RADIO FREQUENCY
INTERFERENCE DESIGN GUIDE
FOR
AEROSPACE COMMUNICATIONS SYSTEMS

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Prepared by:

The Moore School of Electrical Engineering

UNIVERSITY of PENNSYLVANIA

PHILADELPHIA, PENNSYLVANIA 19104

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GREENBELT, MARYLAND

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University of Pennsylvania
THE MOORE SCHOOL OF ELECTRICAL ENGINEERING
Philadelphia, Pennsylvania

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Prepared by

F. Haber

M. Celebiler

C. Weil-Malherbe

for

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TABLE OF CONTENTS

Page

LIST OF ILLUSTRATIONS

Figure		Page
4-3	Illustration of Unwanted Signal Spectrum and the Receiver Tuning Characteristic	73
4-4	Some Receiver Spurious Responses	84
4-5	Relative Response Signal to Interference for Equal Outputs	84
4-6	Mixer Conversion Loss Test Results	94
5-1	Noise Intrusion Paths	97
5-2	Balanced Coupling Circuit	105
5-3	Attenuation of Circular Tubes used in Air Ducting Assembly	109
6-1	Audio Susceptibility Test on Power Line	115
6-2	RF Susceptibility Test, Power Line	117
6-3	Typical Test Set-up, Radiated Susceptibility Measurements	119
6-4	Quasi-Peak Circuit	125
6-5	A Ratio of Quasi-Peak Readings as a Function of RFI Meter Parameters	126
7-1	Relative Noise Levels for a Half-Wave Dipole	130
7-2	Change of Man-Made Noise With Distance from Urban Center	132
7-3	Distribution of Fixed Transmitters in Europe	136
7-4	Distribution of Fixed Transmitters in Africa	137
7-5	Distribution of Fixed Transmitters in South America	138
7-6	Distribution of Fixed Transmitters in North and Central America	139
7-7	Distribution of Fixed Transmitters in Asia	140

ABSTRACT

This publication has been prepared to provide guidance to designers of ground station equipment on potential electrical interference and its control for satellite tracking networks such as STADAN (space tracking and data acquisition network). Described herein are mechanisms of generations of unwanted emissions from radio transmitters, noise generation from electrical equipment not intended to radiate, susceptibility mechanisms of radio receivers, and the possible routes whereby unwanted signals can enter a receiver. The principle is described in each case, typical levels are specified where possible, reduction techniques are described, and examples are given. In addition, a brief account is given of the considerations in selecting low noise sites, and of current practice in interference measurement and instruments properties.

RADIO FREQUENCY INTERFERENCE (RFI) DESIGN GUIDE FOR AEROSPACE COMMUNICATIONS SYSTEMS

1.0 INTRODUCTION

1.1 PURPOSE AND SCOPE

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

1.2 DEFINITIONS OF TERMS

In this publication the terms "electromagnetic noise" and "electromagnetic interference" are viewed as being cause and effect, respectively (the specification 'electromagnetic' will generally be omitted, except where the briefer usage is ambiguous). Noise is any electromagnetic fluctuation, periodic or random, which may have a disturbing influence on devices exposed to it. It may arise from natural or man-made sources; the latter category includes deliberate radiators (transmitters), and electrical devices which radiate unintentionally. The transmission path may be conductive, or the noise may be propagated through space. The port through which it enters may be a normal signal input port, or it may be admitted at a spurious port. Frequent reference will be made to signal emissions in which the receiver is not interested. To avoid the derogatory implication of the term 'noise,' we shall in this case make use of the term 'unwanted signal' instead.

Interference, on the other hand, will be used in a qualitative and/or quantitative sense to describe the deterioration in the normal function of a device exposed to noise. It may take the form of a uniform reduction of normal output, a fluctuation superimposed on the normal output, etc. Common parlance frequently does not distinguish between noise and interference; in fact, the latter is often used to identify the source of the disturbance. This usage will be avoided here.

We shall make use of the notion of susceptibility. By this we mean the degree of sensitivity of a device to exposure to a source of noise. For instance, the level of input noise required to give a specified degradation of the output quality of a receiver is a measure of susceptibility. The susceptibility will vary with the input noise source, with the path through which the noise enters the device, and with the device characteristics.

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

1.3 SUMMARY OF SIGNIFICANT RADIO NOISE SOURCES AND SUSCEPTIBILITY MECHANISMS AT STADAN STATIONS

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

1. The presence of more than one satellite in the region of sensitivity of the ground antenna. The primary mechanism involves two satellites simultaneously in the main beam of a ground antenna and having co-channel frequency assignments. Secondary mechanisms involve the undesired satellite seen by the antenna sidelobes and/or the undesired satellite being in an adjacent channel.

2. The presence of airborne transmitters above the horizon of the ground station antennas. The primary mechanism here is of an adjacent channel nature. Because of the large powers, and the close distances involved, aircraft interference may very well be significant even if seen by the antenna sidelobes. The magnitudes will not ordinarily be so high as to stimulate nonlinear effects at the ground receiver--except when the aircraft is in the main beam of the ground antenna and flying less than 10 miles from the antenna.

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

Possible nonlinear mechanisms whereby such signals find their way into the receiver are (a) spurious responses, (b) intermodulation, (c) cross-modulation, and (d) desensitization. Mechanisms involving inadvertent coupling through an indirect input are often of the following nature: (a) conduction through power and control cables, (b) conduction and induction in ground loops, and (c) field induced pickup through inadequate shields around equipment and cables--the inadequacy resulting from under-design or from imperfections developed with use. It is also possible that the noise source is the transmitter itself. That is, the transmitter may generate unintentionally at the receiver tuned frequency. Mechanisms here are (a) sideband noise generation, (b) sideband splatter, (c) generation of harmonics of output signal, (d) escape of signals intended for internal use, and (e) intermodulation and cross-modulation with an external signal.

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

5. Interference from surrounding ground sources. Significant sources here are (a) electrical noise generated by the many sources operating simultaneously in nearby populated areas (urban noise); (b) noise from automobile ignition systems, particularly if major highways are nearby; (c) discharges from nearby high voltage power transmission lines (corona noise); and (d) emission from ground transmitters for broadcast, commercial, or government use.

The nature of these phenomena has been a subject of study in many laboratories for many years. Numerous publications can be found which treat one or more of these topics. Where more detail is required on interference mechanisms and reduction techniques references 1-3 are recommended.

1.4 QUANTITATIVE EVALUATION OF NOISE LEVELS AND EFFECT

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

It will not, ordinarily, be enough to know the mechanism of entry and the level of unwanted signal admitted in the passband. The degree to which the unwanted signal influences the output is the real question. It will be recognized that the interference effect depends on (1) the output measure of quality itself, (2) the nature of the receiving system, (3) the nature of the desired signal, (4) the nature of the undesired signal, and (5) the admission mechanism. The alternatives are too numerous for detailed individual treatment. Furthermore, the dependence of the effect on its causes is in many cases too complicated for quick estimation. It will be possible, however, to specify some level of acceptability after admission which will virtually guarantee satisfactory operation for most cases. Typically, a desired signal level of 20 db relative to the undesired one will be satisfactory in most non-critical cases.

2.0 GENERATION OF UNWANTED OUTPUT IN TRANSMITTERS

An actual transmitter will radiate a certain amount of unwanted energy outside of its assigned frequency band. In some instances, this is a result of the nature of the modulation process. Mechanisms of spurious generation are, however of equal or even greater importance. The most significant cause of spurious transmitter output is undesired nonlinearity.

The two most serious offenders are: (1) the transmitter modulator, and (2) the final amplifier. The former, which must be a time-varying device, is usually a nonlinear element with a specific required response law. Deviations from this law are the cause of spectrum broadening or "sideband splatter." The final amplifier is required to be a linear device over a wide input range. Deviations here from linearity give rise to transmitter harmonics.

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

There are times when the transmitter is not, by itself, responsible for the spurious output. The output of nearby transmitters may enter the transmitter, usually through the antenna, and mix with the desired signal to form an unwanted output component. Again, here the mechanism depends on nonlinearity--in the final amplifier, in the transmission system between the final amplifier and the antenna, or in the antenna itself. The nature of the effect is either (1) cross-modulation, where the information sidebands of the undesired signal turn up on the desired signal, or (2) intermodulation which is the forming of a third signal containing some version of the information sidebands of both signals.

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

2.1 SIDEBAND SPLATTER

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

2.1.1 Spectral Properties of AM and Related Systems

The fundamental process involved in generating AM, SSB, or DSB signals is multiplication. A low frequency information waveform $x_1(t)$ is multiplied by a sinusoidal carrier frequency $x_2(t) = \cos(\omega_c t + \phi_c)$. The vehicle for generating the product is virtually always a nonlinear device. This is followed by a narrow band filter to eliminate undesired components as shown in Fig.1-1. Ideal nonlinear elements are the square law device and the linear diode (Ref. 4, pp. 97-100 or Ref. 5, pp. 187-192). The latter, to give the required result without distorting the desired sidebands, demands that the sinusoidal carrier be much larger than the information waveform. The ideal characteristics are, however, only approximately realizable. A typical nonlinear device can, for instance, be described by the output-input characteristic

$$y = \sum_{n=0}^{N} a_n x^n \tag{2-1}$$

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

$$y = \sum_{n=0}^{N} a_n (x_1 + x_2)^n$$

$$= \sum_{n=0}^{N} a_n \sum_{k=0}^{n} \binom{n}{k} x_1^{(n-k)} \cos^k(\omega_c t + \phi_c) \tag{2-2}$$

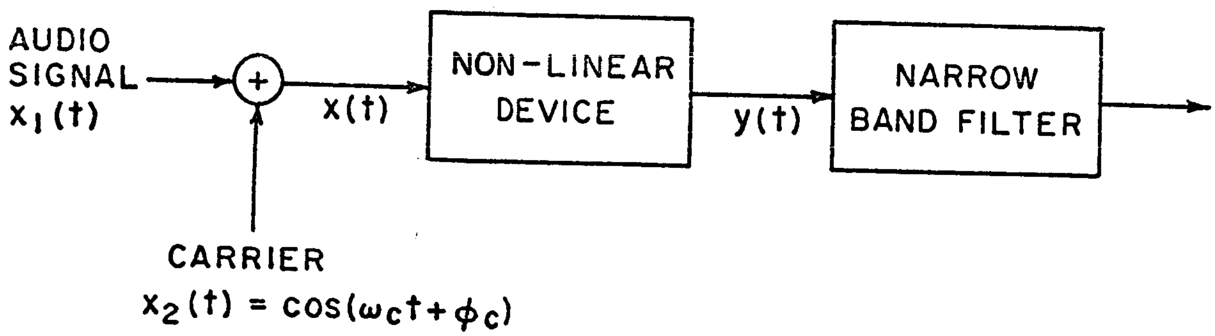


FIGURE 1-1

BASIC ELEMENTS OF PRODUCT MODULATORS

The second form in (2-2) is obtained from the first, using the binomial expansion formula. Assuming that the nonlinear element is followed by a narrow band filter centered at the carrier frequency, ω_c, only terms containing $\cos(\omega_c t + \phi_c)$ are passed. By expanding $\cos^k(\omega_c t + \phi_c)$ (see eq. 4-10a), and retaining only these terms, there is obtained

$$\sum_{n=1}^{N} \sum_{k=1}^{n} a_n \binom{n}{k} x_1^{(n-k)} \cdot \frac{k!}{\frac{k-1}{2}!\ \frac{k+1}{2}!} \cos(\omega_c t + \phi_c) \tag{2-3}$$

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

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

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

The component $x_1^2(t) \cos(\omega_c t + \phi_c)$ in (2-4), contributes sideband energy covering a band equal to four times the modulating frequency rather than two times, as in the ideal case; the $x_1^6(t) \cos(\omega_c t + \phi_c)$ component contributes energy in a band twelve times the original base band of $x_1(t)$.

The exact nature of the resultant spectrum requires that the quantities $x_1^n(t)$ be determined. If $x_1(t)$ is a pure sinusoid of frequency ω_m, $x_1^n(t)$ can readily be determined and (2-4) can be expanded in a harmonic series involving discrete sidebands of the form $\cos\left[(\omega_c - n\omega_m)t + \phi_c\right]$. Furthermore the coefficients a_n will be required. These are not ordinarily available and will have to be found by test (e.g., see Ref. 6). A set of typical values are given below. (Abstracted from Ref. 1. Obtained by polynomial least squares fit to a triode characteristic operated at a bias point of -10 volts on the grid and at 3 ma plate current).

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

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

$$y(t) = A_a \left[1 + m\,x(t)\right] \cos(\omega_c t + \phi_c) \qquad (2\text{-}5)$$

A non-overmodulated waveform is one in which the factor

$$\left[1 + m\,x(t)\right] > 0 \qquad (2\text{-}6)$$

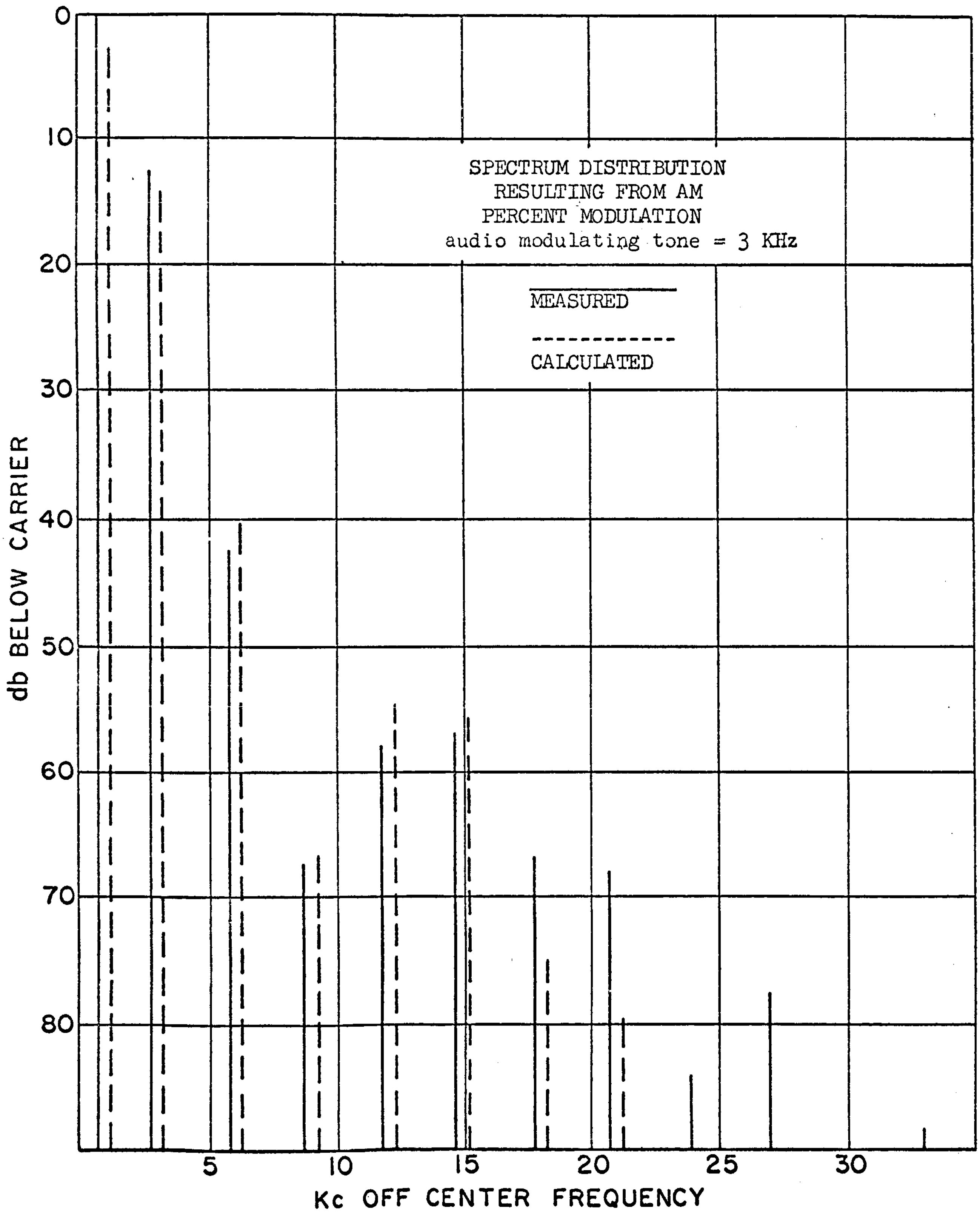
for all values of time. In some AM transmitters, if this factor is negative the transmitter is effectively cut off. In such cases the amplitude of splatter components becomes quite high. Correct practice requires that peaks of $x(t)$ be limited in such a way that (2-6) is always satisfied, and subsequently filtered to eliminate the high frequency components generated in the limiting process.

To minimize the effects described, it is desirable to choose nonlinear elements with characteristics as near as possible to the ideal. For a given device intended to be used as a square law modulator, it will usually be possible to find empirical operating conditions which result in a near square law characteristic; that is, which make the higher order coefficients small. Furthermore, the final tuned circuits in the transmitter will also act to partly reduce the level of the unwanted components. As a rule, however, the Q of these circuits cannot be made high enough to completely eliminate the splatter components. A technique frequently found effective to eliminate the unwanted sidebands associated with even powers of $x_1(t)$ in (2-3) is to use a balanced modulator (Ref. 4, p. 104). (The balanced modulator will be described further in Section 2.2 in connection with harmonic reduction).

Prediction of the magnitude of the unwanted sidebands will ordinarily be difficult. The coefficients, a_n, in equation (2-1) depend on the bias conditions and on the input range, and even vary among electronic devices of the same kind. To determine the actual level measurements are advised.

Measured and calculated results of excessive sideband output are reported by Firestone, et al (Ref. 7). In an AM transmitter with a 3 Kc/s audio modulating signal and with a modulating percentage of 54.7%, the result shown in the bar graph, Fig. 2-1, was obtained. It was pointed out by the authors that the distortion terms together amounted to 3.3% total harmonic distortion--not an excessive figure. A receiver operating in adjacent channel 10-15 KHz from its neighbor will experience splatter components on the order of 60 db below the level of the main components.

FIGURE 2-1



SPECTRUM DISTRIBUTION
RESULTING FROM AM
PERCENT MODULATION
audio modulating tone = 3 KHz
MEASURED
CALCULATED
db BELOW CARRIER
0
10
20
30
40
50
60
70
80
5
10
15
20
25
30
Kc OFF CENTER FREQUENCY

If, as may readily happen, the adjacent transmitter carrier power corresponds to an input voltage of about 100 millivolts (mv) at the receiver, the unwanted sidebands will correspond to about 100 microvolts (μv). This will often be far greater than the level of a desired signal.

2.1.2 Spectral Properties of Angle Modulated Systems

In the case of an FM transmitter there are, as is well known, a large number of sidebands even with a perfect modulation technique and with a single sine wave modulation. An FM signal containing an information waveform $x(t)$ is written

$$y(t) = A_f \cos\left[\omega_c t + \phi(t)\right] \tag{2-7}$$

where

$$\phi(t) = \int x(t)\, dt \tag{2-8}$$

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

$$\omega_i(t) = \frac{d}{dt}\left[\omega_c t + \phi(t)\right] = \omega_c + \frac{d\phi}{dt} = \omega_c + x(t) \tag{2-9}$$

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

$$x(t) = a \cos \omega_m t \tag{2-10}$$

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

$$y(t) = A_f \left\{ J_o(\beta)\cos \omega_c t + \sum_{n=1}^{\infty} J_n(\beta)\left[\cos(\omega_c + n\omega_m)t + (-1)^n \cos(\omega_c - n\omega_m)t\right]\right\} \tag{2-11}$$

where $J_n(\beta)$ is the Bessel function of the first kind and of order n, and

$\beta \equiv \frac{a}{\omega_m}$ is the modulation index. Though the result contains, in principle, components to infinite frequency, the sidebands fall off very rapidly in amplitude beyond a certain point. A rule of thumb frequently used is that all significant components lie inside the range

$$\left[f_c - (\beta + 2)f_m\right] < f < \left[f_c + (\beta + 2)f_m\right] \tag{2-12}$$

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

$$J_n(\beta) \cong \frac{1}{\sqrt{2\pi n}} \left(\frac{\epsilon\beta}{2n}\right)^n \tag{2-13}$$

where ϵ is the base of the natural logarithm and is equal to 2.718. For instance, if $f_m = 3$ KHz, and $\beta = 1$, then according to (2-13), at $nf_m = 6 \times 3 = 18$ KHz, the component has an amplitude, $J_6(1)$, of about -94 db relative to the level of an unmodulated carrier of the same power.

