\frac{.41\,\lambda}{d}\right)^2 \qquad (4\text{-}7)$$

where the wavelength "λ" and the separation distance "d" are both measured in the same units.

The ratio of P_{out} to P_{in} is called the transmission loss between dipoles in free space and can be found in decibels using the nomograph in Figure 4-5. The transmission loss as a function of distance becomes a straight line when plotted on semi-log paper putting distance on the log scale and db loss on the linear scale. Plotting this line is always the <u>first</u> step in any propagation calculation where loss must be found as a function of distance, because it is convenient to make the remaining calculations by referring them to the free space loss. As mentioned above, in some cases the free space loss is sufficient to provide a reasonable answer.

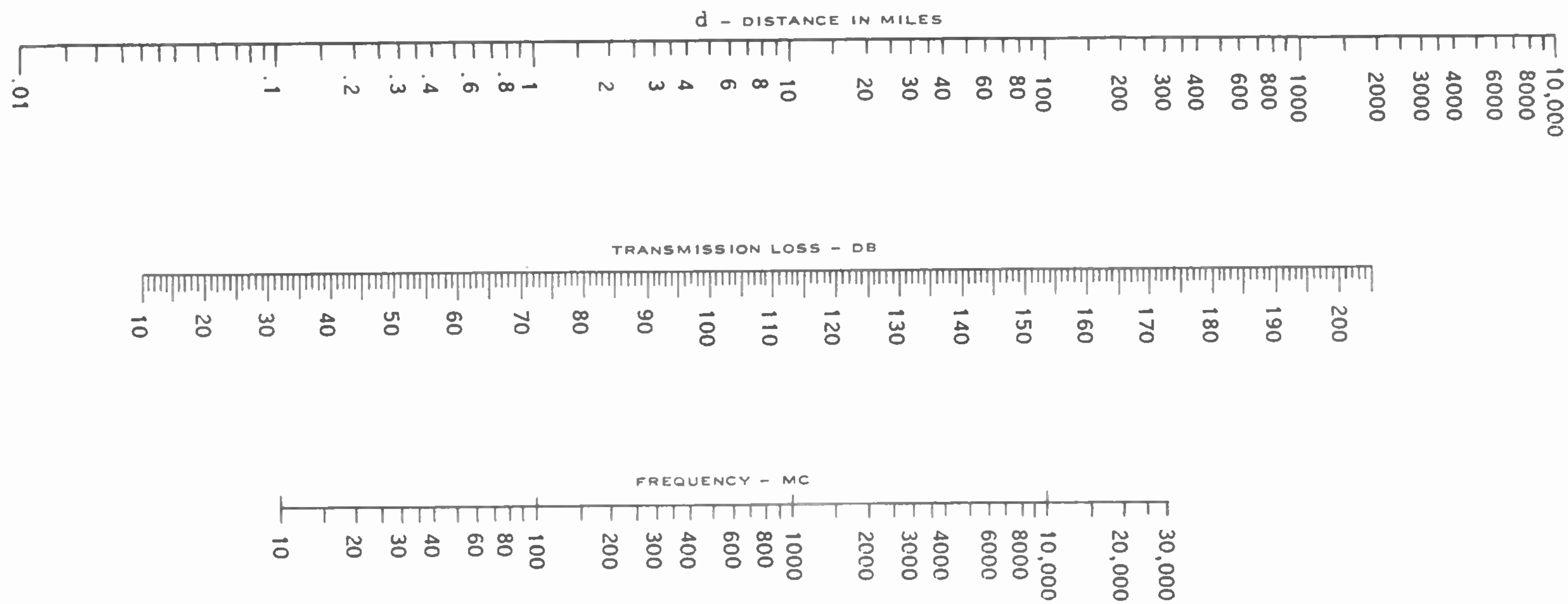


Figure 4-5. Free Space Transmission Loss Between λ/2 Dipoles

1.3.3 METHODS OF ACCOUNTING FOR THE PRESENCE OF THE EARTH[7]

In the event that the boundary conditions indicate the need for more accuracy than is given by free space calculations, it is necessary to account for the presence of the earth. This can be relatively simple or quite complicated, depending on the heights and distances involved. The effect of the earth divides the areas of space into two major sections, one above the radio horizon and one below it. The first is the optical or interference region commonly called the line-of-sight region. The second is the diffraction region in which the earth shadows the receiving antenna and no energy is received by direct propagation. The radio horizon is defined as the point at which a radio wave traveling directly from the transmitter to the receiver first touches the surface of the earth. The distance between antennas, when this is the case, is determined by the expression

$$d_L = \sqrt{2h_1} + \sqrt{2h_2} \quad \text{(the bending of the ray due to atmospheric refraction is accounted for)} \tag{4-8}$$

where: h_1 = transmitting antenna height in feet
h_2 = receiving antenna height in feet
and d_L = radio line of sight distance between antennas in miles (radio horizon).

A nomograph for this calculation is shown in Figure 4-6. In all the formulas that follow, all heights are to be expressed in feet, all distances in miles, and all angles in degrees.

There are three waves that can be considered as radiating from an antenna. The first is the free space wave which travels directly from the transmitting antenna to the receiving antenna and has been discussed above. The second is the reflected wave which travels from the transmitter to the receiver by reflection from the earth. The third is the ground wave which follows the surface of the earth. This last wave contributes very little to the total field in the line-of-sight region but beyond this region it is almost solely responsible for the radiation received.

Several methods of calculation apply as the separation between the antennas is increased and the waves radiated by the transmitting antenna assume varying degrees of importance. It is convenient to denote the region in which a given method of calculation applies by a distinctive name as shown in Figure 4-7. The first region is the <u>plane-earth optical region.</u> In this area, the earth may be treated as a plane and the interference effects created by the direct and reflected waves

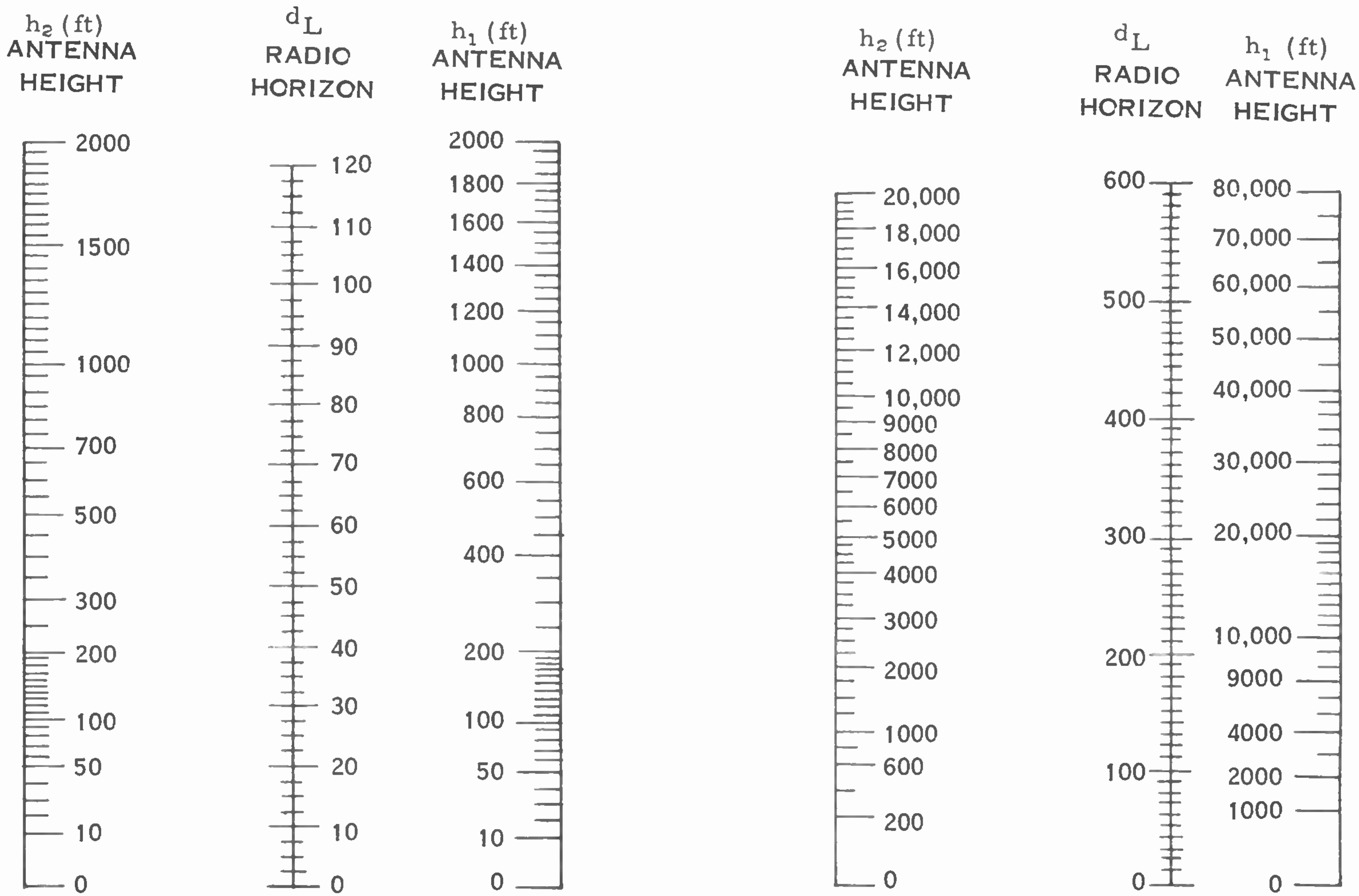


Figure 4-6. Nomographs Giving Radio Horizon Distance

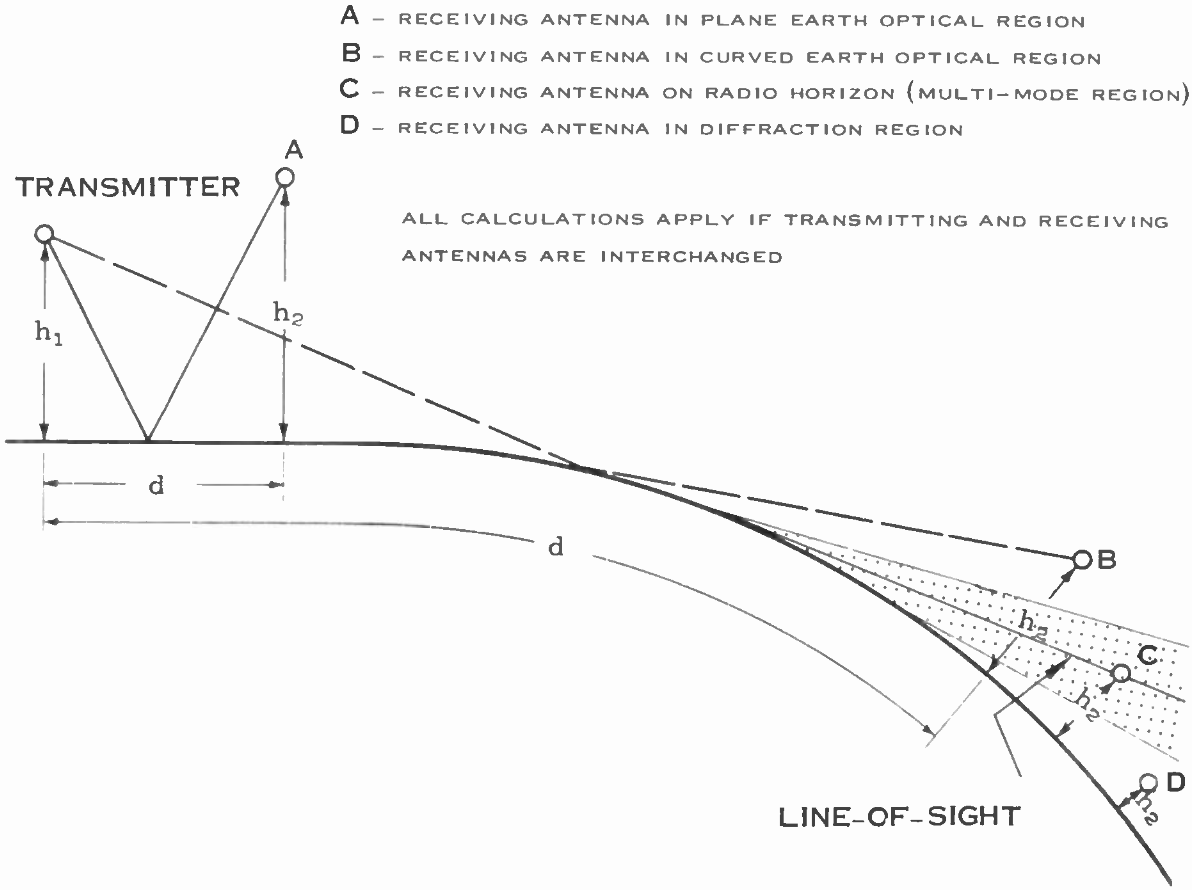


Figure 4-7. Diagram of Propagation Regions

are the sole matter of concern. The receiving antenna is located in this region when the reflected ray makes an angle ψ with the surface of the earth such that:

$$\psi > .113 \left[\frac{h_1 h_2}{h_1 + h_2}\right]^{1/2} = \psi_1 \text{ (degrees)} \tag{4-9}$$

or,

$$d \leq \sqrt{(92\,\psi_1)^2 + 2h_1} + \sqrt{(92\,\psi_1)^2 + 2h_2} - 184\,\psi_1 = d_1 \text{ (miles)} \tag{4-10}$$

These formulas assume that the divergence factor, D, discussed below is such that D >.99.

The second region is the <u>curved earth optical region.</u> For angles of ψ less than the above value but for d still in the line-of-sight region, the curvature of the earth must be accounted for. This is accomplished by using a divergence factor D to decrease the reflected wave to allow for the additional spreading of energy caused by reflection from a spherical rather than a plane earth. This decrease of the reflected wave will increase the total field above that expected and will therefore increase the possibility of RFI. In calculating divergence, the value of the earth's radius "r" is increased to 4/3 r to account for atmospheric refraction which causes the waves to follow a curved path in space. When a radius of 4/3 r is used, a straight line path can be assumed. Calculations for this region are valid as long as:

$$\psi_1 > \psi > \psi_c = \frac{1.02°}{f_{mc}} \tag{4-11}$$

or

$$d_1 < d < d_c = \sqrt{(92\ \psi_c)^2 + 2h_1} + \sqrt{(92\ \psi_c)^2 + 2h_2} - 184\ \psi_c \tag{4-12}$$

f_{mc} = frequency in mc, d, ψ and h are as defined above.

The third region is the multi-mode region around the radio horizon. It is not possible to calculate the field in this region by any precise method because many modes of propagation contribute to the field making the equations extremely involved. If it is desired to know the field in this region, calculations are performed for the two adjacent regions and a smooth extrapolation is made between them.

The fourth region is the diffraction region. This region is located in the area well below the radio horizon such that the separation distance, d, is defined by

$$d > \sqrt{2h_1} + \sqrt{2h_2} + \frac{173}{f_{mc}^{1/3}} \quad \text{(miles)} \tag{4-13}$$

The calculations in this region are considerably more complicated than those in the optical region. For most practical applications, this region may be considered to extend indefinitely as d is increased. For very high-powered transmitters, however, it is possible to achieve somewhat erratic communication at much greater ranges than predicted by diffraction theory alone. These effects should be considered separately as should the consideration of effects which non-standard meteorological conditions have on propagation in this region.

1.3.4 CALCULATIONS IN THE PLANE EARTH OPTICAL REGION

In the plane earth optical region the field at the receiving antenna is the vector sum of the direct and reflected waves. Since the reflected wave must travel the distance from the transmitter to the reflecting surface and then to the receiver, while the direct wave travels the lesser distance from transmitter to receiver, the reflected wave component lags the direct wave component by a phase angle which is proportional to the path length difference. The reflected wave may also undergo a change in magnitude and a shift in phase due to its contact with the reflecting surface. The E field (proportional to voltage) expression which accounts for all these effects is

$$E = \sqrt{1 + \rho^2 + 2\rho\cos(\theta - \phi)} \qquad (4\text{-}14)$$

where: ρ = magnitude of surface reflection coefficient, for given ψ.

ϕ = phase of surface reflection coefficient, degrees, for given ψ.

θ = phase lag due to path length difference, degrees

or:

$$\theta = \frac{1.385 \times 10^{-4}\, h_1 \cdot h_2 \cdot f_{mc}}{d} \text{ (degrees)} \qquad (4\text{-}15)$$

ρ and ϕ are given versus ψ in the graphs in Figure 4-8a through 4-8e where ψ is given by:

$$\psi = \tan^{-1}\left[\frac{h_1 + h_2}{5280d}\right] \text{ (degrees)} \qquad (4\text{-}16)$$

The expression for E has a maximum value of 2 and a minimum of zero. The actual value of the field in space is found by expressing this value in db and adding it to the free space transmission loss for the same point in space.

It will be noted that in addition to the geometric relations of the two antennas, it is also necessary to know the magnitude and phase of the reflection coefficient of the earth at the point of reflection. The graphs shown in Figure 4-8a through 4-8e[7a] give these values as functions of the frequency, polarization, and angle of incidence of the reflected wave. It will be noted that the magnitude and phase of the reflection coefficient for horizontal polarization is substantially constant while that for vertical polarization undergoes large variations and has a minimum value at some critical angle. This angle is called the pseudo-Brewster angle and because of its effect, a vertically polarized field is stronger if the geometry of the path is such that the angle of incidence of the potentially interfering wave is near the Brewster angle at the distances where RFI must be considered. The reflected wave is substantially reduced and the effects of cancellation between direct and reflected waves are greatly diminished.

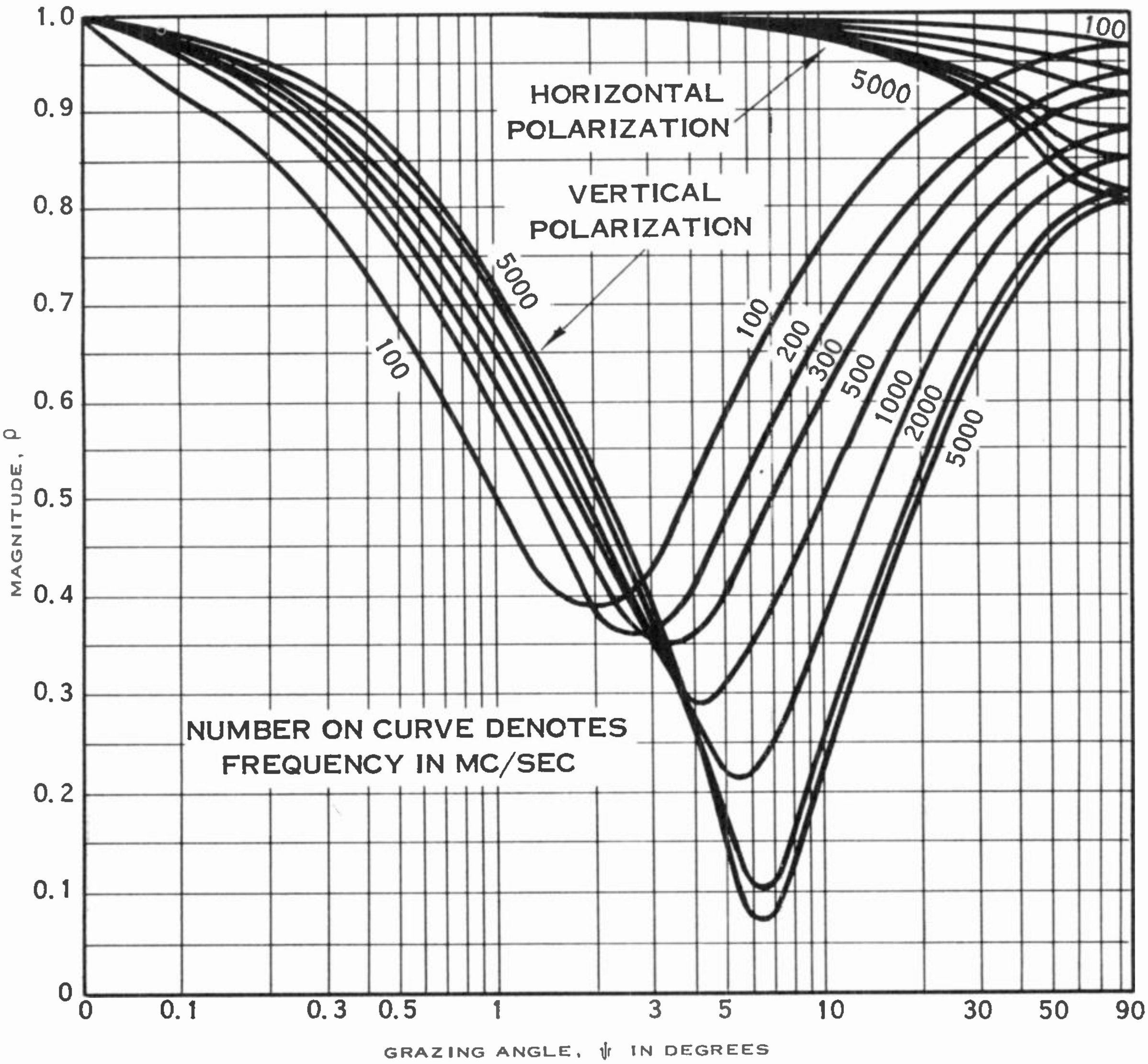


Figure 4-8a. Magnitude ρ of the Reflection Coefficient vs. Grazing Angle ψ for Smooth Sea Water

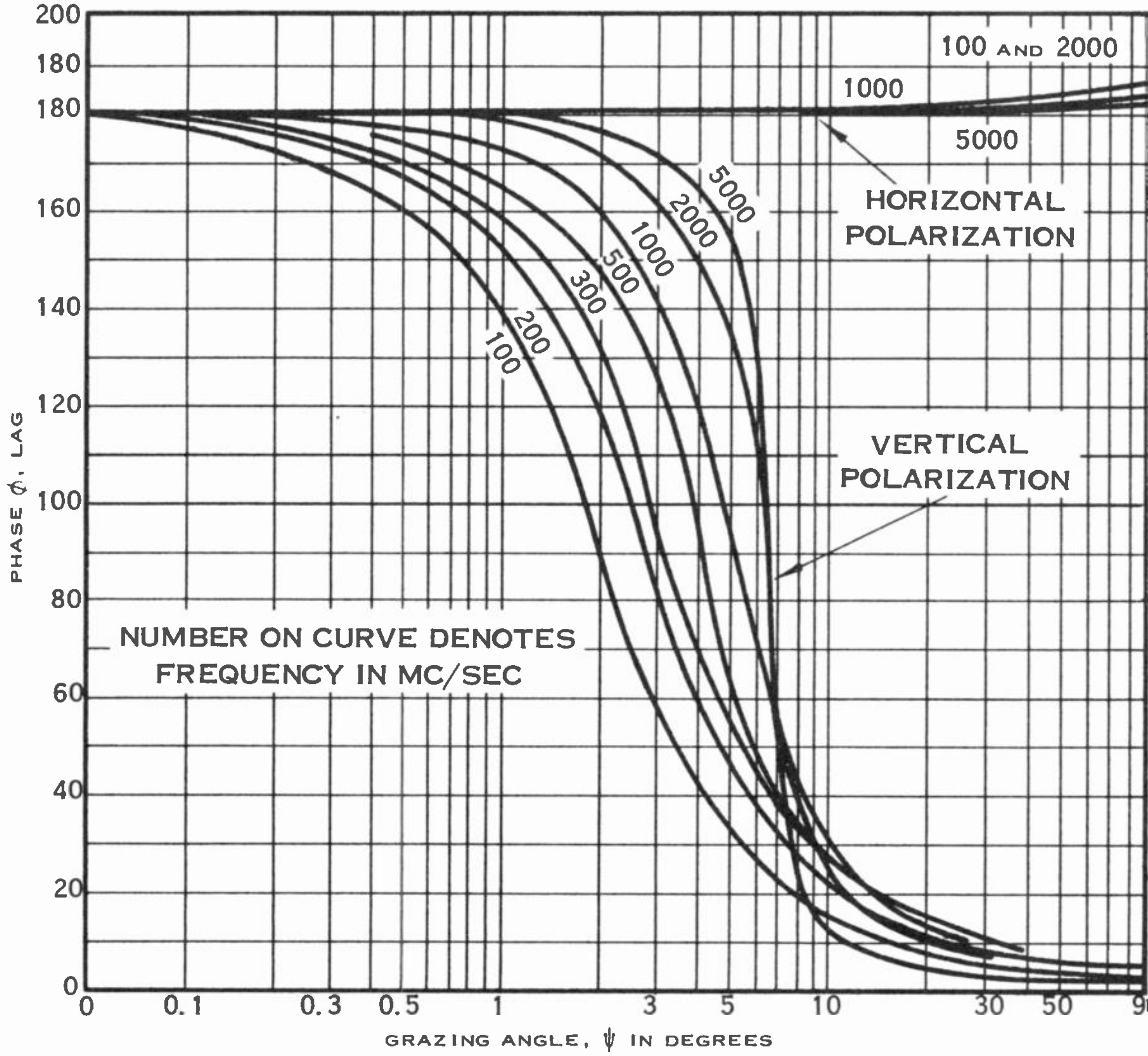


Figure 4-8b. Phase ϕ of the Reflection Coefficient vs. Grazing Angle ψ for Smooth Sea Water

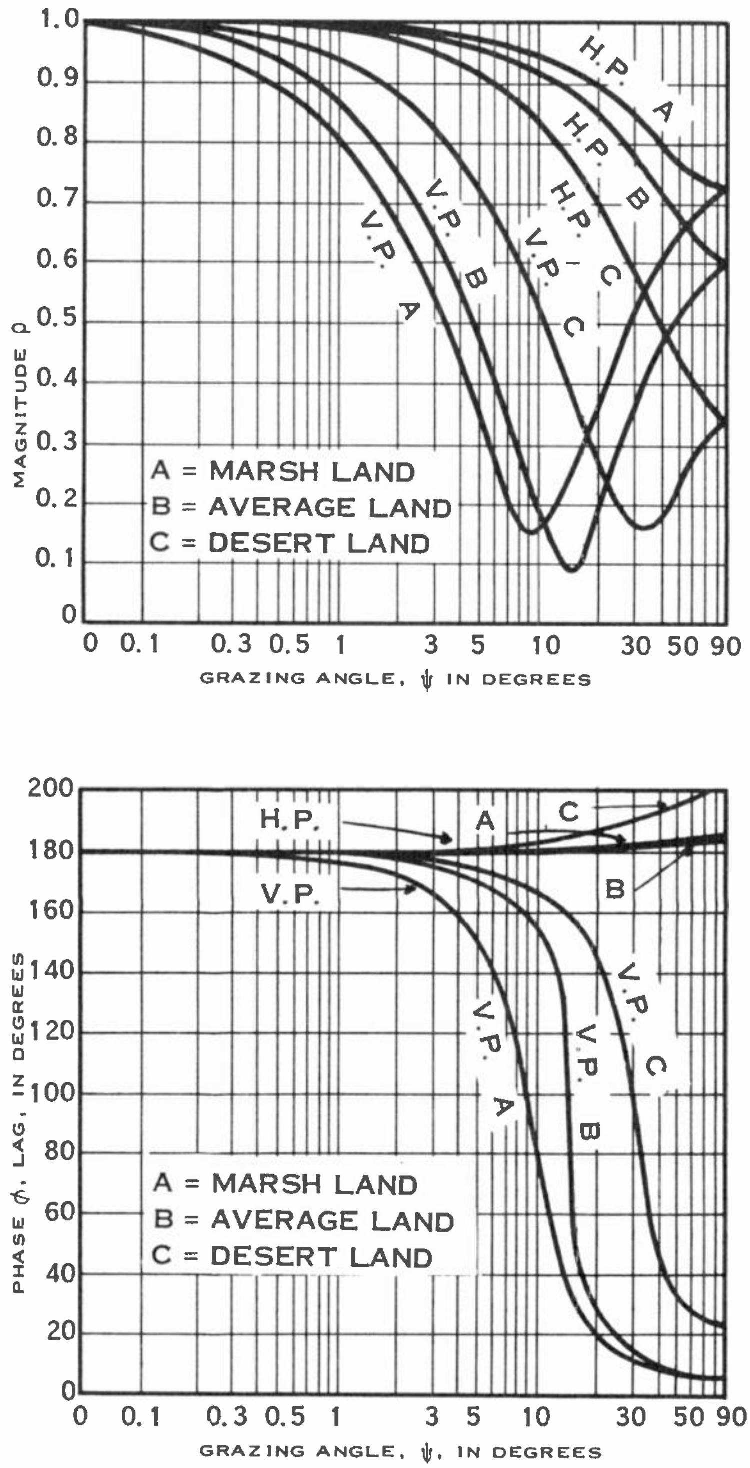
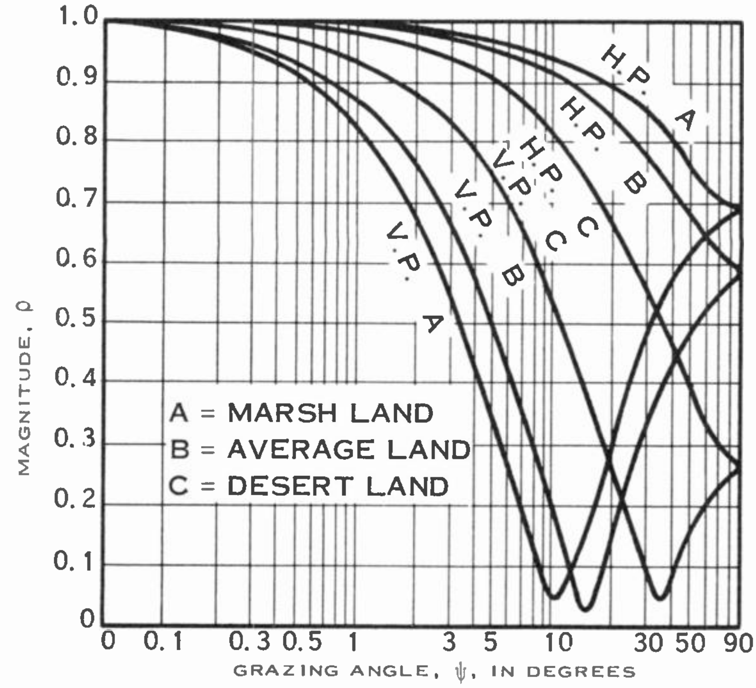


Figure 4-8c. Magnitude ρ and Phase φ of Reflection Coefficient vs. Grazing Angle ψ for Smooth Land Frequency: 100 mc/sec.