As in the AM case, it is also possible for the modulation process to create a form of splatter, though for wide-band modulation it will not increase the bandwidth to the same extent as it does for AM. In the case of narrow-band FM, however, it may give rise to a more significant effect. The basis of this phenomena is once again nonlinearity in the modulation process. To generate FM waves it is necessary to vary the

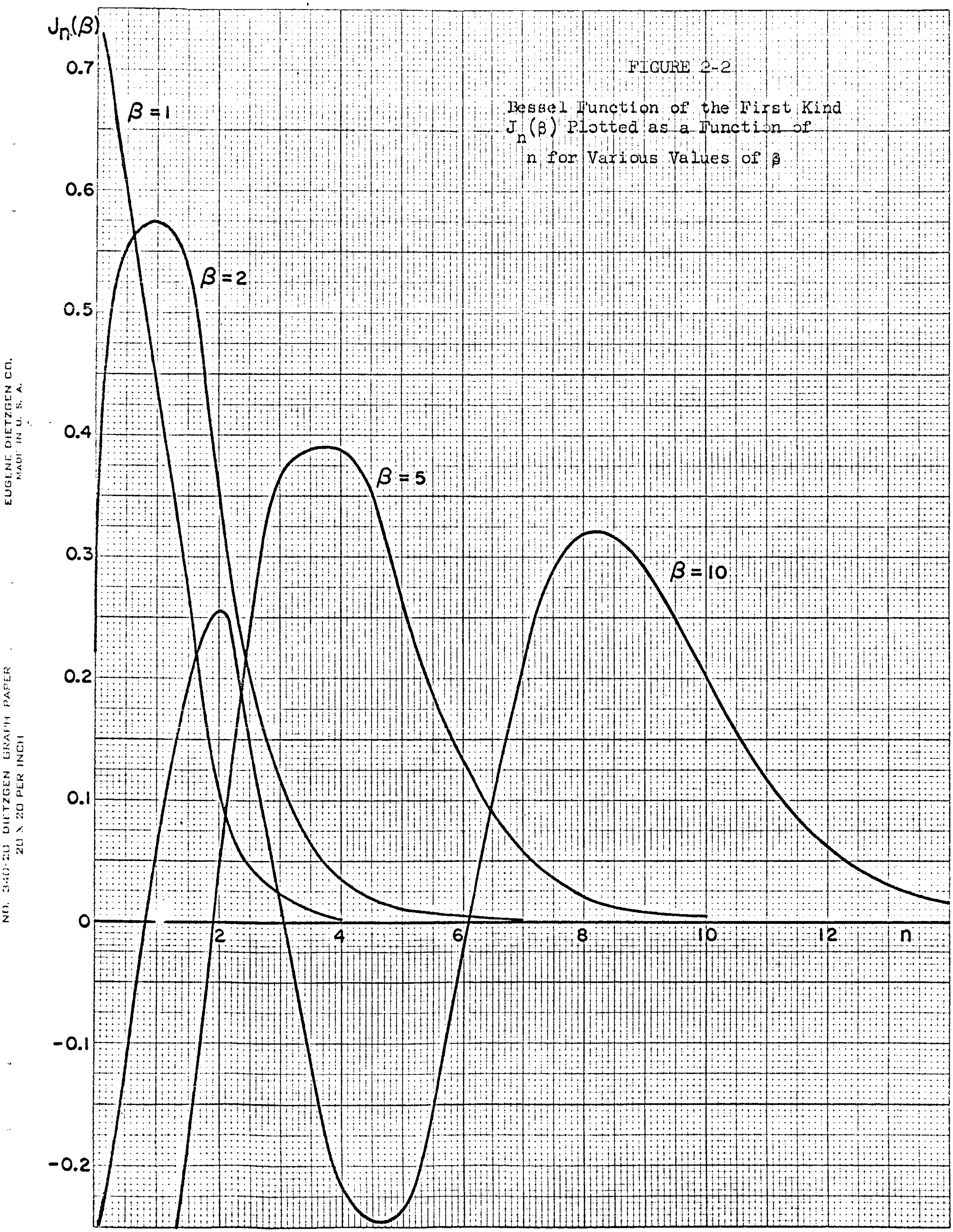


FIGURE 2-2
Bessel Function of the First Kind
$J_n(\beta)$ Plotted as a Function of
n for Various Values of β
$J_n(\beta)$
β = 1
β = 2
β = 5
β = 10
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
-0.1
-0.2
2
4
6
8
10
12
n
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instantaneous frequency, or phase of the carrier, in exact correspondence with the information waveform; that is, the applied voltage vs frequency of carrier characteristic must be perfectly linear. Departures from linearity have the effect of distorting the information waveform and increasing the bandwidth. The modulated output waveform will therefore have a broader bandwidth. As pointed out above, wide band FM waveforms have bandwidths determined mainly by the peak deviation. In this case, for a fixed deviation, the information signal bandwidth will not significantly affect the overall bandwidth. For narrow-band signals, however, the bandwidth is largely determined by the modulating signal, and the effects may be as great as in the AM case. An example involving an indirect FM generation technique will demonstrate some of these points. FM waveforms are sometimes developed by generating, initially, a low modulation index signal using adders and product modulators (Ref. 4, pp. 116-120). If the audio modulation signal is a sinusoid, the form developed is

$$y(t) = A_f \left[\cos \omega_c t - \beta \sin \omega_m t \sin \omega_c t\right] \tag{2-14}$$

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

$$y(t) = A_f\left[1 + \beta^2 \sin^2 \omega_m t \right]^{1/2} \cos\left[\omega_c t + \tan^{-1}(\beta \sin \omega_m t)\right] \tag{2-15}$$

When $\beta \ll 1$ this is approximately

$$y(t) \cong A_f \cos\left[\omega_c t + \beta \sin \omega_m t\right] \tag{2-16}$$

which is an FM wave with modulation index β. When high deviation FM signals are required (2-14) is generated initially, maintaining β sufficiently small. The narrow-band signal is then converted to a wide band signal by frequency multiplication. Even if β in (2-14) is not small, the amplitude factor in (2-15) can be made constant by the use of a limiter. However, if β is not small, the phase factor will carry some distortion components. We will use, for illustrative purposes, the first two terms of an expansion for the arc tangent to give in place of (2-15) (assuming the amplitude factor has been made constant by limiting).

$$y(t) \cong A_f \cos\left[\omega_c t + \beta \sin \omega_m t - \frac{\beta^3}{3} \sin^3 \omega_m t\right]$$

$$= A_f \cos\left[\omega_c t + \left(\beta - \frac{\beta^3}{4}\right) \sin \omega_m t + \frac{\beta^3}{12} \sin 3\omega_m t\right] \qquad (2\text{-}17)$$

The phase, which represents the modulation, now contains a third harmonic term which for β=1 is about 10% of the desired fundamental term. When an FM wave is developed using (2-14), β is kept less than 0.5; for β = 0.5 the third harmonic distortion is in the order of 1%. Thus, the modulation index used in the initial narrow band FM signal should be made as low as possible if excessive bandwidth is not to be consumed.

A recent study by the General Electric Company (Ref. 8) may be found useful in the control of nonlinearity. Two methods recommended here are (1) predistortion, where the input signals are distorted initially in a manner which is inverse to the distortion of the modulator, and (2) feedback to degenerate unwanted components. It should also be noted that to generate the form (2-14) requires a multiplier as in the generation of

AM. The observations made in the AM case about imperfect multipliers is applicable here too.

A final point to be noted concerns modulation limiting. As in AM generation, the audio input to FM systems will be limited to avoid overmodulation. If the limited output is applied directly to the modulator, the high frequency components in the amplitude limited waveform will give rise to unnecessary FM waveform sidebands. It is evident that filtering after amplitude-limiting is essential to reduce the information signal spectrum and the spectrum of the final FM output. This requirement is also more important in the case of narrow-band FM than it is for wide-band FM.

2.1.3 Spectral Properties of Pulse Modulation Systems

Information transmitted by perfectly rectangular pulses will require more spectral space than is ordinarily tolerable. It is well known that a rectangular pulse of width τ_r and height A_r has a Fourier Transform given by (Ref. 5, p. 41).

$$X_r(f) = A_r\tau_r \frac{\sin(\pi\tau_r f)}{\pi\tau_r f} \equiv A_r\tau_r \operatorname{sinc}(\pi\tau_r f) \qquad (2\text{-}18)$$

where

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

(An RF pulse of the same shape and of mid-band frequency f_o will have a spectrum $\frac{1}{2}\left[X(f-f_o) + X(f+f_o)\right]$). The envelope of the sinc function falls off as $\frac{1}{f}$ for higher frequencies; that is, it falls off at the rate of 6db per octave. This rather slow fall off rate implies relatively large levels overlapping into adjacent channels. In practice however, slower rise time pulses will be obtained either by forming such pulses initially or by passing rectangular pulses through a band limiting filter. Filters which absolutely limit the band will cause some crosstalk among adjacent pulses. In general, the requirements of low crosstalk and low level spectral tails are opposed to one another. It is possible, however, to find a pulse shape for which both are adequately small. For instance, pulses in the shape of a cosine cycle with pulse width τ_c and height A_c, are given by

$$x(t) = \frac{A_c}{2}\left(1 + \cos\frac{2\pi t}{\tau_c}\right), \quad -\frac{\tau_c}{2} < t < \frac{\tau_c}{2} \qquad (2\text{-}19)$$

$$= 0 \qquad\qquad , \text{ elsewhere}$$

and have a spectrum given by the Fourier transform

$$X_c(f) = A_c\tau_c \quad \frac{\operatorname{sinc}(\pi\tau_c f)}{2(1 - \tau_c^{\,2}f^2)} \qquad (2\text{-}20)$$

where

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

For frequencies sufficiently larger than $\frac{1}{\tau_c}$, the inverse of the pulse width (i.e., where $(\tau_c f)^2 \gg 1$), the spectrum falls off as $\frac{1}{f^3}$ or 18 db per octave. A comparison of the spectrum of a square pulse and a cosine pulse of two different widths is shown in Fig. 2-3. The spectra have been plotted with A_c chosen so that the energy of all the pulses being compared is the same. For equal energy it turns out that $A_c = 1.635\, A_r \sqrt{\tau_r/\tau_c}$. The rectangular pulse is shown as pulse (a) in Fig. 2-3 and its Fourier spectrum is shown in normalized form as $X_r(f)/A_r\tau_r$ in the figure. Cosine pulse (b) in Fig. 2-3 is chosen to have the same width as the rectangular pulse ($\tau_c = \tau_r$). Its spectrum is defined as $Xc_1(f)$. Its value is given by (2-20) and it is plotted in a normalized form as $Xc_1(f)/A_r\tau_r$ in Fig. 2-3. It will be noted that the spectrum of the cosine pulse is concentrated at low frequencies and has a greater (3 db) bandwidth than the rectangular pulse. At higher frequencies, however, ($f > 2/\tau_r$), the cosine pulse spectral components are comparatively very small. In practice the duration of the cosine pulse might be made longer than the duration of the square pulse. Accordingly, a cosine pulse of twice the duration of the rectangular pulse ($\tau_c = 2\tau_r$) is chosen for illustration and is shown as pulse (c) in Fig. 2-3. Its spectrum is defined as $Xc_2(f)$ and is plotted in Fig. 2-3 in the normalized form $Xc_2(f)/A_r\tau_r$. The spectrum in this case is comparatively quite small for $f > 1/\tau_r$.

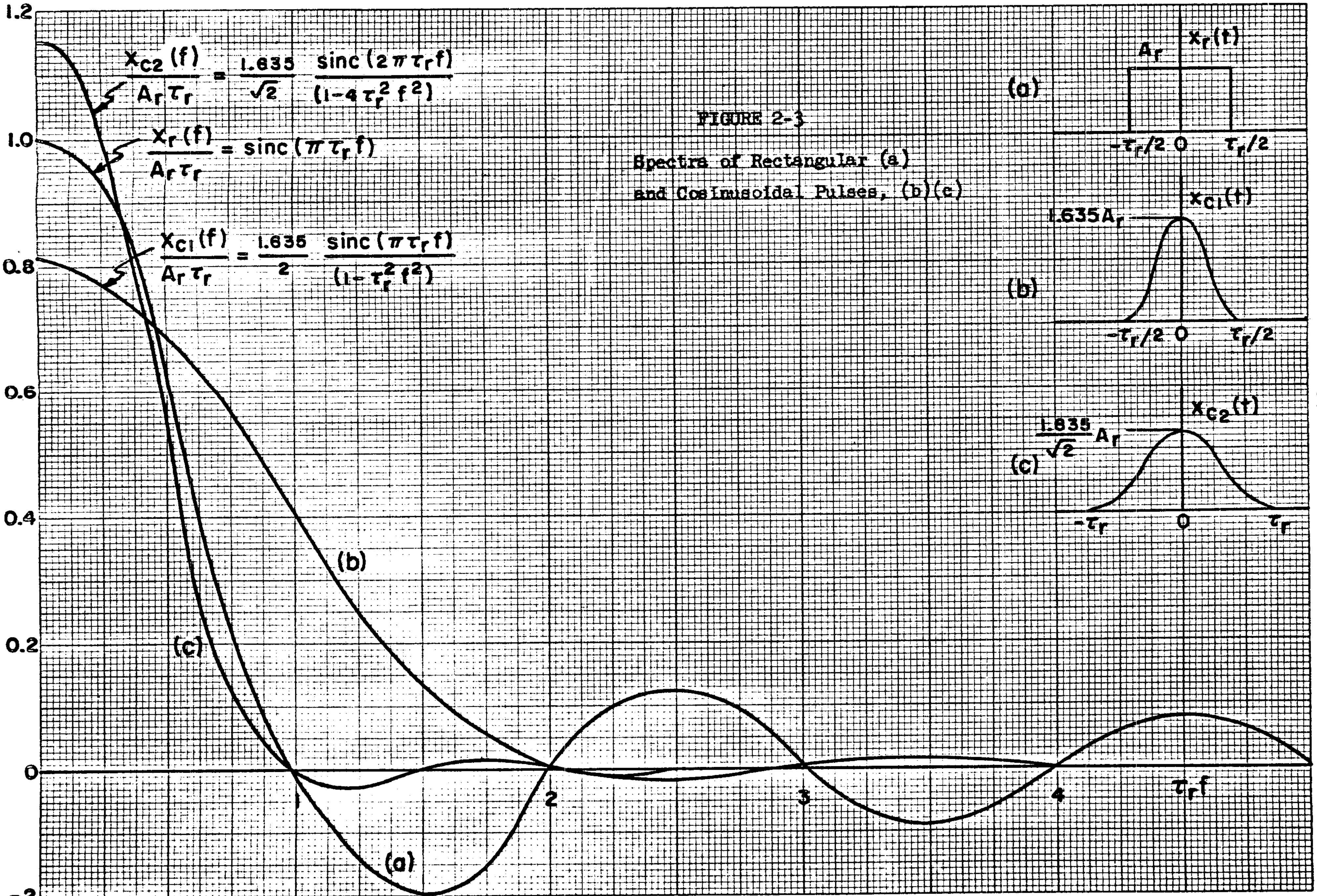


FIGURE 2-3

Spectra of Rectangular (a)
and Cosinusoidal Pulses, (b)(c)

2.1.4 Control of Splatter--Estimates of Levels

Reduction of sideband splatter requires that attention be given to the modulator to assure linear variation of the modulation parameter (the amplitude or angle). Balanced modulators in AM and related systems will be useful to eliminate the sideband components arising from even harmonic distortion of the modulating function. To minimize splatter generally in such systems it is necessary to use modulators which are as nearly square law as possible. Perfect rectifier modulators with very large RF driving voltage can also be shown to give ideal linear modulation. For modulators operating at low levels it should be possible to come reasonably close to the ideal situation. In angle modulation systems where the angle or frequency deviation is obtained by varying a reactive component in an oscillator, or where it is obtained by the indirect method discussed in Section 2.1.2, linearity is better for small initial deviation. Large initial deviation will give rise to nonlinear modulation unless predistortion or feedback is used. It would appear that the methods of angle modulation which involves modulating a carrier with a pulse-position signal developed from the information waveform (see Ref. 5, p. 388) can be made linear over a wide deviation.

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

$$H(f) \doteq \frac{B}{\left[B^2 + 4(f-f_o)^2\right]^{1/2}} \tag{2-21}$$

where $B = \frac{1}{2\pi RC}$, the 3 db bandwidth

$f_o = \frac{1}{2\pi\sqrt{LC}}$, the center frequency.

If the splatter frequency, f, is removed from the bandcenter by several bandwidths (2-21) can be reduced to

$$H(f) \doteq \frac{B}{2|f-f_o|} \tag{2-22}$$

If there are k stages the relative response at the output is

$$H_k(f) = \left[\frac{B}{2|f-f_o|}\right]^k \tag{2-23}$$

For instance, if a receiver operating at 136 MHz is exposed to a command transmitter at 148 MHz which has 3 tuned circuits each of 2 MHz bandwidth following the modulator the effect of the tuned circuits is to reduce any 136 MHz splatter components by a factor

$$H_3(f) = \left[\frac{2}{2(148-136)}\right]^3 \cong \frac{1}{1728} \quad \text{(approximately -65 db)}$$

Where high level modulation is used there will be fewer tuned circuits to attenuate the splatter components unless special attention is given to this matter. Note that a single tuned circuit in the example given above attenuates any undesired 136 MHz sideband by only 21.6 db.

To complete the estimate of the magnitude of undesired splatter which would reach a receiver requires now (a) the level of the splatter component at the point of generation, and (b) the attenuation in transmission. The former was discussed in sections 2.1.1 through 2.1.3. While calculated estimates are possible, in principle, it is better practice, especially for components far from the center frequency, to measure the levels at the point of generation. In this way the elusive parameters which are required for calculation do not have to be determined explicitly (to determine them would require measurement, too). Rough estimates using representative figures would, however, be useful in preliminary work. An estimate of this sort is now made assuming pulse transmission.

Suppose a rectangular RF pulse centered at f_o is generated as described in section 2.1.3, with a spectrum given by $X(f) = \frac{1}{2}\left[X_r(f-f_o) + X_r(f+f_o)\right]$; $X_r(f)$ is given in (2-18). The energy of the pulse is

$$E = A_r^2 \tau_r / 2 \tag{2-24}$$

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

$$\left|X(f_c)\right| \cong \frac{1}{2} \frac{A_r}{\pi\left|\,|f_c| - f_o\right|} \tag{2-25}$$

The energy of the pulse components selected by the receiver is

$$E_r = \frac{1}{2} \frac{A_r^2 B}{\pi^2\left[\,|f_c| - f_o\right]^2} \tag{2-26}$$

This energy relative to the total energy in the pulse is

---

* B as used here is the effective power bandwidth as defined in eq. (4-2). A receiver with rectangular bandpass of width B Hertz has an effective power bandwidth equal to B Hertz.

$$\frac{E_r}{E} = \frac{B}{2\pi^2 \tau_r \left[\left| f_c \right| - f_o \right]^2} \qquad (2\text{-}27)$$

If, for instance $B = 0.25$ MHz, $\tau_r = 10^{-6}$ seconds, $\left| f_c \right| - f_o = 12$ MHz, then

$$\frac{E_r}{E} = \frac{1}{8\pi^2 (12)^2} \cong .875 \times 10^{-4}$$

or

$$\frac{E_r}{E}\,(\text{db}) = 10 \log(1.75 \times 10^{-4}) = -40.6 \text{ db} \quad ;$$

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

If the rectangular pulses were fed through a succession of 3 single tuned circuits prior to transmission as described above, the sideband components in the 1/4 MHz band would be 65 + 41 = 106 db below the pulse power transmitted (the tuned circuits are here assumed to produce no significant effect on the total pulse power).

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

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

2.2 HARMONIC GENERATION

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

It will often be possible to find modes of operation for the active devices which minimize the relative harmonic amplitude--as a rule, at the cost of efficiency. The extent to which these efforts are made depends on the amount of filtering. Low level circuits located far back in the transmitter and followed by many selective circuits are unlikely to result in significant final harmonic output. Some of these mechanisms will now be described in more detail.

Modulators for AM signals and signals related to AM are, as pointed out earlier, multipliers. This function is performed perfectly by a square law device when the sum of low frequency signal and carrier are delivered to it. If $x_1(t)$ is the IF signal and $\cos \omega_c t$ is the carrier, then the output of a square law device to the sum input is

$$\left[x_1(t) + \cos \omega_c t\right]^2 = x_1^2(t) + 2x_1(t) \cos \omega_c t + \frac{1}{2} + \frac{1}{2} \cos 2\omega_c t \qquad (2\text{-}28)$$

The desired term is the term $2x_1(t) \cos \omega_c t$. The term $x_1^2(t)$ is a distorted low frequency signal which is readily rejected in the RF filter following the modulator (the dc term, $\frac{1}{2}$ is obviously also rejected). The second harmonic term may trickle through especially if there is not much selectivity to the output. A solution is the balanced modulator as shown in Fig. 2-4. Assuming each nonlinear device has a power series characteristic given by

$$y = a_o + a_1 x + a_2 x^2 \qquad (2\text{-}29)$$

where y is the output and x is the input, the total output is

$$\begin{aligned} y_a - y_b &= a_1(x_a - x_b) + a_2(x_a^2 - x_b^2) \\ &= 2a_1 x_1(t) + a_2 \left[x_1(t) + \cos \omega_c t\right]^2 - a_2 \left[x_1(t) - \cos \omega_c t\right]^2 \\ &= 2a_1 x_1(t) + 4a_2 x_1(t) \cos \omega_c t \qquad (2\text{-}30) \end{aligned}$$

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

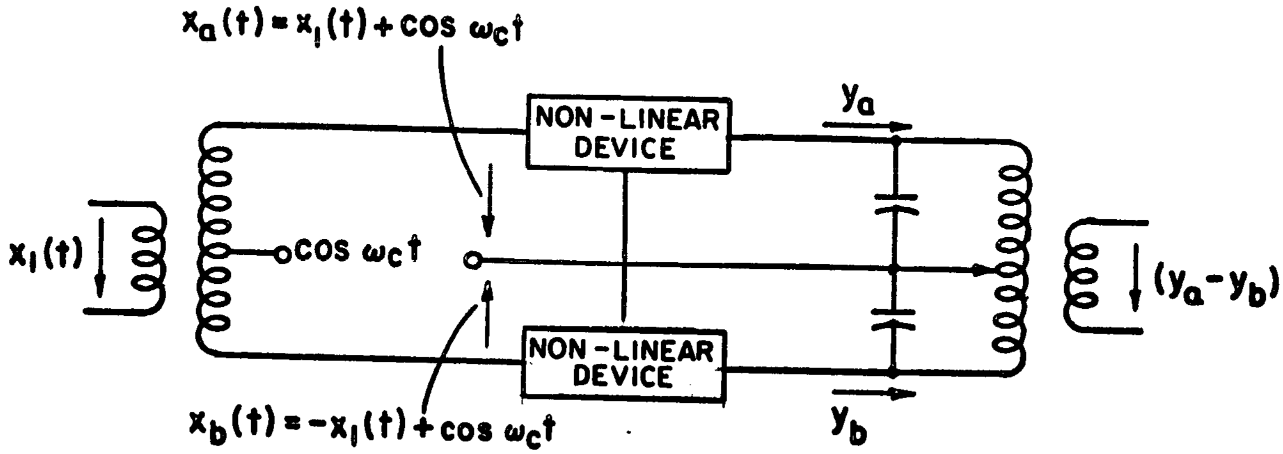


FIGURE 2-4

BALANCED MODULATOR

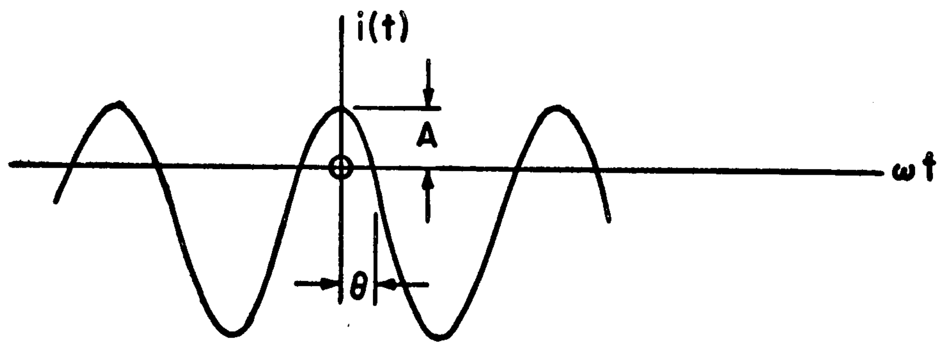


FIGURE 2-5

SINUSOIDAL CAP WAVEFORM

a_3x^3 were added to (2-29), the output would contain a term

$$2a_3 \cos^3 \omega_c t = \frac{3}{2} a_3 \cos \omega_c t + \frac{1}{2} a_3 \cos 3\omega_c t \tag{2-31}$$

which has a third harmonic component (the rejection of splatter components involving even powers of $x_1(t)$ can be illustrated also by adding a_3x^3 and expanding as in 2-30).

It is common practice to use crystal controlled oscillators operating at moderate frequencies (i.e. 30 MHz and below) and to employ frequency multiplication to obtain the needed output carrier frequency. In such systems, as well as in class B and C amplifiers, an electronic device biased at or below cutoff is the source of the output, and it is rich in harmonics. The current pulse in the output is approximately of the form of the cap of a sinusoid as shown in Fig. 2-5 and given by the expression

$$i(t) = \begin{cases} a\left[\cos \omega t - \cos \theta\right], \text{for } 2n\pi - \theta \leq \omega t \leq 2n\pi + \theta, \; n=0, \pm 1, \pm 2, \ldots \\ 0 \qquad , \text{otherwise} \end{cases} \tag{2-32}$$

The Fourier expansion of such a wave is given by

$$i(t) = \frac{A}{\pi(1-\cos \theta)} \left\{ \sin \theta - \theta \cos \theta + \sum_{n=1}^{\infty} \left[\frac{\sin(n+1)\theta}{n+1} + \frac{\sin(n-1)\theta}{n-1} + \frac{2 \sin n\theta \cos \theta}{n} \right] \cos n\omega t \right\} \tag{2-33}$$

The harmonic magnitude determined from (2-33) is plotted in Fig. 2-6 as a function of the conduction angle 2θ. It will often be found that the sinusoidal cap is itself somewhat distorted in passing through the amplifier giving rise to even higher levels of harmonics. Shown on Fig. 2-6 also is the harmonic magnitude of squared sinusoidal caps, given by the expression

$$i(t) = \begin{cases} A^2\left[\cos\omega t - \cos\theta\right]^2, & \text{for } 2n\pi - \theta \leq \omega t \leq 2n\pi + \theta,\ n=0, \pm1, \pm2, \ldots \\ 0 & , \text{ otherwise} \end{cases} \tag{2-34}$$

which may be a better approximation of the output current pulses from some electronic devices than the undistorted caps of Fig. 2-5 (Ref. 9). In general, the larger the conduction angle, the smaller the harmonic output. It is in principle possible to estimate the harmonic output of the transmitter using the levels determined by (2-33) or Fig. 2-6 and the response characteristic of the tuned circuits following the source of harmonics. Such a procedure is described in Ref. 7, but the equations of Section 2.1 provide the necessary forms. For rough approximations, a single tuned circuit with a Q=10 will attenuate the 2nd, 3rd, and 4th harmonic approximately 24, 30, and 33 db respectively. Doubling the Q will add about 6 db of attenuation to each.

It was pointed out earlier that the predominant contribution of harmonic outputs will, in the usual case, be the final amplifier(s). The reason for this is that (1) the final amplifier is ordinarily driven hard in order to get as much efficiency as is possible, and (2) the selectivity of the tuned circuits following the final amplifier, particularly at lower frequencies, is limited.

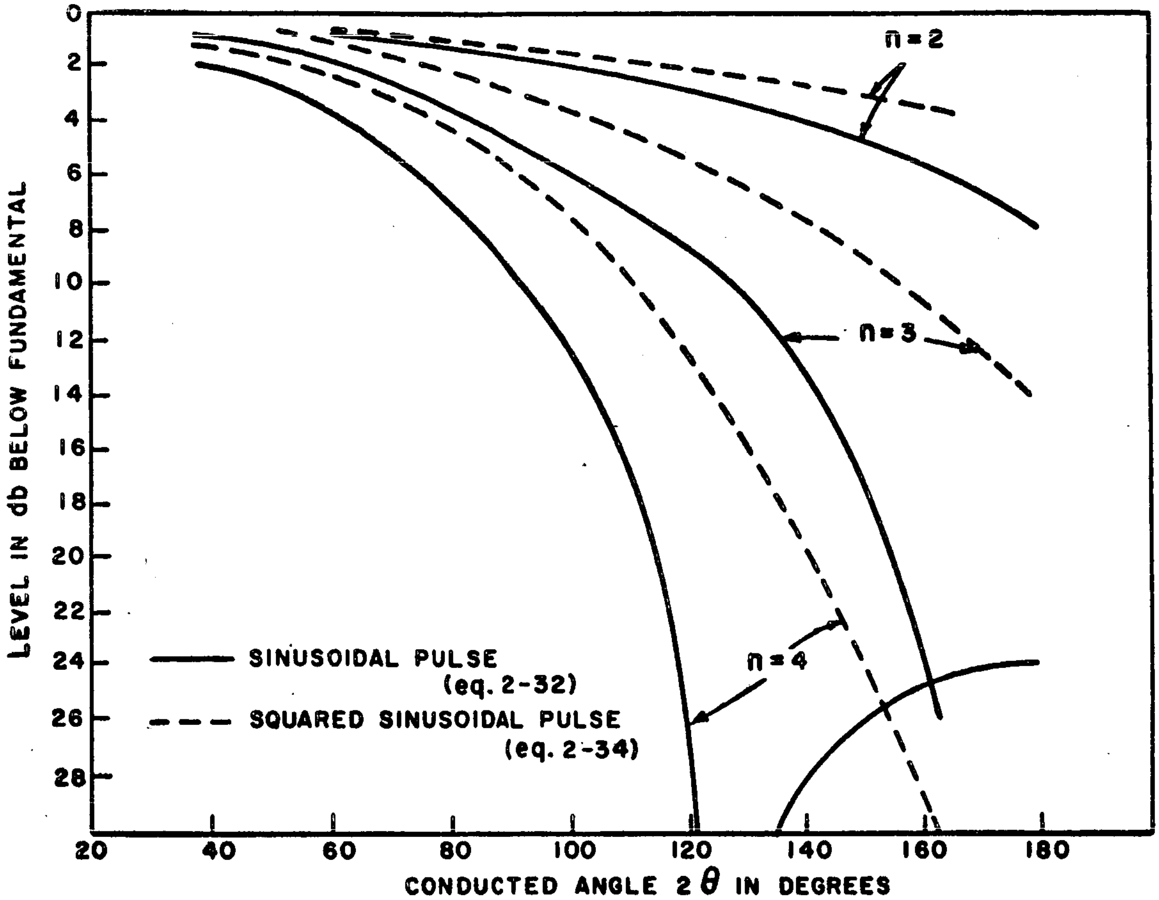
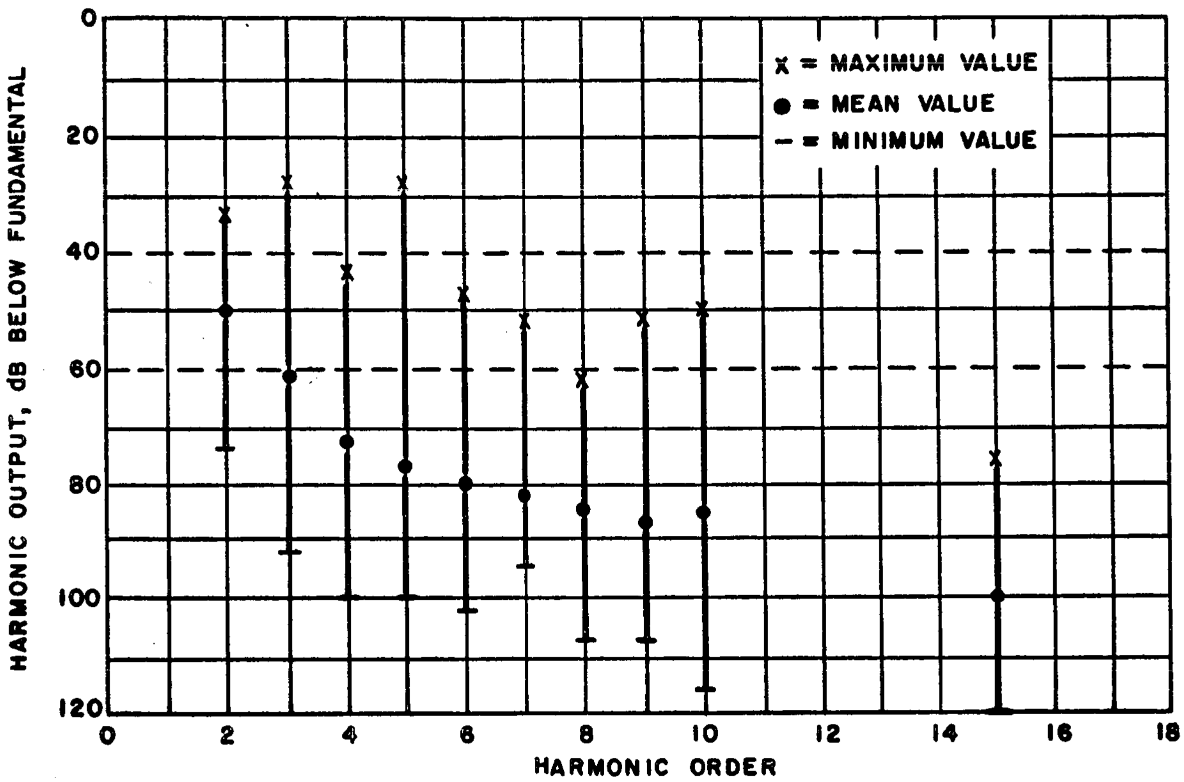


FIGURE 2-6

HARMONIC INTENSITY AS A FUNCTION OF CONDUCTION ANGLE



FCC LIMITS

10 KHz to 30 MHz: -40 db BUT LESS THAN 50 MILLIWATTS

30 MHz to 235 MHz: -60 db BUT LESS THAN 1 MILLIWATT IF $P_0 > 25$ w -40 db BUT LESS THAN 25 MICROWATTS IF $P_0 < 25$ w. P_0 = TRANSMITTER POWER

FIGURE 2-7

A FEW OF THE MANY OUTPUTS OF 14 COMMUNICATIONS TRANSMITTERS. ULTIMATE LEVEL DECREASE TO -100 db IN THE VICINITY OF THE 15th HARMONIC.

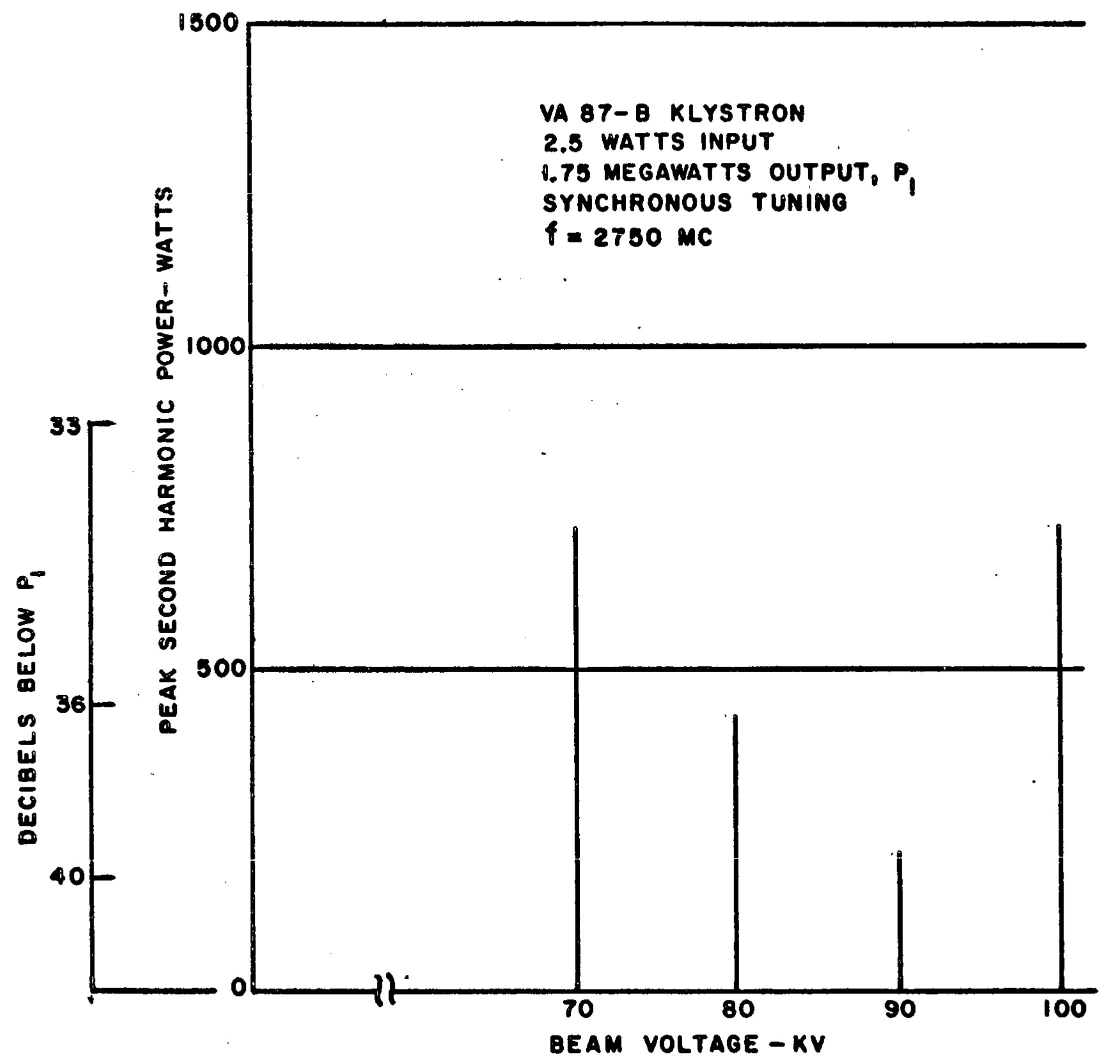


FIGURE 2-8

VA 87-B KLYSTRON HARMONIC OUTPUT AS A FUNCTION OF BEAM VOLTAGE

Tube type amplifiers operate class C, as a rule, and are not different, in principle, in harmonic output from the frequency multipliers discussed above. As the conduction angle is increased, the harmonic output decreases but so does the efficiency. It would obviously be advantageous to use push-pull class B from an interference viewpoint to give the effect of a linear amplifier. It may even be advantageous to use push-pull with class C amplifiers since this would tend to cancel the second harmonic (and other even harmonics). The third harmonic (and other odd harmonics) would be unaffected relative to the fundamental and is also more readily attenuated in the output tuned circuits. Shown in Fig. 2-7 are the harmonic outputs of 14 representative communication transmitters (Ref. 10). Indicated also on the figure are the FCC limits for harmonic output. Note that some of these transmitters have harmonics which are within 30 db of the fundamental; a 50 kw transmitter will be putting out a 50 watt signal at the harmonic frequency--in this case hardly significant.

Amplifiers for microwave frequencies are also generators of harmonic output though the mechanisms are different. The klystron which is an important high power amplifying device today is inherently a harmonic source. The bunching process develops moving charge concentrations which pass the catcher in the form of sharp bursts. The induced current in the catcher cavity is highly non-sinusoidal. It has been shown that the induced current has a harmonic intensity given approximately by (Ref. 11).

$$I_c(n) = 2I_o J_n(nx) \qquad (2\text{-}35)$$

where

n = the order of the harmonic

I_o = the direct current in the tube

J_n = the n'th order Bessel function of the first kind

x = bunching parameters = $\omega s V_1 / 2V_o^{3/2}\ 2e/m$

ω = frequency of excitation

s = spacing between input and output cavities

V_o = beam accelerating voltage

e = electronic charge

m = electronic mass

V_1 = peak gap voltage at buncher

For maximum efficiency (which occurs at $x = 1.84$) the induced current pulses are very non-sinusoidal and the harmonic current is very high. Much depends on the Q of the cavity to avoid large harmonic outputs.

Loaded Q's ranging from 500-1000 are obtainable, and even with equal fundamental and second harmonic current in the tube it should be possible to get harmonic outputs more than 60 db below the fundamental. Measurements reported on a particular pulsed klystron (Ref. 12) are reproduced in Fig. 2-8. Variation of beam voltage which alters the bunching parameters is seen to markedly alter the second harmonic. Even at the value of beam voltage for which the second harmonic is minimum, it is only about 40 db below the fundamental output here.

The traveling wave tube as a microwave power amplifier would appear to be a good choice from the interference generation viewpoint. The fundamental phenomenon does not involve impulsive re-enforcement of a field as in the klystron but a uniform re-enforcement of a traveling wave. Yet reports (Ref. 12) indicate second harmonic output in particular cases ranging from 20 to 40 db below the fundamental. It is possible that these conditions obtain when maximum efficiency is sought, and when the input is driven heavily. The traveling wave tube input-output characteristic, for specific tubes, was found to be linear over a range of inputs and to saturate moderately slowly beyond the linear range. Operating into saturation will give rise to maximum output but the harmonic content will increase too.

High power magnetrons are generally used as sources of pulsed RF energy coupling directly to the antenna. In the magnetron the energy of a rotating cloud of electrons is imparted to the field in tuned cavities on the periphery of rotation. The mechanism which selects electrons emitted from the cathode, where they are roughly in proper phase, tends to maintain them in bunches as in the klystron. The spatial distribution of charge is not sinusoidal and, again, pulses are induced in the cavities rather than sinusoids. The magnetron cavities can also be excited in an harmonic mode so that a large number of output components are possible in addition to the easily identifiable harmonics. Level and frequency are both difficult to predict. A figure of 40 db below the fundamental has been quoted (Ref. 12) for the second harmonic, but the third harmonic was said to be able to reach -20 db.

From the foregoing it is evident that harmonic intensities can be reduced by (1) choosing linear operating regions and thereby working at reduced efficiencies, (2) cancelling nonlinear components in balanced circuits, and (3) filtering. The first of these involves an increase in power consumption and heat generation. As a result more effort must be spent on heat removal. The other two involve additional circuitry and the balanced circuit eliminates only even degree harmonics. Filtering is, by far, the most popular method for suppressing harmonics and we discuss some techniques below.

One way of suppressing harmonics is by utilization of the wave trap or bypass principle. In the case of a transmission line system, the filter elements may take the form of shunt or series stubs. See Fig. 2-9 for an indication of how this is accomplished.

The difficulty with such techniques as that just mentioned is that the harmonic power is reflected back to the generator which may misbehave because of the presence of these extra reactive volt-amperes. One

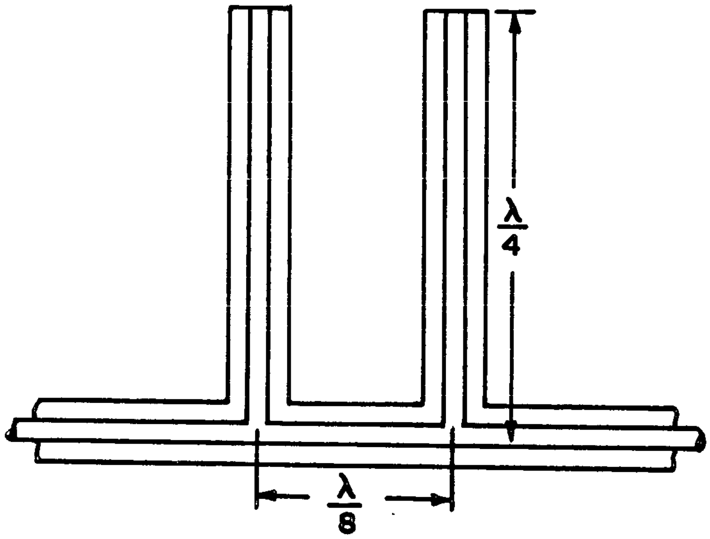


FIGURE 2-9

TWO QUARTER-WAVE SHORT-CIRCUITED STUBS SPACED ONE EIGHTH OF A WAVELENGTH AT THE FUNDAMENTAL ARE EFFECTIVE IN ATTENUATING THE SECOND HARMONIC

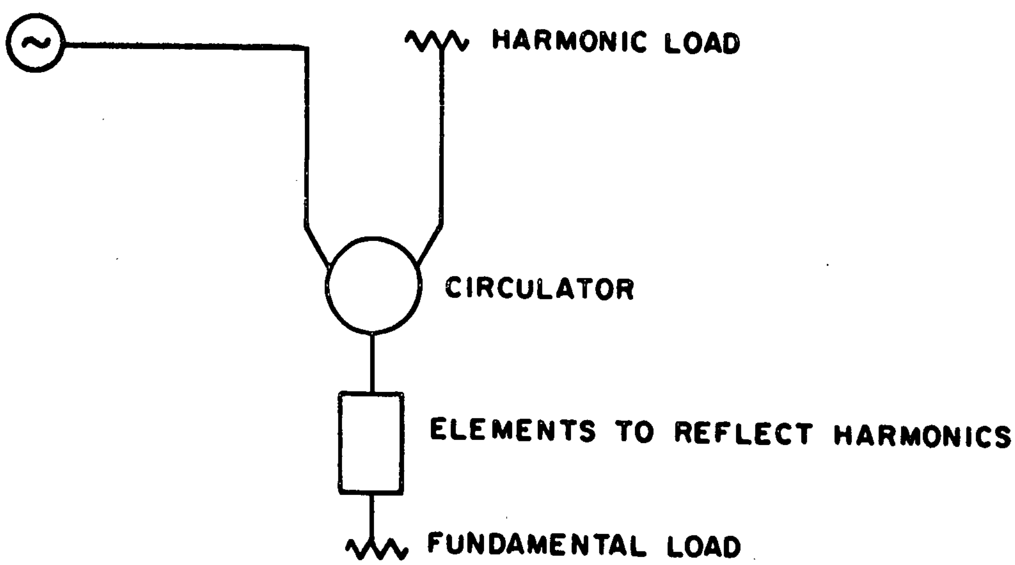


FIGURE 2-10

CONCEPT OF USING CIRCULATOR TO SEPARATE HARMONICS way of avoiding this difficulty in theory would be to use an isolator between the generator and filter. Unfortunately, most isolators are designed to work as isolators at the fundamental frequency, and their performance at harmonic frequencies is unspecified. This becomes increasingly true at the higher microwave frequencies. Assuming for a moment that an isolator technique would be used, it is noted that the dissipative properties of the device would need to be such that all reflected components (fundamental plus harmonics) could be absorbed without exceeding the rating.

It might be possible to utilize a ferrite circulator in the harmonic suppression problem. If, for example, the configuration of Fig. 2-10, is used, the reflected energy could be diverted into a resistive load. Unfortunately, the circulator has some of the same bandwidth problems as does the isolator, and such a technique cannot be realized everywhere in the spectrum.

Another harmonic filtering and absorbing technique which has received some attention is that of directionally coupling the unwanted signals into a second waveguide where they are absorbed. What can and cannot be done in this regard is highly dependent upon the specific problem at hand, and further generalization is pointless.