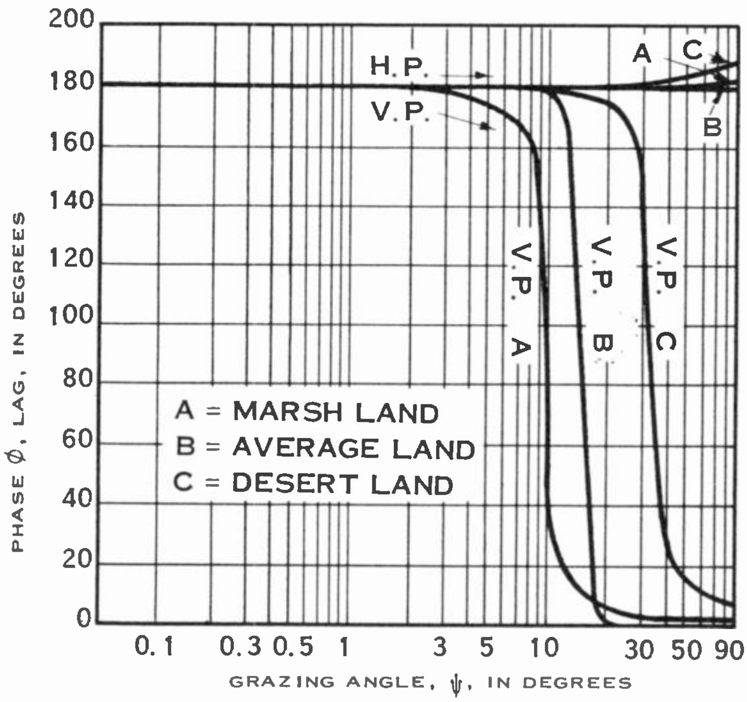
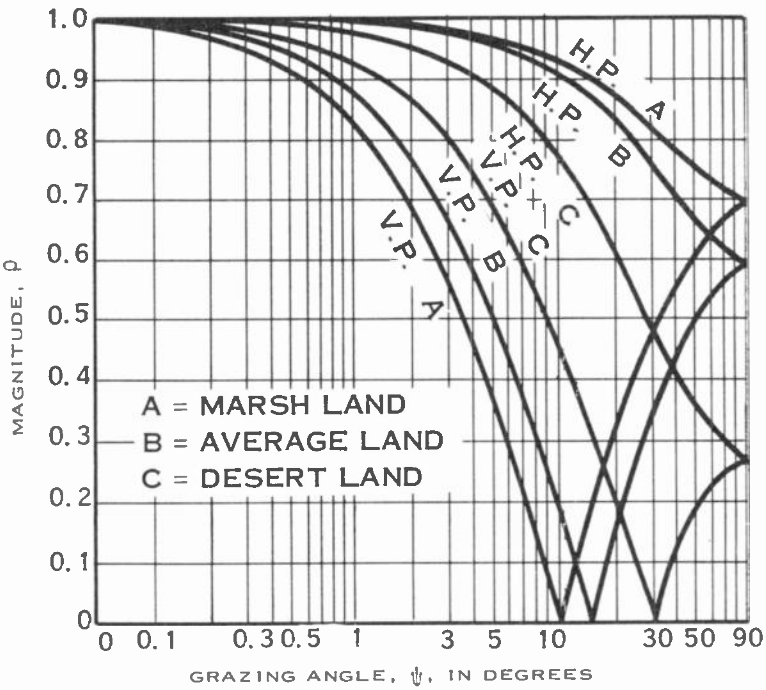


Figure 4-8d. Magnitude ρ and Phase ϕ of Reflection Coefficient vs. Grazing Angle ψ for Smooth Land Frequency: 400 mc/sec.



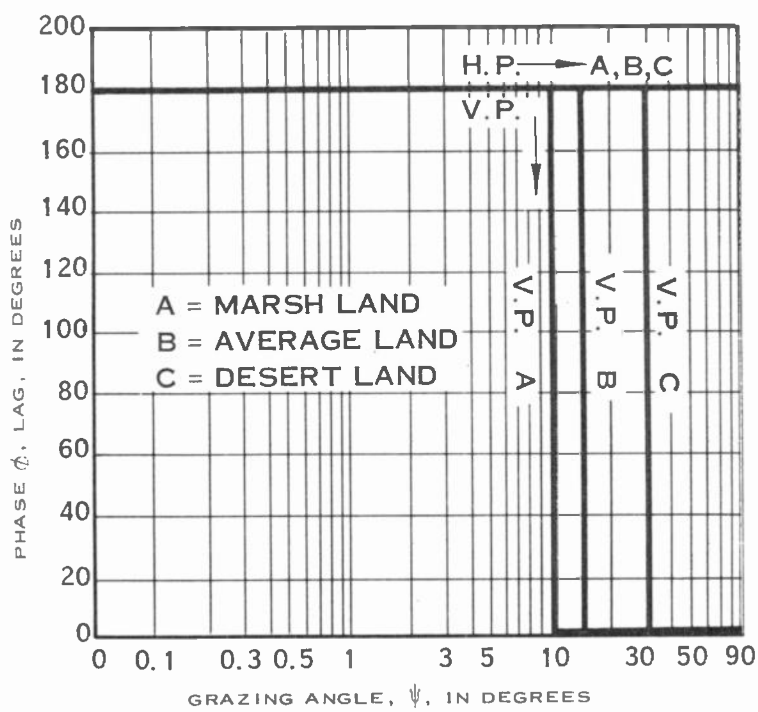


Figure 4-8e. Magnitude ρ and Phase Angle ϕ Reflection Coefficient vs. Grazing Angle ψ for Smooth Land Frequency: >5000 mc/sec.

1.3.5 CALCULATIONS IN THE CURVED EARTH OPTICAL REGION[7]

When the angle, ψ, and the distance, d, are such as to satisfy equations (4-11) and (4-12), the formula for the E field given in (4-14) must be modified to account for the bulge of the earth between the two antennas. This changes the expression for the value of ψ and the value of θ and necessitates multiplying the magnitude of the reflection coefficient by the divergence factor, D. In practical calculations, it is found that for the most commonly encountered antenna heights, the curved earth optical region is not very great in extent and calculations for this region can frequently be omitted. Since D is always equal to or less than unity, its effect on the magnitude of the reflection coefficient will always be to reduce it. This means that when the effect of D is appreciable, the reflected wave will be very small and the propagation loss will be approximately equal to the free space loss.

In calculating the RFI which can result in a given case, little harm is done if the calculations are slightly pessimistic; that is, if the field strength which might cause interference is estimated to be somewhat higher than it actually is. Using free space calculations in the curved earth optical region will almost always give a value of field that is somewhat higher than is actually the case, so in general the special calculations necessary for this region can be neglected.

If it is found that the curved earth optical region is quite extensive or a critical equipment is located in this region, then calculations should be made. The mathematical expressions are quite complex and involve the solution of non-linear equations; however, a set of graphs has been plotted which allows the problem to be solved with relative ease. These are shown in Figures 4-9 through 4-12[7b]. The value of, θ', the modified phase lag due to path length difference is calculated from Figure 4-9 as follows: two parameters S and T are calculated to find J from the graph,

$$S = \frac{d}{\sqrt{2h_1} + \sqrt{2h_2}} \leq 1 \qquad (4\text{-}17)$$

$$T = \frac{h_2}{h_1} \leq 1 \qquad (4\text{-}18)$$

The intersection of the ordinates of S and T determine the value of J then:

$$\theta' = f_{mc} \cdot \frac{h_1 h_2}{d} \cdot J \cdot 1.385 \times 10^{-4} \qquad (4\text{-}19)$$

and,

θ' is in degrees

note that

$$\theta' = J \cdot \theta \qquad (4\text{-}20)$$

From Figure 4-10 the value of ψ', the modified angle that the reflected ray makes with the surface of the earth, is found by using the same parameters S and T in a similar manner to find K, then:

$$\tan \psi' = \frac{h_1 + h_2}{5280d} \cdot K \qquad (4\text{-}21)$$

Note that

$$\tan \psi' = K \cdot \tan \psi \qquad (4\text{-}22)$$

From Figures 4-11 and 4-12 the value of the divergence factor, D, is found by again using S and T and reading the value of D directly from the graph. The second graph in this case is used for very small values of T.

The E field is then found for the curved earth optical region by using the formula for the plane earth optical region modified as stated above.

$$E = \sqrt{1 + \left[D\rho^{2}\right] + 2D\rho \cos(\theta' - \phi)} \qquad (4\text{-}23)$$

where:

ρ = magnitude of surface reflection coefficient for given ψ'

ϕ = phase of surface reflection coefficient for given ψ' (degrees)

θ' = modified phase lag due to path length difference (degrees) ρ and ϕ are given by the same graph used for the plane earth case, that is Figure 4-8.

$$\theta' = f_{mc} \cdot \frac{h_1 h_2}{d} \cdot J \cdot 1.385 \times 10^{-4} \qquad (4\text{-}24)$$

$$\psi' = \tan^{-1}\left[\frac{h_1 + h_2}{5280d} \cdot K\right] \qquad (4\text{-}25)$$

D = divergence factor determined from Figures 4-11 and 4-12.

As in the plane earth case, the actual value of the field is found by expressing E (voltage) in db and adding it to the free space transmission loss for the same point in space.

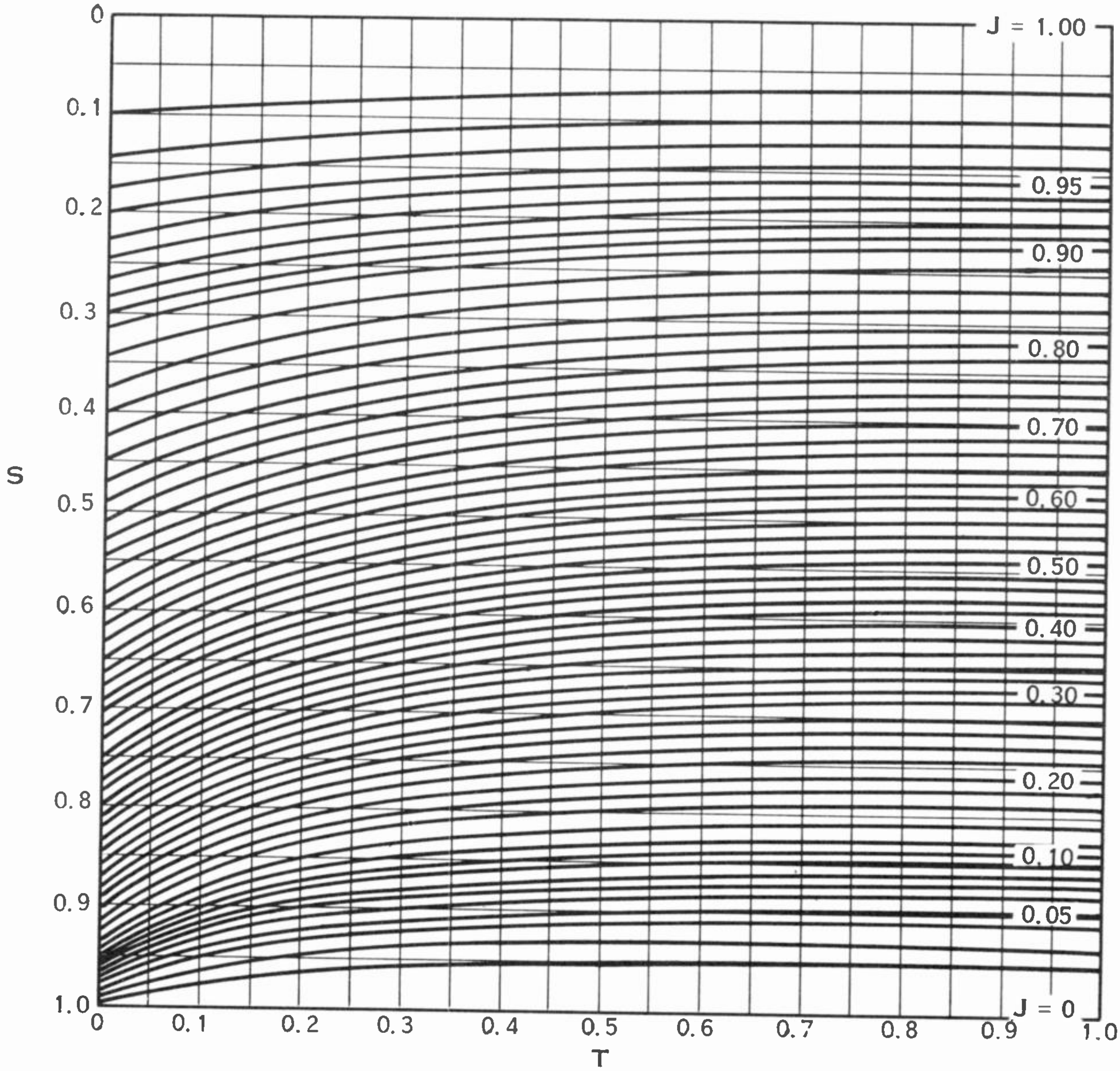


Figure 4-9. J Versus S and T for Determining θ'

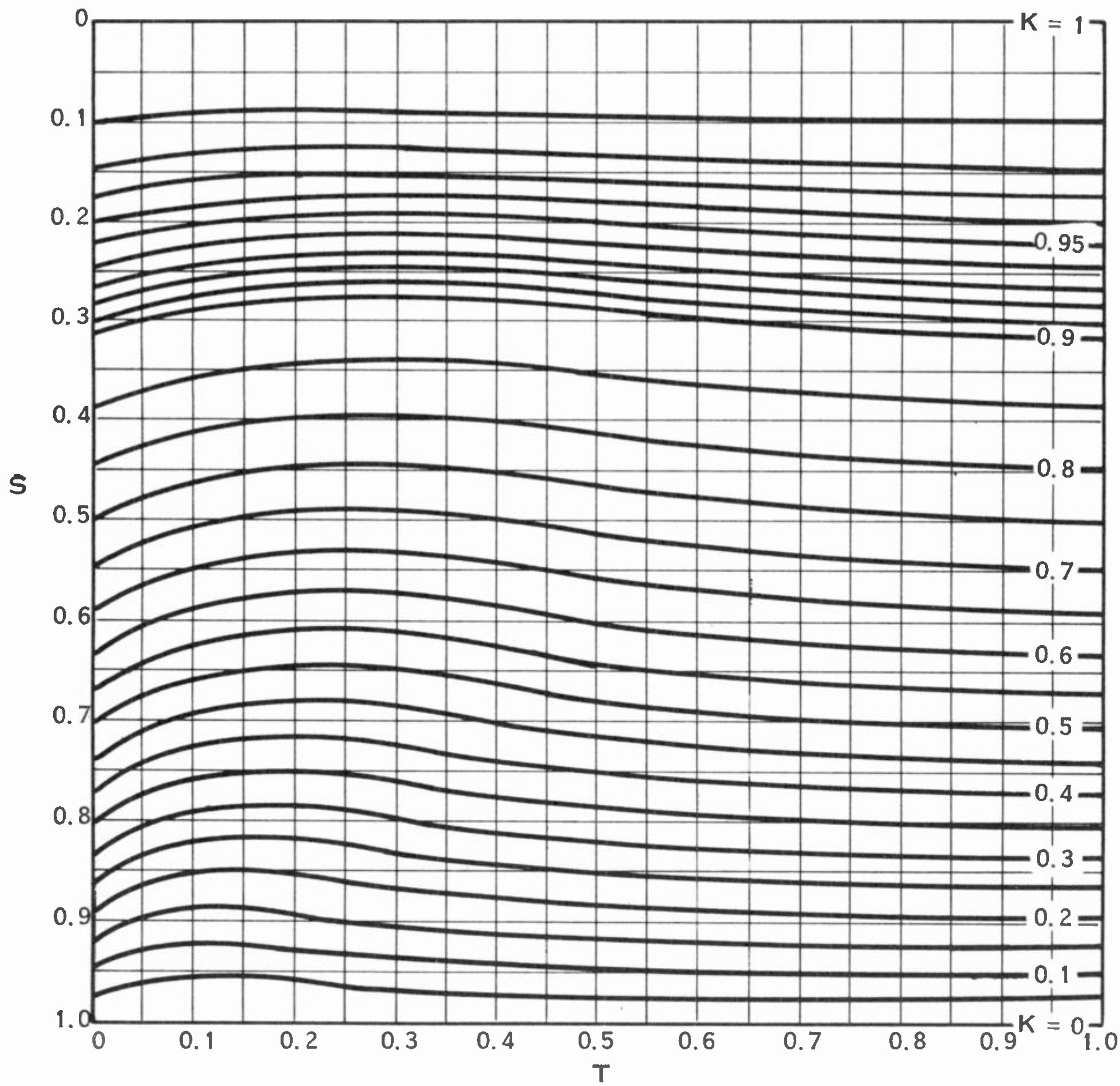


Figure 4-10. K Versus S and T for Determining ψ'

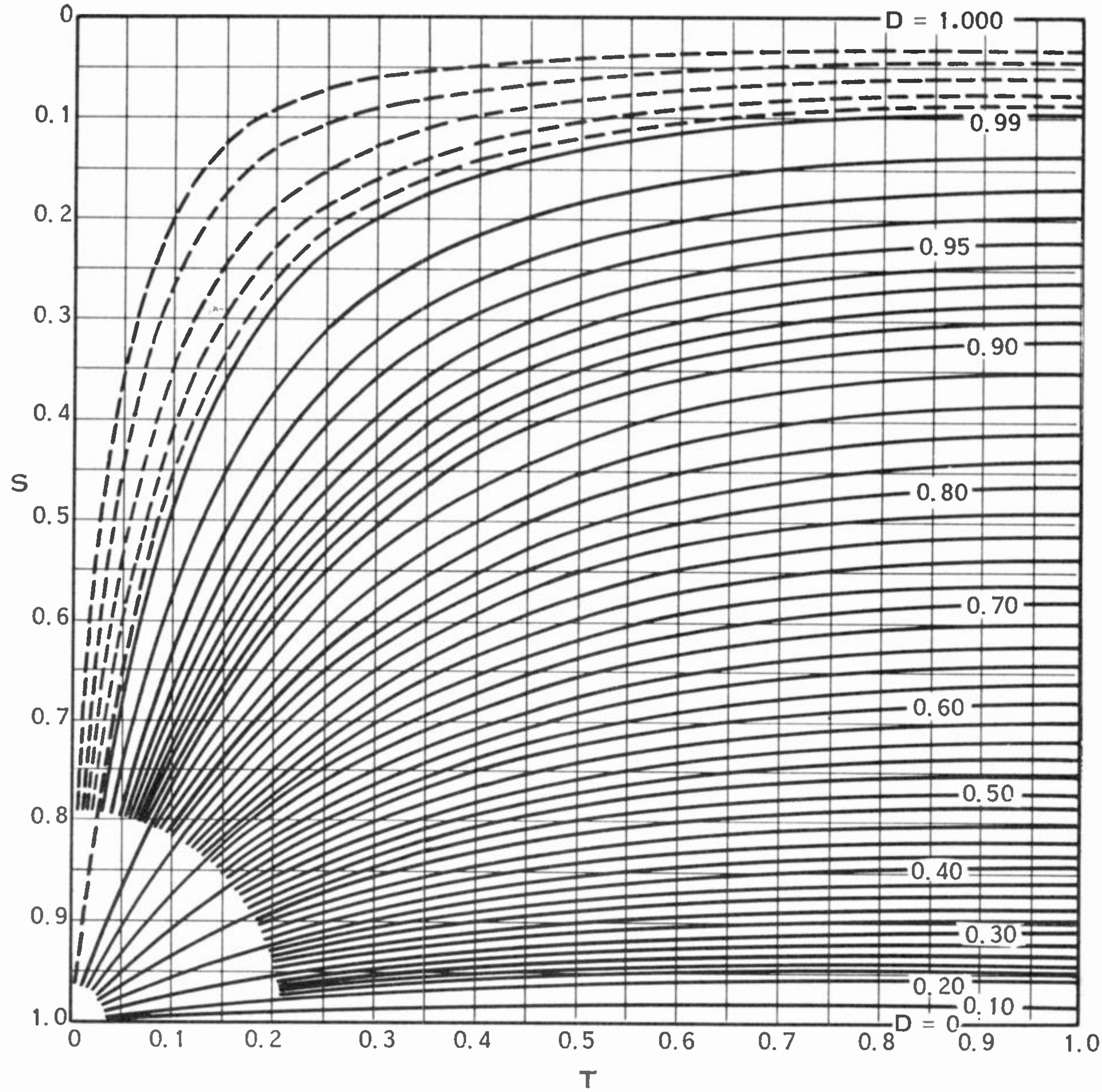


Figure 4-11. D Versus S and T

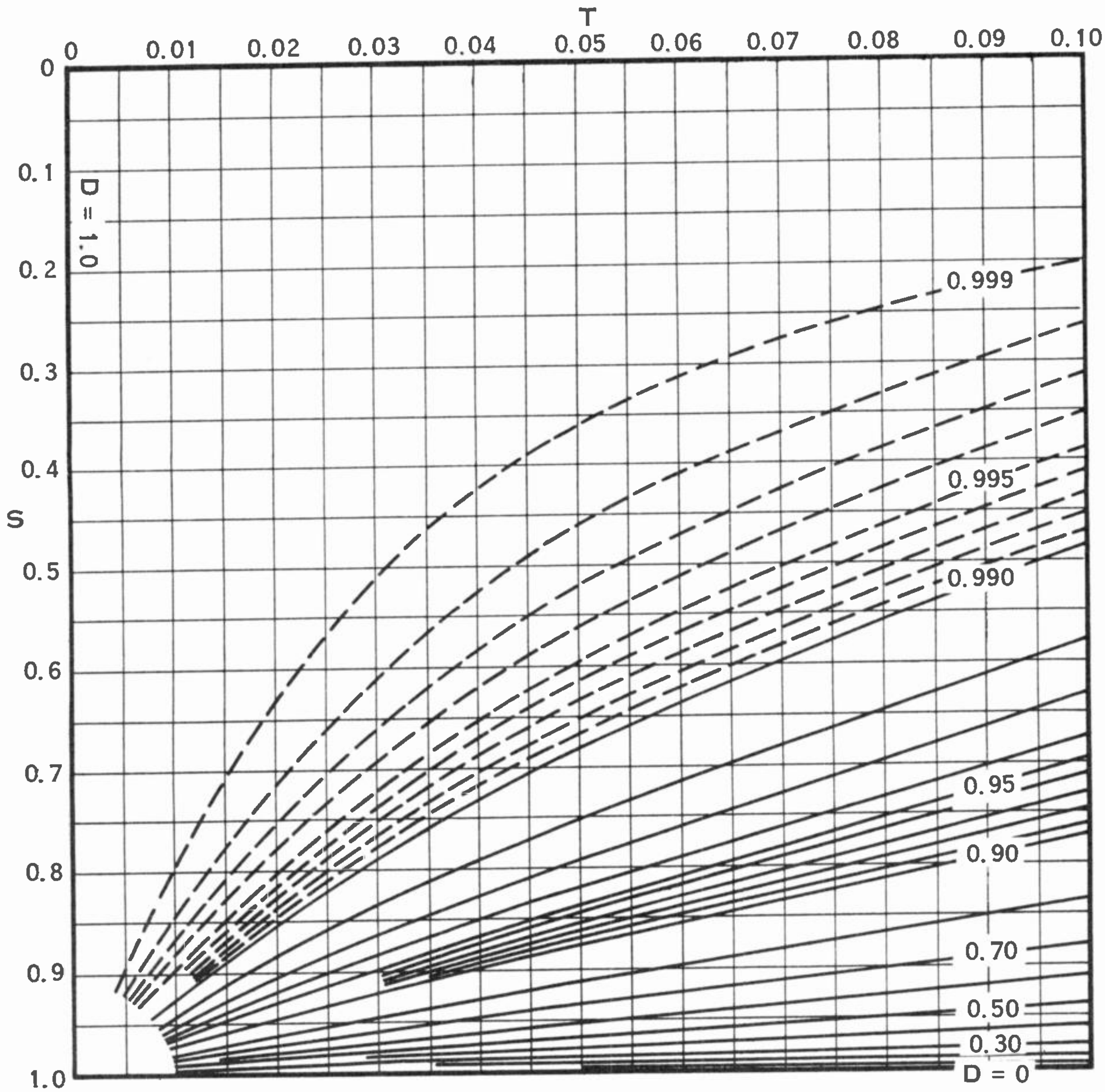


Figure 4-12. D Versus S and T for Low Values of T

1.3.6 CALCULATIONS IN THE DIFFRACTION REGION[8]

Whenever the distance between antennas is such that they are well below the line of sight and meet the conditions of equation (4-13), then they are in the diffraction region and the calculations are made on the basis of surface wave propagation. Exact calculations in this region are exceedingly complex and the literature dealing with them is not too well organized; however, Bullington gives a nomograph that can be used in most cases for RFI calculations above 30 mc.

For the use of Bullington's nomograph, in Figures 4-13a and 4-13b, the total distance, d, between antennas is divided into three parts- (1) the distance between the lower antenna and its radio horizon, d_1 ; (2) the distance between the higher antenna and its radio horizon, d_2; and (3) the remaining distance d_3 , which is:

$$d_3 = d - (d_1 + d_2) \tag{4-26}$$

The loss associated with each of these distances is then found from the nomograph which gives L_1 , L_2, and L_3, in db. These losses are then totaled and added to the free space loss for the total distance, d, and the result is the propagation loss for antennas in the diffraction region.