It should be pointed out that merely coupling a smaller size waveguide, which is below cutoff for the fundamental but above cutoff for all harmonics, to the main waveguide may provide an alternative path for harmonic absorption and some reduction of their level in the main load.

Many useful filtering techniques will be found discussed in depth in references 1 and 13.

2.3 TRANSMITTER NOISE

While with care in the design of the modulator the spurious sidebands discussed in Section 2.1 can be reduced, it is frequently still found that a significant background level around the center frequency exists. Now, however, it is the result of random noise modulating the carrier in amplitude, frequency, or phase. The noise is often associated with the oscillator itself which, compared to amplifiers, is quite noisy, and to some extent with the amplifiers as well. It is possible for corona to form or arcing to occur at high voltage points in the final amplifier. Subsequent filtering in the output tank circuit results in a pair of noise sidebands surrounding the carrier frequency. Estimates of levels will be difficult to make. Measurements are more appropriate though they are difficult to make because of the presence of the much larger main transmitter output. Some measurements have been reported on VHF transmitters (Ref. 14) and the results of tests on a 45 MHz and a 160 MHz transmitter are shown in Fig. 2-11. The basic crystal frequencies were 3.75 MHz and 6.67 MHz, respectively, and 3 or 4 frequency multiplier stages were required before the final amplifier. The tests were carried out on an unmodulated transmitter and the effect of modulation on the noise sidebands was not determined. It was determined, however, that the power amplifier alone contributed in the order of 20 db per KHz to the noise output. It will be noted that the levels are quite low in this typical case; the mechanism is significant when transmitter and receiver antennas are very close to one another.

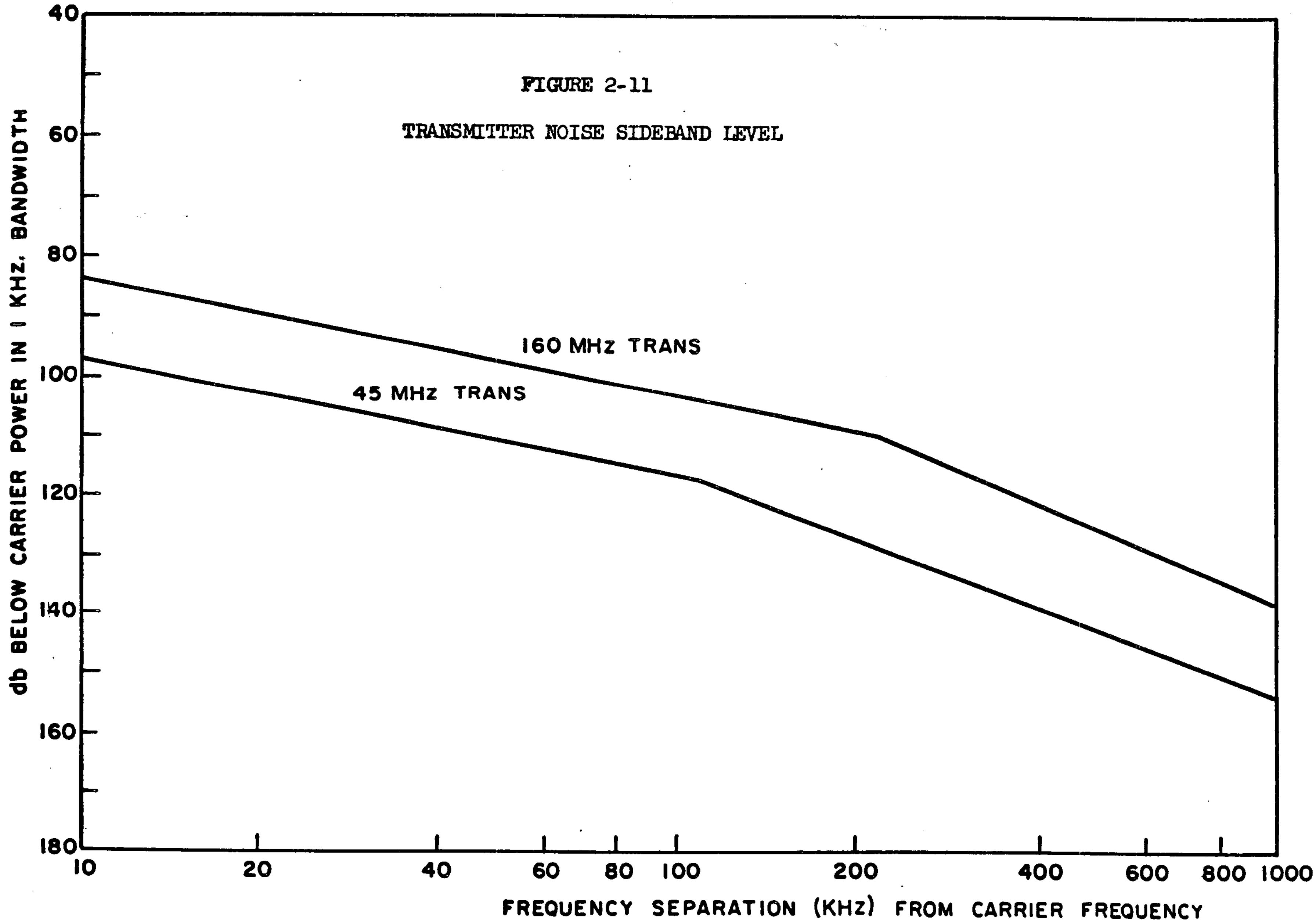


FIGURE 2-11
TRANSMITTER NOISE SIDEBAND LEVEL
160 MHz TRANS
45 MHz TRANS
db BELOW CARRIER POWER IN 1 KHz, BANDWIDTH
40
60
80
100
120
140
160
180
10
20
40
60
80
100
200
400
600
800
1000
FREQUENCY SEPARATION (KHz) FROM CARRIER FREQUENCY

2.4 INTERMODULATION AND CROSS-MODULATION

The phenomena of intermodulation and cross-modulation imply the mixing of two or more signals in a nonlinear element in such a way that multiplicative mixtures of the two signals arise. The mechanism arises in receiver input circuits as well as in transmitters, and sometimes in some nonlinear element in the channel. As far as transmitters are concerned, the process involves picking up an unwanted signal by way of the transmitting antenna, delivering it back to the output electrode of the final amplifier, and there having it mix with the transmitted signal. In effect, the unwanted signal appears as a variation in voltage across the final amplifier output. In some situations the output amplifier will not respond significantly to such variations as in the case of high-μ vacuum tubes. More generally there is a dependence and it is nonlinear. Since the output circuit is selective, there is a tendency to reject unwanted signals when they are far removed in frequency from the transmitter frequency. Furthermore, if the nonlinear action generates a component far removed from the transmitter frequency, it also tends to be rejected in the output tuned circuits. The process is therefore of greatest significance when both the unwanted signal and the nonlinear product are near the pass-band of the final amplifier. This combination of events can occur if the nonlinearity is of odd degree. For instance, suppose the output signal has a component which depends on the third power of the voltage at the output. Two components of voltage at the output are the output signal voltage itself,

$$x_1(t) = v_1(t) \cos\left[\omega_1 t + \phi_1(t)\right] \qquad (2\text{-}36)$$

and the unwanted signal voltage

$$x_2(t) = v_2(t) \cos\left[\omega_2 t + \phi_2(t)\right] \tag{2-37}$$

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

$$y_2(t) = v_1^{\,2}(t)\, v_2(t) \cos\left[(2\omega_1 - \omega_2)t + 2\phi_1(t) - \phi_2(t)\right] \tag{2-38}$$

which is termed an intermodulation component, and a component

$$y_b(t) = v_1^{\,2}(t)\, v_2(t) \left[\cos \omega_2 t + \phi_2(t)\right] \tag{2-39}$$

termed a cross-modulation component. The former is at a different center frequency than either ω_1 or ω_2; but, if the two frequencies are close to one another, the new term is at a frequency not too different. The modulation on this term is a combination of the modulations on each of the causative signals. The cross-modulation component is at the frequency of one or the other of the original signals and it contains an amplitude factor which is a combination of the original amplitude coefficient. The intermodulation is usually the more serious since it covers an entirely new frequency slot. The cross-modulation term shown in (2-39) implies a re-radiation of the unwanted signal with a distorted envelope. As a rule it is much smaller than the original unwanted signal and might be viewed as a small new splatter or noise component.

2.5 OTHER SPURIOUS OUTPUTS

Parasitic oscillations are known in low and high frequency amplifiers. In low frequency devices they are the result of stray external and internal capacities and inductances which form into spurious feedback loops. The cures are often very simple, sometimes involving inserting resistors in the leads to the grid and plate of the amplifier to make the spurious tuned circuit lossy. The methods of handling these situations are well enough described in standard reference books (e.g., Terman's Handbook).

At microwave frequencies corresponding phenomena exist. Often, the phenomenon is completely internal to the tube. Transit time effects are sources of negative resistance at some frequencies. If a tuned circuit exists somewhere in the structure which is coupled to the negative resistance, parasitic oscillations occur. An harmonic output resulting from excitation of unwanted modes in magnetrons was mentioned in Section 2.2. A number of these mechanisms are described more fully in Ref. 10.

3.0 UNINTENTIONAL NOISE GENERATION

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

The majority of these sources of noise are naturally concentrated in urban areas of population and industrial centers. By siting ground stations away from urban population centers, it is possible to virtually eliminate most of the noise originating from such sources. However, there will very probably be highways, containing vehicular traffic, as well as power lines, in the immediate vicinity of a ground station. Consequently, these forms of noise are of greater significance to station operations and will be considered in more detail. The remaining sources of interference referred to above, will also be considered briefly.

Before examining the different types of noise sources in detail, it is worth briefly examining the susceptibility of the different antenna systems used at the sites to these forms of interference. Since these noise sources are located on the ground, they will not significantly affect the fixed interferometer arrays used at the Minitrack interferometer sites owing to the fact that the main beam of these arrays is fixed and is pointing vertically upwards. This will not, of course, be true for any of the mechanically-movable antennas such as the SATAN (Satellite Automatic Tracking Antenna) array located at Minitrack sites and the 85 ft. parabolic dish located at Data Acquisition Facilities. These antennas may be affected by noise sources on the ground either when the antenna mainlobe is pointed at the horizon or whenever the sidelobes or backlobe are pointed at the horizon. However, the energy entering the antenna through its sidelobe will obviously be attenuated relative to that collected by the main lobe.

3.1 NOISE FROM VEHICLE IGNITION SYSTEMS

Noise from automobile and truck ignition systems undoubtedly represents the largest contribution to the overall man-made noise level at VHF and UHF frequencies. A number of workers (references 15-22) have made quantitative measurements on the noise levels radiated by automobiles and trucks and have attempted to evolve comprehensive theories for the explanation of radiation from ignition systems. Measurements made by George (Ref. 17) at 180 MHz and 450 MHz are reproduced in Figs. 3-1 and 3-2. Although this data is now 25 years old, it is probably still applicable to modern day automobiles and trucks. Most modern automobiles,

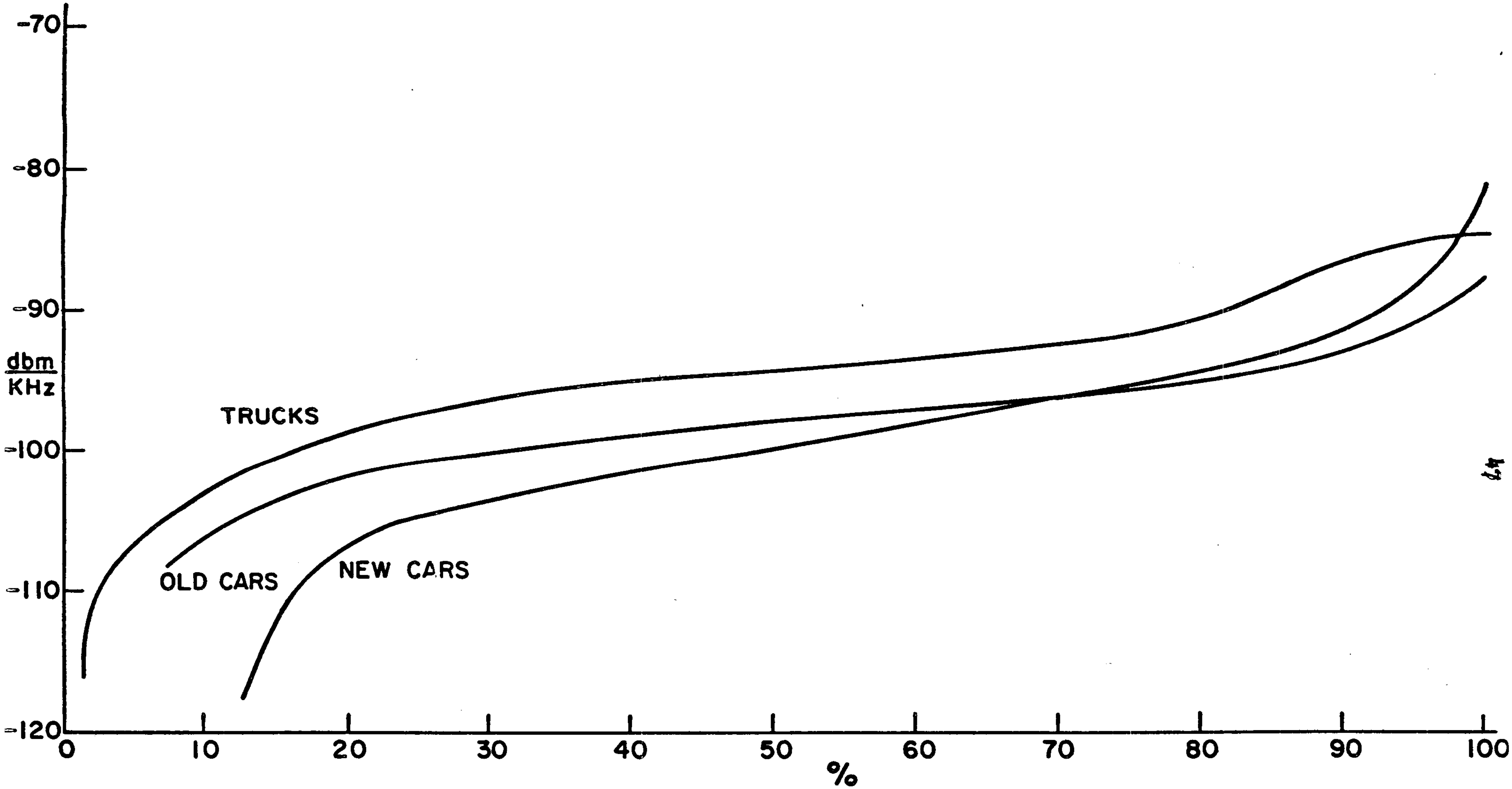


FIGURE 3-1

IGNITION NOISE AT 180 MHz, HORIZONTAL POLARIZATION. CURVES SHOW PERCENT OF CARS OR TRUCKS RADIATING LESS THAN THE INDICATED POWER. RECEIVING ANTENNA IS A $\lambda/2$ DIPOLE, 37' HIGH AND 100' FROM ROAD.

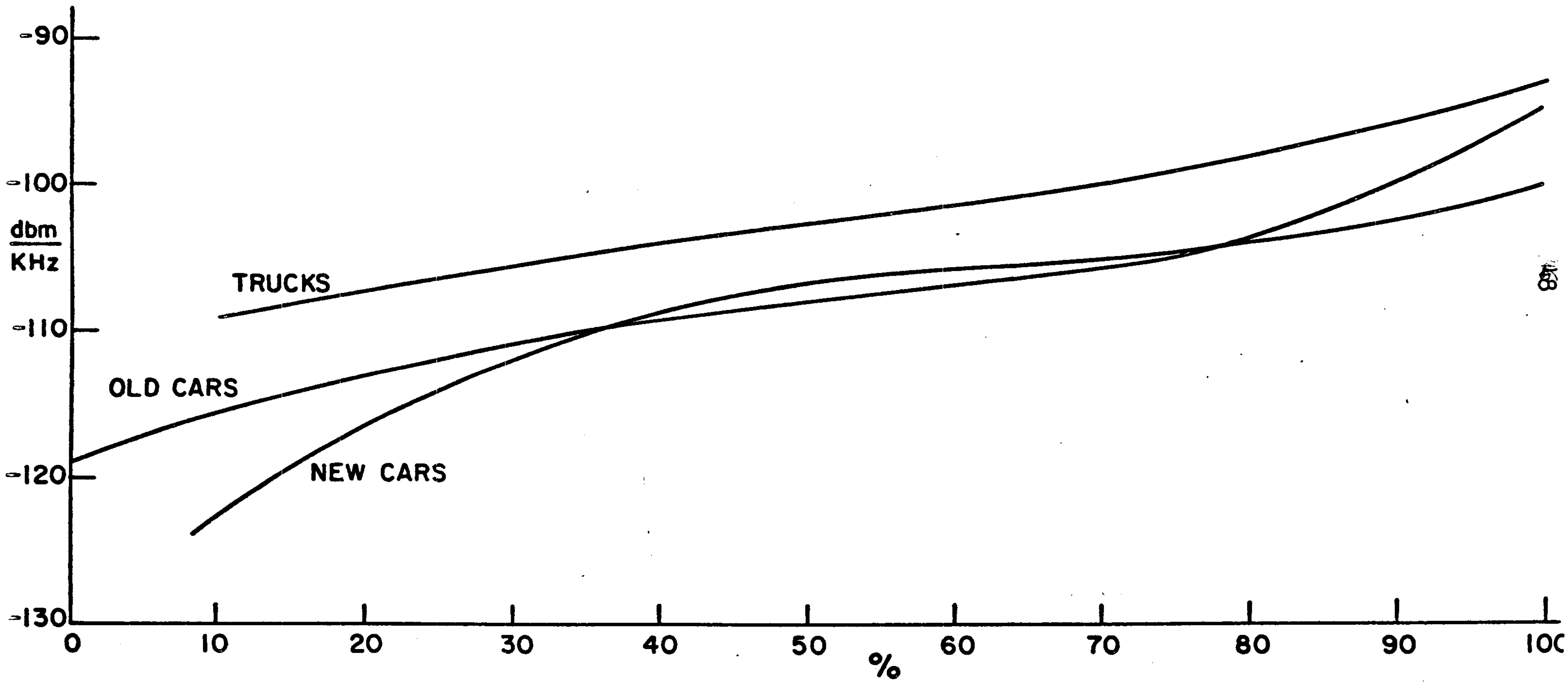


FIGURE 3-2

IGNITION NOISE AT 450 MHz, HORIZONTAL POLARIZATION. CURVES SHOW PERCENT OF CARS OR TRUCKS RADIATING LESS THAN THE INDICATED POWER. RECEIVING ANTENNA IS A $\lambda/2$ DIPOLE, 35' HIGH AND 100' FROM ROAD.

when delivered from the manufacturer are equipped with devices for the suppression of ignition interference, so that the levels of ignition interference for new cars is probably lower nowadays. On the other hand, the number of cars on the highways has greatly expanded since 1940, so that the overall noise level has probably increased significantly. From George's data, noise levels emanating from trucks is seen to be the most serious problem and this is still true today. Figure 3-3 shows the theoretical propagation curves for horizontal polarization over level ground. In practice, noise levels will probably fall off with distance at a faster rate due to absorption and screening by vegetation, foliage and trees, etc.

Ignition noise is created by the oscillatory current which is set up in the high tension leads when the energy stored in the self-capacitance of the spark plug, the high tension cables and the ignition coil, is rapidly discharged through the spark plug gap. The rapidly changing current then radiates energy from the high tension cables and also from coupled low voltage circuits. This current, called the capacitive component of the ignition spark, has a peaked wave shape and lasts for only a very short time (of the order of 4 nanoseconds for a 1 mm spark gap). This creates a continuous frequency spectrum distribution which covers a very wide part of the VHF/UHF frequency spectrum. The spectrum remains relatively flat up to at least 600 MHz, while some energy is usually still measurable at 1000 MHz. The actual spectrum distribution radiated by an automobile and the overall noise levels will actually depend on the duration of each spark discharge, the layout of the electrical equipment and wiring within the engine compartment, and the shielding effectiveness of the hood and

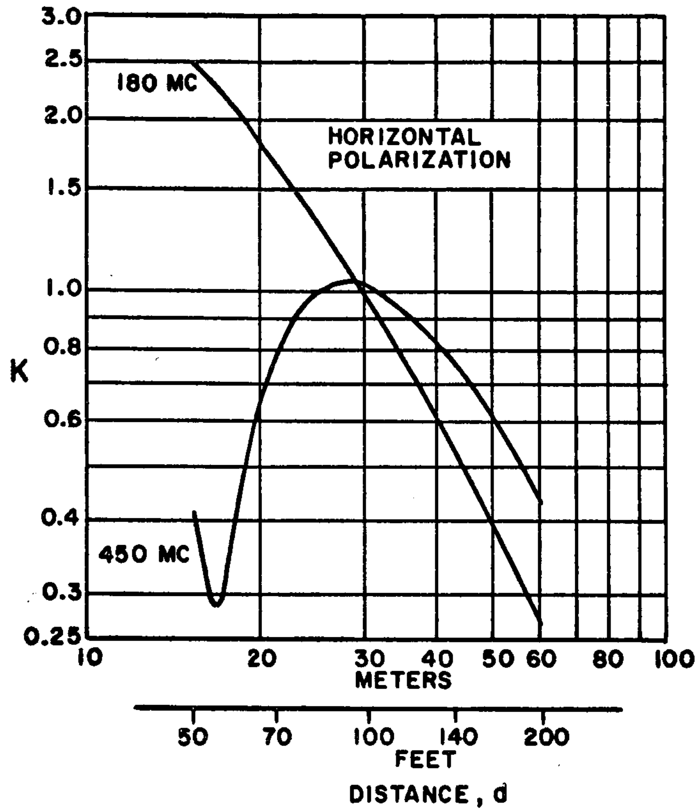


FIGURE 3-3

THEORETICAL PROPAGATION CURVES FOR HORIZONTAL POLARIZATION. FIELD STRENGTH AT DISTANCE d EQUALS THE FIELD STRENGTH MEASURED AT A DISTANCE OF 100 FEET, TIMES THE INDICATED FACTOR K. TRANSMITTING AND RECEIVING ANTENNAS 2 FEET AND 35 FEET HIGH. PROPAGATION OVER GROUND HAVING A DIELECTRIC CONSTANT OF 10 AND NEGLIGIBLE CONDUCTIVITY.

body work which surrounds the engine. For this reason, the interference levels radiated by an ignition system varies widely with different makes and models of automobiles and it is very difficult to effectively predict the noise levels which can be expected from a particular car or truck. The hood which covers the automobile engine has been found to be especially important in reducing ignition interference.

A number of relatively effective means of suppressing ignition interference have been developed. The most effective method consists of completely shielding the engine compartment of an automobile. However, this is generally considered to be impractical and too expensive. The most widely used system involves resistive suppression in which the resistance used is either lumped or distributed. The suppression resistance effectively reduces the amplitude and frequency of the oscillatory capacitive current. In the lumped system, a high-temperature 10 w carbon resistor of between 5KΩ and 15KΩ is placed close to the distributor end of the coil-distributor high tension cable. Additional suppression can be obtained by adding similar resistors at the spark plugs. (Special spark plugs in which the resistor in incorporated into the body of the plug are also available today.) In the distributed system, the metallic conductor in the high tension leads is replaced by braided textile fibres which have a total resistance of between 4KΩ and 10KΩ over their entire length. Such cables are more effective than lumped resistors and have now been adopted by all U.S. automobile manufacturers for use as standard ignition cables.

Unfortunately, although suppressive devices are fitted to nearly all new cars, they frequently get removed during normal servicing and maintainance and generally are not replaced. This is largely due to the general ignorance of automobile mechanics in these matters and is also due to the mistaken belief that suppressive devices will degrade engine performance. Any service vehicles, construction vehicles or gasoline engines, etc., which are stationed at or near ground sites, should be fitted with effective and adequate suppressive devices. Care must also be taken to see that these devices have not been removed following any maintainance work on such vehicles or engines.

3.2 NOISE FROM HIGH POWER TRANSMISSION LINES

High voltage transmission lines are a source of broadband noise due to the presence of corona discharges at various points along the line. Noise may also be caused by the presence of arcing over dirty or wet insulators. The noise levels propagated by power lines becomes very quickly attenuated with distance; at a frequency of about 1 MHz, the corona noise from a 200 kV power line is no longer detectable at distances greater than two hundred feet from the line, when using a receiver with a half-wave dipole. Corona noise levels are relatively significant in the broadcast and HF bands, but decrease rapidly at frequencies above about 10 MHz. At VHF frequencies above 100 MHz, the levels appear to be virtually insignificant. Most of the measurements taken during on-site tests have been at frequencies in the broadcast band and below. These tests have shown that the levels of corona noise can also fluctuate greatly from day to day. Conditions of humidity and wind can cause large and rapid fluctuations of noise levels. These levels also tend to change over long periods of time, as the transmission line ages. Newly installed lines tend to be noisier than older lines.

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

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

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

Noise from corona discharges has been investigated by many workers (References 23-28) in both the lab and the field. Owing to the large number of variables involved, it is difficult to effectively correlate measurements obtained at one field site with those obtained at another site. Figure 3-4 shows some data extracted from Reference 25. As can be seen, the power levels radiated at VHF frequencies are very low (only about 10 db above the receiver level). It should be remembered that these measurements were made with the dipole antenna located only 40 feet from the power lines; receiving antennas will not normally be located this close to a power line. As would be expected, the 400 kv line radiates more noise than does the 138 kv line. However, part of the noise measured at these frequencies probably represents the overall man-made noise background, which is continuously present at VHF/UHF frequencies.

No data has been found on measurements of corona noise at 400 MHz and it can probably be concluded that such effects may be ignored at all UHF frequencies.

In summary, it may be concluded that corona noise from power lines does not represent a serious problem to ground site operations, since

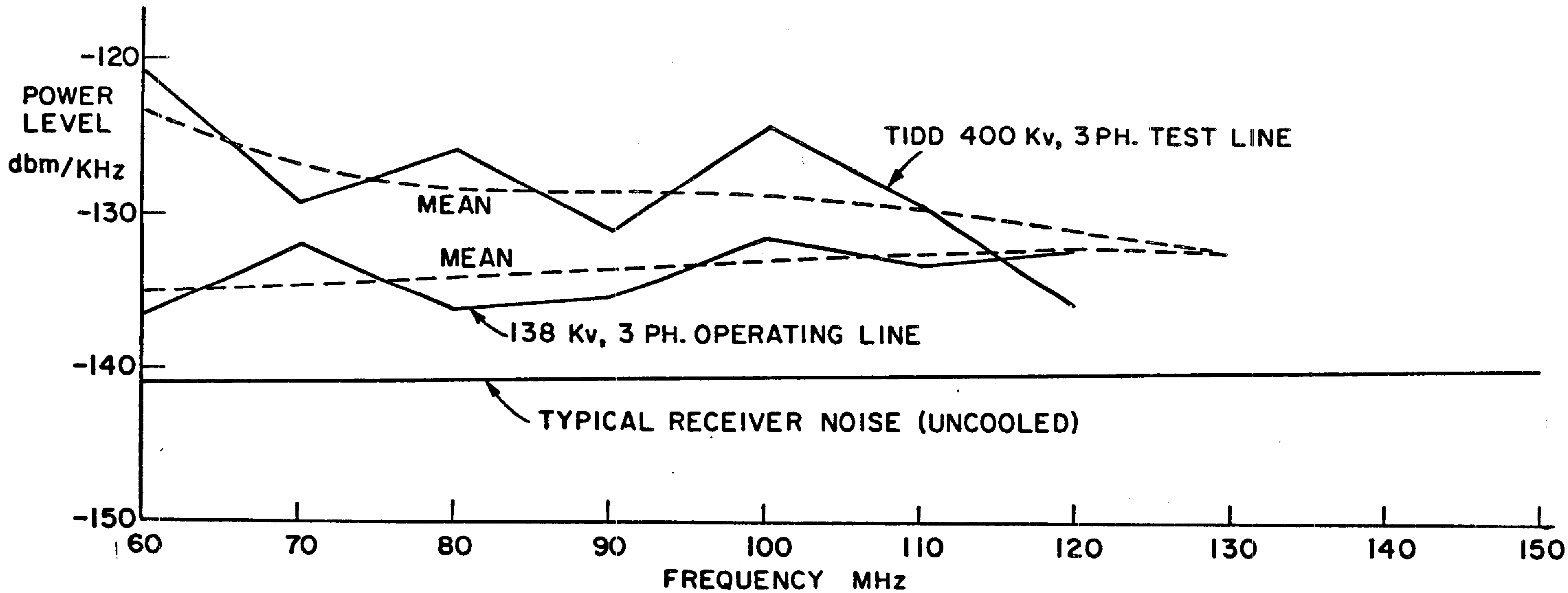


FIGURE 3-4

VARIATION OF CORONA NOISE LEVEL WITH FREQUENCY;
PEAK MEASUREMENTS USING HALF-WAVE DIPOLE LOCATED
40 FEET FROM OUTSIDE LINE CONDUCTOR.

the noise levels generated cease to be significant at the telemetry frequencies used in these operations.

It should be emphasized that, while corona noise levels radiated from a good power line are not expected to cause serious problems under normal conditions, a bad or faulted power line can cause serious noise problems. In this case, the noise is usually caused by arcing somewhere along the line and is not due to corona. Such effects probably occur only very seldom and can be prevented by regular maintenance of the power line.

3.3 OTHER SOURCES OF NOISE

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

Of the different noise sources, mentioned above, relatively the most serious, is the fluorescent lamp. These lamps, in common with all plasma operated devices, are generators of radio noise. The noise, originating from plasma devices is caused by: (1) the presence of periodic fluctuations in the ion or electron density in the area immediately surrounding the electrodes, and (2) the periodic switching of the polarity on ac operated devices. Figure 3-5 which is taken from Reference 29, shows a typical spectrum radiated by a fluorescent lamp. As can be seen, the noise level radiated at frequencies of 100 MHz and above is virtually insignificant. The major part of the radiated spectrum occurs in the VLF region, below 100 KHz. Also, as with all of these noise sources, the noise levels fall off rapidly with distance.

It may, therefore, be assumed that none of the noise sources mentioned in this section pose any kind of serious problems to ground station operations. If any of these devices is found to be particularly noisy and does cause some disruption of signal reception, it will probably be a relatively easy matter to suppress the source by effectively shielding it and by adding interference suppression filters, where possible. On the other hand, it should be emphasized that any of these devices which are being used outside the operations buildings and close to the receiving antennas may well cause some difficulties. This might involve the use of hand tools such as electric drills or saws for any outside construction work or the use of arc welders. Such problems can, of course, be avoided by halting such construction work and outside activities during the schedules satellite passes.

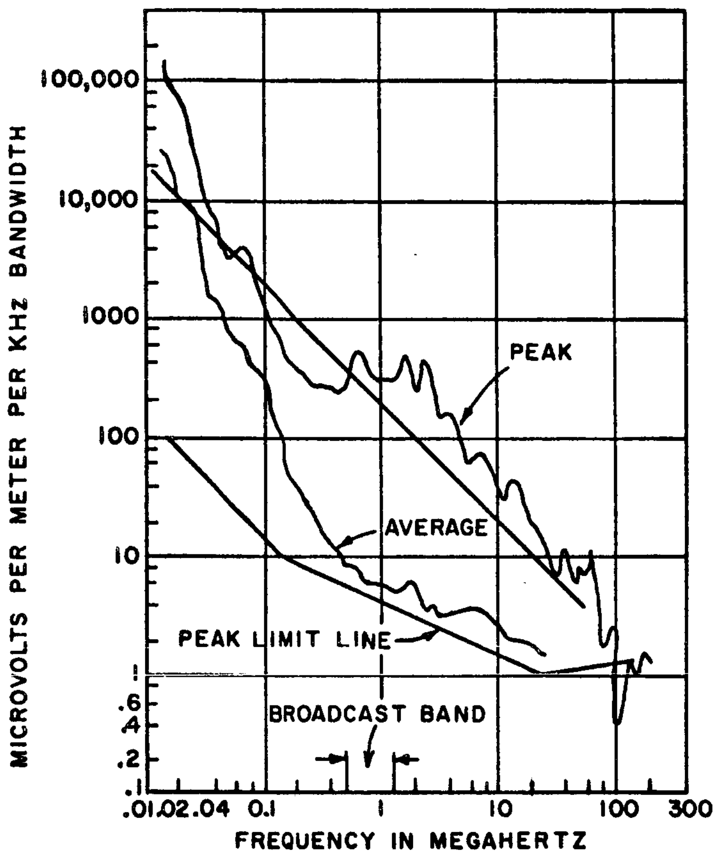


FIGURE 3-5

PEAK AND AVERAGE INTERFERENCE RADIATED
FROM LOW NOISE STANDARD 40-WATT FLUORESCENT LAMPS.

4.0 LINEAR AND NONLINEAR ADMISSION MECHANISMS VIA NORMAL INPUT TERMINALS

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

(1) linear intrusion via normal input terminals;

(2) nonlinear intrusion via normal input terminals;

(3) intrusion through ports not intended as signal inputs.

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

The block diagram of Fig. 4-1 shows the essential elements of a receiver. In the linear intrusion mode the receiver acts as a normal bandpass filter which accepts any input containing frequency components in the receiver passband. Unwanted inputs whose spectrum is centered at, or near, the tuned frequency of the RF filter, originating from communication sources or from noise sources, are the usual interference sources. The second mode, called the nonlinear mode, acts through a nonlinear element and is usually characterized by the acceptance of unwanted signal energy which lies outside the normal passband of the receiver. The RF filter in Fig. 4-1 is a preselector network which limits the band to which the succeeding active elements in the receiver are exposed. The latter have nearly always some residual nonlinear properties which play a significant role when the input amplitudes are large. When the RF filter is inadequate to limit large out-of-band inputs to a satisfactory low level, the devices (vacuum tubes, transistors, diodes, etc.) will, by means of their nonlinearity, generate frequency components not originally present. When these new components are within the passband of the portion of the receiver following the electronic device, interference can occur. Phenomena typical of this mode are single

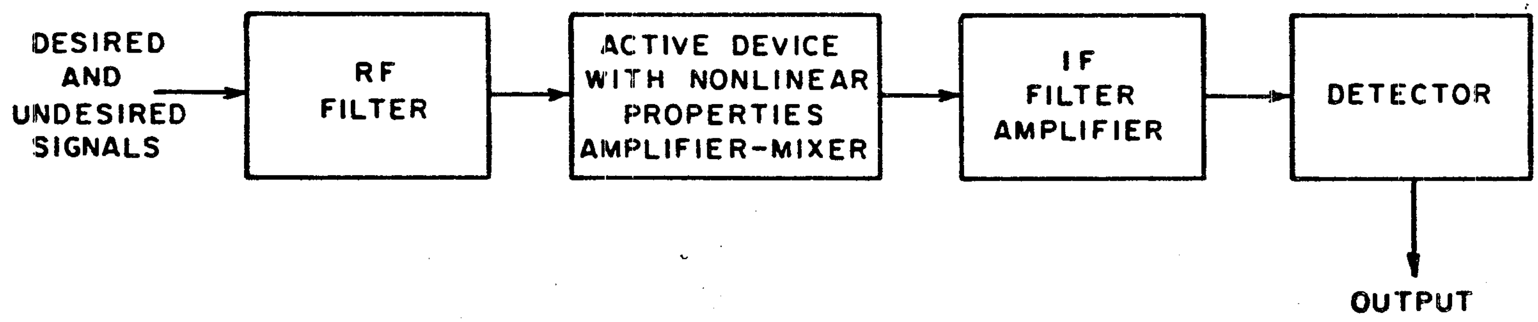


FIGURE 4-1

BLOCK DIAGRAM OF BASIC RECEIVER ELEMENTS

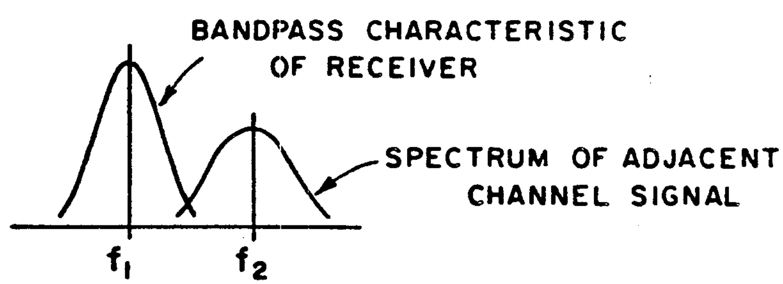


FIGURE 4-2

ADJACENT CHANNEL INTERFERENCE

spurious response, multiple spurious response (intermodulation), sideband transfer (cross-modulation), etc.

Ultimately, it will be necessary to know the degree to which the undesired signals interact with the desired signal. This turns out to be a difficult problem for several reasons. First, the unwanted signals, after being operated on by the linear and nonlinear receiver circuits, have their original properties altered. Second, the reaction of the detector to the combination of desired signal and the undesired remnants is an involved problem of analysis. Third, the loss of information to the ultimate information received is difficult to evaluate. When analytical or experimental results are not available judicious guesses of the interaction are often necessary.

4.1 LINEAR INTRUSION

The linear admission mechanism is a straightforward matter of the response of the receiver bandpass filter. The unwanted signals will fall into one of the following categories:

a. Broadband noise arising from natural or man-made sources which radiate inadvertently, or which are intended to radiate. Over the bandpass of a typical receiver, the spectrum of such noise is essentially flat.

b. Signals from communications sources assigned to a frequency at, or near, the receiver center frequency (cochannel interference). More particularly, when the center frequencies are separated by an amount less than the receiver bandwidth the signals are said to be co-channel.

c. Signal from communications sources assigned to a frequency separated from the receiver center frequency by more than the signal bandwidth (adjacent-cnannel interference).

d. Signals from communications sources assigned to frequencies within one of the internal pass frequencies.

A receiver usually contains one or more mixers for translating the input to a frequency more convenient for amplification. In this discussion, it is assumed that no fundamental change takes place in the nature of the signal with this translation. The signals, both wanted and unwanted, are therefore viewed as being at IF amplifier frequency, or near it, once they are admitted.

Broadband noise sources of the nonperiodic variety, sometimes also referred to as non-coherent sources, are characterized by irregularity of waveform and unpredictability of detail. Typical natural sources of this kind are thermal noise, shot noise, galactic noise, solar noise, and atmospheric noise. Typical man-made sources are discharges on high-voltage lines and devices, noise from automobile ignition systems, commutator noise, noise in complex switching systems, and noise generated by fluorescent lamps. In some of these sources, a certain amount of regularity exists. Atmospheric noise bursts are sometimes found to have some coherence because of multiple propagation paths (Reference 32). Furthermore, there are long time fluctuations, depending on the time of day, season, and sunspot cycle, which are roughly predictable (Reference 33). Corona noise on high-voltage lines and fluorescent lamp noise are usually modulated by the power-line frequency (References 26, 27). A single, stationary ignition noise source is more or less periodic; noise from one or more randomly passing vehicles or from many stationary vehicles is not. A more detailed discussion of some of these sources appears in Section 5.1.

Nonperiodic broadband sources are best characterized by their power spectral densities (Reference 34). The mean square value of the noise admitted through the IF amplifier is:

$$\langle e_o^2 \rangle = \int_0^\infty N_1(f) \left| G(f) \right|^2 df \qquad (4\text{-}1)$$

where:

$N_1(f)$ = the one-sided power spectral density at the receiver input in volts^2 per Hertz.

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

It was pointed out earlier that the noise spectrum is usually flat over the receiver passband. If this is so, and if the center frequency of the IF amplifier is written f_1, the mean square value of the IF amplifier output is:

$$\langle e_o^2 \rangle = N_1(f_1) \left| G(f_1) \right|^2 \int_0^\infty \frac{|G(f')|^2}{|G(f_1)|^2} df = N\, G^2 B_p \qquad (4\text{-}2)$$

where:

$N = N_1(f_1)$ and is the input power spectral density at the band center.

$G^2 = \left| G(f_1) \right|^2$ and is the gain at the band center.

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

By periodic broadband sources is meant sources of regularly spaced, constant amplitude, short pulses (or those which are nearly so). Two cases can be distinguished. In the first, the pulses are sufficiently separated in time so that each pulse acts as a separate transient exciting the receiver. In the second, the pulses are close to one another so that the responses tend to overlap. The overlapping case arises when pulse spacing is of the order of the inverse of the receiver bandwidth or less. Since typical impulse sources (e.g., radar transmitters and modulators) are of relatively low repetition frequency compared to receiver bandwidths, attention is directed here to the nonoverlapping case exclusively.

The instantaneous output of the IF amplifier to an input which has a Fourier spectrum given by S(f) is:

$$e_o(t) = \int_{-\infty}^{\infty} S(f)\, G(f)\, e^{j2\pi ft}\, df \qquad (4\text{-}3)$$

where:

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

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

$\left| G(f) \right|$ = the amplitude versus frequency characteristic of the IF amplifier.

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

$$e_o(t) = S(f_1) \int_{-\infty}^{\infty} \left|G(f)\right| e^{j\left[2\pi ft + \phi(f)\right]} \, df \qquad (4\text{-}4)$$

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

$$e_o(t) = 2S(f_1) \int_{0}^{\infty} \left|G(f)\right| \cos\left[2\pi ft + \phi(f)\right] df \qquad (4\text{-}5)$$

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

$$e_{om} = 2S(f_1) \quad G(f_1) \int_{0}^{\infty} \frac{\left|G(f)\right|}{\left|G(f_1)\right|} df = 2S \; G \; B_e \qquad (4\text{-}6)$$

where:

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

$B_e = \int_{0}^{\infty} \frac{\left|G(f)\right|}{\left|G(f_1)\right|} df$, and is the effective impulse bandwidth of the amplifier.

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

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

Remedies for broadband noise which overlaps the receiver band must take advantage of differences which are known to exist between the form of the signal and the noise in the time domain. A review of methods of dealing with noise of a discrete impulse nature appears in Reference 35. The most common methods used are (1) limiting and (2) blanking, both done before the broadband pulses have been filtered in the IF amplifier. The principle here is that this sort of noise, having short duration and large peak value, can be limited above the level of the desired signal, or can be totally blanked out for its brief duration. In either case the signal is eliminated for the duration of the impulse, but this hardly ever causes any loss of information. In the case of the limiter the original large amplitude pulse is replaced by a smaller one thereby reducing the impulsive spectral magnitude in the desired passband. It is evident that a filter preceding the limiter or blanker will widen the interference pulse and make these processes less effective as noise reducers.

Blanking systems require information on the location of the impulses. Systems for eliminating the periodic pulses of a nearby radar system may use direct synchronization from the radar source. Where there is no access to the source, the receiver itself must sense the pulse in one branch in order to eliminate it in a second branch. The limiter is much simpler, frequently using a manual setting of the clipping level by the operator for best reception. A technique cited in Reference 35 (ch. 11, 19) for counteracting impulse noise is to pre-smear the information prior to transmission and to re-assemble it after detection. The re-assembling process smears the impulse over a wide-time interval thus reducing its effect.

The rejection of broadband incoherent noise which is dense and uniformly distributed in time, e.g., Gaussian noise, is a well-treated problem of communication theory and will not be pursued here in any detail. Briefly, an advantage is obtained by using a modulation process which builds distinguishable characteristics into the signal which the detector then uses to help in identification. A matched filter detector, for instance, uses a priori information about the shape of the signal pulses to discriminate between signal and noise (Reference 35). In general, broadband modulation methods designed to achieve better output quality in the face of Gaussian noise at the price of increased bandwidth will often do the same job for interference.

The term co-channel interference is reserved for interference situations involving communications systems which have been assigned equal, or nearly equal, carrier frequencies. Co-channel assignments are ordinarily made when the probability of an encounter between the two systems is insignificant. Such systems are either separated physically by large distances or do not operate at the same time. Sometimes, because of unusual propagation conditions, or as in the case of satellite transmitters, where co-channel sources do exist but are not often expected to be in view, simultaneously, interference on this account will arise.

Naturally, the linear portion of the receiver will handle both signals. If one of these is ultimately suppressed, it will be because of the nature of the detector.

The term adjacent-channel interference is reserved for the case of interference between communication systems which have been assigned neighboring channels. Channel spacing policy varies; but, for our purposes, adjacent will be viewed as meaning separation by an amount greater than the average of the two signal bandwidths. An exaggerated situation of this nature is indicated in Fig. 4-2 where energy on the skirt of the adjacent-channel signal spectrum overlaps the bandpass characteristic of the receiver. As a rule, assignments of immediately adjacent channels in the same locality are avoided. While in the typical case the sensitivity in this mode is low compared to the in-band sensitivity, receivers located close to an adjacent-channel transmitter are exposed to very large magnitudes of unwanted signals.