1.3.7 SAMPLE PROPAGATION CALCULATION

If it is required to determine the potentially interfering signal strength at a receiving site a specific distance from the transmitter, it is necessary to determine in which of the four regions this site is located. Equations (4-9) through (4-13) establish the boundaries for these regions, dictating which of the equations (4-14), (4-23) or (4-26) and Figure 4-13 should be employed to calculate the transmission loss of the radiated energy due to the presence of the earth. When this result is added to the free-space transmission loss for the particular distance, the number is the theoretical loss in db occurring over the transmission path in question.

In the following example a specific propagation problem is set up, the four regions are located, and hypothetical distances are chosen to demonstrate the mechanics of all the calculations.

Consider an antenna forty feet above the ground, transmitting a vertically polarized 200 mc signal to a receiving antenna some distance away, sixty feet above the ground. First, find the line of sight distance from Figure 4-6 or Equation (4-8); then find the location of the four regions dictated by the equations mentioned above.

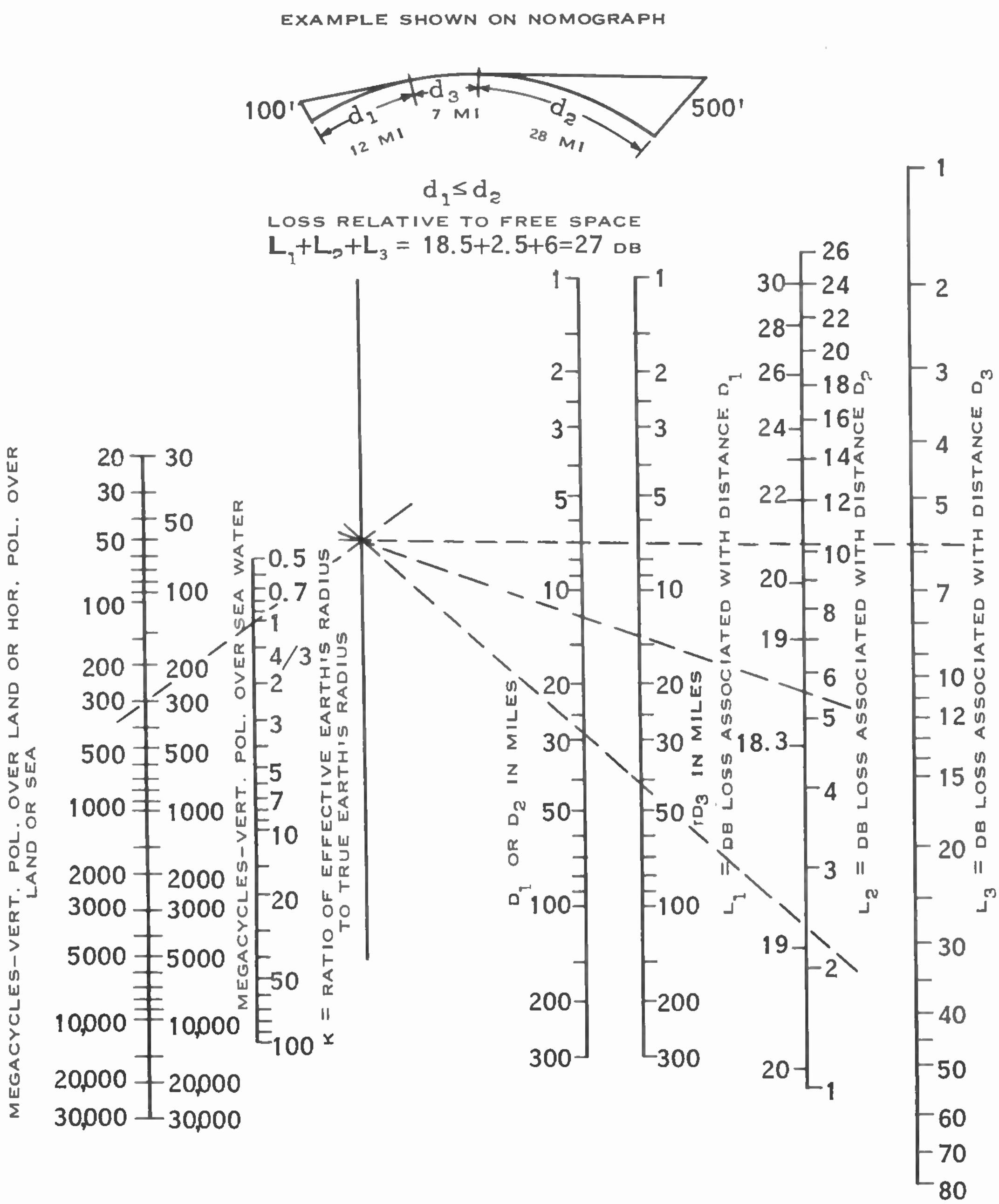


Figure 4-13a. Decibel Loss Relative to Free-Space Transmission at Points Beyond Line-of-Sight Over a Smooth Earth

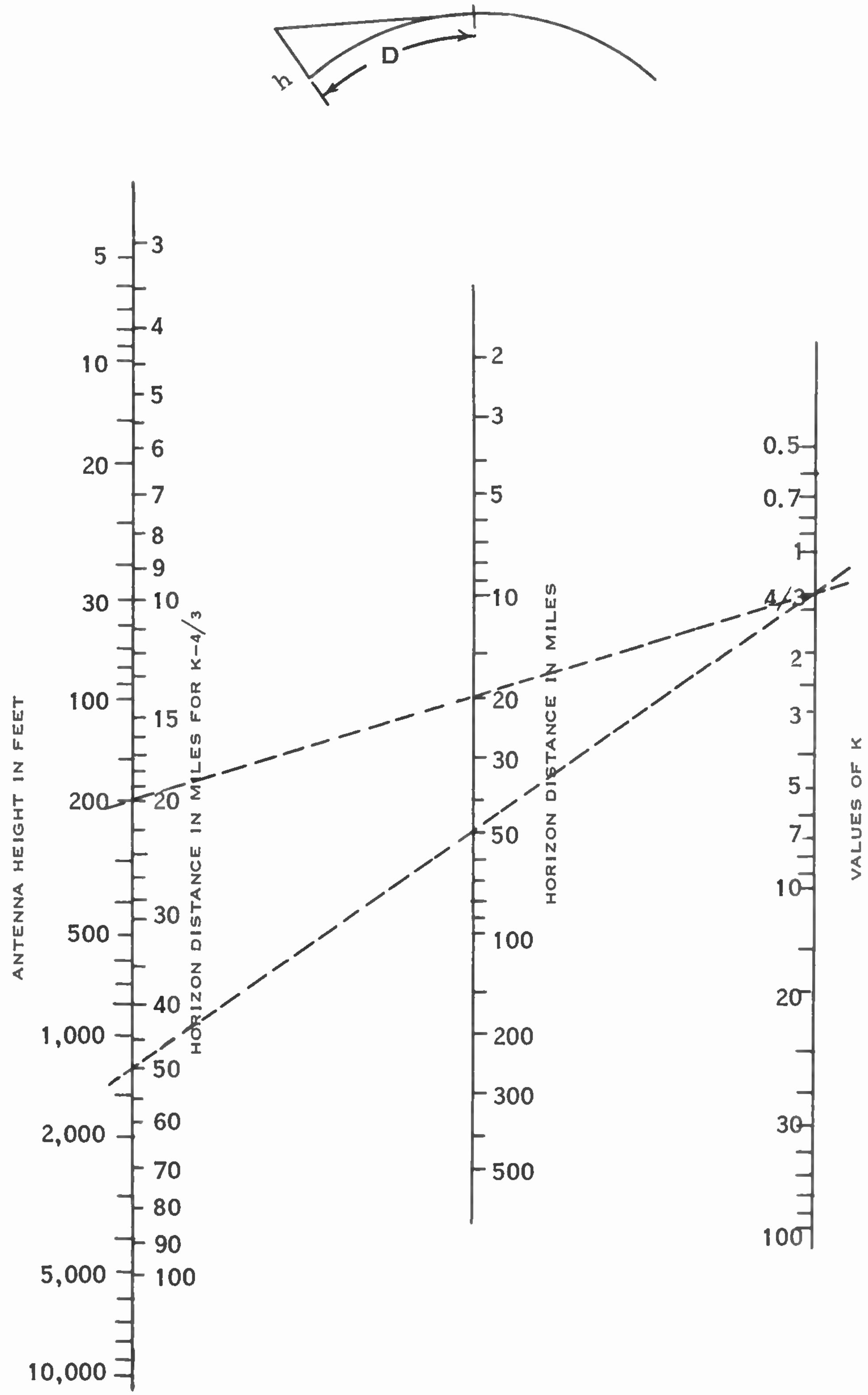


Figure 4-13b. Distance to the Horizon

a. Location of Region Boundaries

The outer boundary of the plane earth optical region is defined by ψ_1 and d_1 (Equations 4-9 and 4-10); they are applied here:

$$\psi_1 = .113\left(\frac{h_1 h_2}{h_1 + h_2}\right)^{1/2} = .113\left(\frac{40\cdot 60}{40+60}\right)^{1/2} = .554° \qquad (4\text{-}9)$$

$$d_1 = \sqrt{(92\psi_1)^2 + 2h_1} + \sqrt{(92\psi_1)^2 + 2h_2} - 184\psi_1 \qquad (4\text{-}10)$$

$$= \sqrt{(92\cdot .554)^2 + 2\cdot 40} + \sqrt{(92\cdot .554)^2 + 2\cdot 60} - 184\cdot .554$$

$$= 2 \text{ miles}$$

The outer boundary of the curved earth optical region is defined by ψ_c and d_c (equations 4-11 and 4-12); they result in

$$\psi_c = \frac{1.02°}{f_{mc}^{1/3}} = \frac{1.02°}{(200)^{1/3}} = 072°, \qquad (4\text{-}11)$$

$$d_c = \sqrt{(92\psi_c)^2 + 2h_1} + \sqrt{(92\psi_c)^2 + 2h_2} - 184\psi_c \qquad (4\text{-}12)$$

$$= \sqrt{(92\cdot .072)^2 + 2\cdot 40} + \sqrt{(92\cdot .072)^2 + 2\cdot 60} - 184\cdot .072$$

$= 10.2$ miles. (Note: For some heights this region does not exist.)

Finally, the outer boundary of the multi-mode region, as defined by Equation (4-13), results in the following:

$$d = \sqrt{2h_1} + \sqrt{2h_2} + \frac{173}{f_{mc}^{1/3}} \qquad (4\text{-}13)$$

$$= \sqrt{2\cdot 40} + \sqrt{2\cdot 60} + \frac{173}{(200)^{1/3}}$$

$$= 49.5 \text{ miles}$$

These figures can be summarized:

(a) out to 2 miles: plane earth optical region
(b) 2 to 10. 2 miles: curved earth optical region
(c) 10. 2 to 49. 5 miles: multi-mode region
(d) from 49. 5 out: diffraction region

These regions are displayed on the graph in Figure 4-14a, which shows the transmission loss curves calculated for a number of locations, including the three representative points calculated below (there is no calculation for the multi-mode region and the dotted line represents an extrapolation).

b. A Plane Earth Optical Region Sample Calculation

Let us choose d = 1 mile as the sample case for this region. By use of the nomograph (Figure 4-5), the free space transmission loss for this distance is found to be 78 db.

The E field due to the earth in this region is:

$$E = \sqrt{1 + \rho^2 + 2\rho \cos(\theta - \phi)} \qquad (4\text{-}14)$$

The quantities ρ and ϕ are functions of the angle ψ and can be read off the graph in Figure 4-8 once ψ is known.

$$\psi = \tan^{-1}\left(\frac{h_1 + h_2}{5280d}\right) = \tan^{-1}\left(\frac{40 + 60}{5280}\right) = 1.08°, \qquad (4\text{-}16)$$

Therefore,

$$\rho = 0.567,$$
$$\phi = 150°$$

The angle θ is calculated with:

$$\theta = \frac{1.385(10^{-4}) h_1 h_2 f_{mc}}{d} = 1.385(10^{-4}) \cdot 40 \cdot 60 \cdot 200$$

$$= 66.5°$$

Substitution of these values into Equation (4-14) yields an E field gain over the free-space transmission loss of 1. 5 db. Therefore, the field at one mile is reduced from the field at the transmitting antenna by 76. 5 db. The data for these calculations is shown in Figure 4-14b.

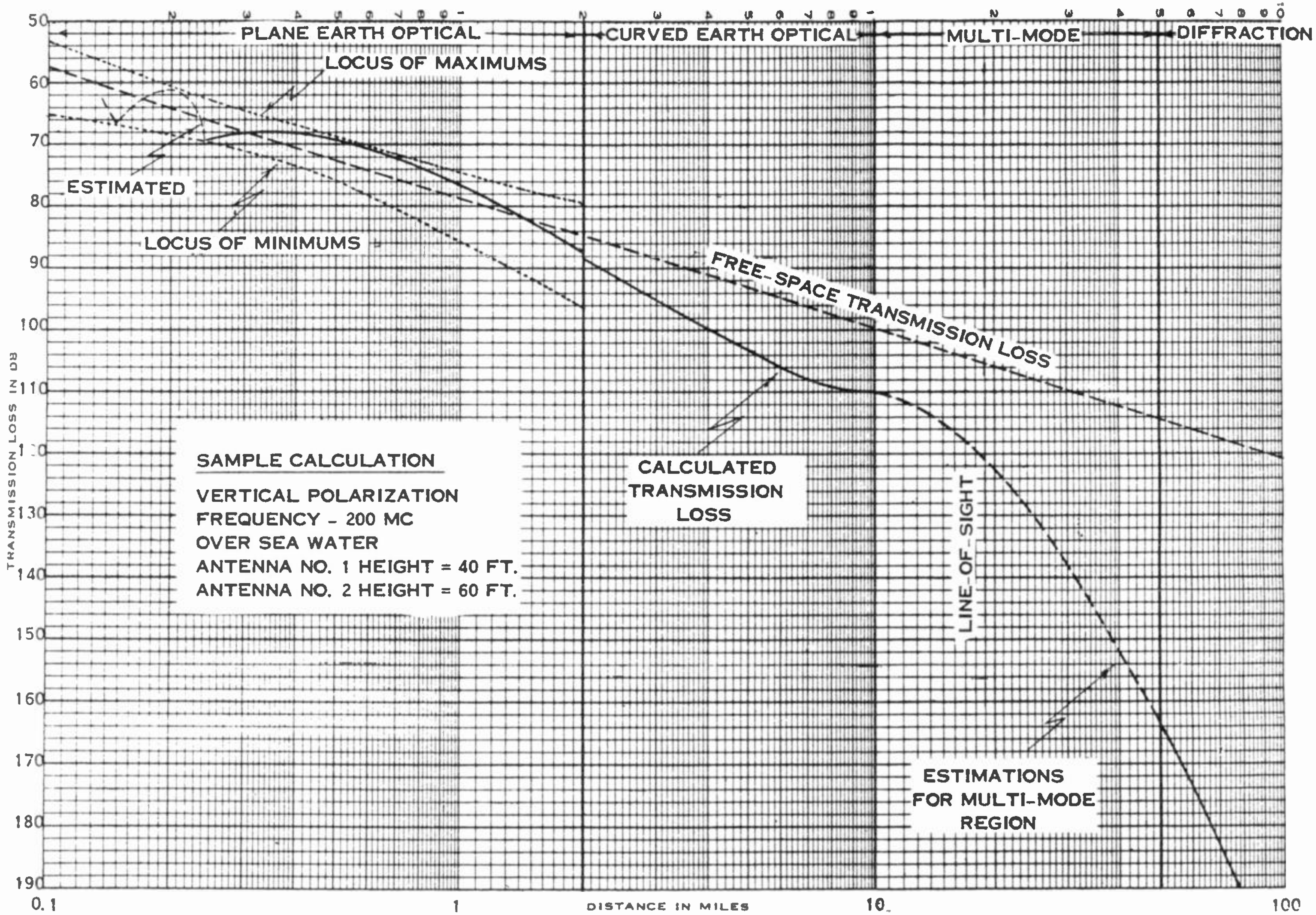


Figure 4-14a. Transmission Loss Between Half-Wave Dipoles for the Sample Calculation

For calculations in this region, it is helpful to establish the loci of maximums and minimums for the envelope of the E field values. Figure 4-14a shows them for this case.

They represent the highest and lowest values of Equation 4-14, which are given by:

$$E_{max} = 1 + \rho \qquad (4\text{-}27)$$

$$E_{min} = 1 - \rho \qquad (4\text{-}28)$$

c. A Curved Earth Optical Region Sample Calculation

For this region let the distance be five miles. Figure 4-5 shows that the free-space transmission loss for this point is 92 db.

The E equation here is similar to that for the preceding calculation except for the inclusion of three factors to allow for the curvature of the earth. Figures 4-9 through 4-12 yield values for factors J, K, and D, each of which are functions of S and T, inputs to the graph.

For an example,

$$S = \frac{d}{\sqrt{2h_1} + \sqrt{2h_2}} = \frac{5}{\sqrt{2 \cdot 40} + \sqrt{2 \cdot 60}} = 0.251 \qquad (4\text{-}17)$$

$$T = \sqrt{\frac{h_1}{h_2}} = 0.817 \qquad (4\text{-}18)$$

The resulting values for the factors are:

$$J = 0.880$$

$$K = 0.935$$

$$D = 0.941$$

Inserting these into their places in Equations (4-19), (4-21), and (4-23), and calculating as in the plane earth case, the E field is found to be -10.6 db, which means that the resulting loss figure for this distance is 102.6 db. The data for these calculations is shown in Figure 4-14b.

d. A Diffraction Region Sample Calculation

The Bullington nomographs in Figures 4-13a and 4-13b allow the determination of loss in terms of the three distances d_1, d_2, and d_3.

PLANE EARTH OPTICAL REGION					
d (Miles)	θ (Degrees)	ψ (Degrees)	ρ	ϕ (Degrees)	Loss (db)
.25	266	4.3	.405	58	-3.5
.50	133	2.2	.385	114	+2.7
.75	88.7	1.4	.471	139	+2.6
1.00	66.5	1.1	.567	152	+1.5
1.25	53.2	0.87	.625	158	+0.3
1.50	44.3	0.72	.676	163	-1.0
1.75	38.0	0.62	.713	166	-1.5
2.00	33.3	0.54	.745	168	-3.0

CURVED EARTH OPTICAL REGION										
d (Miles)	S	T	J	K	D	θ' (Deg.)	ψ' (Deg.)	ρ	ϕ (Deg.)	Loss (db)
2.0	.100	.817	.98	.99	.99	32.5	.539	.772	169	-3.3
3.0	.151	.817	.96	.98	.98	21.2	.353	.828	173	-6.3
4.0	.201	.817	.92	.96	.96	15.3	.248	.877	175	-9.0
5.0	.251	.817	.88	.94	.94	11.7	.213	.896	176	-10.6
6.0	.301	.817	.83	.91	.92	9.3	.164	.917	177	-11.8
7.0	.352	.817	.78	.87	.89	7.4	.133	.934	177	-12.7
8.0	.402	.817	.72	.83	.86	6.4	.114	.942	178	-12.3
9.0	.452	.817	.64	.80	.82	4.8	.096	.951	178	-12.2
10.2	.513	.817	.54	.73	.77	3.2	.078	.960	178	-11.4

Figure 4-14b. Table Showing Results of Sample Calculation

DIFFRACTION REGION							
d (Miles)	d_1 (Miles)	d_2 (Miles)	d_3 (Miles)	L_1 (db)	L_2 (db)	L_3 (db)	Loss (db)
49.5	9.0	11.0	29.5	-21.2	-8.5	-21	-50
55	9.0	11.0	35.0	-21.2	-8.5	-24	-54
60	9.0	11.0	40.0	-21.2	-8.5	-28	-58
70	9.0	11.0	50.0	-21.2	-8.5	-37	-67
80	9.0	11.0	60.0	-21.2	-8.5	-44	-74
90	9.0	11.0	70.0	-21.2	-8.5	-53	-83
100	9.0	11.0	80.0	-21.2	-8.5	-61	-91

Figure 4-14b. Table Showing Results of Sample Calculation (Continued)

Selecting a distance of 55 miles, (free-space transmission loss = 114 db) our antenna heights of 40 and 60 feet yield from the nomograph

d_1 = 9 miles

d_2 = 11 miles

d_3 = 35 miles

At the selected frequency, these represent losses of

L_1 = 21.2 db

L_2 = 8.5 db

L_3 = 24 db

for a total of 53.7 db. The result, then, is a total loss figure at this distance of 167.7 db. The data for these calculations is shown in Figure 4-14b.

1.3.8 EFFECTS OF TERRAIN IRREGULARITIES

The preceding discussion assumes that the earth is a perfectly smooth sphere. The modification in these results caused by the presence of hills, trees, and buildings is difficult or impossible to compute, but the order of magnitude of these effects may be obtained from a consideration of the other extreme case, which is propagation over a perfectly absorbing knife edge.

The diffraction of plane waves over a knife edge or screen causes a shadow loss whose magnitude is shown in Figure 4-15. The height of the obstruction H is measured from the line joining the two antennas to the top of the ridge. It will be noted that the shadow loss approaches 6 db as H approaches zero (grazing incidence), and that it increases with increasing positive values of H. When the direct ray clears the obstruction, H is negative, and the shadow loss approaches zero db in an oscillatory manner as the clearance is increased. In other words, a substantial clearance is required over line-of-sight paths in order to obtain "free-space" transmission.

There is an optimum clearance, called the first-Fresnel-zone clearance, for which the transmission is theoretically 1.2 db better than in free space. Physically, this clearance is of such magnitude that the phase shift along a line from the antenna to the top of the obstruction and from there to the second antenna is about 1/2 wavelength greater than the phase shift of the direct path between antennas. When this phase difference is 1 wavelength, the path clears the first two Fresnel zones, and there is theoretically a loss of about 1 decibel relative to free space. Similarly, when the phase difference is 3/2 wavelengths, the path clears the first three Fresnel zones and this is a gain of about 0.8 db relative to free space. The locations of the first three Fresnel zones are indicated on the right-hand scale on Figure 4-15, and by means of this chart the required clearances can be obtained. At 3000 megacycles, for example, the direct ray should clear all obstructions in the center of a 40-mile path by about 120 feet to obtain full first-zone clearance. The corresponding clearance for a ridge 100 feet in front of either antenna is 4 feet. Should the ridge project above the direct path by 4 feet, the shadow loss is about 15 decibels. It will be noted that the effective clearance obtained on a particular path will vary with the weather conditions, since the effect of atmospheric refraction is neglected in Figure 4-15.

The problem of two or more knife-edge obstructions between the transmitting and receiving antennas, such as is shown in Figure 4-16, has not been solved rigorously. However, graphical integration indicates that the shadow loss for this case is equivalent within two or three decibels to the shadow loss for the knife edge represented by the height of the triangle composed of a line joining the two antennas and a line from each antenna through the top of the peak that blocks the line-of-sight from that antenna.

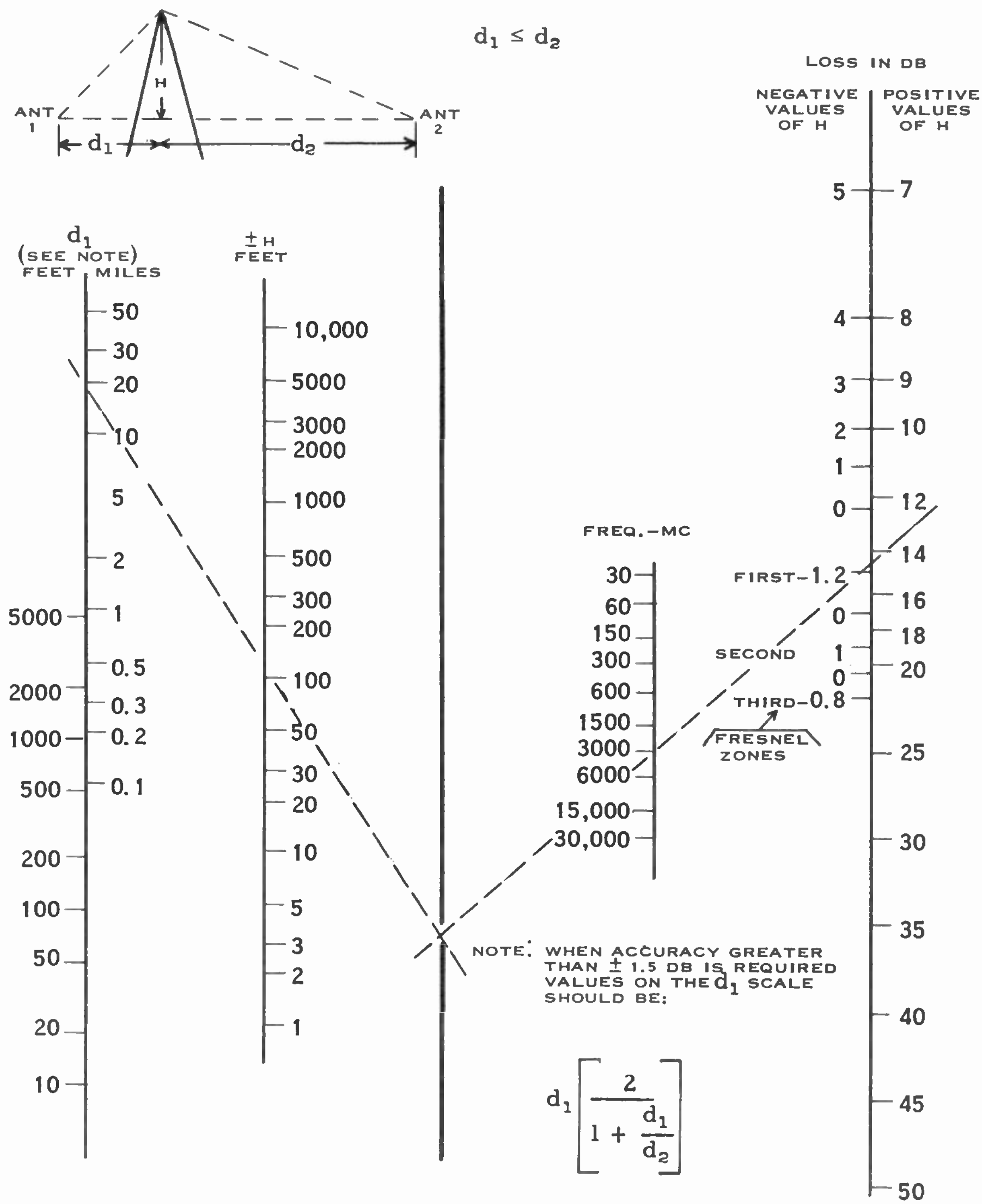


Figure 4-15. Shadow Loss Relative to Free Space

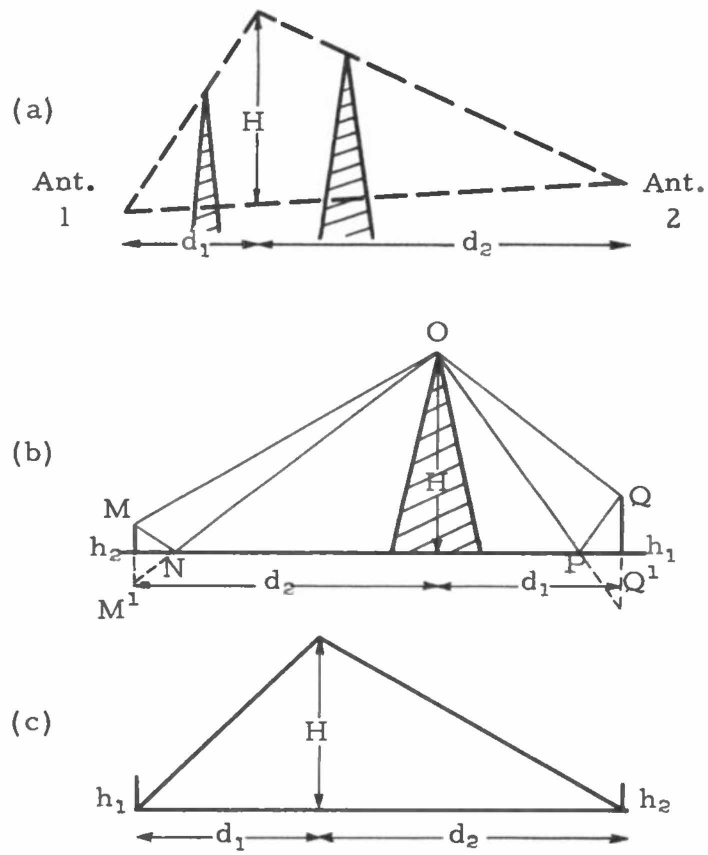


Figure 4-16. Ideal Profiles Used in Developing Theory of Diffraction over Hills

1.4 PROPAGATION OVER VERY SHORT DISTANCES WHICH ARE BEYOND THE FRESNEL REGION

When propagation over very short distances, say two miles or less, is considered, three mechanisms assume primary importance. The first is the transfer of energy by means of the direct and the ground reflected waves indicated in Figure 4-7. The second is the guidance of electromagnetic waves by the surface of the earth, somewhat analogous to transmission by a Goubeau surface wave line or a dielectric waveguide. This mechanism is the previously referred to surface wave transmission. The third is the propagation by reflection from man-made structures and terrain features not normally considered.