Estimates are apt to be made in one of three ways--by treating the unwanted signal as a pure sinusoid, or as a broadband waveform perfectly centered in the band, or as a broadband waveform centered far enough away from the tuned frequency so that the unwanted spectrum is nearly constant over the receiver band. The first of these will be useful for estimating both co- and adjacent-channel effects. The second is appropriate for co-channel situations and the third is appropriate for adjacent-channel interference. In the latter case, since the bandpass filter is exposed to a portion of the one sideband of the unwanted signal, the noise output is usually in unintelligible garble. It would appear to be a fair guess that the effect is not much different from the effect of thermal noise of equal mean square value. We therefore estimate the spectral density of the unwanted signal at the receiver band center and, as in the case of non-periodic broadband noise discussed above, determine the mean square value of the IF amplifier output by use of (4-2).

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

4.1.1 Adjacent Channel Interference

Two satellites, with closely-spaced adjacent channel frequency assignments, can result in interference within a given receiver. To illustrate, two examples of such situations will be analyzed assuming, respectively, interfering unmodulated and modulated (broadband) signals. In the latter case, random noise modulation will be assumed.

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

$$f_a = f_1 + \Delta f_1$$

$$f_b = f_2 + \Delta f_2$$

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

f_2 = carrier frequency of the signal transmitted by satellite B

Δf_1 = Doppler shift of signal transmitted by A, and

Δf_2 = Doppler shift of signal transmitted by B.

The receiver selective circuits are assumed to consist of n single-tuned parallel RLC type circuits. From equation (2-21), the relative response of these circuits is written

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

where B^* is the 3 dB bandwidth of one single tuned circuit. $H_n(f)$ becomes, in terms of the receiver passband 3 dB bandwidth, $B_{3\ dB}$,

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

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

$$P_i = \left| H_n(f_b)^2 \right| \overline{X^2_b}$$

To give a numerical example the parameters are assumed as follows

$n = 3$

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

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

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

$\overline{X_a^2} = 10^{-16}$ watt

$\overline{X_b^2} = 3 \times 10^{-15}$ watt

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* The 3 dB bandwidth of n cascaded single-tuned circuits is given by

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

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

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

P_i in this case is given by

$$P_i = \left[1 + \frac{4(136.227 - 136.202)^2 \left[2^{1/3} - 1\right]}{(30)^2}\right]^{-3} \times 3 \times 10^{-15}$$

$$P_i \cong 5.9 \times 10^{-16} \text{ watt.}$$

The signal to interference ratio is

$$\frac{S}{I} = \frac{\overline{X_a^2}}{P_i} = \frac{10^{-16}}{5.9 \times 10^{-16}} \cong 0.17 \quad \text{or} \quad -7.7 \text{ dB}$$

So low a signal to interference ratio is viewed as representing an interfering situation.

When the interfering signal is modulated with broadband noise the power $\overline{X_b^2}$ is assumed to be uniformly distributed over an ideal bandwidth, ΔB, centered about the carrier frequency f_b. The power spectrum of the undesired signal and the receiver response curve are shown in Fig. 4-3, where the power spectral density of the interfering signal is $P(f) = \phi_i$ watt/unit bandwidth. When the receiver is tuned to the frequency f_a, the unwanted interference power that gets into the receiver passband is given by

$$P_i = \int_0^\infty \phi_i \left| H_n(f) \right|^2 df$$

Assuming $\phi_i = 0$ except where $f_b - \frac{\Delta B}{2} < f < f_b + \frac{\Delta B}{2}$, P_i is then

$$P_i = \phi_i \int_{f_b - \frac{\Delta B}{2}}^{f_b + \frac{\Delta B}{2}} \left[1 + \frac{4(f - f_a)^2 (2^{1/n} - 1)}{B^2_{3 \text{ dB}}}\right]^{-n} df$$

The integral above, by an appropriate change of variable, can be reduced to a standard integral (e.g., ref. 43, p. 1068, eq. 13) of the form

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

where $c = \Phi_i B_{3\ dB}$, $a = 4\left[2^{1/n} - 1\right]$, $b = 1$, $y^2 = \frac{(f-f_a)^2}{B^2_{3\ dB}}$

$$a_f = \frac{f_b + \frac{\Delta B}{2} - f_a}{B_{3\ dB}}, \qquad b_f = \frac{f_b - \frac{\Delta B}{2} - f_a}{B_{3\ dB}}$$

The receiver selective circuits are assumed to consist of three single tuned stages, i.e., n = 3, making $a \cong 1.0$. For n = 3, integrating, changing the variable y back to f, and inserting the upper and lower integration frequency limits result in

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

$$P_i = \Phi_i B_{3\ dB} K$$

In the example below, the parameters of the first example and $\Delta B = 30$ kHz are assumed. The following values then can be written from their definitions above.

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

$a_f = 1.33, \qquad b_f = 0.33, \qquad K \cong 0.267$

$$P_i = \Phi_i \, B_{3\ dB} \, K = 10^{-16} \times 30 \times 0.267$$

$$P_i = 8 \times 10^{-16} \text{ watt } (-121 \text{ dBm})$$

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

$$\frac{S}{I} = \frac{\overline{X_a^2}}{P_i} = \frac{10^{-16}}{8 \times 10^{-16}} = 0.125 \qquad \text{or } -9 \text{ dB}$$

This, too, represents an interfering situation.

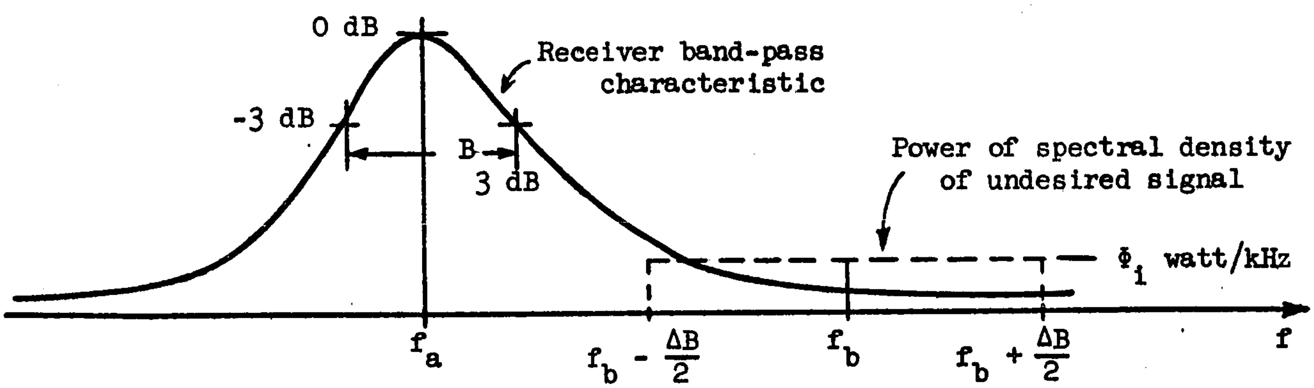

Figure 4-3

Illustration of unwanted signal spectrum and the receiver tuning characteristic.

4.2 NONLINEAR INTRUSION

Nonlinear mechanisms arise as a result of inadequate rejection of the unwanted signal in the input filter circuits of the receiver that is followed by some nonlinear process in an electronic device. Less common nonlinear admission mechanisms involved imperfect joints between conductors prior to the receiver filter circuits, which give rise to nonlinear junction effects. Except in unusual cases, the level of input signal required to create an effect in the presence of a desired signal is several tens of decibels greater than the desired signal. That is, the undesired signal level required to register an effect is such that communication transmitters are usually the only significant sources. Furthermore, the spurious transmitter outputs are rarely large enough to be significant so that only the main signal needs to be considered. Natural noise sources and incidental sources of man-made noise are rarely found to be important in these modes. There are exceptions, however, so that such possibilities must not be dismissed without some examination of the environment.

4.2.1 Spurious Responses

The spurious responses of a receiver will be the result of nonlinearity in an early stage which gives rise to harmonics of incoming signals, nonlinearity in the mixer which results in oscillator and signal harmonics, and frequency multiplication in the local oscillator and its related circuits. At each frequency of tuning another set of possible spurious response frequencies exist and each of these sets has its own level of significance.

Spurious responses arising in mixers can often be described by the following mechanism. The nonlinear device (a transistor diode or vacuum tube) has an output-input characteristic specified by a power series:

$$y = \sum_{n=0}^{N} a_n x^n \qquad (4\text{-}7)$$

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

$$x_s(t) = v_s(t) \cos (\omega_s t + \phi_s) \qquad (4\text{-}8)$$

$$x_o(t) = A \cos \omega_o t \qquad (4\text{-}9)$$

Then

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

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

(4-10)

and $\cos^p x$ which is a periodic function can be expanded* (Reference 36) in a Fourier series so that when the above equation is written out as a sum of individual cosine terms, all frequencies,

$$\left| m\omega_s \pm n\omega_o \right| , \qquad (4\text{-}11)$$

where,

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

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

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

may be found in the result.

Whenever one of these frequencies coincides with the intermediate frequency, a potential spurious response frequency is said to exist. That is,

---

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

$$\cos^p x = \frac{1}{2^{p-1}}\left[\cos px + p \cos (p-2)x + \frac{p(p-1)}{2!} \cos(p-4)x + \ldots + \frac{1}{2} \frac{p!}{\left(\frac{p}{2}!\right)^2}\right] \qquad (4\text{-}10a)$$

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

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

any input frequency which satisfies the equation

$$\omega_s = \frac{(\pm n\omega_o \pm \omega_{if})}{m} \tag{4-12}$$

with any combination of the signs, is a potential spurious response frequency.

Frequently, the signal component, $x_s(t)$, is small compared to the oscillator component, $x_o(t)$, and terms involving $x_s^k(t)$ with $k > 1$, can be ignored. The significant portion of (4-10) is then,

$$y_1(t) = \sum_{n=1}^{N} na_n\, x_o^{(n-1)}(t) \cdot x_s(t) = g(t)\, x_s(t) \tag{4-13}$$

The quantity $g(t)$ is the transconductance as a function of time when an oscillator voltage, $x_o(t)$, is applied. That is, the transconductance is the derivative of (4-7)

$$g = \frac{dy}{dx} = \sum_{n=1}^{N} na_n\, x^{(n-1)} \tag{4-14}$$

Then writing $x = x_o(t) = A \cos \omega_o t$, we obtain

$$g(t) = \sum_{n=1}^{N} na_n A^{(n-1)} \cos^{(n-1)} \omega_o t = \sum_{n=0}^{N-1} g_n \cos n\omega_o t \tag{4-15}$$

The last term on the right of (4-15) is the form which would be obtained if $\cos^{(n-1)} \omega_o t$ is expanded according to (4-10a) and all terms of like harmonic were collected. Thus from (4-13)

$$y_1(t) = \frac{v_s(t)}{2} \sum_{n=0}^{N-1} g_n \left\{ \cos\left[(\omega_s - n\omega_o)t + \phi_s\right] + \cos\left[(\omega_s + n\omega_o)t + \phi_s\right] \right\} \tag{4-16}$$

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

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

$$i = I_s(\epsilon^{av} - 1) \tag{4-17}$$

where I_s is the reverse saturation current

'a' is a constant which in theory is e/KT ($\cong 40$)

e is the electronic charge $= 1.602 \times 10^{-19}$ coulomb

K is the Boltzmann's constant $= 1.38 \times 10^{-23}$ Joules/^oK

T is temperature in degrees Kelvin

Therefore,

$$g = \frac{di}{dv} = aI_s\epsilon^{av} \tag{4-18}$$

When

$$v = A \cos \omega_o t$$

$$g(t) = aI_s\epsilon^{(aA \cos \omega_o t)}$$

$$= aI_s\left[I_o(aA) + 2\sum_{n=1}^{\infty} I_n(aA) \cos n\omega_o t\right] \tag{4-19}$$

I_n (aA) is the modified Bessel function of the first kind of order "n" (n=0,1,2..) and of argument (aA) (Reference 36). The conversion transconductance as defined in (4-15), is

$$g_n = 2aI_sI_n(aA) \tag{4-20}$$

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

The relative levels of interference to signal magnitude are not so easily calculated for the reason that the gains up to the mixer of all possible interfering signals at all frequencies of tuning are not ordinarily known. Furthermore, the oscillator level and harmonic and subharmonic content as a function of tuned frequency are not ordinarily known. It is more common to measure the intensity of the spurious response than to calculate them. The usual procedure is to set the tuning control to three points in each band, the center and the vicinity of the band extremes, and to tune an input signal generator through the regions of potential response at each point. The ratio of signal to interference carrier-voltage levels at the input required to give equal outputs is the observed quantity. The ratio may depend on the input level. The intensity of the spurious responses nevertheless can be calculated in many cases. Especially when only approximate values are needed. An example of calculation of the spurious response intensities is given below.

Let N = 3 in (4-7). Then expanding (4-10) we get,

$$
\begin{aligned}
y(t) = {} & a_o + a_1 v_s(t) \cos(\omega_s t + \phi_s) + a_1 A \cos \omega_o t \\
& + \frac{1}{2} a_2 (v_s^2(t) + A^2) + \frac{1}{2} a_2 v_s^2(t) \cos 2(\omega_s t + \phi_s) \\
& + \frac{1}{2} a_2 A^2 \cos 2\omega_o t + a_2 v_s(t) A \cos \left[(\omega_s - \omega_o)t + \phi_s\right] \\
& + a_2 v_s(t) A \cos \left[(\omega_s + \omega_o)t + \phi_s\right] \\
& + \left[\frac{3}{4} a_3 A^3 + \frac{3}{2} a_3 A v_s^2(t)\right] \cos \omega_o t \\
& + \left[\frac{3}{4} a_3 v_s^3(t) + \frac{3}{2} a_3 A^2 v_s(t)\right] \cos(\omega_s t + \phi_s) \\
& + \frac{1}{4} a_3 v_s^3(t) \cos 3(\omega_s t + \phi_s) + \frac{1}{4} a_3 A^3 \cos 3\omega_o t \\
& + \frac{3}{4} a_3 v_s^2(t) A \cos \left[(2\omega_s - \omega_o)t + 2\phi_s\right] \\
& + \frac{3}{4} a_3 v_s^2(t) A \cos \left[(2\omega_s + \omega_o)t + 2\phi_s\right] \\
& + \frac{3}{4} a_3 v_s(t) A^2 \cos \left[(\omega_s - 2\omega_o)t + \phi_s\right] \\
& + \frac{3}{4} a_3 v_s(t) A^2 \cos \left[(\omega_s + 2\omega_o)t + \phi_s\right]
\end{aligned}
\tag{4-21}
$$

From the collection of components in (4-21) we consider the ones wherein the undesired signal, now written ω_{SI}, is such that (see (4-12))

$$\omega_{SI} - 2\omega_o = \pm \omega_{if}$$

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

$$\omega_{SD} - \omega_o = \omega_{if}$$

Suppose for instance that $\omega_{SD} = 140$ MHz, $\omega_o = 110$ MHz, $\omega_{if} = 30$ MHz. Then

$$\omega_{SI} = 2\omega_o \pm \omega_{if} = \begin{matrix} 250 \text{ MHz} \\ 190 \text{ MHz} \end{matrix}$$

From (4-21) the desired component at the mixer output is

$$D = a_2 \, v_s(t) \, A \cos \left[(\omega_s - \omega_o)t + \phi_s\right]$$

and its peak is

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

where V_{SD} is the peak of $v_s(t)$ and the signal input is the desired signal. The undesired component is

$$I = \frac{3}{4} a_3 \, v_s(t) \, A^2 \cos \left[(\omega_s \mp 2\omega_o)t + \phi_s\right]$$

and its peak is

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

where V_{SI} is the peak of $v_s(t)$ and the signal input is the undesired signal. The ratio of desired to undesired levels at the mixer output is therefore

$$\frac{D}{I} = \frac{4a_2 V_{SD}}{3a_3 V_{SI} A} \tag{4-22}$$

V_{SD} and V_{SI} are the peak levels at the MIXER input. However, this ratio ought to be obtained at the receiver input and this depends upon the preselector gain factor. If we write V_D, the desired <u>receiver</u> input peak, and V_I, the undesired <u>receiver</u> input peak, then

$$\left(\frac{D}{I}\right)_{input} = \frac{4a_2 V_D}{3a_3 AKV_I} \tag{4-23}$$

where K is the gain of the undesired signal relative to the desired one in the RF circuits preceding the mixer. For instance, if two single tuned circuits, each of bandwidth, B = 15 MHz, are used in the RF stage then from (2-23) for the 190 MHz interference component (k = 2),

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

If now we use the values of a_2 and a_3 given in Section 2.1.1 and assume the oscillator level A = 10 volts we obtain from (4-23),

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

For equal output to desired and undesired components, $\frac{D}{I} \doteq 1$, the input ratio must be

$$\frac{V_I}{V_D} = 192.5 = 45.7 \text{ db}$$

Thus the spurious response is said to be 45.7 db below the response to a desired signal.

Figure 4.4 shows a plot of equation (4-12) relating tuned frequency, f_{SD}, and potential frequency of response, f_{SI}, for several values of m and n for a receiver covering a range from 100 to 200 MHz, and having an oscillator frequency, f_o, set 30 MHz below the tuned frequency. Measured or computed values of strength of response may be indicated on the diagram at the appropriate points as shown on the line labeled $f_{SI} = 2f_o - f_{if}$ at $f_{SD} = 140$ MHz. Or, for each spurious response line on Fig. 4-4 a corresponding curve can be plotted as shown in Fig. 4-5, showing the relative response at each tuned frequency.

The reduction method that applies to the nonlinear mechanisms of generation and admission is filtering in appropriate places and

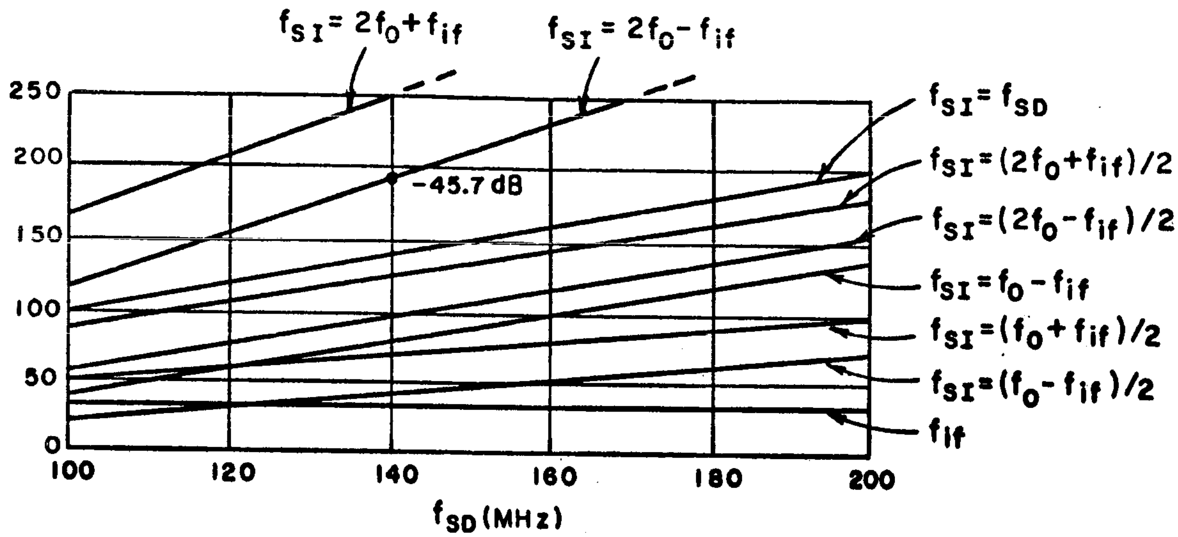

FIGURE 4-4

SOME RECEIVER SPURIOUS RESPONSES
OSCILLATOR FREQUENCY, f_o, IS 30 MHz (f_{if})
BELOW DESIRED FREQUENCY, f_{SD}

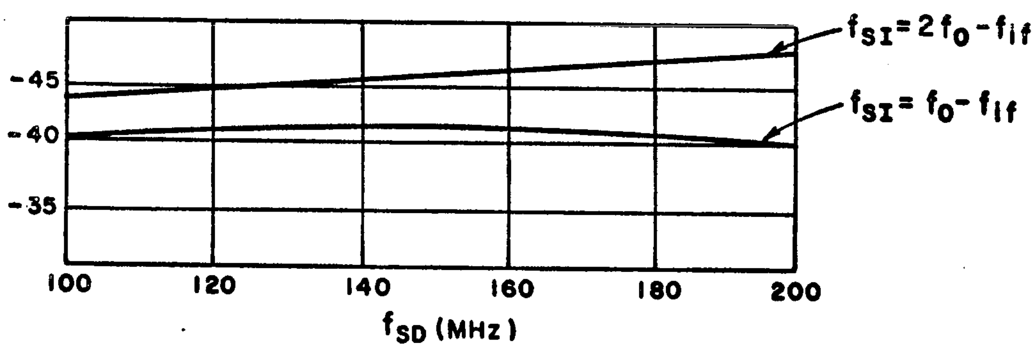

FIGURE 4-5

RELATIVE RESPONSE SIGNAL TO INTERFERENCE FOR
EQUAL OUTPUTS

reduction of the input levels. For instance, the spurious response of a receiver is increased when the local oscillator is made to drive the mixer over a wider range. While the desired conversion efficiency might be increased somewhat, the "harmonic conversion" efficiency is also increased and, beyond a certain point, proportionately more than the desired effect. To keep the spurious response down, additional filtering prior to the mixer can be used. There is a trade-off then between cost of additional filtering and the increased gain which ought to be considered as a general rule. The region over which the mixer is operated should be as near square-law as possible.

4.2.2 Inter- and Cross-Modulation

Inter- and cross-modulation, discussed under transmitter mechanisms (Section 2.4) are important also in receivers. The mechanisms are essentially the same. Intermodulation in receivers is, however, the result of two or more unwanted signals present simultaneously at the input. Cross-modulation is the transfer of information from an undesired carrier onto the desired one. In either case it is the nonlinearity in a circuit near the input which is the cause.

Intermodulation is the more important of these mechanisms and is discussed further here. It becomes especially important when a range of frequencies is subdivided into separate communication channels and when a number of these channels, which are closely spaced must be used simultaneously. Then two unwanted signals of the form (2-36) and (2-37) of Section 2.4, give rise to a component of the form

$$y_I(t) = v_1^2(t)\, v_2(t) \cos\left[(2\omega_1 - \omega_2) + 2\phi_1(t) - \phi_2(t)\right] \qquad (4\text{-}24)$$

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

It will similarly be found that with three channels at frequencies f_1, f_2, and f_3, intermodulation products of the kind discussed above (that is, those products near in frequency to the original generating frequencies but not coincident with them), are:

a. $f_1 + f_2 - f_3$
b. $f_1 - f_2 + f_3$
c. $2f_1 - f_2$
d. $2f_1 - f_3$
e. $2f_2 - f_1$
f. $2f_2 - f_3$
g. $2f_3 - f_1$
h. $2f_3 - f_2$

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

a. $f_1 + f_2 + f_3 - f_4 - f_5$
b. $2f_1 + f_2 - f_3 - f_4$
c. $f_1 + f_2 + f_3 - 2f_4$

d. $2f_1 + f_2 - 2f_3$

e. $3f_1 - f_2 - f_3$

f. $3f_1 - 2f_2$

These are only representative forms; the subscript on the frequencies above may be permuted in any way among the assigned frequencies. Thus, if there are 10 frequencies, $3f_7 - 2f_{10}$, or $f_1 - f_2 + f_3 - f_4 + f_5$ are frequencies which can be significant. Techniques for channel selection to avoid interference are given by Babcock (Reference 38) and also by Beauchamp (Reference 39).

Tests of susceptibility to intermodulation in actual receivers have been described in detail. McLenon (Reference 40) applied signals to commercial grade receivers to give potential intermodulation at 5.1 MHz. He obtained a resultant interference carrier level of 0.5μv for inputs ranging from 0.01 volt to 0.1 volt. The highest input was required in a receiver which had two tuned circuits before the first amplifier tube.

A sample calculation of the magnitude of intermodulation interference will now be given. Third-degree nonlinearity is assumed. Carrying out an expansion similar to that given in (4-10) but with $x_s(t)$ and $x_o(t)$ replaced by two incoming signals $x_1(t)$ and $x_2(t)$ as given by (2-36) and (2-37), and with $n = 3$, an output interference component

$$y_I(t) = \frac{3a_3}{4} v_1^2(t)\, v_2(t) \cos\,(2\omega_1 - \omega_2)t, \qquad (4\text{-}25)$$

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

The signal-to-interference ratio, $\frac{S}{I}$, is defined as the ratio of the coefficients of these two components, or,

$$\frac{S}{I} = \frac{4a_1\, v_s(t)}{3a_3\, v_1^{\,2}(t)\, v_2(t)} \tag{4-26}$$

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

$$v_1 = \left(\frac{4a_1\, v_s}{3a_3}\right)^{\frac{1}{3}} \tag{4-27}$$

For instance, if

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

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

$$a_3 = 5 \times 10^{-5} \text{ ampere/volt}^3$$

then $v_1 = 0.11$ volt. At VHF a spacing of about 150 feet between transmitting sources and the receiving antenna, assuming 50-watt transmitters, will give such a value.

v_1 is the amplitude at the input to the nonlinear element but the amplitude at the antenna terminals can be greater than this figure. The selectivity of the input circuit may not be sufficient to cause much attenuation to the unwanted signals. The input unwanted signal voltage can thus be about 0.11 volt, also.

In the case of cross-modulation arising from third-degree nonlinearity the interference component, involved is (again using

(2-36), (2-37), and expanding in a form such as 4-10 with $v_2(t) \cos\left[\omega_2 t + \phi_2(t)\right]$ viewed as the desired signal),

$$y_I = \frac{3a_3\, v_1^2(t)\, v_2(t)}{2} \cos\left[\omega_2 t + \phi(t)\right] \qquad (4\text{-}28)$$

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

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

the signal-to-interference ratio, defined as the ratio of coefficients of $y_s(t)$ and $y_I(t)$ is

$$\frac{S}{I} = \frac{2a_1}{3a_3 v_1^2(t)} \qquad (4\text{-}29)$$

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

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

For example, if values of a_1 and a_3 given in page 88 are used

$$v_1(t) = 8.165 \text{ volt}$$

4.2.3 Desensitization

Desensitization refers to a reduction in over-all receiver gain, or sensitivity or both, when a large unwanted signal enters the receiver. The interference signal alone may not even be observed if it is either unmodulated or modulated in a way to which the receiver is nonreceptive. Typical mechanisms of desensitization are as follows.

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

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

Low duty cycle pulsed signals, such as radar emissions, can be especially troublesome because of their large peak amplitudes. Microwave radar interference to low frequency communications systems by overload of an early receiver stage is not uncommon, largely because the lumped input tuned-circuits are virtually useless as filters to microwave energy. Once the unwanted pulse appears at the input amplifying device it will act according to one of the mechanisms described in the previous paragraph. Even low duty cycle unwanted pulse signals, without charging networks being present, may be a source of interference. Two mechanisms are discussed in the following paragraphs.

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

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

In Section 4.2.1 it was pointed out that diode mixers act naturally as harmonic mixers to create spurious responses. They are also subject to desensitization effects (Reference 41). The effect is found to arise in microwave receivers, e.g., radar receivers, where the mixer is the first electronic device following the input terminals. It can be shown that the conversion transconductance (defined by $g(t)$ in eq. (4-19)) is altered by the presence of a large unwanted signal. A more important effect, however, appears to be connected with mismatch; the effective impedance of the IF output of the mixer is altered by the unwanted signal. Assuming the IF amplifier input were matched to the mixer in the absence of unwanted signals, it will become unmatched when the unwanted signal appears. Tests rep orted in Reference 41 (pp. 28-29) show a drop in conversion efficiency by approximately 3 db for an unwanted sinusoid equal in amplitued to the local oscillator signal (Fig. 4-6); the larger the local oscillator input power to the mixer the larger the tolerable unwanted signal. It was pointed out above, however, that with increasing local oscillator inputs the harmonic conversion transconductance g_n becomes significant for higher values of $\underline{n}$. A compromise is therefore needed between high local oscillator power to minimize desensitization potential, and low local oscillator power to minimize spurious response potential.

It is evident that with adequate filtering prior to the active elements in the receiver the effects of nonlinearity in these elements can be obviated. Ideally, the bandwidth ahead of a potentially nonlinear element should be equal to the IF amplifier bandwidth but this will generally be impractical and difficult to accomplish. Unwanted signals relatively near the desired band will therefore not always be easy to reject in the RF amplifier. Where such interference is expected as would seem to be the case when aircraft signals in the 135-136 MHz band are encountered it is desirable to use input circuits with large dynamic range to avoid such effects as overload and desensitization. The rejection filter would then be the IF amplifier. However, sharp rejection filters (wave traps) especially for rejecting fixed frequency unwanted signals in an adjacent channel have been devised (References 1, 12).

4.2.4 Interference Reduction by Electrical and Mechanical Separation

In addition to interference reduction by proper choice of the electronic device and its operating point, and by filtering, certain evasion techniques should be considered. Electromagnetic signals can be maintained separately by the use of different frequencies, time sharing, the use of codes which make different signals separable, spacial separation, shielding, and the use of different polarizations. For the case of potential satellite interference to a ground station studies are underway for establishing methods of frequency assignment to minimize interactions (Reference 42). In aerospace applications it is not unreasonable to assign the same frequency to several satellites if they are not expected in the same place at the same time too often. The use of address codes will further aid in separating signals. The beacon, for instance, is usually an unmodulated sinusoid. It could be a pulsed sine wave with a distinctive coded pulse pattern. The pulse bandwidth requirement is not expected to be very great and it is likely that the band normally allocated is sufficient.

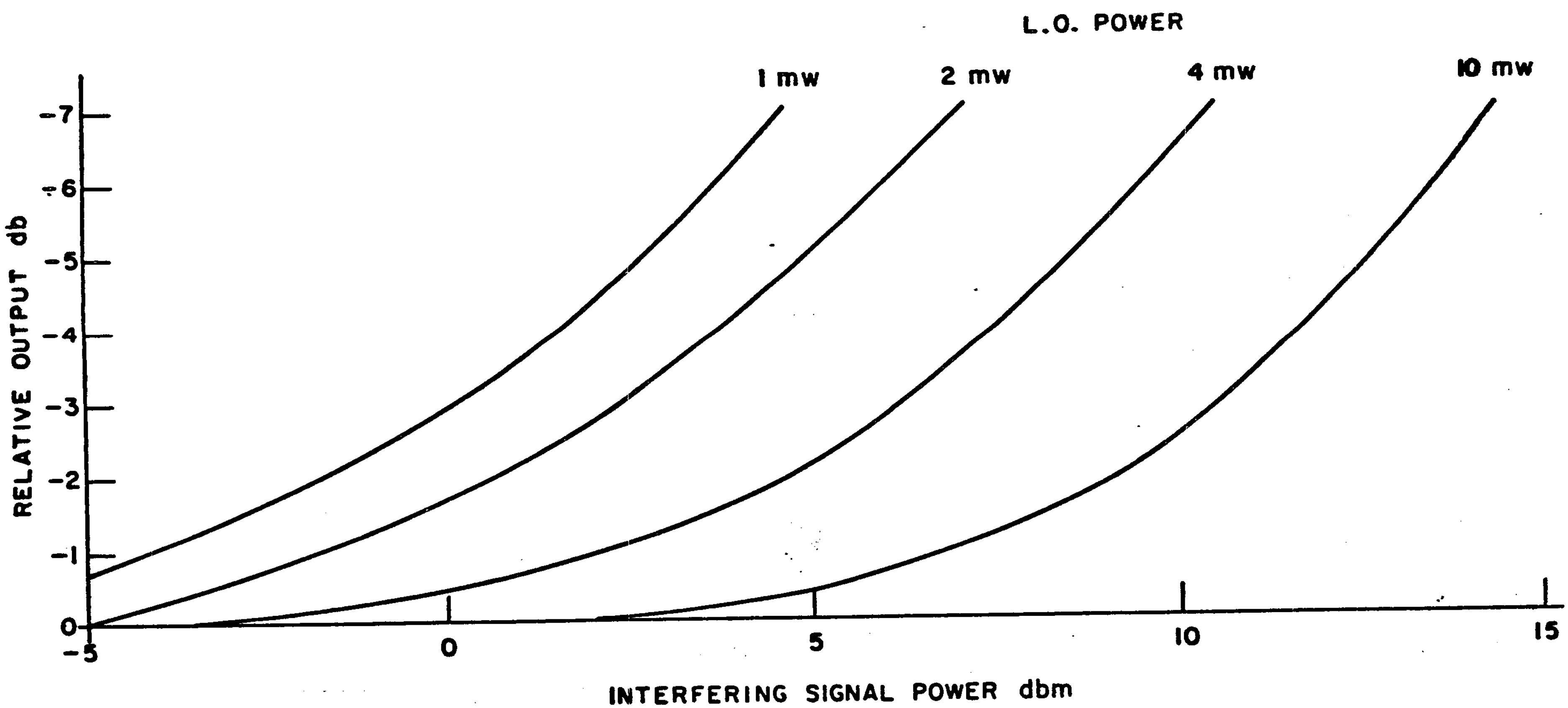

FIGURE 4-6

CRYSTAL MIXER CONVERSION LOSS TEST RESULTS

The use of polarization as a means of separation is an ultimate possibility though it may not be readily implemented at the moment. If linearly polarized ground antennas would have a cross-polarization rejection in the order of 40 db then two satellites using the same power and frequency but having orthogonal polarizations could have their transmissions separated in the ground antennas. To accomplish this would however require attitude control of the spacecraft antenna. There is, furthermore, the question of polarization rotation which may arise when the field passes through the earth's magnetic field and the ionosphere (Faraday Effect).

5.0 SPURIOUS PATH ADMISSION MECHANISMS

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

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

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

5.1 PENETRATION THROUGH CABLES

The basic mechanisms by which noise may penetrate into a receiver via the input cables are illustrated in Figs. 5-1a, b, and c. In Fig. 5-1a, noise is able to enter the receiver directly through conductive paths. In Figs. 5-1b and 5-1c, however, the noise is coupled into the input cable through impedance and inductive coupling respectively. The mechanism of 5-1b is referred to as "common-mode" interference, while that of Fig. 5-1c is referred to as "differential-mode" interference. These basic mechanisms will now be further discussed.

5.1.1 Conductive Path

Figure 5-1a is an example of conducted noise transfer from one unit to another via a common power supply. Since the power source has a finite internal impedance, the noise existing in one unit gets transferred to the other unit. Predicting the magnitude of such effects is difficult since it requires a knowledge of the noise transfer characteristics of the system as well as the susceptibility of the unit being interfered with. It is a common practice in government or industry to set maximum limits on the allowable conducted noise output from any electrical device or equipment connected to a power line.*

5.1.2 Common-Mode Path

This is illustrated in Fig. 5-1b. The ground return is seen to be common to both loops and also contains some impedance. Consequently, the two loops tend to couple into each other through the common impedance. If the impedance of the common branch were zero, there would obviously be

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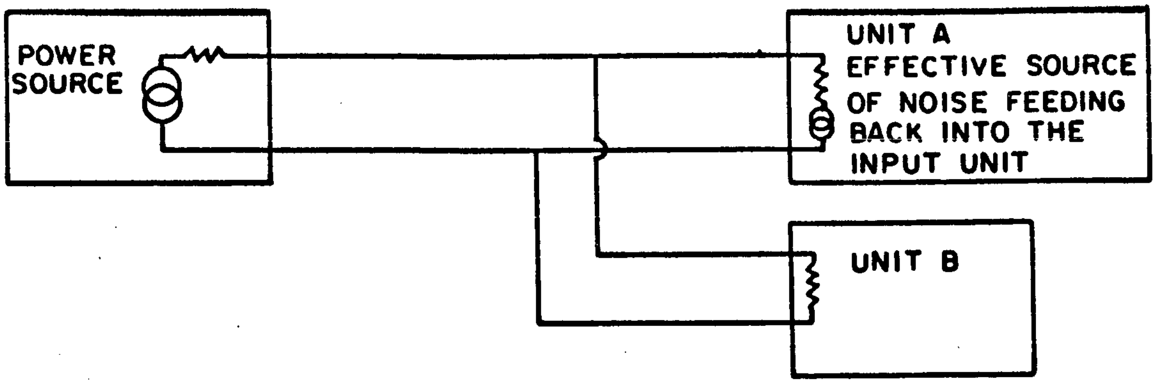
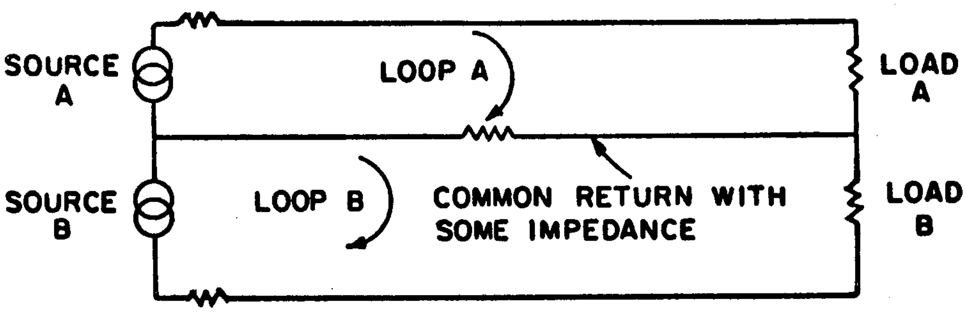
* See for example, MIL. Spec., MIL-I-6181D, Interference Control Requirements, Aircraft Equipment.

FIGURE 5-1

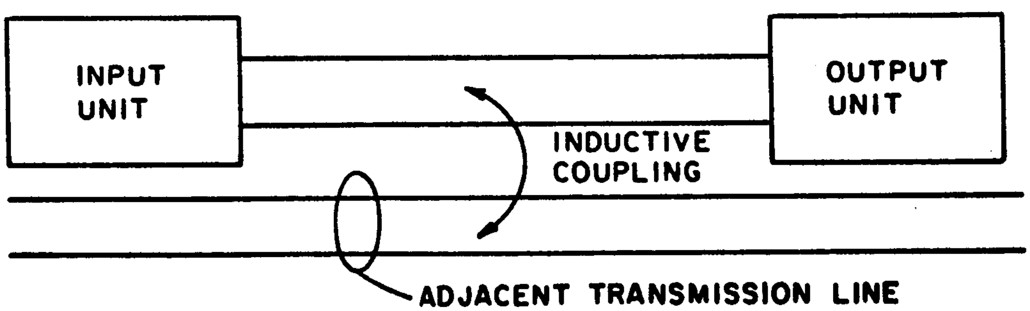
NOISE INTRUSION PATHS



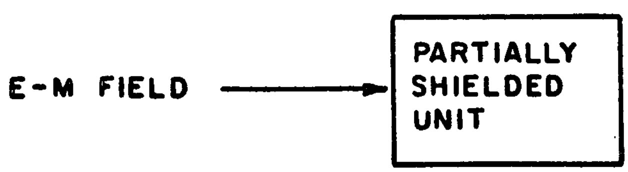
(a) Simplified Circuit for Conducted Noise Through Common Source.



(b) Simplified Circuit for Conducted Noise Through Common Path (common mode interference path).



(c) Simplified Circuit Showing Differential Mode Path



(d) Direct Irradiation of Imperfectly Shielded Unit.

no noise transfer. The obvious remedy to this problem is to avoid the use of common ground returns. A designer will frequently use separate grounds, tied to a single common point or he will use a common ground bus made of thick low impedance material. It should be noted that while resistance in a common ground return is a very important factor in common-mode coupling, inductive reactance in the ground return is equally important at high frequencies. In such a situation, the coupling can only be remedied by the use of filters.

5.1.3 Inductive Coupling

Inductive mode coupling is shown in Fig. 5-1c. Due to the coupling, noise is induced in series with the desired input and flows in the same direction.

At low frequencies (below 50 KHz) the coupling between an identical pair of cables can be accurately determined on either a theoretical or experimental basis or both. The types of cable commonly used include the coaxial and the shielded and unshielded pair. At these frequencies, electrostatic shielding, when it exists, will usually eliminate any electric coupling between cables. Magnetic coupling depends upon spacing between conductors and is proportional to the square of this distance. Inductive coupling can therefore be reduced to tolerable levels by maintaining an adequate spacing between cables or by twisting the cables, in the case where these are open unshielded lines.

A coaxial cable with a solid outer conductor and a perfectly concentric cross section will not exhibit any coupling. However, in practice, coaxial cables contain braided outer conductors, so that some coupling generally does exist. A typical value of mutual impedance is 10^{-6} ohms (1 μV induced for one amp flowing).

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

5.2 FILTERING

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

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

5.2.1 Power Line Filters

Power line filters are used to prevent unwanted high-frequency signals from being coupled between equipments by a common power connection. A common technique is to utilize a "brute force" filter; i.e., several ladder sections in which the choice of components is not critical. Sometimes it is necessary for such filters to have extremely broadband performance, e.g., filters for screened rooms. In these cases multiple sections are used, each being designed to reject different portions of the spectrum. For stringent requirements multiple section filters combining constant k and m-derived sections may be required. Filters of this kind may be useful in other applications where lumped parameter filters are needed. An example of the design of such a filter will be found in Reference 43, p. 184.

There are a number of constructional precautions which must be followed in all filters and some particular ones for this class of filters.

Since substantial amounts of power-line current may be required, the filter inductors may require rather large sized wire. This places severe limitations on how much inductive reactance can be secured.

5.2.2 Bypassing and Feedthrough Capacitors

The objective in bypassing is to provide a low impedance shunt path to effectively "short circuit" unwanted signals. Typically, a capacitor is used where impedance at the frequency in question is less than a tenth of that of the circuit being bypassed. It is to be noted that a 10:1 impedance ratio may not be enough in some cases. Unfortunately, it does not always suffice to merely use a larger capacitor, since these elements have internal and lead inductance which provide a frequency limit above which their capacitive properties cannot be realized. (See, for instance, Reference 13, figure 6.27 for the resonance properties of capacitors.) The self-resonant property is sometimes utilized to obtain nearly perfect bypassing in a narrow band of frequencies. Bypassing of input and output leads can be effectively accomplished using feedthrough capacitors, which provides a conductive path from one terminal to the other, but gives at the same time a high capacity (up to 2300 pf) to a bulkhead or mounting panel. They are available in such small sizes that they can be incorporated into electrical connectors. In special cases transmission line sections may be used instead, making use of the impedance transforming property of a quarter-wave line. However, each such contemplated use must be considered individually.

Reference 13 provides useful data on modern synthesis and design of electrical filters. Many examples are given on the design of low-pass, high-pass and band-reject filters. Reference 1, Secs. 7.5.4.2

and 7.5.4.3 give details of filter design at VHF/UHF frequencies and at microwave frequencies.

5.3 GROUNDING

Conducted noise on signal leads and cables is frequently created by poor grounding practices. Such problems can only be avoided by very careful design of the grounding system. (see References 44 and 45)

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

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

5.3.1 Single Point vs Multiple Point Ground

Generally, multiple point grounds are preferred by electronic designers for two reasons:

(1) convenience--circuit construction can be simplified by returning circuit elements to the nearest appropriate point;

(2) circuit efficiency--at the higher frequencies, the lead lengths (connections to components) must be kept short in order to keep the self-inductance of the leads small.

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

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

When circuits are separated by long distances, cables are generally used to interconnect the system. The use of cables can create noise problems, as follows:

(1) The length of the cable can be such as to be resonant. If so, the cable can act as an efficient radiator of energy; it can also receive energy radiated from some other source. The possibility of such radiation can be substantially reduced by connecting the cable at a large number of points to earth grounds.

(2) The ground connection of the cable provides a common mode path for coupling of currents that may flow in the earth or reference grounds. The solution is to use balanced signal circuits, so that undesired voltages or currents are balanced out of the system.

5.3.2 Balanced Circuits

Figure 5-2 shows the arrangement of a balanced coupling circuit. Current flowing in the loop formed by the cable shield and the ground will induce voltages across the twisted pair signal leads. If these leads are perfectly balanced, the output voltage at the receiver will be cancelled out. The use of a balanced cable has therefore eliminated the noise pick-up. Current in the shield of the cable can be minimized by connecting only one end of the shield to the ground, as shown, but then a substantial potential may appear between the end of the shield and the receiver chassis. However, the balanced cable helps to reduce the effects of this. In particularly

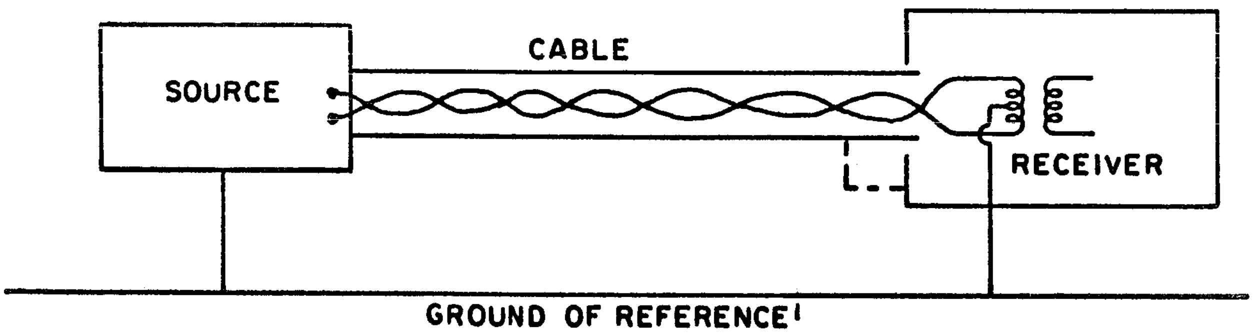


FIGURE 5-2

BALANCED COUPLING CIRCUIT

sensitive circuits the balance may be made adjustable.