Computational procedures for the space wave have been discussed previously and the computations for the surface wave were discussed for the diffraction region beyond the line of sight.

The ratio of space wave field strength to surface wave field strength versus frequency is shown in Figure 4-17. By examining it, several facts become apparent. First, surface wave transmission is most important at low frequencies. For example, if the receiving and transmitting antennas are 5 feet above an average ground and are 300 feet apart and if the polarization is vertical, the amplitude of the space wave is only 6 db above the amplitude of the surface wave at 30 mc; at 300 mc, the corresponding figure is 28 db. Thus at higher frequencies, the surface wave becomes negligible and the calculations for very short distances can be computed on the same basis as those for great distances. (One should, of course, make sure that the distance is great enough that Fresnel zone considerations do not apply; that is, the distance must be greater than the square of the largest aperture dimension divided by the wave length.) Second, Figure 4-17 shows that the surface wave field strength of a vertically polarized wave, at a fixed distance from a given source is much larger than the field strength of a horizontally polarized wave, if it is assumed that all other circumstances, such as antenna elevation and frequency, remain unchanged. This accounts for the phenomenon that low frequency interference from sources placed close to the earth's surface, such as ignition noise or radar modulator radiation, can be expected to appear vertically polarized. The more detailed explanation is as follows: it can be shown that the received free space field strength is almost always very low if the transmitting and receiving antennas are located, in terms of wavelength, very close to the earth's surface. Therefore, surface wave transmission is the primary vehicle for low frequency, low elevation interference; and since the horizontally polarized surface wave is attenuated more heavily than the vertically polarized surface wave, vertically polarized interference predominates at the receiver location even if the signal was randomly polarized at its source.

The third fact which can be noticed by considering Figure 4-17 is important enough to be emphasized separately. Comparison of the two sets of curves for different elevations shows that the importance of surface wave transmission increases as the antenna height is decreased.

Since surface wave propagation over a lossy earth requires a component of the Poynting vector to be directed into the ground to account for losses, surface wave propagated energy from a purely vertically polarized source will exhibit wave tilt. This effect accounts for the well-

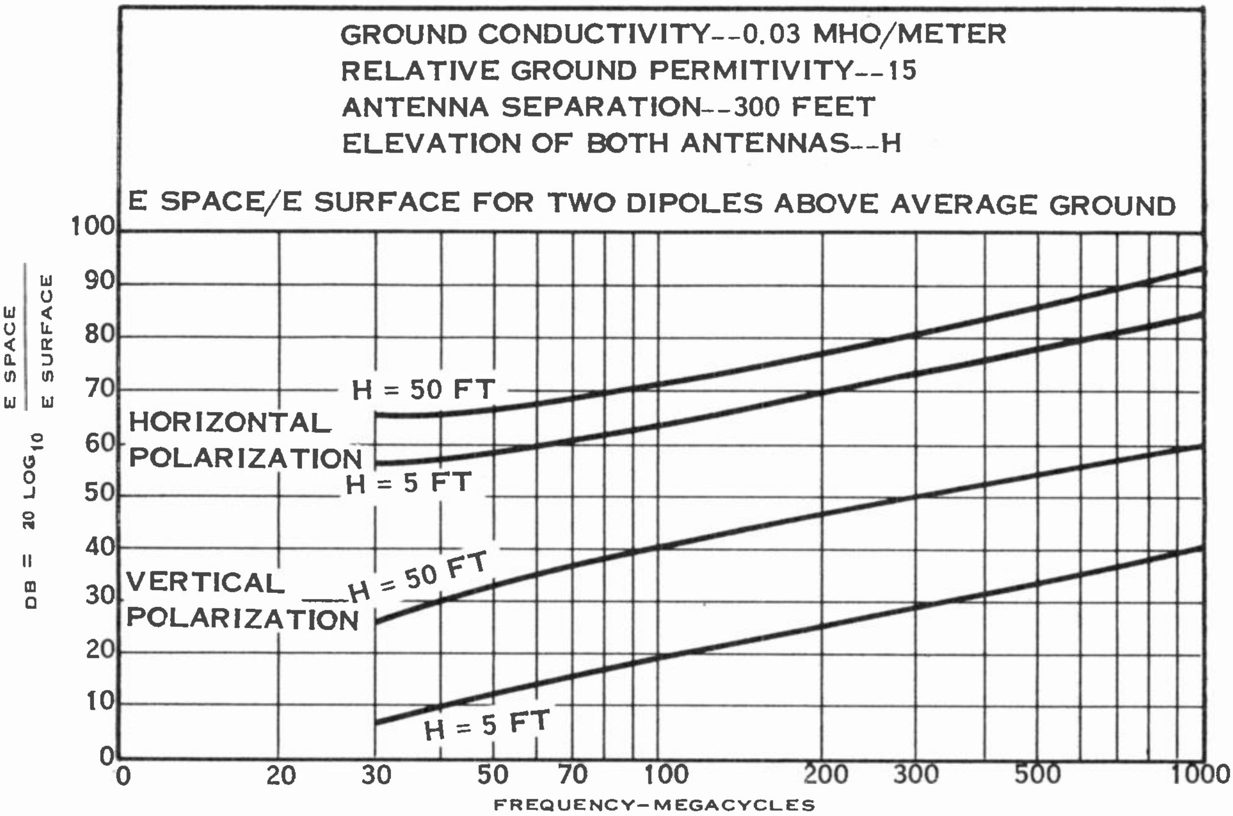


Figure 4-17. Relative Magnitudes of Surface and Space Waves

known fact that a horizontal antenna will pick up broadcast frequencies transmitted from a vertical tower. Equations and curves for this aspect of surface wave propagation are given in Terman's Handbook.[10]

From the above, it has been shown that the polarization radiated from a source may be changed because of the characteristics of surface wave propagation. The entire subject of polarization conversion is an extremely interesting one and much experimental and theoretical research remains to be done. These studies would be very useful for the problem of interference prediction.

Quite frequently, because of the close proximity of the antennas, the source of interference is differently polarized than the antenna of the disturbed equipment; for example, a horizontally polarized search radar versus a vertically polarized UHF receiver. In such cases, it would be desirable to determine how much attenuation, in free space, can be expected at a given frequency due to this difference in polarization.

Figure 4-18 shows the additional attenuation over normal free space attenuation resulting from a difference in polarization between an arriving wave and the receiving antenna. Since the data was based on a small statistical computation from measurements on polarization losses in radar antennas, the information serves as a guide for the attenuation that may be expected to occur in most cases. As an example, assume that the receiving antenna is aligned to receive horizontal polarization and the interference source is transmitting vertical polarization; therefore from the table, 20 db of attenuation is provided by the fact that the antennas are cross polarized. Also, if the receiving antenna is vertically polarized and the interference source is transmitting circular polarization, 3 db of attenuation is provided.

Reflection from man-made structures and unusual terrain features causes polarization rotation which has effects similar to those just discussed. Polarization changes and the generation of a cross-field component must be considered whenever the electric vector of the incident wave is not parallel to the reflecting surface. Figure 4-19 shows the situation for a perfect conductor. With horizontal polarization, reversal

TRANSMITTER-RECEIVER POLARIZATION ALIGNMENT FACTORS*

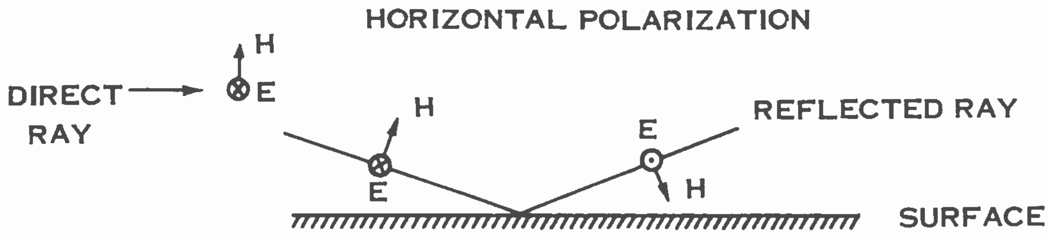
(Expressed in Units of DB Loss)

	Horizontal	Vertical	Diagonal (45°)	Circular RH	Circular LH
Horizontal	0	20	3	3	3
Vertical	20	0	3	3	3
Diagonal (45°)	3	3	0-20	3	3
Circular RH	3	3	3	0	25
Circular LH	3	3	3	25	0

*For harmonic radiations from the transmitter, use 10 db for cross-polarized conditions.

Figure 4-18. Transmitter-Receiver Alignment Factors

1. E-Vector Parallel to Reflecting Surface (Perfect Conductor).



2. E-Vector Not Parallel to Reflecting Surface (Perfect Conductor).

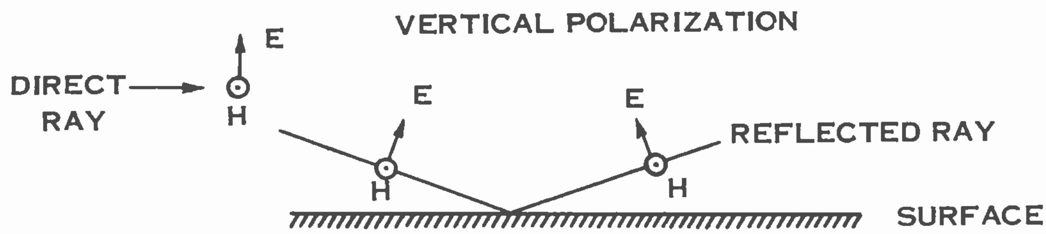


Figure 4-19. Polarization of Direct and Reflected Waves.

of the E vector occurs upon reflection from a horizontal surface; however, the inclination of the vector remains zero degrees. In the case of vertical polarization, the inclination, with respect to some reference direction of the E vector in the reflected wave, is different from the inclination of the E vector in the direct wave. If the reflecting surface is inclined with respect to the horizontal, then, depending upon the slope of this reflecting surface, a wide range of linear polarizations of the reflected wave becomes possible. Limiting factors are the beam widths of the transmitting and receiving antennas. Furthermore, if the direct and reflected waves are not in time phase, the resultant E vector will rotate in the plane of the paper as time progresses. This is called elliptical cross field.

In general, it would be desirable for interference calculations to know at least the maximum possible amount of polarization conversion, given a specified transmitting antenna and assuming finite, not infinite, ground conductivity and a finite dielectric constant. Further analytical work in this direction, as well as a great deal of experimentation, is required.

1.5 PROPAGATION CONSIDERATIONS IN THE FRESNEL REGION

Frequently, it will be necessary to calculate the RFI potential between two large aperture antennas when the largest aperture size L is such that the separation distance, d, requires that Fresnel zone corrections be applied to the propagation calculations. The Fresnel zone is taken to be any distance between antennas such that

$$d < \frac{L^2}{\lambda} \qquad (4\text{-}29)$$

where:

d = distance between antennas

L = largest linear dimension of the antenna

λ = wavelength

In the Fresnel region, the antenna gain and pattern is a function of d. Calculations of gain in the Fresnel zone are made by calculating the gain by the formula given in Equation 4-5, Paragraph 1.1.3 and modifying this figure by correction factors which vary with distance.

1.5.1 GAIN CORRECTION FOR FRESNEL REGION OPERATION

Figures 4-20, 4-21, 4-22, 4-23a and 4-23b are the gain correction factors for antennas having uniform, cosine, cosine squared, cosine cubed and cosine fourth aperture illumination. The abcissa is the distance from the antenna in wavelengths and the ordinate is the gain reduction in db due to Fresnel region operation. Each graph has a family of curves corresponding to different aperture dimensions. The largest linear aperture dimension L_λ is in wavelengths.

The following example illustrates the use of those curves:

H = 7.5 feet = Horizontal Dimension of Rectangular Aperture
V = 3.75 feet = Vertical Dimension of Rectangular Aperture
Far Field Gain: 24.5 db
Distribution in H plane is cosine
Distribution in V plane is uniform
Frequency: 1310 mc

Find the gain of the antenna at a distance of 7.5 feet and 75 feet.

$$\lambda = \frac{C}{f} = \frac{3 \times 10^{10}}{1310 \times 10^{6}} = 22.29 \text{ cm} = 0.75 \text{ feet} \qquad (4\text{-}30)$$

where C, the velocity of light = $3 \times 10^{10} \frac{\text{cm}}{\text{sec.}}$

$$H = \frac{7.5}{.75} = 10\lambda \qquad (4\text{-}31)$$

$$V = \frac{3.75}{.75} = 5\lambda \qquad (4\text{-}32)$$

7.5 feet = 10λ

75 feet = 100λ

From Figure 4-21 on the $L_\lambda = 10$ curve, the gain reduction is read as 6.1 db and 0.15 db at distances of 10 and 100 wavelengths, respectively, from the antenna. From Figure 4-20 on the $L_\lambda = 5$ curve, the gain reduction is read as 1.6 db and 0 db at a distance of 10 and 100 feet, respectively, from the antenna.

At a distance of 7.5 feet,

$$G = 24.5 - 6.1 - 1.6 = 16.8 \text{ db}$$

At a distance of 75 feet,

$$G = 24.5 - .15 - 0 = 24.35 \text{ db}$$

The gain in the Fresnel region is always lower than in the far-field region.

Unfortunately, for many antennas the aperture illumination is unknown. Since the gain reduction in the Fresnel region is dependent upon the illumination, it is necessary to estimate the illumination from other properties of the antenna. The following method uses the beam width at the half power points to determine the illumination.

Let: θ_H = full beam width at half power points in H direction in degrees

θ_V = full beam width at half power points in V direction in degrees

Calculate a constant R defined as follows:

$$R = \frac{\pi}{180} \cdot \frac{\theta_H H}{\lambda} \cdot \frac{\pi}{180} \cdot \frac{\theta_V V}{\lambda} \qquad (4\text{-}33)$$

H and V are as defined on page 4-48.

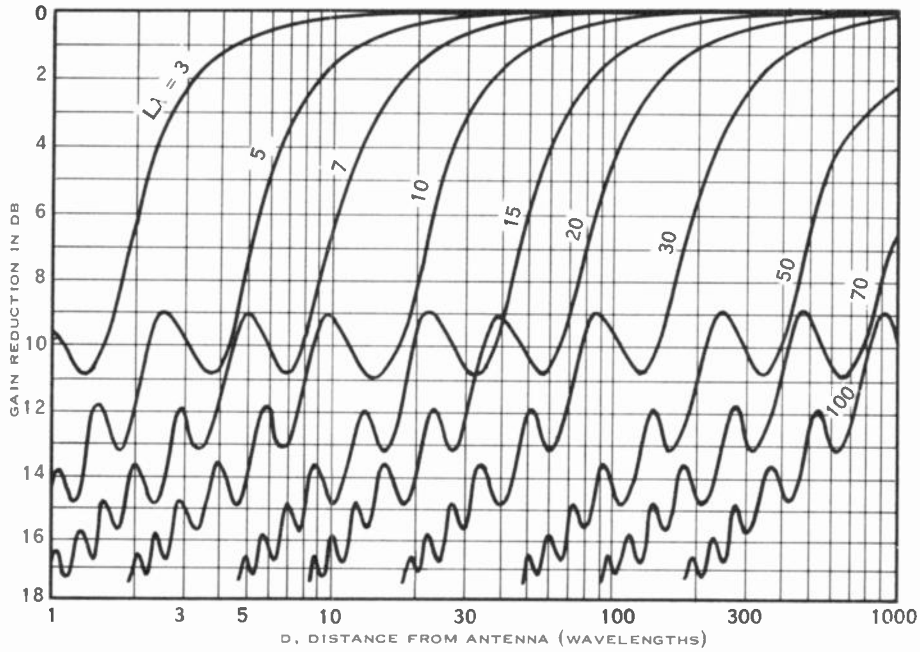


Figure 4-20. Fresnel Region Gain Correction for Uniform Illumination

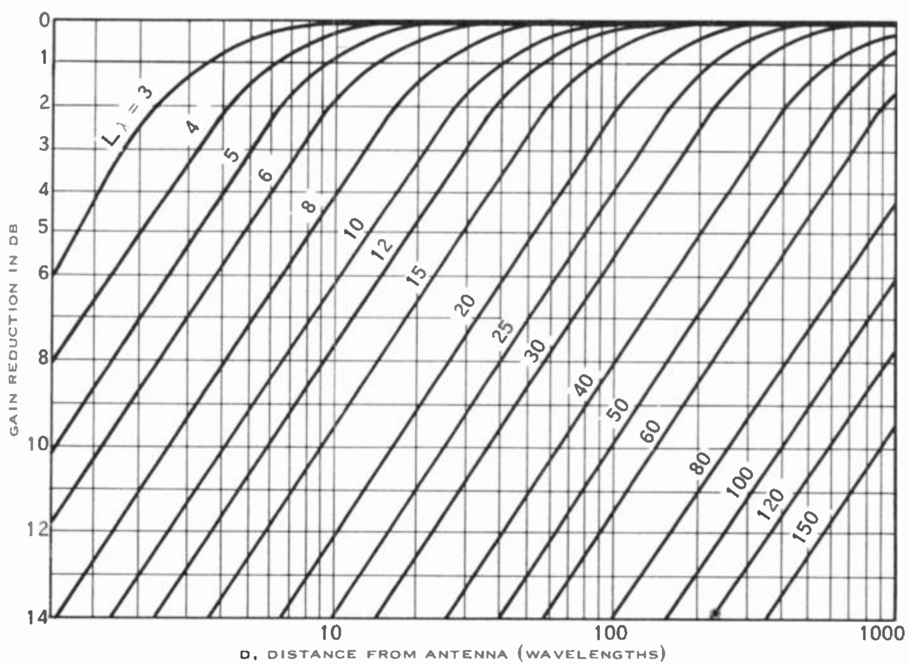


Figure 4-21. Fresnel Region Gain Correction for Cosine Illumination

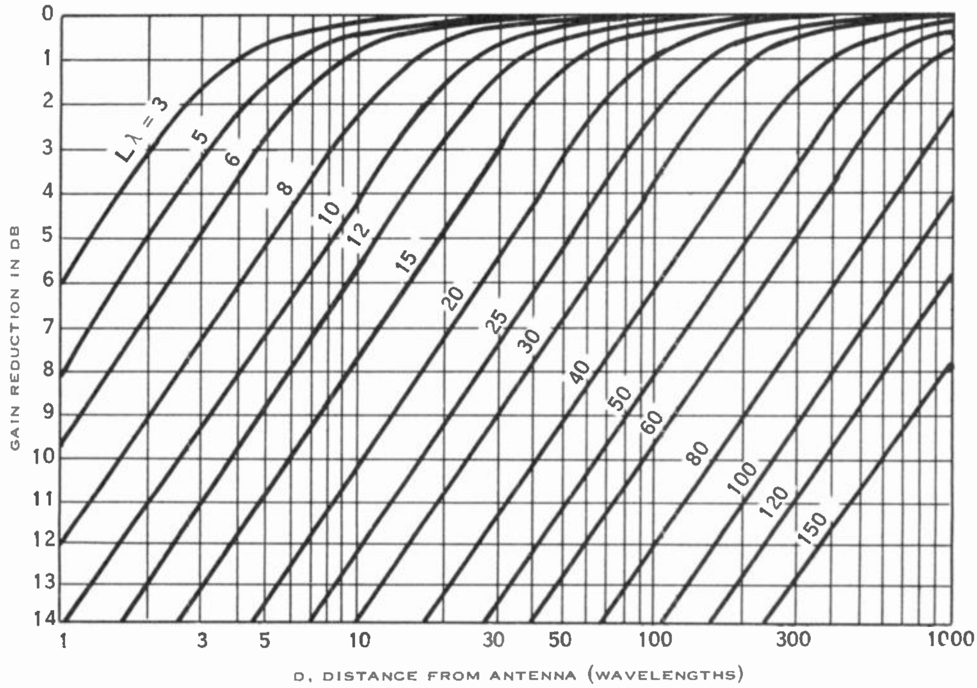


Figure 4-22. Fresnel Region Gain Correction for Cosine Square Illumination

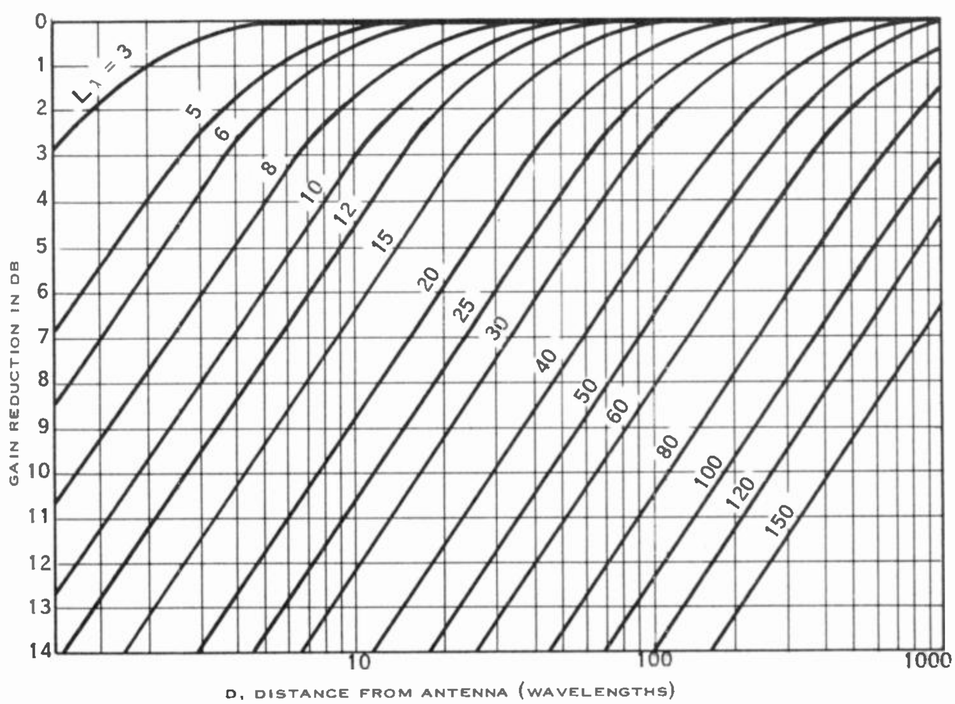


Figure 4-23a. Fresnel Region Gain Correction for Cosine Cubed Illumination

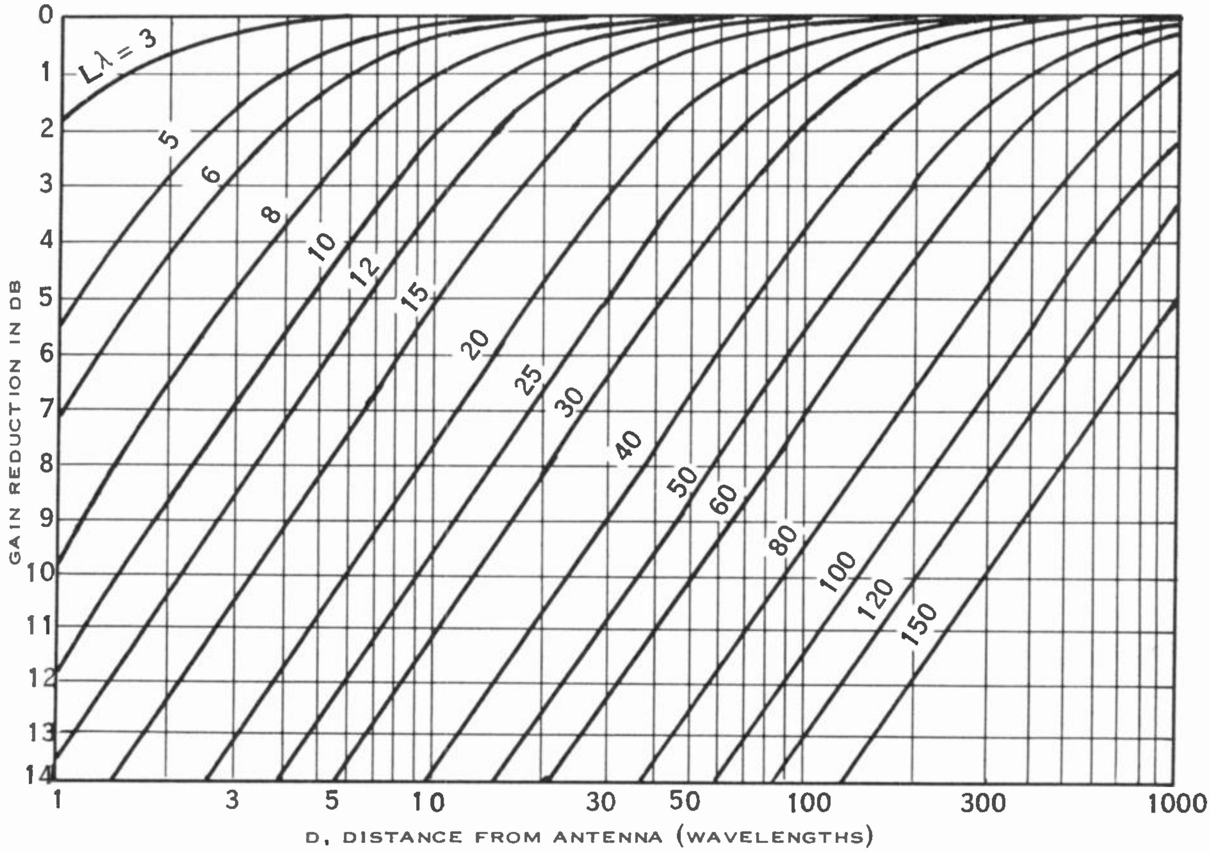


Figure 4-23b. Fresnel Region Gain Correction for Cosine Fourth Illumination

Limits of R	Estimated Illumination
$0.88 \leq R < 1.2$	Uniform
$1.2 \leq R < 1.45$	Cosine
$1.45 \leq R < 1.66$	Cosine Square
$1.66 \leq R < 1.93$	Cosine Cubed
$1.93 \leq R < 2.03$	Cosine Fourth

Table II

Figure 4-24. Limits of R for Given Illumination

In the table in Figure 4-24, limits are given for R to use in estimating the illumination. The following check can be made to determine whether the aperture illuminations estimated above are reasonable. Rewriting Equation 4-5, we have an expression for the antenna efficiency

$$K = \frac{G \lambda^2}{4 \pi A_P F_H F_V} \tag{4-34}$$

The efficiency of the antenna is then computed. If the computed antenna efficiency is reasonable, then it is safe to assume that the determined aperture illuminations are satisfactory. Reasonable efficiency can be defined as, generally $0.9 \geq K \geq 0.5$. Of course, this criterion can be modified for individual antennas when there is efficiency information specifying otherwise.

1.5.2 FRESNEL REGION PATTERNS[11]

Figures 4-25, 4-26, 4-27, 4-28 and 4-29 are the Fresnel region power patterns in a plane parallel to the aperture for uniform, cosine, cosine squared, cosine cubed and cosine fourth distributions. Since the distribution considered in this paper are symmetrical, the patterns are symmetrical. Therefore, the plots presented are only one-half of the antenna pattern. This is all that is necessary because the pattern on the other side of the center axis is exactly the same as the side shown. (Figure 4-30)

The abscissa is in units of normalized distance from the center axis defined as follows:

$$\overline{X} = \frac{2x}{a} \tag{4-35}$$

See Figure 4-30 for definition of symbols. The parameter P is the normalized distance from the antenna defined as follows

$$P = \frac{d_1}{a_1^2}$$

where d_1 = distance from antenna in wavelengths

a_1 = aperture dimension in wavelengths

Shown dotted on each of these figures is the theoretical power density at the aperture of the antenna. It must be remembered that in

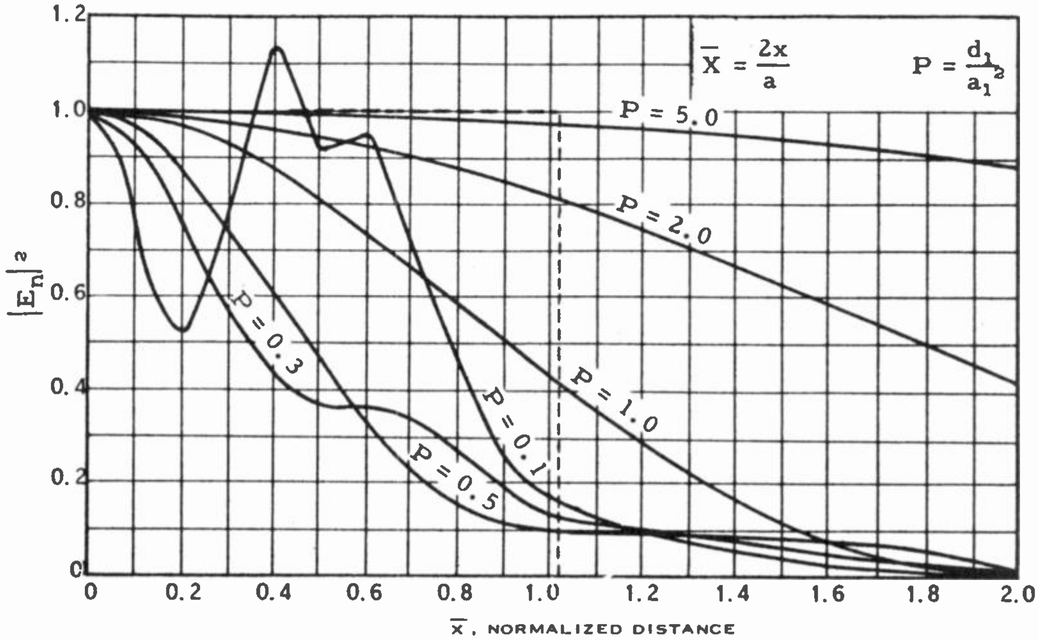


Figure 4-25. Fresnel Region Patterns for Rectangular Aperture with Uniform Illumination

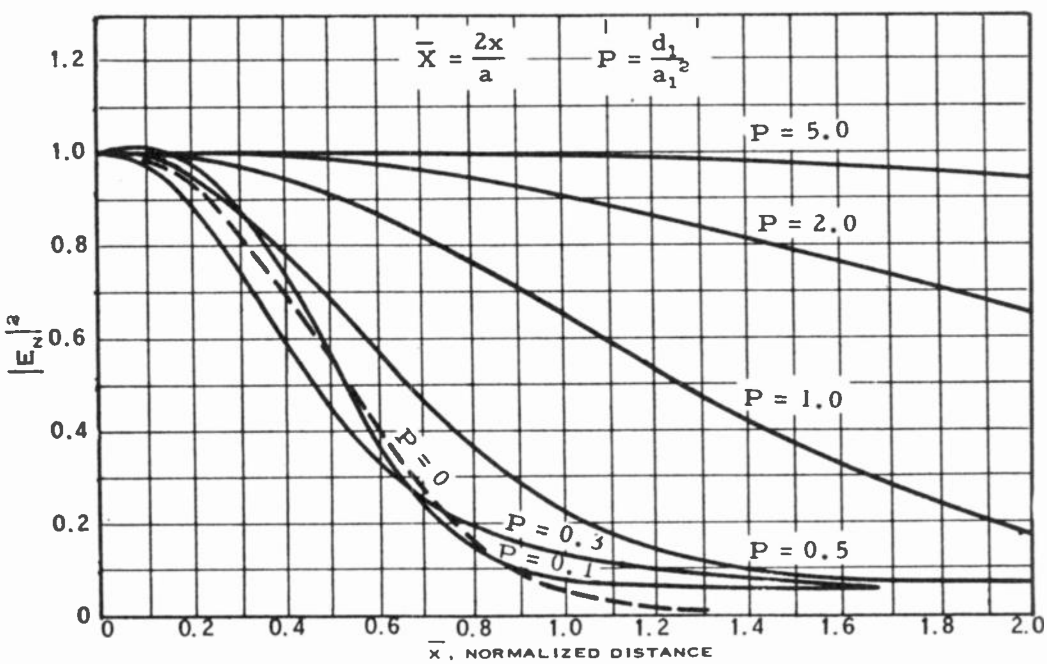


Figure 4-26. Fresnel Region Patterns for Rectangular Aperture with Cosine Illumination

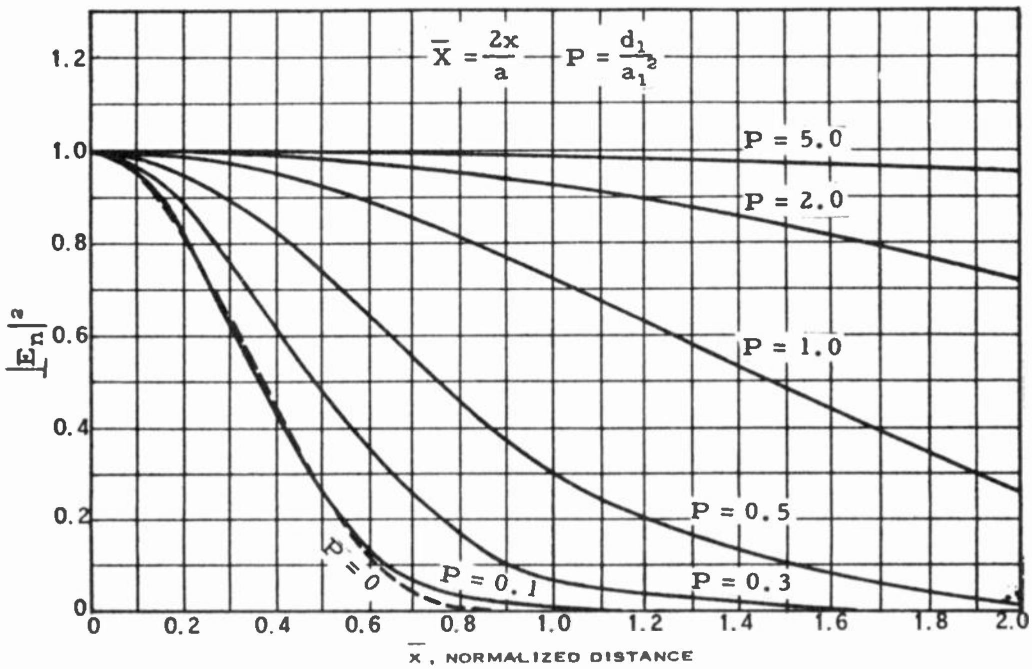


Figure 4-27. Fresnel Region Patterns for Rectangular Aperture with Cosine Squared Illumination

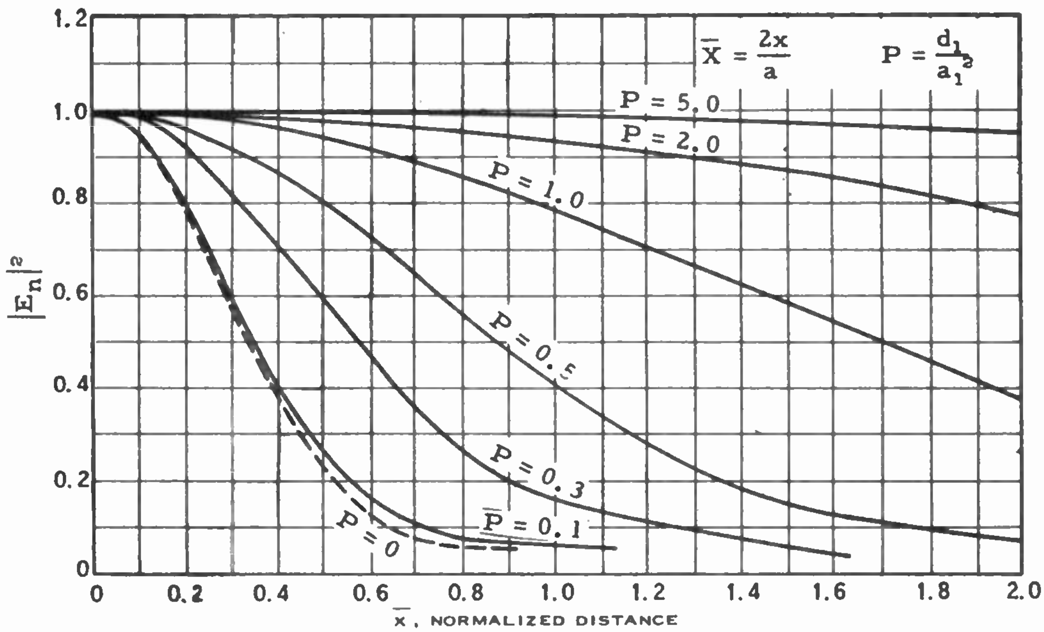


Figure 4-28. Fresnel Region Patterns for Rectangular Aperture with Cosine Cubed Illumination

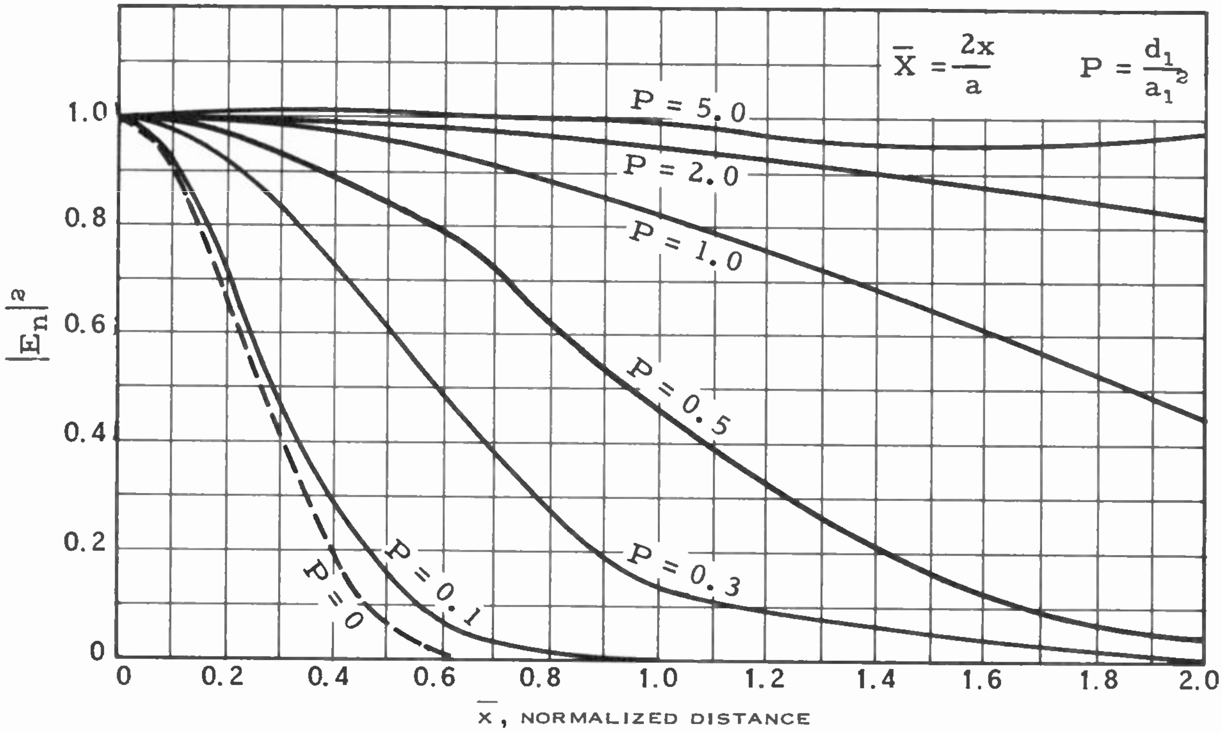


Figure 4-29. Fresnel Region Patterns for Rectangular Aperture with Cosine Fourth Illumination

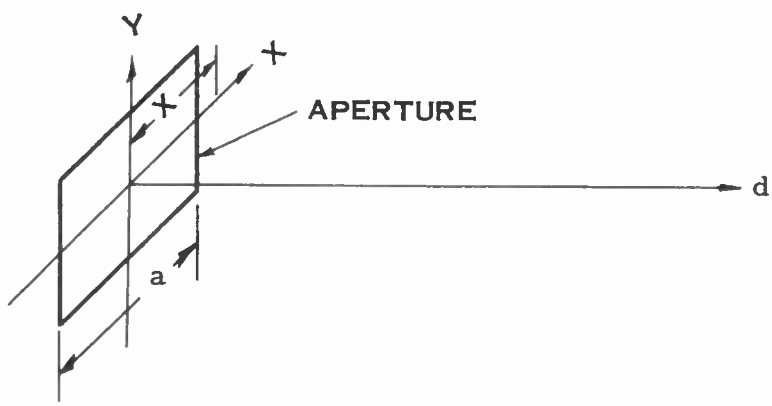


Figure 4-30. A Definition of Symbols for Fresnel Region Figures

specifying the aperture distribution, the field distribution is implied. Therefore, the theoretical power density at the aperture is the square of the specified aperture distribution.

All the pattern figures are normalized to the same amplitude of one at the center of the antenna. The relative amplitude of the patterns can be obtained from the corresponding gain correction curves.

1.5.3 POWER TRANSFER BETWEEN TWO ANTENNAS (FRESNEL REGION)

When d, the distance apart between two antennas is $\geq \frac{L_1^2}{\lambda}$, the power transfer is given by the Friis transmission formula.

$$\frac{P_R}{P_T} = \frac{G_{TO} G_{RO} \lambda^2}{16\pi^2 d^2} \tag{4-37}$$

where:

G_{TO} = Far-Field gain of transmitter

G_{RO} = Far-Field gain of receiver

d = Distance apart

L_1 = Largest linear dimensions of larger antenna

L_2 = Largest linear dimension of smaller antenna

By making a slight modification, the Friis transmission formula can be extended to include the Fresnel region.

$$\frac{P_R}{P_T} = \frac{G_L G_S \lambda^2}{16\pi^2 d^2} \tag{4-38}$$

where:

G_L = gain of larger antenna at distance d

G_S = Fraunhofer gain of smaller antenna

d = distance apart

In the Frauhofer region $\left(d \geq \frac{L_1^2}{\lambda}\right)$ where Equation (4-37) is valid, it is equivalent to Equation (4-38) and thus there is no change in results. However, Equation (4-38) is also correct when one antenna is operating in its Fresnel region; i.e., the restriction now becomes $\left(d \geq \frac{L_2^2}{\lambda}\right)$. When both antennas are operating in the Fresnel region $\left(\lambda < d < \frac{L_2^2}{\lambda}\right)$ Equation (4-38) can be used as an approximation to predict the order of magnitude of the power transfer.

It must be remembered that these data can only be used in situations involving rectangular apertures of uniform, cosine, cosine square, cosine cubed and cosine fourth amplitude distribution. All cases are for antenna with uniform phase distribution. It is believed that, at the present state of the art, a large number of the rectangular antennas can be approximated by one of the above illuminations.

This method gives a good approximation of Fresnel region power transfer. It should only be used when the antennas under question are of the same polarization and directed toward each other.

1.6 TRANSMISSION LINE LOSSES

When an antenna is used as a transmitting device, the fact that its impedance (Z_a) does not match the characteristic impedance (Z_o) of the transmission line feeding it, causes standing waves on the line. Because the standing waves cause current nodes to exist, the power dissipation at these points is increased above what it would normally be on a perfectly matched line. To calculate the total line loss under these conditions, it is necessary to add to the normal loss for a matched line the additional loss due to the standing waves. The extra loss is found in the following way; first, the attenuation for a perfectly matched line of the given length and type is found. If coaxial cable is employed, the plots in Figure 4-31 can be used, and if waveguide is employed, the table in Figure 4-33 can be used to find the matched attenuation as a function of frequency. Second, when this value is determined, the graphs of Figure 4-34 are used to find the additional loss due to the voltage standing wave ratio (VSWR). To use these plots, it is necessary to know the VSWR as well as the normal attenuation. The VSWR can be determined in many ways but one way uses the relation:

$$\text{VSWR} = \frac{1 + \left|\dfrac{Z_a - Z_o}{Z_a + Z_o}\right|}{1 - \left|\dfrac{Z_a - Z_o}{Z_a + Z_o}\right|} \qquad (4\text{-}39)$$

The standard Smith chart is also a very convenient way of determining VSWR. A copy of this chart can be found in Electronics, January 1939 and January 1944.

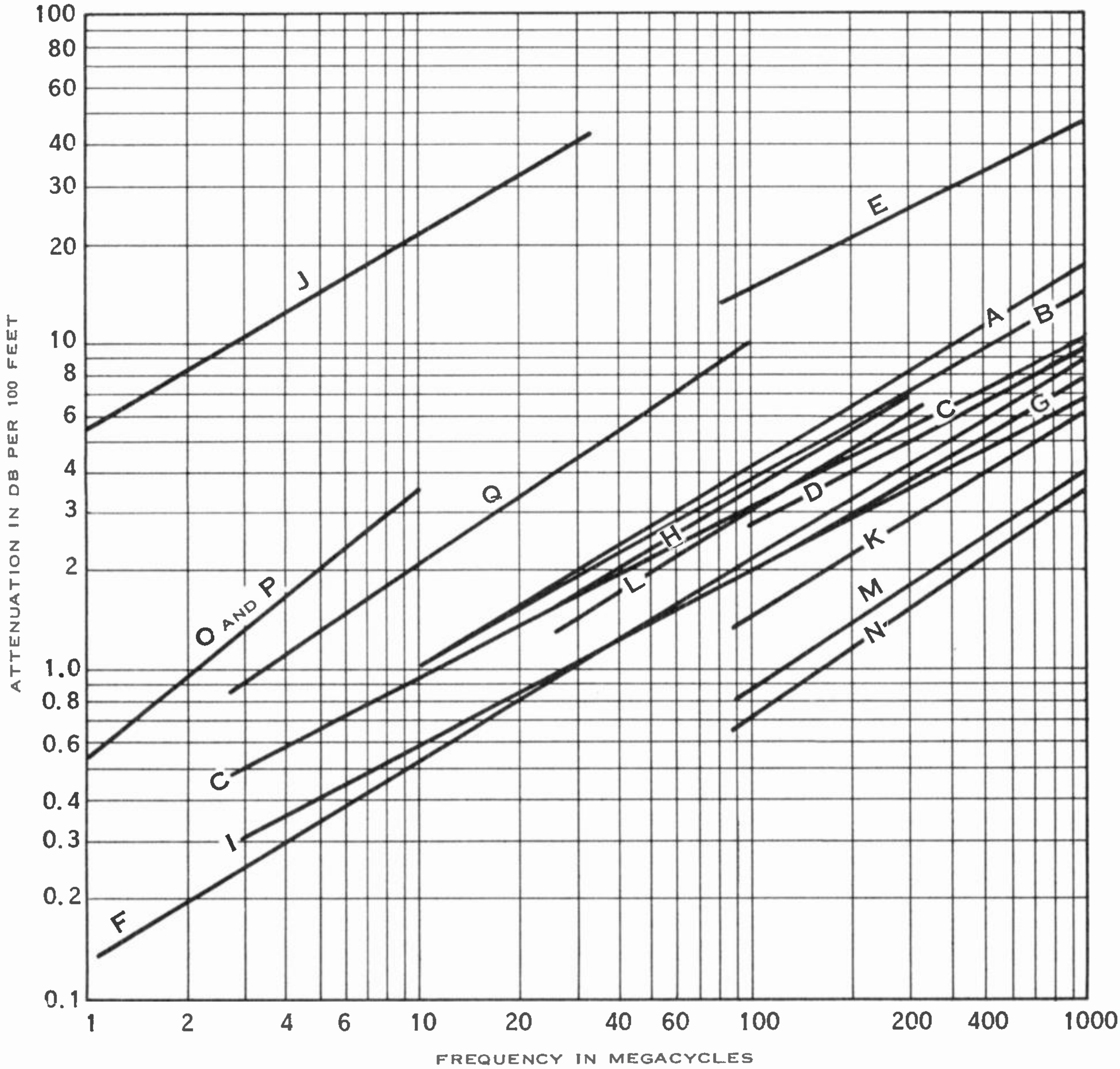


Figure 4-31. Attenuation of Standard RF Cables vs. Frequency
(See Figure 4-32 for identification of curves.)

Using the graphs in Figure 4-34, the additional line loss is found as a function of the normal line loss and the VSWR. The total attenuation is then the normal loss plus the additional loss. These factors are items b and c in Paragraph 1.3.1.

CURVE	RG-( )/U	CURVE	RG-( )/U
A	55	I	63
A	58	J	65
B	59	K	14
C	62	K	74
C	71	L	57
D	5	M	17
D	6	M	18
E	21	N	19
F	8	N	20
F	9	O	25
F	10	O	26
G	11	O	64
G	12	P	27
G	13	P	28
H	22	Q	4

Figure 4-32. Identification of Curves in Figure 4-31

When an antenna is used as a receiving device, the situation is slightly more complicated than for the transmitting case because reflections normally occur at two points in the system. First power is reradiated from the antenna due to the mismatch between the antenna impedance and the characteristic impedance of the transmission line. The graph in Figure 4-35 shows this loss in db below a perfectly matched antenna when the VSWR of the antenna is known. (This is the VSWR that would result on the transmission line if the receiving system were treated as a transmitting system.) Second, power is reflected from the receiver terminals due to the mismatch between the receiver impedance and the characteristic impedance of the transmission line. This causes a situation similar to the case of the transmitting antenna and the calculations are carried out in the same manner. The standing wave on the receiver transmission line caused by the receiver, causes a loss above the normal line attenuation. This loss is found by using Figures 4-31 and 4-33 as in

FREQUENCY (GC) TE_{10} MODE	EIA DESIGNATION WR( )	THEORETICAL ATTENUATION: LOWEST TO HIGHEST FREQUENCY (DB/100 FT.)		
		BRASS	ALUMINUM	SILVER
.320 - .490	2300		0.039-0.027	
.350 - .530	2100		0.046-0.031	
.410 - .625	1800		0.056-0.038	
.490 - .750	1500		0.069-0.050	
.640 - .960	1150		0.128-0.075	
.750 - 1.120	975		0.137-0.095	
.950 - 1.500	770		0.201-0.136	
1.120 - 1.700	650	0.424-0.284	0.269-0.178	
1.450 - 2.200	510	0.606-0.398	0.388-0.255	
1.700 - 2.600	430	0.788-0.516	0.501-0.330	
2.200 - 3.300	340	1.046-0.728	0.669-0.466	
2.600 - 3.950	284	1.478-1.008	0.940-0.641	
3.300 - 4.900	229	1.862-1.320	1.192-0.845	
3.950 - 5.850	187	2.79 - 1.93	1.77 - 1.22	
4.900 - 7.050	159	2.89 - 2.24	1.84 - 1.42	
5.850 - 8.200	137	3.85 - 3.08	2.45 - 1.94	
7.050 - 10.00	112	5.51 - 4.31	3.50 - 2.74	
8.200 - 12.40	90	8.64 - 6.02	5.49 - 3.83	
10.00 - 15.00	75	10.07-7.03	6.45 - 4.50	
12.40 - 18.00	62	12.76-11.15	8.13 - 7.10	6.14-5.36
15.00 - 22.00	51	17.3 - 12.6	11.05-8.05	8.37-6.10
18.00 - 26.50	42	27.7 - 19.8	17.6 - 12.6	13.3-9.5
22.00 - 33.00	34	33.3 - 23.1	21.3 - 14.8	16.1-11.2
26.50 - 40.00	28			21.9-15.0
33.00 - 50.00	22			31.0-20.9
40.00 - 60.00	19			38.8-27.2
50.00 - 75.00	15			52.9-39.1
60.00 - 90.00	12			93.3-52.2
75.00 - 110.00	10			100 - 70.4
90.00 - 140.00	138*			152 - 99
110.00-170.00	136*			163 - 137
140.00-220.00	135*			308 - 193
170.00-260.00	137*			384 - 254
220.00-325.00	139*			512 - 348

* Jan Type RG-( )/U

Figure 4-33. Normal Waveguide Loss

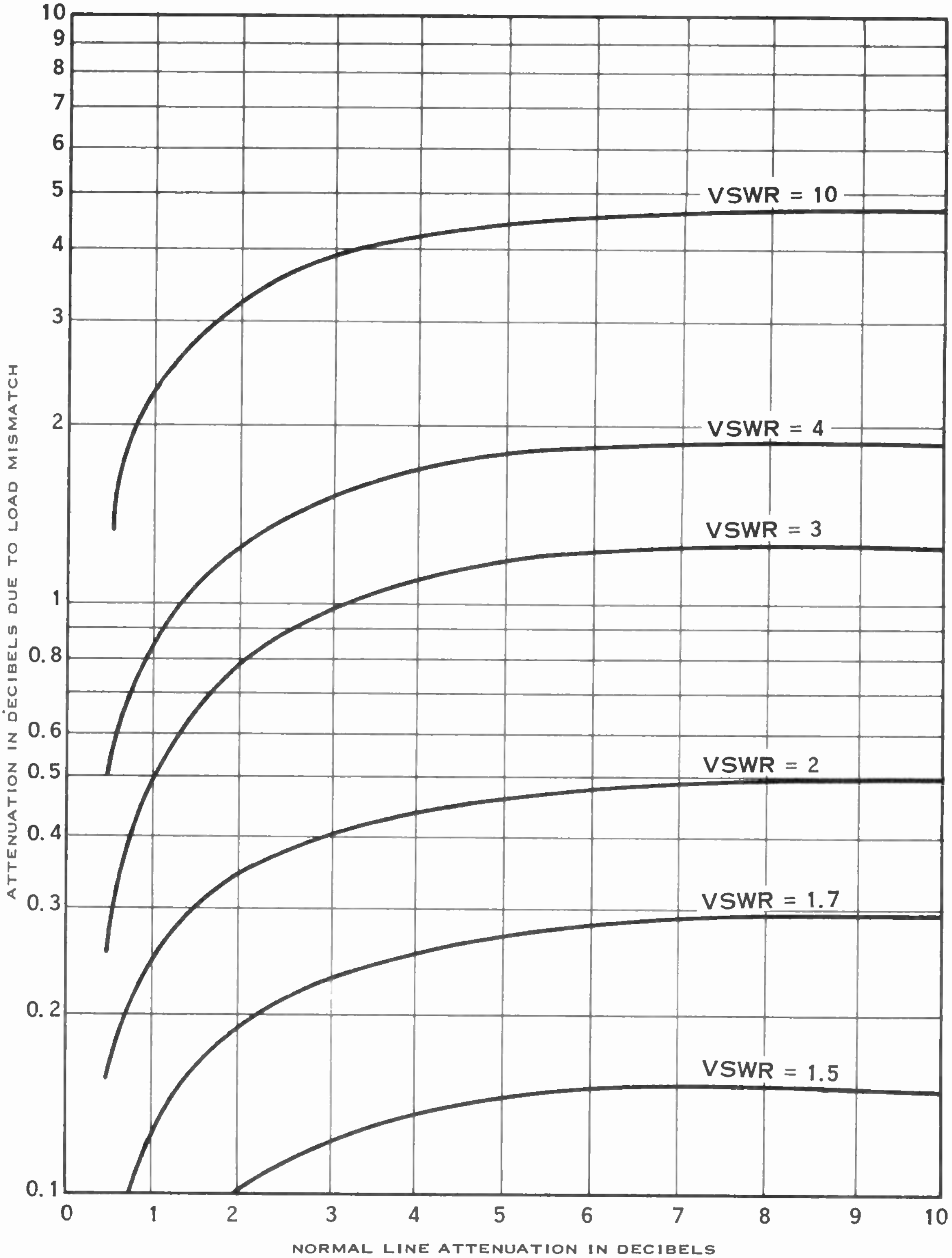


Figure 4-34. Transmission Line Attenuation Due to Load Mis-Match

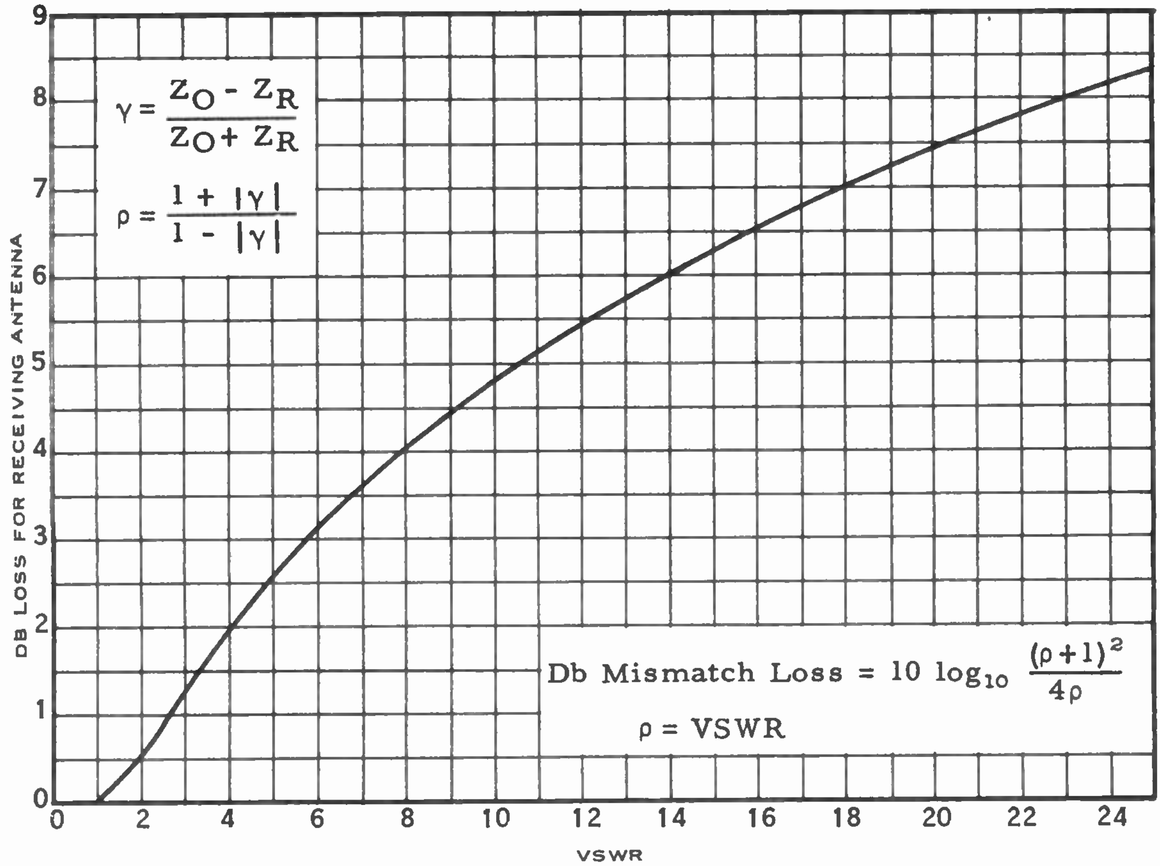


Figure 4-35. Receiving Antenna Mismatch Loss vs. VSWR

the transmitting case and Figure 4-34. The total loss in a receiving antenna-transmission line-receiver system is then (1) the antenna mismatch loss from Figure 4-35 plus the normal line loss from Figures 4-31 or 4-33 and the additional line loss due to the VSWR from Figure 4-34. These factors are referred to as items "g" and "h" of Paragraph 1. 3. 1.

2. SITING CONSIDERATIONS

In many installations, it is necessary to locate several potentially interfering systems in close proximity so it is necessary to consider the effects of natural and man-made objects in the immediate vicinity of the antenna. If a choice is possible in the location of an antenna which has not yet been installed, then especial attention should be given to siting considerations not only in regard to potential interference to and from other systems, but also in regard to the optimum performance of the system at its design frequency. Locating an antenna on a hill which slopes downward toward the area of potential RFI usually increases the received power.[8] The magnitude of this increase can be estimated by assuming that the effective antenna height is the difference in elevation between the antenna and the bottom of the hill, providing that first-Fresnel-zone clearance is obtained over the immediate foreground. At 30 megacycles this means a clearance of about 20 feet at a distance of 20 feet, 40 feet at a distance of 100 feet, 90 feet at a distance of 500 feet, etc. The required clearance decreases as the square root of the wavelength, and may be obtained from Figure 4-15. When these clearances are not met, it is convenient to assume that the effective antenna height is the difference in elevation between the antenna and the point where the actual profile intercepts the curve of required clearance (first Fresnel zone).

2.1 EFFECT OF MAN-MADE OBJECTS[8]

Built-up areas have little effect on radio transmission at frequencies below a few megacycles, since the size of any obstruction is usually small compared with the wavelength, and the shadows caused by steel buildings and bridges are not noticeable except immediately behind these obstructions. However, at 30 megacycles and above, the absorption and reflection of a radio wave in going over an obstruction is not negligible, and both types of losses tend to increase as the frequency increases. The attenuation through a brick wall, for example, may vary from 2 to 5 db at 30 megacycles and from 10 to 40 decibels at 3000 megacycles, depending on whether the wall is dry or wet. Consequently, most buildings are rather opaque at frequencies of the order of thousands of megacycles. Shadow losses at street level in the downtown area of large cities may be of the order of 30 db or more at frequencies in the 30 to 150 megacycle range, and the received power may vary 15 to 20 db within a few feet because of wave interference caused by multi-path transmission. As the frequency increases, the number of possible multiple paths also increases, so that there is some tendency to fill in the deep shadow regions. This means that the average shadow loss at street level may not increase as rapidly with frequency as the shadow loss behind an isolated ridge.

2.2 EFFECT OF NATURAL OBSTRUCTIONS[8]

When an antenna is surrounded by moderately thick trees and is below tree-top level, the average loss at 30 megacycles resulting from the trees is usually 2 or 3 decibels for vertical polarization and is negligible with horizontal polarization. However, large and rapid variations in the received field intensity may exist within a small area, resulting from the standing-wave pattern set-up by reflections from trees located at a distance of as much as 100 feet or more from the antenna. At 100 megacycles, the average loss from surrounding trees may be 5 to 10 decibels for vertical polarization and 2 or 3 decibels for horizontal polarization. The tree losses continue to increase as the frequency increases, and above 300 to 500 megacycles they tend to be independent of the type of polarization. Above 1000 megacycles trees that are thick enough to block vision present an almost solid obstruction, and the diffraction loss over or around these obstructions can be obtained from Figure 4-15.

3. NEAR FIELD RADIATION

The term "radiation," technically speaking, is used to describe the phenomenon of electromagnetic waves spreading out in space from a source, according to the laws of wave propagation. Nevertheless, the term "radiated noise" has become so commonly used to mean any interfering signal detected through the medium of an electric or magnetic field, that it would not be possible to restrict the words "radiation" and "radiated" to their technical meaning without rewriting much of the literature on radio interference and interference measurements. It is clear that the transmission of interfering signals through mutual inductances and mutual capacitances, as discussed in Chapter 2. Paragraphs 2. 1 and 2. 2 would be classified as radiated interference in the loose sense of the word.

Examples of true radiation in the sense defined above, within any vehicle are extremely rare. An example of true radiation outside a vehicle, would be the transmission of an interfering signal from a transmitting antenna to a receiving antenna mounted several wavelengths away. When the receiving antenna picks up an interfering signal which is apparently radiated from a nearby opening in the skin of the vehicle, the transmission is usually not through true radiation but through capacitive coupling from the antenna to the outside surface of the vehicle, which has acquired a charge due to currents flowing around the edges of the opening.

Radiation in the strict sense is not amenable to treatment with the methods of ordinary circuit analysis because radiation is caused entirely by the effects of retardation, the phenomenon specifically neglected in the transition from field to circuit concepts. If an electromagnetic disturbance could be propagated in space with infinite velocity, the fields that give rise to the effects of mutual inductance and capacitance would be unchanged, but the radiation field would be absent, a fact that can easily be demonstrated mathematically.

3.1 CALCULATION OF NEAR FIELD RADIATION FOR VEHICLE SYSTEMS

The mathematical expression for the total fields in the vicinity of a conductor carrying time-varying currents is very complicated and cannot, in general, be found analytically, except in very simple cases. The cases where it is possible to find such expressions are the ones involving very simple geometric arrangements, such as a thin straight wire or a thin circular loop. It can be shown, however, that the mathematical expression for the field due to any source can be developed into a series of terms, each of which represents the field due to certain simple arrangements of conductors. The first term in this series gives the field due to an electric dipole; the second, that due to a magnetic dipole and an electric quadrupole, etc. This expansion is useful only if the mathematical series converges rapidly, so that the first few terms give a good approximation to the total field. The rapidity of convergence depends on the ratio of r, the distance of the point of observation to the center of the source, to a, the radius of the smallest sphere that can be placed so as to enclose all parts of the source. (See Figure 4-36a.) When this ratio is very large, only the first term in the expansion is important (unless this term happens to be zero, in which case, only the first nonvanishing term is important). The result is dipole radiation. When this ratio is less than unity, the expansion does not converge at all. This possibility of "multiple expansion" explains why the consideration of very simple, but nonpractical cases nevertheless, has much significance for the analysis of the more complicated, practical cases.

The two simplest cases are those of an oscillating electric and magnetic dipole. The electric dipole is represented physically by a very short length of very thin wire carrying a uniform current that varies sinusoidally in time. The magnetic dipole is represented physically by a very small circular loop carrying a uniform current that varies sinusoidally in time. For the electric dipole, the components of the fields surrounding the conductor of length L are given in spherical coordinates by the following expressions, assuming $L<r$ and $L<<\lambda$:

$$E_\theta = \frac{LI\sin\theta}{4\pi\epsilon}\left[\frac{\sin\omega(t-r/v)}{r^3\omega} + \frac{\cos\omega(t-r/v)}{r^2 v} - \frac{\omega\sin\omega(t-r/v)}{rv^2}\right] \quad (4\text{-}40)$$

$$E_\varphi = 0 \quad (4\text{-}41)$$

$$E_r = \frac{2LI\cos\theta}{4\pi\epsilon}\left[\frac{\sin\omega(t-r/v)}{r^3\omega} + \frac{\cos\omega(t-r/v)}{r^2 v}\right] \quad (4\text{-}42)$$

$$H_\theta = 0 \quad (4\text{-}43)$$

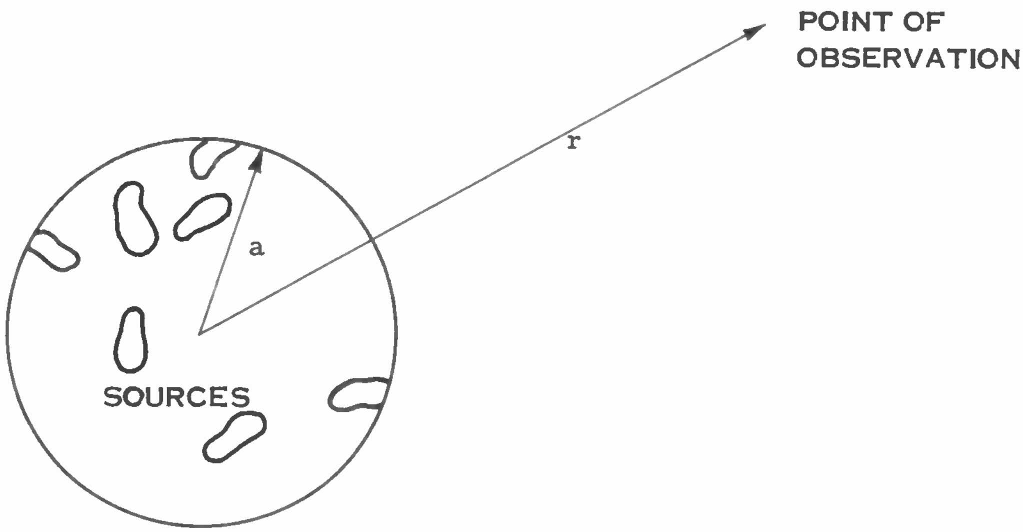


Figure 4-36a. Quantities of Length in Radiation Problems

$$H_\varphi = \frac{L I \sin\theta}{4\pi}\left[\frac{\cos\omega(t - r/v)}{r^2} - \frac{\omega \sin\omega(t - r/v)}{rv}\right] \qquad (4\text{-}44)$$

$$H_r = 0 \qquad (4\text{-}45)$$

where $v = 3 \times 10^8$ meters per second is the velocity of electromagnetic waves in free space, $\epsilon = 8.854 \times 10^{12}$ farads per meter is the absolute permittivity of free space, and the other symbols are explained in Figure 4-36b. In the case of the magnetic dipole, the components of the fields surrounding the small loop of area A and carrying a current $I \cos\omega t$, assumed to be at the origin in the x-y plane, are given by the following expressions:

$$E_\theta = 0 \qquad (4\text{-}46)$$

$$E_\varphi = \frac{I A \sin\theta}{4\pi}\,\omega\mu\left[\frac{\sin\omega(t - r/v)}{r^2} + \frac{\omega\cos\omega(t - r/v)}{rv}\right] \qquad (4\text{-}47)$$

$$E_r = 0 \qquad (4\text{-}48)$$

$$H_\theta = \frac{I A \sin\theta}{4\pi}\,\omega\left[\frac{\cos\omega(t - r/v)}{r^3\,\omega} + \frac{\sin\omega(t - r/v)}{r^2 v} - \frac{\omega\cos\omega(t - r/v)}{rv^2}\right] \qquad (4\text{-}49)$$

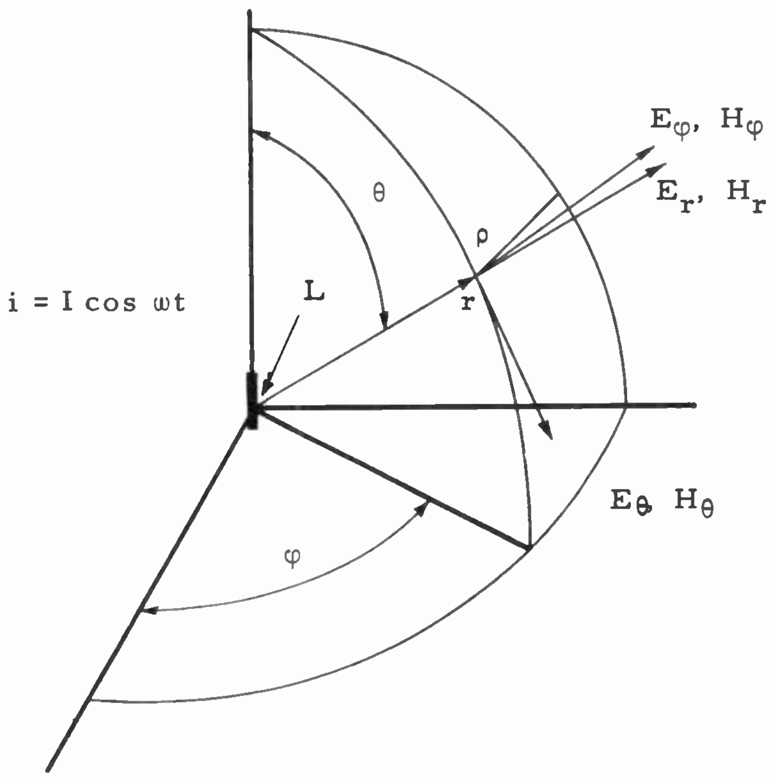


Figure 4-36b. Dipole Radiation

$$H_\varphi = 0 \tag{4-50}$$

$$H_r = \frac{2IA\cos\theta}{4\pi}\,\omega\left[\frac{\cos\omega(t-r/v)}{r^3\omega} - \frac{\sin\omega(t-r/v)}{r^2 v}\right] \tag{4-51}$$

It is seen that in the above equations there are three kinds of terms which differ in their variation with r. The terms that vary as $1/r^3$ represent an electrostatic dipole. The terms that vary as $1/r^2$ represent what is called the induction field. Finally, the terms that vary as $1/r$ represent what is called the radiation field. The relative importance of the various kinds of terms depends on the ratio of r to λ, where $\lambda = 2\pi v/\omega$ is the wavelength of the radiation. When r/λ is much smaller than unity, the static and induction terms are negligible. When $r = \lambda/6$, approximately, the induction and radiation fields are equal.

Comparison of the case of the magnetic dipole (current loop) with that of the electric dipole shows that the structures of the two fields are the same except that the poles of the electric and magnetic fields are reversed.

Based on the structure of the electromagnetic field in the vicinity of an oscillating dipole, the statement is sometimes made that for any radiating system there is an induction field, varying as $1/r^2$, and a radiation field, varying as $1/r$. This statement, however, is misleading. Although it may be true if properly applied, generally it is false when applied to most problems arising in the propagation of interfering signals within a vehicle.

To determine the correct applications of this statement, a clear distinction must be made between the three quantities of length entering every problem of this kind, illustrated in Figure 4-36a. These are: (1) r, the distance to the point of observation from the center of the source; (2) a, the radius of the smallest sphere that can enclose all of the source; and (3) λ, the wavelength. This last quantity, the wavelength, does not have a unique meaning unless the time variation is sinusoidal. For some other variation, say an interference pulse, each frequency component must be considered separately.

The point that is frequently forgotten is that the validity of the statement made above depends entirely on the relation between r and a, not r and λ. The statement holds whenever $r \gg a$, and the configuration of the source is such that there is dipole radiation. In practice, neither condition is ever satisfied within a vehicle except, possibly, at ultra-high and super-high frequencies. The first condition fails because the dimensions of the potential radiators (the vehicle structure itself is an important example) are usually of the same order of magnitude as the distances between the radiator and the receiver or pick-up lead, due to the proximity of electrical equipment in the vehicle. The second condition fails because there are present within the vehicle so many absorbers, reflectors, and systems capable of reradiation, that the resultant field at any point will consist of several waves of different amplitudes and phases superimposed, and the laws of simple dipole radiation would never apply.

It must also be remembered that the separation of the total field into an induction and a radiation field is by no means physical, but entirely analytical. Any possible physical measurement will always yield a measure of the total field, not its analytical components. When the basic conditions mentioned above are not fulfilled, even the analytical separation is no longer possible. The total field must then be considered and its law of variation with distance will, in general, be very complicated. It cannot even be said that the field will always decrease as the distance from the prime radiator increases. Because of the possibilities of reflection and reradiation, standing waves may be set up and the field may actually increase in certain regions, as the distance from the prime source is increased, a condition which is called resonance excitation.

In general, the conditions in a vehicle are almost always such that the fields of interest are those observed very close to the source. In the frequency range from about 30 to 300 megacycles per second, the distances of interest are of the same order of magnitude as a wavelength. Even if there were true dipole radiation, no approximation could be made in this region and the total field would have to be considered. The actual radiation is likely to be much more complicated than dipole radiation, and such simple approximations will be even less valid.

True radiation is more likely to be important at frequencies above 300 megacycles per second. The rather thorough treatment of radiation in this paragraph is justified by the increased use of the ultra-high frequency and super-high frequency regions in many vehicles. Moreover, most of the above statements about the relative unimportance of true radiation were based on the fact that, in order for any body to be an efficient radiator, its dimensions must be at least of the order of magnitude of a wavelength. But because of the extremely high sensitivity of modern receivers, even a very inefficient radiator could produce a troublesome radiation field at the receiver.

In considering the interfering fields, both close to and far from the source, it is sometimes important to know the significance of the electric field, relative to that of the magnetic field. Confusion may arise from the fact that, according to Maxwell's equations: a varying electric field is always accompanied by a magnetic field; a varying magnetic field is always accompanied by an electric field; and the ratio of the two, called the impedance of the medium, is a constant (377 ohms for air) for a plane wave. A careless interpretation of these facts leads to the conclusion that it makes no difference which is considered for all possible conditions, since the two always occur together and bear a fixed ratio to each other. This reasoning is in error, because what exists within a vehicle is usually not a plane wave. The impedance for more complicated types of waves, such as cylindrical or spherical waves or combinations of these, is not, in general, equal to 377 ohms. It may take on almost any value, and often varies rapidly from point to point in the vicinity of the source. A very small impedance means that the electric field is small as compared to the magnetic field. A very large impedance means that the magnetic field is small as compared to the electric field. Since some receivers are more sensitive to one type of field than the other, a knowledge of the impedance would be of great practical value in many cases. Unfortunately, the evaluation of the impedance is practically impossible for the complicated field configurations encountered in the interior of modern vehicles.

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RADIATION HAZARDS TO PERSONNEL AND EQUIPMENT — CHAPTER 5

1. GENERAL ASPECTS OF THE RF HAZARD PROBLEM

A relatively new problem to radio frequency technology has arisen in the last few years. This is the problem of hazards created by the new high powered electromagnetic devices operating at radio frequencies. It fits logically into a definitive study of radio frequency interference problems for two reasons. First, it represents in its most elemental forms RFI on a scale so magnified that the effects may become irreversible. When an interfering source leaves the air, an equipment which has previously been totally unusable because of RFI regains its full capability in a short time and no permanent damage has been done; however, when a hazardous RF field exists, damage may be done which is not self-correcting when the field is removed. It might be said in a broad sense that the greatest distinguishing feature between RFI and RF hazards is the presence of irreversible effects.

A second reason that RF hazards are important to the person engaged in RFI work is that measurement personnel frequently must work in the immediate vicinity of high powered radiating systems and in order to insure their safety, the RF hazard situation must be determined at their operating stations.

1.1 DEFINITION OF AN RF HAZARD

An RF hazard is said to exist in a given area when there is sufficient electromagnetic energy present to produce a physical effect such that a dangerous or destructive reaction is highly probable. This definition has purposely been made broad enough to include all the known dangerous effects as well as any which may be discovered. The known hazards include dangers to personnel, ordnance, and fuel. Dangers to other devices as the result of internal heating or explosive burning must also be considered. There may be unknown effects which involve danger or destruction by other mechanisms not yet discovered. Attention should always be directed to this possibility.

1.2 FIELDS EFFECTIVE IN CREATING AN RF HAZARD

An RF hazard is usually caused by one of two effects: (1) heating from RF currents caused by the magnetic component of the field, and (2) sparking due to the electric component of the field. The H field is the oscillating magnetic field which generates circulating currents in objects or loops. The E field is the oscillating electric field which is perceived when arcs are drawn from an object in an RF field.

As was discussed in Chapter 4, Section 3, the relation

$$E = 377\ H$$

only holds true in free space, and therefore in areas of RF hazards it may be quite different. Thus, in some cases the E field may be very low compared to the H field, and in others the H field may be very low compared to the E field. If a particular configuration presents a high impedance to the field, then the E field will predominate and the primary effect of the field will be to cause sparking when objects in the field engage or disengage. If a particular configuration presents a low impedance to the field, then the H field will predominate and the primary effect of the field will be to cause circulating currents which may result in heating.

1.3 FREQUENCIES EFFECTIVE IN CREATING AN RF HAZARD

Most RF hazards discovered to date are caused by simple ohmic heating due to the reaction of electromagnetic energy and a lossy material. This implies that the frequency of the electromagnetic energy which can cause a hazard is not bounded. Any frequency which is well enough matched to the lossy material to cause appreciable heating must be considered as a possible hazard. In any natural environment, the lossy material will not be subject to just a single frequency but to frequencies covering the entire spectrum in use. This is especially true in military installations where many strong emitters are present. The heating effect which occurs, unfortunately, cannot usually be considered as the result of the sum of the individual effects from each frequency but must be considered on the basis of the time-varying vector sum of all the frequencies present. Obviously, in most cases, this cannot be done and, therefore, in the practical case, approximations must be made.

If the various frequency components at a given location can be assumed to be uncorrelated, then the net electromagnetic field can be said to resemble random noise, and the total power is essentially the sum of the powers at the individual frequencies. If some correlation is likely to exist, such as with synchronized radars on several frequencies, then the vector addition of the fields is necessary.

2. HAZARDS TO PERSONNEL

2.1 GENERAL EFFECTS

Most of the experimental work to date supports the belief that the chief effect of radio frequency energy on living tissue is to produce heating. Consequently, exposure should probably represent no hazard unless overheating can occur. Within carefully prescribed limits, the heating effect of radio energy may actually be beneficial as witness the use of diathermy which has long been employed therapeutically.

Heating is a function of the strength of the electromagnetic energy; i.e., the average power flow per unit area (milliwatts per square centimeter). It is also a function of time. The heating may take place near the surface of the body or deep within it. The depth of penetration is related to the frequency of the energy with the lower frequencies (200-900 mc) penetrating more deeply than higher frequencies (1500-11,000 mc). Frequencies used by most high powered radars usually produce heating at the surface of the skin.

The effects of body heating depending on whether the frequency is such that surface or interior heating occurs are: (1) a general rise in body temperature, similar to fever, or (2) something more localized akin to the cooking process in a radar oven. The human body can compensate for a certain amount of heating of the first type through perspiration, if the temperature rise is not too sudden. Consequently, the hazard may be somewhat less in cool weather than on an extremely hot day when the body's cooling mechanism is already working at full capacity. Compensating mechanisms for coping with the second type of heating are less adequate.

Circulating blood acts as a coolant, so that localized heating is least serious in parts such as muscle tissue, which are well equipped with blood vessels. Heating is more of a danger to the brain, the testes, and the hollow viscera. The most widespread publicity has related to the effect of electromagnetic energy on the eyes. The viscous material within the eyeball is affected by heat in much the same manner as the white of an egg, which is transparent at room temperature, but becomes opaque white when warmed slightly. In the eyes, as in the egg white, the process is irreversible.

As the surface of the human body is more generously supplied with sensory nerves than the interior, a feeling of warmth may give a warning in case of over-exposure of frequencies which produce surface heating. If the frequencies are such as to cause a general rise in body temperature, the resulting sensation of discomfort may or may not be perceived in time to provide adequate warning. In the case of localized heating deep within the body, it is still less likely that any warning sensations would be noted before damage was done. Hence, it is important to establish limits and to delineate the areas in which a potential health hazard could exist.

2.2 FREQUENCIES HAZARDOUS TO HUMANS

Electromagnetic energy hazards may exist at any radio frequency capable of being absorbed by the body. Some differentiation should, of course, be made between what normally is referred to as electromagnetic radiation (up to infrared) and all frequencies above this which, while technically electromagnetic radiation, have effects quite different from the lower frequency energy. For example, X-rays are a form of electromagnetic radiation and are very dangerous to humans. Exposure to X-rays may occur if a person opens the shield around a transmitter tube in a high power radar, or if a person is present where there are leaks in the shielding. Ordinarily, however, X-rays are not normally radiated from C-E antenna systems. If X-rays should result from the interaction of high-powered RF energy and appropriate materials then they would be considered a secondary result of the hazard presented by the RF energy.

The cumulative effect of all the energy from all the frequencies present must be considered as discussed in Section 1.3. At the present time, it is not possible to forecast the effects upon personnel of energy at one frequency and power density based upon investigations conducted at a greatly different frequency.

2.3 POWER LEVELS HAZARDOUS TO HUMANS

Sufficient factual data is not, at present, available to determine the safe exposure level for each frequency. Thus, in the interim, one level believed safe for all frequencies, has been chosen. Past research has indicated that a power density level of 0.2 watts/cm^2 was required to produce damage. The accuracy of the methods and instrumentation used was somewhat questionable and there was some evidence that damage may have been produced at 0.1 watts/cm^2. These tests concerned only a single frequency and did not account for the cumulative effect that might result from the effects of many sources of electromagnetic energy being superimposed on the human body. Since it is impractical to measure the power density at each of these frequencies separately, a safety factor of 10 has been selected and so the presently accepted maximum safe level of electromagnetic energy to which the human body should be exposed at any frequency is taken to be 0.01 watts/cm^2.

This level of 0.01 watts/cm^2 is an average power level and not peak power, since available data indicates the only detrimental effects are thermal in nature and these depend only on average levels. Sufficient data has not been collected at present to indicate a relationship between length of exposure and power density. The present level is the maximum for either continuous or intermittent exposure and precautions should be taken to avoid personnel exposure to ambient power levels greater than this for any period of time.

2.4 HAZARDOUS AREAS

Both radar and communication systems may contain hazardous power levels of electromagnetic energy. High powered long range search radars and tropospheric scatter communication systems are examples. Areas within the Fresnel region of the transmitting antennas used with these systems are usually hazardous. The inside of waveguides used with these systems or any location near an open waveguide of a high powered system must be considered hazardous. The tops of buildings or other tall structures near a scanning high powered radar require special precautions if the main beam of the radar is set near zero degrees. This is primarily because the radar may be pointed in another direction and be stationary when the area is entered and it is, therefore, not a dangerous area at that time. Later, however, the antenna may begin scanning or may come to rest on the area in question, thereby making it a very dangerous area. Fences of open mesh wire or insulating material may be placed around areas that are normally dangerous with appropriate warning signs to prevent personnel from entering these areas.

2.5 NON-THERMAL EFFECTS OF VERY HIGH ENERGY ELECTRO-MAGNETIC FIELDS ON THE HUMAN BODY

Experimental work has indicated that very high electromagnetic energy intensities may cause aberrations in chromosomes, may affect micro-organisms and may cause certain changes in blood cells. In addition, it also appears that there may be a possible effect from pulsed power which is different than the heating effect of average power. There is insufficient information on these matters at present to provide useful conclusions.

3. HAZARDS TO ORDNANCE

For many years there has existed a concern among the handlers and users of explosives that electromagnetic energy from radio and radar transmissions might cause accidental explosions. Signs along the highway near blasting operations warn of this danger. It was logical that this concern should extend to military operations. Most shells, torpedoes, bombs, missiles, rockets, and other explosive devices are designed for electrical actuation so the question naturally arose as to whether electrical currents generated by electromagnetic energy might set off these devices. It was shown that theoretically under the proper conditions this could indeed happen.

The squib which is used to fire explosives electrically uses a "hot" wire which can be energized just as effectively by high frequency currents as by the direct currents for which it is designed. This has been proven experimentally in various places by several experimenters. Squibs and fuzes have been actuated by being placed in a radar beam or by being fed currents from high frequency generators. As the result of these and other tests, the possibilities of RF hazards existing for ordnance are being studied intensively. A number of incidents of accidental firings of rockets have been reported in which electromagnetic energy was the suspected or the verified cause. As a result of these studies and occurrences, several military programs have been undertaken to increase the knowledge of the RF hazard to ordnance.

3.1 BASIC HAZARD MECHANISM

A common igniter (squib) which has fired prematurely on occasion uses a bridge wire of resistive material immersed in a special flash charge (primary) explosive material. This flash charge is used to set off a charge of black powder which serves as the basic igniter. In general, primary explosives may be initiated by heat, shock, light, and ionizing radiation energy. Heating is the simplest mode for initiation and requires by far the smallest input energy. Once the chemical reaction has been started, an explosion results when heat is liberated by the reaction at a rate greater than it is lost in the region of the reaction. This effect will sustain and grow into an explosion, provided a small nucleus (hot spot) of decomposition exceeding a minimum size and a minimum temperature has been attained. As stated above, RF currents can cause the generation of a hot spot and result in premature firing of the squib.

A fundamental problem is to devise procedures and criteria by which the degree of hazard may be estimated. This problem is complicated by the wide range of frequencies involved, and the complex geometries associated with the weapon in its transportation, storage, training operations and ultimate deployment and utilization.

Every type of electroexplosive device (EED) has a nominal fire-energy level, usually expressed in millijoules or ergs. However, individual EED's of one type will have fire-energy levels which deviate from the nominal level by as much as 50%. Statistical data has been accumulated on many EED's by various commercial and military laboratories throughout the country. Additional but rather limited data has been collected on the susceptibility of these various devices to RF fields. Experimentation has been done both with the EED alone and the EED installed in normal fashion with firing circuits attached. The results of this work indicate rather clearly that EED's are susceptible to low frequency and high frequency RF energy and thus must be protected from it. Data lacking to date is a spectrum analysis of each type EED, showing its susceptibility, as installed with firing circuitry, to radiated frequency, power or field intensity, and modulation of the radiated energy. Considerable effort is being expended by military and commercial laboratories to get this kind of information; however, progress is slow because of the complexities involved with the selection or development of acceptable measurement instrumentation.

3.2 HAZARDS FROM COMMUNICATIONS TRANSMITTERS

By far the greatest threat of RF hazards to ordnance results from the use of electroexplosive devices (EED) in close proximity to high frequency communications transmitters and not from microwave radars as is sometimes thought. This is not to imply that microwave radars do not or cannot cause hazards but instead, it is intended to emphasize that at most locations where hazards occur, the measured field strengths in the high frequency regions (2-30 mc) are usually of much greater magnitude than those from microwave radars. This, of course, is not true if the main radar beam of a very high powered radar is pointed directly at an exposed EED.

Many cases have been found where the presence of a human body was sufficient to complete a path to ground through the EED for high frequency RF energy and cause a premature explosion. The question is often raised as to how a small device such as a squib can have enough capture area to collect sufficient energy to cause ignition. This is readily answered when the large area of the device to which the EED is usually attached is accounted for. When this is done, it will be found that on occasion astonishingly high voltages or currents are built up between ground and the body of the EED. As an example of how sufficiently large currents may pass through an EED, consider the following: an efficient pickup (for its size) in feeding low-resistance devices is a loop antenna. A loop of 0.1 wavelength on a side has a radiation resistance of 3 ohms and one of 0.05 wavelengths has less than 1 ohm. Thus 10 volts/meter at 10 mc (a field strength readily obtainable near the antenna) will drive better than 1 amp through a 1.5 square meter loop. Since loops are non-resonant below 0.1 wavelength on a side, they act as wide band pickup devices over the entire low frequency range.

3.3 HAZARDS FROM MICROWAVE RADARS

While it has been demonstrated that a relatively low powered microwave radar can set off an EED which is in the open and unshielded, it is more difficult to prove from theoretical calculations and measurements that radars have, in fact, caused premature detonation under field conditions. In many instances, radars are prohibited from illuminating a missile when the EED is being installed and this is undoubtedly a wise procedure. Present radars are unlikely to cause malfunctioning of an EED when the missile is airborne.

Radars planned for the future especially those of the phased array type will, if they actually deliver the power proposed, constitute an RF hazard of considerable magnitude since the field strength even at appreciable distances from the antenna will be very high. In this case, it will be necessary to make calculations, and measurements at reduced power, in the Fresnel region of the antenna where complicated phase and spatial relations exist. This means that simple estimates of the RF hazard threat will not be possible.

4. MISCELLANEOUS HAZARDS

In addition to obvious hazards to personnel and ordnance, intense RF fields can also create other hazardous effects, some of which probably have not yet been traced to the presence of intense electromagnetic energy in the area. One of the most serious of the miscellaneous hazards is the explosion of fuel such as gasoline or kerosene during refueling operations around aircraft or missiles. The aircraft makes an excellent pickup antenna and intense fields can be generated between the fueling nozzle and the aircraft. Obviously, any sparking between these points is quite likely to result in a very disastrous explosion.

Another hazard which is more dangerous because of possible secondary reactions is the setting off of photographic flash bulbs by intense RF fields. Ignition of a large quantity at once could cause a fire but solitary ignition would probably be more of a nuisance than a danger.

APPLICATION OF FOURIER INTEGRAL ANALYSIS TO SIMPLE TRAPEZOIDAL PULSES

APPENDIX I

Let f(t) represent any trapezoidal pulse as follows:

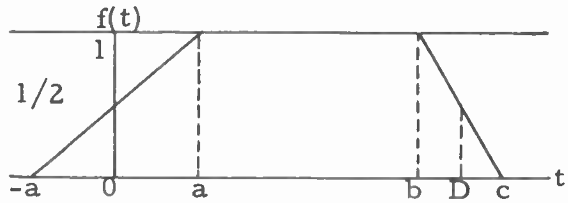
$f(t) = 0; \qquad t < -a$

$f(t) = 1/2 + k_1 t; \qquad -a < t < a$

$f(t) = 1; \qquad a < t < b$

$f(t) = 1/2 - k_2 (t - D); \qquad b < t < c$

$f(t) = 0 \qquad c < t$



The parameters, k_1 and k_2, are the slopes of the rising and falling portions of the pulse, respectively. The points at which f(t) equals 1/2 for t = 0 and t = D are fixed so that the area under the pulse given by the integral of f(t) is a constant independent of the values of k_1 and k_2. Since the total energy of the pulse is proportional to the area, the energy also is independent of k_1 and k_2. The Fourier transform f(t) is:

$$F(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} f(t)\, e^{-j\omega t}\, dt$$

$$= \frac{1}{2\pi} \left\{ \int_{-a}^{a} \left[\frac{1}{2} + k_1 t\right] e^{-j\omega t}\, dt + \int_{a}^{b} e^{-j\omega t}\, dt + \int_{b}^{c} \left[\frac{1}{2} - k_2 (t - D)\right] dt \right\} \qquad (1)$$

After carrying out the integrations, and making use of the fact that $a = 1/(2k_1)$, $b = D - 1/(2k_2)$, and $c = D + 1/(2k_2)$, one obtains

$$F(\omega) = \frac{1}{2\pi\omega^2} \left[k_2 \sin \omega D \sin \frac{\omega}{2k_2} + j \left(k_2 \cos \omega D \sin \frac{\omega}{2k_2} - k_1 \sin \frac{\omega}{2k_1}\right) \right] \qquad (2)$$

The quantity of interest is $|F(\omega)|$, the absolute value of the frequency function, which is

$$\left| F(\omega) \right| = \frac{1}{2\pi\omega^2} \sqrt{k_2^2 \sin^2 \frac{\omega}{2k_2} + k_1^2 \sin^2 \frac{\omega}{2k_1} - 2k_1 k_2 \cos \omega D \sin \frac{\omega}{2k_1} \sin \frac{\omega}{2k_2}} \quad (3)$$

As it stands, this is too complicated to allow an easy interpretation. If $\cos \omega D$ varies much faster than the other terms, i.e., $D >> 1/k_1$ and $D \gg 1/k_2$, an envelope of $F(\omega)$ may be established by noting that for $\cos \omega D = \pm 1$,

$$\left| F(\omega) \right| = \frac{1}{2\pi\omega^2} \left(k_1 \sin \frac{\omega}{2k_1} \pm k_2 \sin \frac{\omega}{2k_2} \right) \quad (4)$$

When $\omega \ll k_1$ and $\omega \ll k_2$, $\sin(\omega/2k_1) \simeq \omega/2k_1$ and $\sin(\omega/2k_2) \simeq \omega/2k_2$, so that

$$\left| F(\omega) \right| = \frac{1}{4\pi\omega} (1 \pm 1) \quad (5)$$

Hence, the low frequency response is independent of k_1 and k_2. At high frequencies, another envelope is obtained by setting $\sin(\omega/2k_1) = \pm 1$ and $\sin(\omega/2k_2) = \pm 1$. Hence,

$$\left| F(\omega) \right| = \frac{1}{2\pi\omega^2} (k_1 \pm k_2) \quad (6)$$

and the high frequency components increase linearly with both k_1 and k_2.

In order to plot $\left| F(\omega) \right|$, it is assumed, for simplicity, that $k_1 = k_2 = k$. Then,

$$\left| F(\omega) \right| = \frac{1}{\pi\omega^2} \quad k \sin \frac{\omega}{2k} \quad \sin \frac{\omega D}{2} \quad (7)$$

The curves of Figure 1-4 in Chapter I of this volume show how this function is affected by a variation of k and D.

The purpose of this appendix is to stimulate thought and promote discussion on the applicability of the "impedance concept" to radio-interference problems. The method of the "impedance concept" is anything but new, but its application to radio interference problems has lagged due to the lack of sufficient basic data. This lack severely limits the practical application of the method at this time. It is hoped, however, that design and research engineers will be stimulated into developing this method sufficiently to make it a true, practical aid in the solution of radio interference problems. Difficulties that may arise are no argument against the method itself, nor do they detract from its potential fertility. As a first step towards tapping this fertility, this appendix will point out what information will have to be obtained in order to make practical use of the impedance concept, how some of the difficulties in obtaining this information may be overcome, and how this information, if it were obtained, could be used to establish practical methods of reducing radio interference.

The "impedance concept" is the foundation of engineering transmission theory. Authorities in this field have stated that components of transmission systems must have their properties expressed in terms of appropriately chosen impedances, or else a new transmission theory must be developed.

The "impedance concept" was defined in the text as the idea of considering the source of radio interference, the network or medium through which it is transmitted, and the receiver whose effectiveness it finally impairs, as one single entity. The term "impedance" is a general term defined as the ratio of the cause to the effect. In electrical systems, it is the ratio of the electromotive force to the current that flows as a result of that force. The same idea may be applied in other fields. For example, in mechanical systems the impedance is defined as the ratio of the force to the velocity. It is therefore quite common for the design engineer to refer to mechanical impedances and acoustic impedances, as well as to mixed mutual impedances and even the "impedance of free space."

This approach may at first appear "impractical" to some engineers engaged in "interference research." But careful study of the subject brings out the fact that interference problems are basically of the same nature as the transmission problems which, in the fields of telephony, telegraphy, and other types of communication, have been solved successfully by the application of the rigorous analytical methods of circuit analysis. In this process, the source is characterized solely by its generated electromotive force and internal impedance (each assumed to be a known function of frequency over the entire range of interest), the transmitting network or medium by its three image parameters (assuming that the network is linear), and the receiver simply by its impedance. Thus, it takes just four impedances and one image transfer constant to specify such a system completely. This is the justification for the term "impedance concept." The impedances are simply the internal generator impedance, the image impedances looking into the input and output of the transmission network or medium, and the load or receiver impedance. The transfer constant gives the attenuation and the phase shift of the currents or voltages under consideration.

The definition of the term "impedance" as given above requires further elaboration. For the impedance to have a definite, single value, the cause and the effect of which the ratio is being taken must both vary sinusoidally in time. The case of being constant in time is included as that of varying sinusoidally with zero frequency. The impedance is, in general, a complex quantity, its magnitude being equal to the ratio of the magnitudes of the cause and effect, and its phase angle giving the difference in phase between the two. Next, the definition is extended to quantities that vary periodically, but not necessarily sinusoidally, in time. In this case, use is made of a Fourier series, as explained in Section 1, Chapter I, of this volume. The periodic variation is represented as a sum of an infinite number of sinusoidal terms. With each term there is associated a definite frequency and a definite impedance. The impedance of the system has no longer a single, definite value, but is now a function of frequency, assuming a definite value at each discreet frequency contained in the Fourier series. Finally, the definition is extended to nonperiodic quantities. Now a Fourier-integral expansion (see Section 1) is required, and the impedance becomes a continuous function of frequency. It should be pointed out that such a Fourier-series or Fourier-integral expansion need not be carried out explicitly. The knowledge that such an expansion is possible suffices to permit the use of the impedance of a system (considered as a continuous, complex function of frequency) in formulating and analyzing any transmission problem.

In most practical transmission problems, the initial datum is the signal to be transmitted. An analysis is then performed of the various types of transmission systems including the terminal equipment at both ends so that the most suitable system may be chosen and a basis for design may be established. The interest usually centers on the proper transmission of the desired signal. Sometimes, however, for example in the design of filters, the attenuation of signals other than the one to be transmitted is a consideration also. Even then, the impedances of the source and the load into which the filter operates usually need to be considered only at the frequencies to be passed. There is the possibility, of course, that the source or load impedances may become very low or high, or go to zero or infinity, at some important frequencies to be attenuated, and, therefore, the designer should have some idea of the frequency characteristics at all frequencies in order to avoid the possibility of sharp dips in the attenuation characteristics of the filter or network being designed.

The impedance point of view is less helpful in the modification of existing equipments and installations (i.e., "field fixes") than in the incorporation of interference-reducing practices into the original designs. For, once a particular piece of equipment has been designed and manufactured, its impedance at all frequencies is fixed and very difficult to change. If the design was carried through without giving any consideration whatever to the impedances at the radio frequencies likely to be encountered, then one cannot expect that by slight changes in the transmitting network - such as the addition of capacitors or filters, or the rerouting of wiring - an efficient overall system is produced having all the desired and none of the undesired properties. To produce such a system, the impedance point of view must be adopted from the very beginning and must be taken into consideration in the design of terminal equipment at both ends as well as of the interposed transmission network. Nevertheless, this does not mean that the existing equipment is useless from the impedance point of view. No design can proceed without a wealth of pertinent basic data. Measurements on existing equipments can contribute a great deal to such data, and, in fact, until other and better equipments become available, existing equipments are the only sources from which such data can be obtained.

The RF problem can be formulated in the following way: Given a piece of apparatus - say a direct current motor - which is known to be a potential generator of interfering voltages or currents, given a receiver which may be affected by these interfering currents or voltages, and given finally a transmitting network, which will in this case be an entire aircraft with all its electric and electronic equipment and associated wiring. The problem is to find the dependence of the six parameters of this system on all variables under the control of the designer. Once this dependence is known, transmission theory can be used to determine the optimum values of the system parameters, and the design can proceed to obtain the proper values of the controllable variables.

If impedances are to be measured in the conventional manner, terminals must be designated between which the measurements are to be made. At very high and ultra-high frequencies such terminals may not be available, and the very word "impedance" must be redefined to acquire a meaning applicable to such situations. This introduces a new type of difficulty, which will be discussed later. For the time being it will be assumed that the frequencies are low enough to insure meaningfulness of the expression "impedance between a set of two terminals."

After the statement of the problem, the first step is to choose two pairs of terminals, one to be the junction between the source and the transmitting network, and the other one to be the junction between the transmitting network and the receiver. Since the interfering signal may leave the source by any one of several routes, and may enter the receiver at any one of several places, this choice itself is not an easy one to make and involves certain simplifications at the outset. If the source is completely surrounded by an effective shield so that radiated interference may be neglected and conduction through the power lead is the only way by which interfering signals can leave, then the approximation involved in considering only one pair of terminals at the source is a very good one. Similarly, if the receiver is adequately shielded and employs a shielded lead-in from the antenna, and if direct coupling to the antenna is negligible, so that again the power lead is the only path of entry, then all terminals but two at the receiver may safely be neglected. In other cases these approximations may be quite poor. At any rate, as in all complicated problems, it is certainly permissible to start with the simplest conditions and make as many simplifications as possible in the beginning. As the analysis progresses, refinements and extensions may be able to take care of the complications that were neglected initially.

The next step is to break the connections at the output terminals of the source in an actual installation and make measurements to obtain information about the internal impedance and the generated electromotive force of the source. Here a practical difficulty is encountered which is caused by the fact that, by disconnecting the source at its terminals, usually its normal operating conditions are destroyed so that the resulting information about the electromotive force and impedance is irrelevant. For example, in a direct-current motor, the generated interference voltages are functions of the direct current. If the power leads were disconnected, the motor would not run and the generated interference voltages would be zero at all frequencies. Therefore, it is necessary to design a terminal network which will allow the equipment to function as nearly as possible in a normal manner, yet permit measurements equivalent to open-circuit measurements at the frequencies of the interference voltages.

In measuring the internal impedance of the source, indirect methods must usually be used. Direct methods, such as the use of alternating current bridges, are not suitable for the measurement of the internal impedance of active networks. One possible way of obtaining at least the absolute value of the internal impedance of a source is to measure the open-circuit voltage and the short-circuit current and take the ratio of the two. In the case of a direct-current motor, a network that can be used to obtain these two measurements might consist of a battery in series with a parallel combination of inductance and capacitance, which can be tuned to anti-resonance at the frequency of measurement. The battery must be by-passed by a capacitor large enough to make its impedance negligible at all interfering frequencies. In addition, a series resonant circuit is connected in parallel with the battery branch through a switch as shown in Figure II-1. A high-impedance voltmeter is connected so as to measure the voltage either across the anti-resonant circuit or across the capacitance in the series-resonant circuit. The voltmeter is protected from direct currents by means of a large series capacitor. The voltage across the capacitance of the series-resonant circuit is used as a measure of the current through the series-resonant branch since the current would normally be too small to be read directly with a radio-frequency milliammeter. To minimize errors, the Q of both inductance coils should be as high as possible, the ratio of inductance to capacitance in the parallel-resonant circuit should be as low as possible, and the ratio of inductance to capacitance in the series-resonant circuit should be as high as possible. These are the conditions for maximum impedance at anti-resonance of the anti-resonant circuit, minimum impedance at resonance of the resonant circuit, and most rapid decrease (or increase) of the impedance on both sides of anti-resonance (or resonance).

In making measurements with the circuit of Figure II-1, the following procedure is used: First the parallel combination is tuned to resonance and the series combination is tuned to resonance at the desired frequency. Then, with switch S_1 open and switch S_2 in position A, the voltage across the parallel resonant circuit is measured. Since this circuit offers a very high impedance at its resonant frequency, this reading may be taken as the generated voltage at that frequency. Next, switch S_1 is closed and switch S_2 thrown to position B. The source is now effectively short-circuited through the series-resonant circuit, and the voltage across the capacitor is proportional to the short-circuit current. If the capacitance, the voltage, and the frequency are known, the current can be computed.

This procedure must be repeated for each frequency of interest. The frequency may be determined in each case from the inductance and

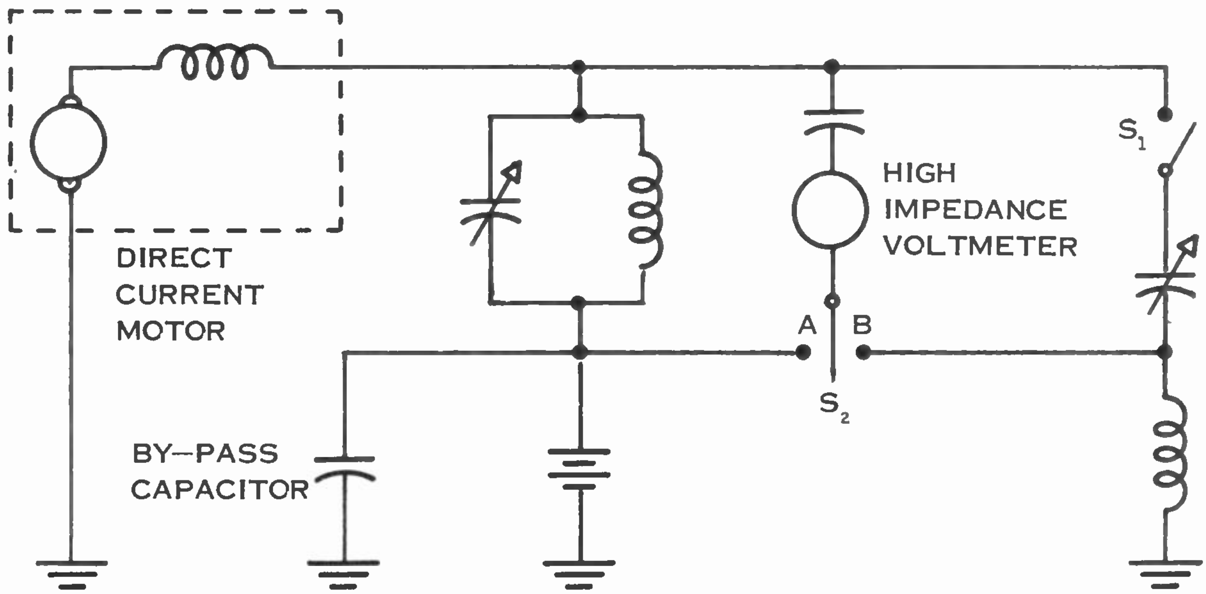


Figure II-1. Measurement of Open-Circuit Voltage and Short-Circuit Current Using High-Impedance Voltmeter

capacitance in the circuit, or by substituting a calibrated signal generator for the source and observing which frequency gives maximum voltage across the parallel circuit with switch S_1 open and maximum voltage across the capacitor of the series circuit with switch S_1 closed.

A "noisemeter" or calibrated receiver can be substituted for the voltmeter provided that it has a high input impedance. If such an instrument is available, the anti-resonant circuit may be replaced by a radio-frequency choke and the series-resonant circuit by a small resistance in series with a large fixed capacitor since the input circuit or first stage of the "noisemeter" or receiver provides the means of tuning. Such an arrangement is shown in Figure II-2. The resistance must, of course, be small as compared to the internal impedance of the source so that it can act as an effective short circuit. The procedure of making measurements is the same as before except that the tuning is now done with the meter and the frequency may be read directly.

The measurements described above provide information about the absolute values of the generated electromotive force and the short-circuit current. The ratio of the two yields the absolute value of the internal impedance of the source. To determine its phase angle, the circuit of Figure II-2 may be modified by adding a standard inductor or

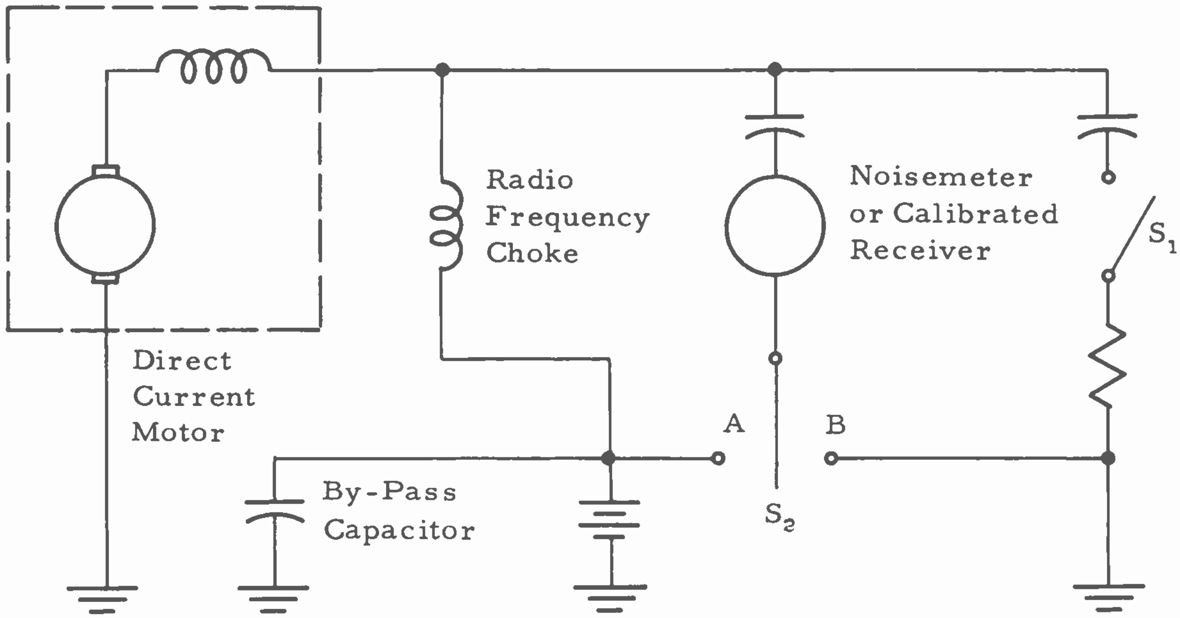


Figure II-2. Measurement of Open-Circuit Voltage and Short-Circuit Current Using "Noisemeter" or Calibrated Receiver

capacitor in series with the resistor. With all other settings the same as for the short-circuit-current measurement, the standard is varied until the current is a maximum as indicated by maximum voltage across the resistor. The reactance of the standard is then equal and opposite to the reactance of the source. From a knowledge of the reactance and the absolute value of the impedance the internal impedance of the source may be determined completely. Thus a set of curves of the generated electromotive force and the absolute value and phase angle of the internal impedance of the source as functions of frequencies may be prepared from this data.

The above method of measuring the short-circuit current is applicable only when the interference voltages are comparatively high. It may happen that the short-circuit current is too small to produce a measurable indication on the meter. Then an alternate method may be employed using a signal generator with calibrated output and known internal impedance as shown in Figure II-3. The signal generator is connected in parallel with the battery, a large series capacitor being used to protect the generator from the direct current. The output of the signal generator must be much larger than the interference voltages generated by the source so that the latter can be neglected. From a knowledge

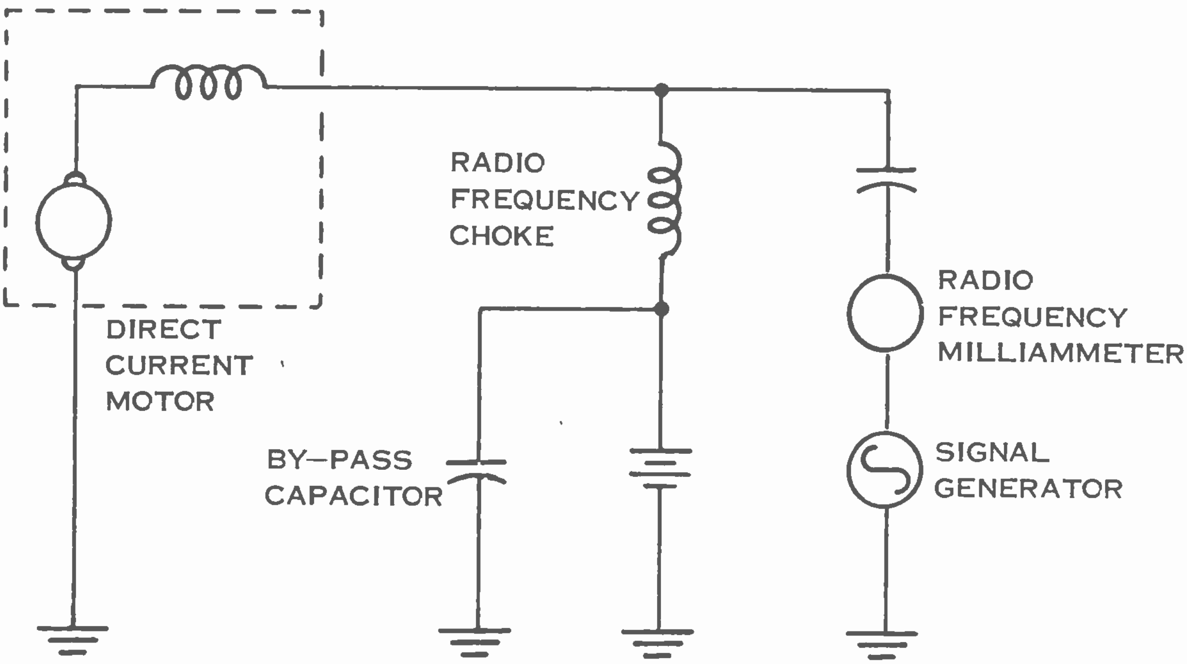


Figure II-3. Measurement of Internal Impedance with External Signal Generator

of the voltage as read on the signal generator, its internal impedance (usually a pure resistance), and the current, the absolute value of the impedance may again be determined. The reactance may again be found by connecting a standard inductance or capacitance in series and resonating it with the reactance of the source. This second method is not subject to frequency and waveform errors provided that the output of the calibrated signal generator is a pure sine wave.

Similar measurements must also be made at the input of the receiver. Again the receiver must operate normally if the measurements are to be meaningful. However, since the receiver is a passive network, bridge methods may be used provided precautions are taken to prevent the power currents (normally at 400 cycles per second in aircraft) from damaging the bridge or affecting the measurements. A possible circuit is shown in Figure II-4. Alternately a signal generator and milliammeter may be used in the circuit of Figure II-5.

To avoid complexities of the impedance measurements just described, the impedance of the source and the receiver may be measured without supplying any power. Recent tests have shown that fair results may be expected using this method, and data so obtained may well serve as a good first approximation.

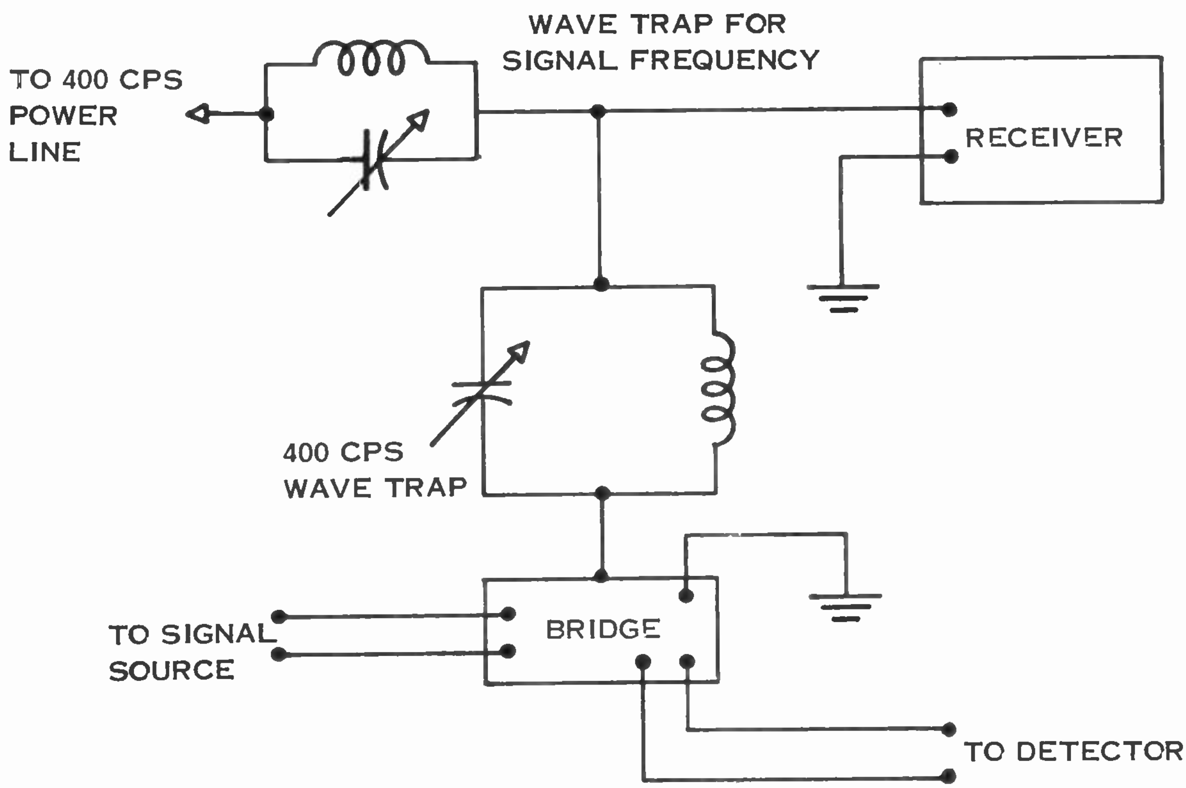


Figure II-4. Bridge Measurement of Receiver Input Impedance

Finally, the parameters of the transmission network must be measured. These may be obtained from measurements of the input impedances at each end with the other end open-circuited and with the other end short-circuited. If the open- and short-circuit impedances at the input are called Z_{oc} and Z_{sc}, respectively, and the corresponding impedances at the output are called Z'_{oc} and Z'_{sc}, then the image impedances Z_{I_1}, and Z_{I_2}, and the image transfer constant, θ, are obtained from:

$$Z_{I_1} = \sqrt{Z_{oc} Z_{sc}} \tag{1}$$

$$Z_{I_2} = \sqrt{Z'_{oc} Z'_{sc}} \tag{2}$$

$$\theta = \tanh^{-1} \sqrt{\frac{Z_{sc}}{Z_{oc}}} = \tanh^{-1} \sqrt{\frac{Z'_{sc}}{Z'_{oc}}} \tag{3}$$

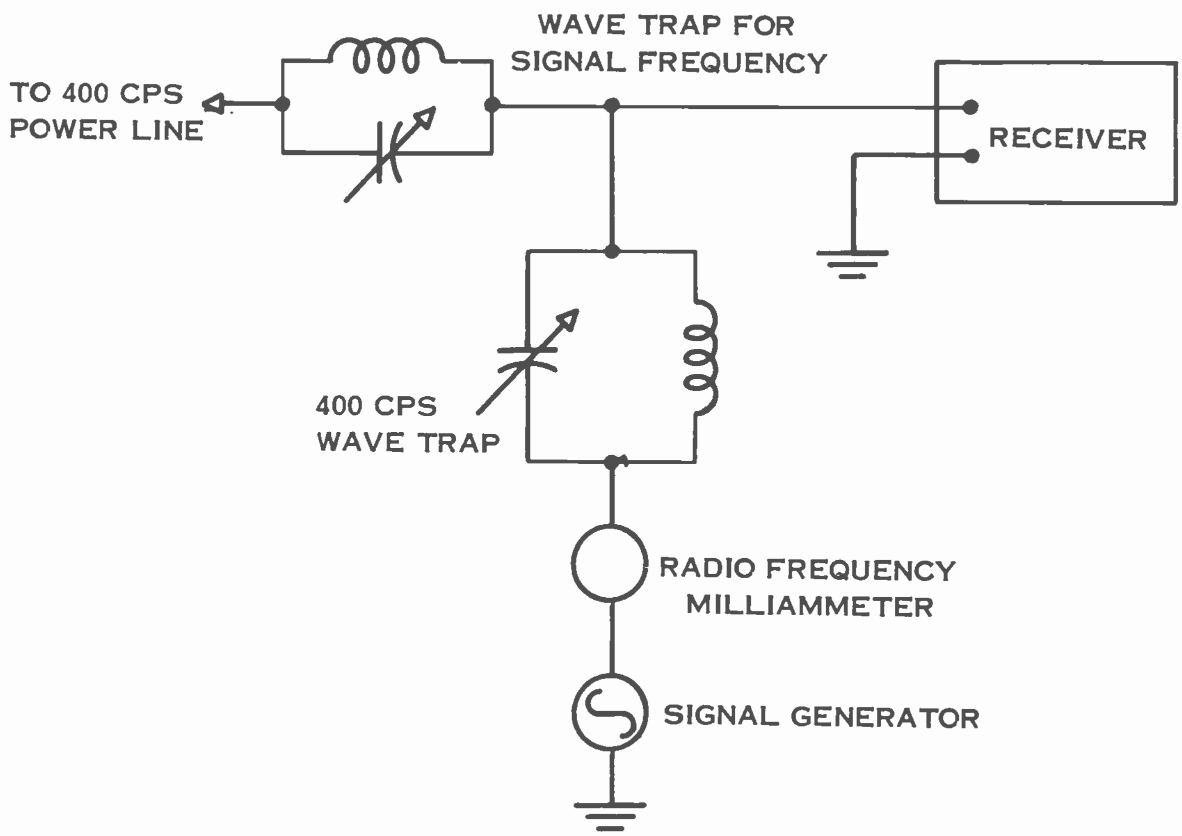


Figure II-5. Alternate Method of Measuring Receiver Input Impedance

The open- and short-circuit impedances may be measured either by bridge methods or by means of a signal generator and milliammeter as before.

After measuring methods for these six quantities have been established, a large number of sets of measurements must be taken to determine the effect of the variation of various controllable design quantities such as size of wires, position of ground leads, shape of frames, brush materials, and many others. This is a huge task, yet one that is necessary if practical results are expected. Only if the effect of each design detail is known can one hope to be able to control the impedances and generated voltages by proper design before the equipment is actually built.

To be of practical use to the designer, the information obtained must be presented in the form of charts or curves. For example, one such curve would give information about the variation of the image impedances of aircraft wiring with the size of wires used. Another might present data on the dependence of generated interference voltages on brush pressure in a motor. A third would give the input impedance of a receiver (at the power leads) as a function of ground lead location. It is quite obvious that the number of charts and curves required is very large.

Measurements of the type just described may never have been made in actual installations on an extensive scale. Isolated attempts are on record of obtaining significant data in a few special cases. Not many conclusions can be drawn from the results of these attempts, yet they bring out clearly the difficulties as well as the definite possibility of overcoming these difficulties. In addition, certain observations are of interest even though they require thorough checking in view of the special conditions under which they were made. These observations are enumerated below:

a. The generated electromotive forces decrease very rapidly for frequencies above one megacycle per second for most motors. Most of their energy is concentrated at the lower frequencies.

b. The impedance of aircraft wiring is such that the frequencies above one megacycle per second are accentuated by the resonances of the wiring system. This results in several high peaks of voltages or currents, their number depending on the length and the complexity of the wiring.

c. For the frequencies of interest (from 0.15 to 1000 mc) the actual attenuation of the wiring system is very small. Apparent attenuation in any one wire may be large because of the spreading of the interference energy over the entire wiring system.

d. Transmission and spreading of the interference energy is almost entirely by conduction and induction and capacitive coupling. Transmission by radiation is negligible below 150 megacycles.

e. Measurements of the impedance of the wiring system show wide and rapid variations of both the resistive and reactive components with frequency. Observed maxima and minima bear no apparent harmonic relationship.

f. Serious impedance mismatches are present at the junction points of the wiring system and at the points where the wiring is terminated in electrical or electronic equipment. These mismatches lead to reflections and standing waves on the wires resulting, in turn, in heavy concentrations of interference energy in certain regions of the system. Broadband matching would eliminate these concentrations, but would also deliver more of the energy to the terminal equipment. At any rate, an attempt must be made to dissipate the energy before it can do any harm rather than simply redistribute it in space or in the frequency spectrum.

No information is available on the effect of controllable factors on the measured parameters. Even if such information were available, an investigation must first be made into which values of the parameters are most desirable from the point of view of eliminating radio interference.

Ideally, the requirement is that no interfering currents or voltages shall reach the receiver. Assuming that the generation of interference electromotive forces cannot be prevented (though the measurements suggested here would form a basis for designing the equipment so that their generation is minimized), the problem then is to attenuate the interference as much as possible. As pointed out in Paragraph 1.4, Section I of this volume, a signal undergoes modification in four distinct ways as it is transmitted from a source to a receiver. There is the attenuation proper in the transmitting network, there are the reflection losses at the input and output of the transmitting network, and there is the interaction loss due to multiple reflections. Of these, only the first, the attenuation proper, is always a true attenuation (or zero). Any one of the other three "losses" may be negative, and hence may be a gain rather than a loss. If the observations enumerated above are trustworthy, then the attenuation proper cannot be relied upon to provide a substantial reduction of interference (see item d above). Therefore, the reflection and interaction losses must be made to produce the necessary reduction unless it is possible to redesign the wiring system so that true attenuation is obtained within the transmitting system. (One suggestion to achieve this has been to use lossy cables for all the major wiring in the aircraft.)

Figure II-6 shows how the reflection loss varies with the magnitude and phase angle of the "mismatch ratio", i.e., the ratio of the impedances looking to the right and to the left of a pair of terminals. This indicates that this loss is positive for all magnitudes of the ratio only when the phase angle, φ is zero. For all other phase angles, the loss is negative for a portion of the range plotted, and this portion as

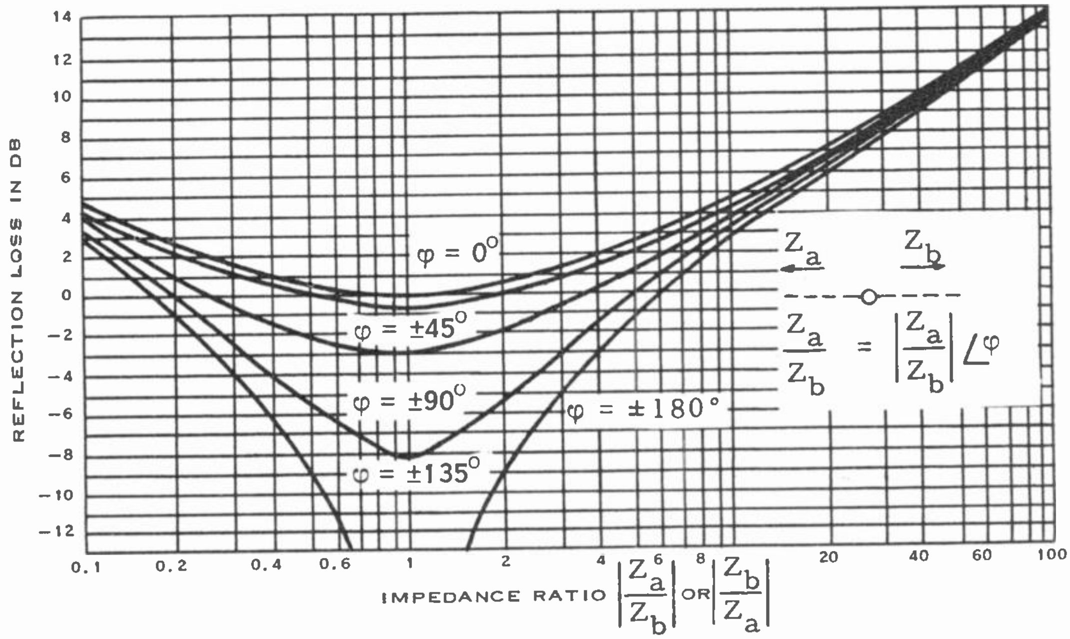


Figure. II-6 Reflection Loss Resulting from Mismatch of Impedances at a Junction

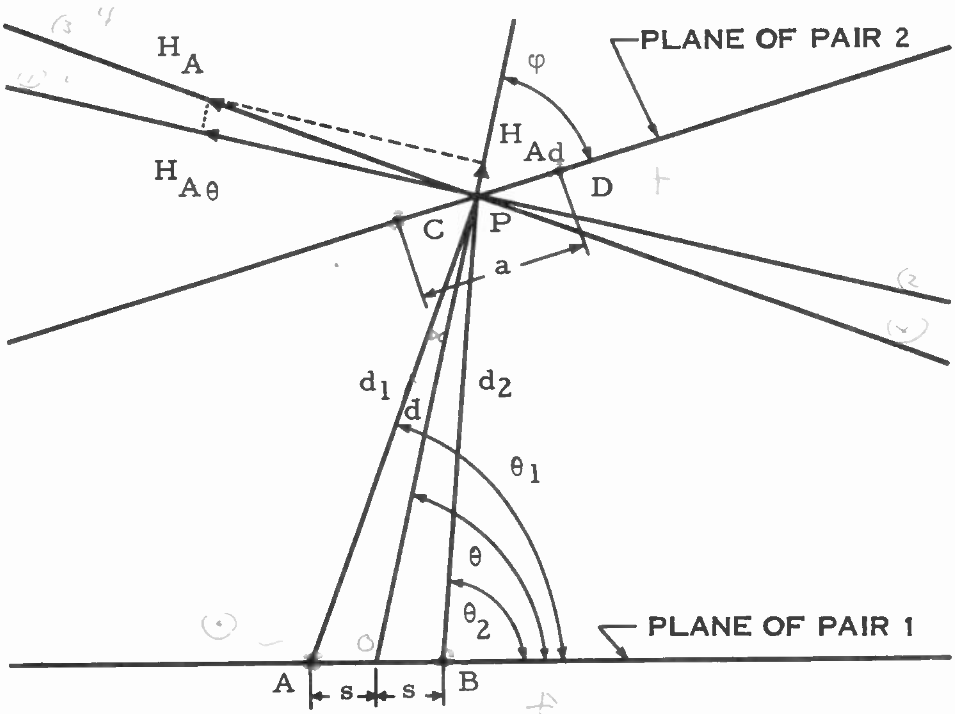
well as the magnitude of the gain both increase substantially as the phase angle varies towards 180° either through positive or through negative values. This figure would have to be used carefully in determining what impedances would be desirable for maximum attenuation. A similar chart should be worked out for the interaction loss.

If the properties of the system itself cannot provide sufficient attenuation, a filter could be inserted between the output terminals of the source and the input terminals of the transmitting network. The most efficient filter providing the maximum attenuation with the minimum expenditure of weight and space can be designed only when the impedances between which it is to operate, as well as the amplitudes and frequencies of the currents it is to transmit and attenuate, are precisely known. The weight and space requirements for most radio interference filters could be reduced substantially without sacrificing effectiveness if full information about the impedances were available. In fact, this information would not even be required over the entire frequency range from 0.15 to 1000 mc, but only for the lower frequencies from 0.15 to about 3 or 4 mc because the generated interference has by far the largest amplitudes in this lower region. The drawback of trying to utilize this saving in weight and space is that a filter so designed would be effective only in the particular position and for the particular system for which it was designed. In the absence of more detailed information, the slightest change in design, such as a change in the size of wires or in the material of the brushes, or even the slightest change in location of wiring or equipment, will have a large and entirely unpredictable effect on the impedances of the system, and hence may make the filter ineffective. Until ways are found of controlling the impedances and predicting their changes with changes in design parameters, filters must obviously be designed so as to remain effective under a large variety of conditions.

If this general method is to be extended to very high and ultra-high frequencies, additional difficulties must be overcome. Above 100 megacycles, radiation will become important, and if the circuit concept is still to be used, the radiation resistance must be included in the impedances. In certain cases, the term "impedance" itself must be given a new meaning. Instead of dealing with the impedance of a network, defined as the ratio of voltage to current, it may be necessary to deal with the impedance of a transmitting medium, which is defined as the ratio of the electric to the magnetic field intensity. In defining this ratio, the type of wave that gives rise to the fields, i. e., its geometrical configuration and polarization, must be specified in addition to the properties of the medium itself. Since several types of electromagnetic waves may be present simultaneously, the additional complexity introduced is clearly evident.

THE MUTUAL INDUCTANCE BETWEEN TWO SETS OF INFINITELY LONG PARALLEL PAIRS OF STRAIGHT WIRES

Consider the cross section of the four wires A, B, C, and D shown below.



Assume that $d \gg s$, so that

$$d_1 \simeq d + s \cos \theta \tag{1}$$

$$d_2 \simeq d - s \cos \theta \tag{2}$$

Let the left wire of Pair 1 carry a current I out of the plane of the paper and the right wire the current I into the plane of the paper. The two components of the magnetic field H (at Point P), produced by the current in wire A of Pair 1, are then

$$H_{A\theta} = \frac{I \cos (\theta_1 - \theta)}{2(d + s \cos \theta)} \tag{3}$$

FRC

$$H_{Ad} = \frac{I \sin(\theta_1 - \theta)}{2(d + s \cos\theta)\pi} \tag{4}$$

H_{Ad} $H_A \sin\alpha$

In the same way, for the field H_B set up by wire B of pair 1,

$$H_{B\theta} = \frac{I \cos(\theta - \theta_2)}{2(d - s \cos\theta)\pi} \tag{5}$$

$$H_{Bd} = \frac{I \sin(\theta - \theta_2)}{2(d - s \cos\theta)\pi} \tag{6}$$

The total field at P is then

$$H_\theta = H_{A\theta} + H_{B\theta} \simeq \frac{I}{2\pi}\left[\frac{-2 s \cos\theta}{d^2 - s^2 \cos^2\theta}\right] \simeq \frac{-I s \cos\theta}{\pi d^2} \tag{7}$$

because $\cos(\theta - \theta_1) \simeq \cos(\theta - \theta_2) \simeq 1$ and $d >> s$, and

$$H_d = H_{Ad} + H_{Bd} \simeq \frac{I}{2\pi}\left[\frac{2d \sin(\theta_1 - \theta)}{d^2 - s^2 \cos^2\theta}\right] \simeq \frac{I s \sin\theta}{\pi d^2} \tag{8}$$

because $\sin(\theta_1 - \theta) \simeq \sin(\theta - \theta_2) \simeq \frac{s \sin\theta}{d}$.

The flux density, B, at Point P is $4\pi \times 10^{-7}$ H, and its component perpendicular to the plane of Pair 2 is

$$B_P = 4\pi \times 10^{-7} (H_d \sin\varphi + H_\theta \cos\varphi) \tag{9}$$

$$= \frac{4Is}{d^2} \times 10^{-7} (\sin\theta \sin\varphi - \cos\theta \cos\varphi) \tag{10}$$

$$|B_P| = \frac{4Is}{d^2} \times 10^{-7} \cos(\theta + \varphi) \tag{11}$$

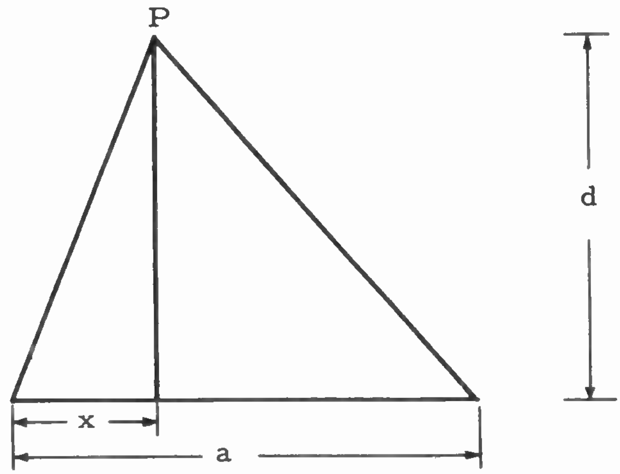
Assuming that $a << d$, the flux density B is practically constant between the wires C and D of Pair 2. Hence, the total flux for a unit length of wire is

$$\Phi = B_P a = \frac{4Isa}{d^2} \times 10^{-7} \cos(\theta + \varphi) \tag{12}$$

and the mutual inductance (henries/meter) is

$$M = \frac{\Phi}{I} = \frac{4sa}{d^2} \times 10^{-7} \cos(\theta + \varphi) \qquad (13)$$

This is a maximum when $\theta + \varphi = 0$, or $\theta + \varphi = 180°$, and zero when $\theta + \varphi = 90°$



Now consider the above diagram. Here $\theta = \varphi = 90°$ and $2s = a$, for simplicity; but it is no longer assumed that $d \gg a$. It is assumed, however, that the cross section of the wires is small as compared with the distances between them. Then, at P,

$$H_P = \frac{I}{2\pi}\left[\frac{x}{x^2 + d^2} + \frac{a - x}{(a - x)^2 + d^2}\right] \qquad (14)$$

$$B_P = 4\pi \times 10^{-7} H_P \qquad (15)$$

$$\Phi = \int_0^a B_P \, dx = 4\pi \times 10^{-7} \int_0^a H_P \, dx \qquad (16)$$

$$= 2I \times 10^{-7} \log\left(1 + \frac{a^2}{d^2}\right) \qquad (17)$$

$$M = \frac{\Phi}{I} = 2 \times 10^{-7} \log\left(1 + \frac{a^2}{d^2}\right) \qquad (18)$$

If $d \gg a$, this expression for M reduces properly to that on the previous page since, for $x \ll 1$,

$$\log(1 + x) \simeq x \qquad (19)$$

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