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

5.3.3 Ground Loops

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

It may be noted that in a balanced twisted pair signal coupling can be used. The effects of such loops can be made negligible in the signal circuit.

5.4 PENETRATION THROUGH SHIELDS

Radiation may be coupled into a susceptible device through a shield of inadequate thickness, through holes provided for ventilation and for other purposes, and through imperfectly joined shield sections. Precise calculation of shielding effectiveness, even for solid perfectly joined shields, depend on the form of the shield and the field structure. Coupling can be both of an electric and magnetic nature, but normally it is easy to provide electric shielding so that this is usually not a serious problem. Magnetic shielding is more difficult to provide, particularly at frequencies below 100 KHz. To avoid uncertainties in critical situations tests will have to be made of the shielding effectiveness. Such tests require establishing a known field and measuring the insertion loss.

5.4.1 Nonlinear Effects in Shielding Materials

Interference may also be created by certain nonlinear characteristics in the metal which forms the equipment shield. This gives rise to cross-modulation components which become manifested in the receiver output. Such effects are usually only created when the equipment is in a relatively strong RF field, such as that which exists in the vicinity of broadcast and radar transmitters. Nonlinear effects are known to arise in magnetic materials, such as steel, nickel, mu-metal, etc., and in corroded metals and corroded joints (especially loose joints). Rough or oxidized surfaces on steel tend to increase the nonlinear effects. Consequently, if the surface is coated with a non-magnetic conducting material, such as copper, these effects become significantly reduced. Welded joints have also been found to be superior to riveted or bolted joints. A recent paper (Reference 46) gives further details of these effects.

5.5 SHIELDING AND BONDING

For VHF equipment shielding against stray fields is mostly a problem of providing adequate bonding at the seams and of maintaining adequate shielding at the ventilation louvres. It is unlikely that the shield will be too thin; if the shield is adequate mechanically, then it will also be adequate electrically at these frequencies.

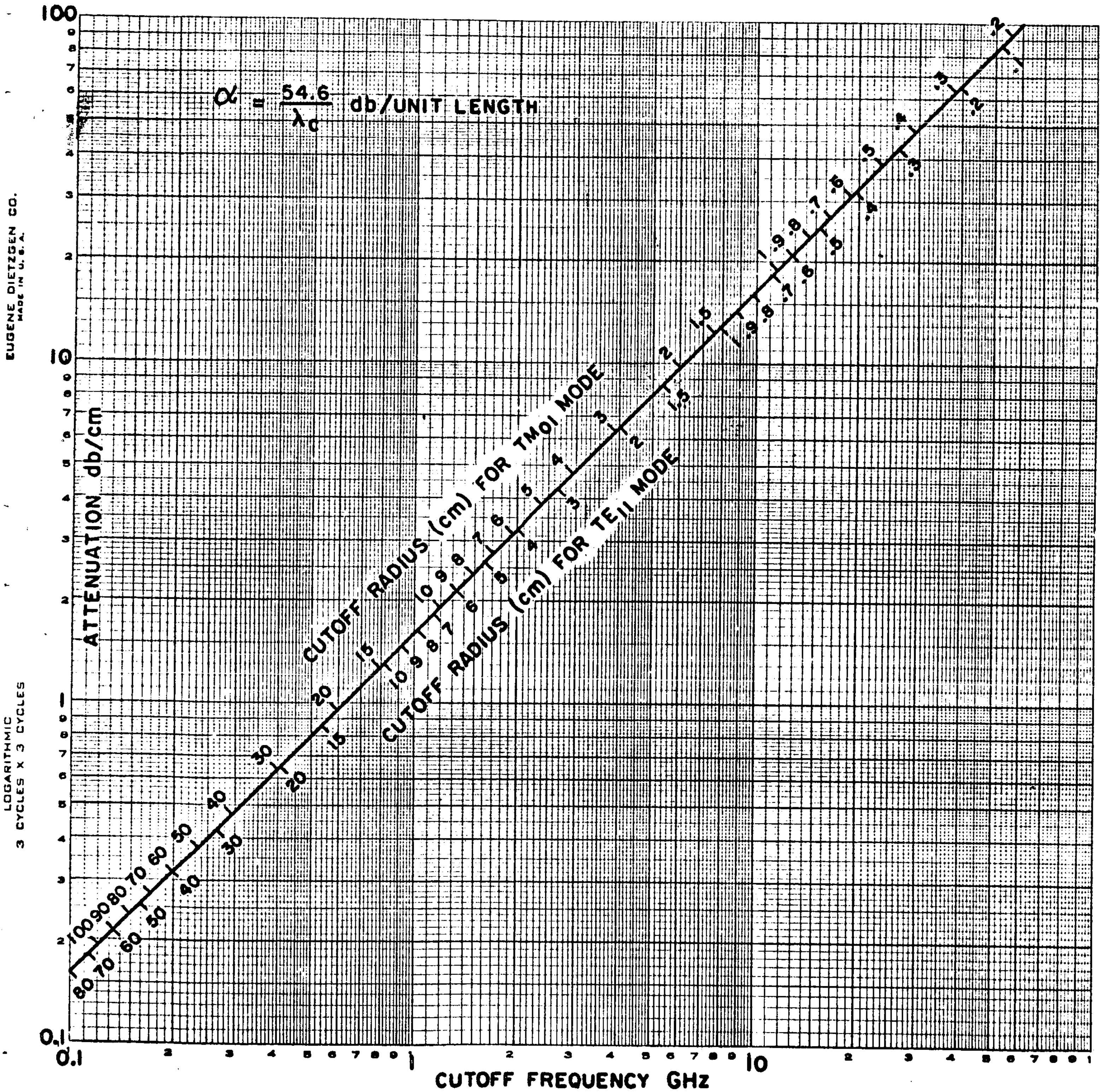
The function of a shielded enclosure is to terminate the RF fields which exist both inside and outside the shield. The terminated fields give rise to surface currents which flow on the inside and outside surfaces of the metal shield. In a perfect shield, these surface currents are able to circulate in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

However, if there are conductive imperfections at a seam joining sections of the shield together then part of the inside surface current will be able to flow to the outside surface and similarly part of the outside surface current flows to the inside surface. Consequently part of the field inside the equipment becomes propagated outside the shield and, in the same manner, part of the outside field is set up inside the shield. Leakage has therefore resulted due to imperfect bonding in the metal shield surrounding the equipment.

Perforations, louvres, etc., are obviously essential for adequate ventilation in the equipment. Consequently, it is generally not possible to have an ideal solid shield. Small perforations in the metal or sections of conductive screening usually provide satisfactory shielding at ventilation louvres, since the surface currents will flow around the openings without appreciably penetrating them. It is mainly at seams, long louvres and at imperfectly joined shielded cables that the surface currents are significantly forced onto the opposite surface. Whenever conducting seams are required, soldered or welded joints are preferred. Pressure joints must be clean; abrasive gasket material that

FIGURE 5-3

ATTENUATION OF CIRCULAR TUBES USED IN AIR DUCTING ASSEMBLY



is also conductive, may be useful. Further details on shielding in general are given in Ch. 5 of Reference 3.

5.5.1 Duct and Shaft Filters

Duct filters are sometimes used to improve the shielding inside ventilation louvres. They consist of an array of parallel closely spaced tubes. The tubes are conductive and have a narrow cross section. Consequently, the tubes act like waveguides below cutoff and possess a relatively high attenuation to any fields incident on the filter. When the tube is far below cutoff ($\lambda \gg \lambda_c$) the attenuation of the tube is given approximately by:

$$\alpha = \frac{54.6}{\lambda_c} \text{ dbs/unit length} \tag{5-1}$$

Figure 5-3 shows the attenuation properties of a circular waveguide operating well below cutoff. It can be seen that the smaller the cross section of the tube, the greater is its attenuation. For example, suppose that a tube attenuation of 30 db/cm is required in a certain ventilation assembly. Using Fig. 5-3, it can be seen that a tube of about 4 mm radius is needed, using the TE_{11} mode as the "worst case" (lowest attenuation) mode. The cutoff frequency of the tube is 20 GHz. The desired attenuation can therefore be obtained at frequencies up to 10 GHz. A tube length of 4 cms will therefore provide a total shielding effectiveness in the filter of about 120 db, which is generally considered to be more than adequate for most applications.

The use of such ventilation tubes makes the effective attenuation of these assemblies much more readily calculable than in the case of holes or louvres in a panel. "Edge" effects tend to introduce large errors in the calculation of the attenuation of louvres.

6.0 INTERFERENCE CONTROL

6.1 INSTRUMENTS AND MEASURING METHODS

The measurement of noise, and susceptibility of equipment to noise, is a vast field which, in a brief account, is apt to be oversimplified. A number of military specifications will be found to be germane; e.g., MIL-I-6181D, "Interference Control Requirements, Aircraft Equipment," MIL-I-11748B (Sig. C), "Interference Reduction for Electrical and Electronic Equipment," MIL-I-16910A (Ships), "Military Specification, Interference Measurement, Radio, Methods and Limits, 14 kc to 1000 Mc," MIL-I-26600 (USAF), Military Specification, "Interference Control Requirements, Aeronautical Equipment." There are other applicable specifications among which are certain ones which are preliminary at this writing, but which reveal a more determined attempt to discover the interference properties of equipment. Among these are MIL-STD-826(USAF), "Electromagnetic Interference Test Requirements and Test Methods." There are also some recent publications of non-government groups which apply to tests and test equipment. These are American Standards Association publications, C63.2-1963, "Radio-Noise and Field Strength Meters 0.015 to 30 Mc/s;" C63.3-1964; "Radio-Noise and Field Strength Meters, 20 to 1000 Mc/s," and C63.4-1963, "Radio-Noise Voltage and Radio-Noise Field Strength 0.015 to 25 Mc/s." A publication of the National Electrical Manufacturers Association, No. 107-1964 on "Methods of Measurement of Radio Influence Voltage of High Voltage Apparatus," is also germane.

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

The tests are intended to reveal both the susceptibility of sensitive devices (Section 6.2.1) and the noise generating properties of electrical equipment--in radiation or induction modes, and in conduction modes (Section 6.2.2). Well established standards have been set by the military for frequencies up to 1000 MHz; beyond this frequency the procedures and limits are still tentative.

6.2 SUSCEPTIBILITY TESTS

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

For testing purposes, typical impulse sources generate unidirectional short impulses lasting anywhere from 10^{-9} to 10^{-10} seconds. Correspondingly, the energy is spread from zero frequency to the 1-10 GHz range. An RF pulse generator of say, 10^{-7} seconds pulse duration, will have the same spectral density as the impulse, over a band of about 10 MHz. Since many potentially interfering pulse sources are, in fact, better simulated by the latter, RF pulse testing is, in many cases, advisable.

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

6.2.1 Conducted Susceptibility Tests

Conducted susceptibility tests are carried out on electronic components which are connected to power sources or to other system components. Such tests ideally are made on all external leads. For conducted noise on power lines standard procedures have been devised (see the specifications enumerated above). Typically, (MIL-I-6181D), for measuring susceptibility to audio voltages in series with the power line a setup as shown in Fig. 6-1 is used. The capacitor C acts to insure low impedance at the power supply at all frequencies of test and the test is carried out over the entire audio range (50-15,000 Hz). The effect of the test is to insert an audio voltage source of a certain magnitude and impedance in series with the power supply. The various standard specifications invoke slightly different conditions. One, for instance, requires that the effective open circuit voltage inserted in this way be three volts and the impedance be less than 0.6 ohms. Under these conditions "no change in indication, malfunctioning, or degradation of performance shall be produced."

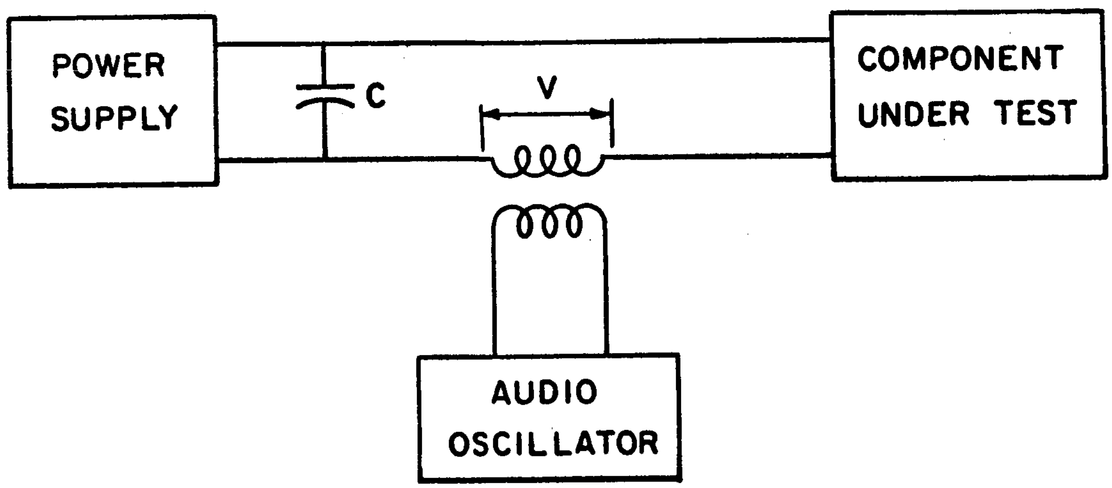


FIGURE 6-1

AUDIO SUSCEPTIBILITY TEST ON POWER LINE

Tests of RF sine wave susceptibility on power lines is also fairly well standardized. Each ungrounded line is tested sometimes using the arrangement shown in Fig. 6.2. In effect, the noise voltage from the RF signal generator is applied between a "hot lead" on the power line and ground. The "impedance stabilization network" is used to standardize the impedance looking back from the unit under test into the power source. The impedance looking into terminals 1-3 on the network varies from about 5 ohms to about 50 ohms for the conditions shown over the range of frequencies for which the device is used. Specifications vary on the conditions to be employed. Typically, (MIL-I-6181D) the RF sine wave is amplitude modulated at 400-1000 Hz, and is applied as shown with a carrier level of 1/10 volt over a range of frequencies from 0.15 KHz to 10 GHz. Under these conditions no indication on the output or malfunction is permitted. Various deviations from this procedure are quoted, some merely applying the signal generator from a line to ground with any device to isolate the power line (MIL-STD-826).

Similar procedures are sometimes prescribed for impulse testing. One specification MIL-I-11748B requires that the circuit of Fig. 6.2 be used with a standard impulse generator replacing the RF signal generator. Proper functioning is demanded with an impulse spectral density of 90 db above one microvolt per MHz.

When leads other than the power lead are to be examined similar methods are used. The specifications listed above do not prescribe any fixed methods and procedures will have to be tailored to the requirements of the equipment and to the environment. Broadband inputs to control leads and to input leads will ordinarily result in some effect since the circuits

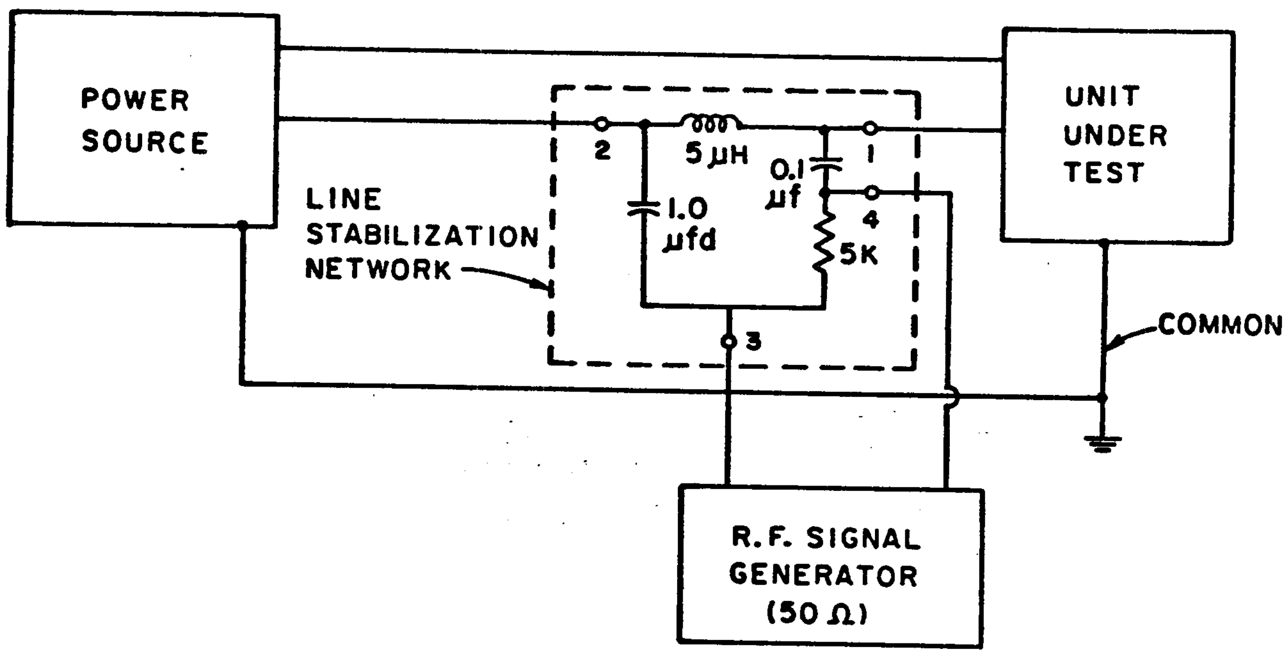


FIGURE 6-2

RF SUSCEPTIBILITY TEST, POWER LINE

to which the leads are connected are meant to accept some of the frequency components in the broadband source. Either input levels of noise must be appropriately limited, or filters must be used to eliminate noise in the band to which the device is normally receptive.

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

6.2.2 Radiated Susceptibility Tests

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

A typical test setup is shown in Fig. 6.3 showing the use of a rod antenna placed some standard distance from the equipment under test. This distance is specified as one foot in MIL-I-6181D, one meter or 7.6 meters depending on the equipment class in MIL-STD-826, and 25 feet

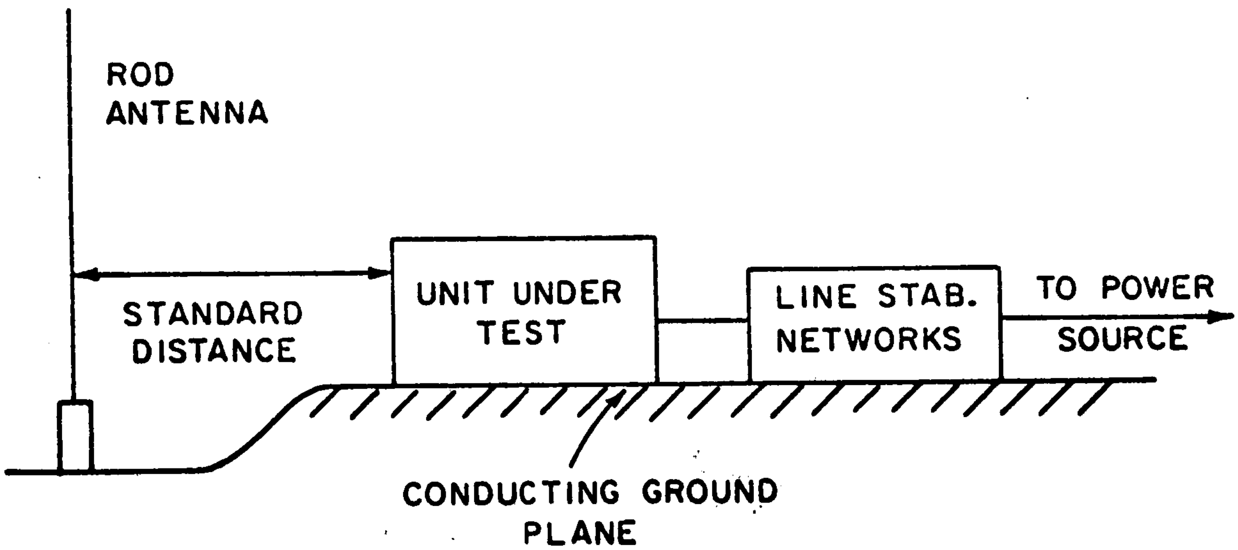


FIGURE 6-3

TYPICAL TEST SET-UP, RADIATED SUSCEPTIBILITY MEASUREMENTS in MIL-I-11748B. The latter specification actually allows reduction in the distance as frequency is increased; the distance quoted is the one specified for low frequencies and is apropos for measurements with a rod antenna.

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

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

6.2.3 Other Tests

The mechanisms of spurious signal susceptibility described in Section 4.2 are examined experimentally using various standard test procedures. A set of such procedures is fully described in MIL-STD-826 and will therefore not be repeated here. Tests are specified in the aformentioned specification for such mechanisms as:

1. two signal intermodulation;
2. spurious response to a single unwanted signal;
3. cross modulation.

In each case the methods simply involve the use of standard signal generators as calibrated sources representing both wanted and unwanted signals. Generally, however, a great deal of care is required in setting up and performing the tests, since most signal generators, themselves produce spurious outputs which can cause confusion in interpreting test results. For tunable receivers it may be necessary to make tests at many points of tuning since the sensitivity to unwanted signals is not constant throughout. Some standard procedures make compromises here by requiring tests at a few points in each band of tuning; e.g., points near the edge of each continuously tuned band and one point near the middle of the band. MIL-STD-826 requires scanning throughout a frequency range. This naturally is a time consuming operation. As a rule the responses will be found where they are expected and it may save much time, particularly for testing at the production stage, to examine only those points at which responses are expected.

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

6.3 TESTS OF NOISE OUTPUT

Tests of noise output are carried out using set-ups which are not too different from those used for susceptibility testing. The important distinction being that a Radio Frequency Interference (RFI) measuring instrument replaces the signal (or impulse) generator. RFI measuring instruments are available for frequencies ranging from low audio into the microwave range. Though manually tunable instruments are used most often there is a tendency today to utilize either mechanically tuned devices to speed up data collection, or electronically swept spectrum analyzers.

The two major problems of measurement are the following. First, measurements of electromagnetic fields at distances close to the source are variable, depending heavily on antenna location and on surrounding objects. Unlike measurements taken in the far field of a source, measurements in the near field are almost impossible to extrapolate to give the field level at other distances. However, if measurements are attempted at a distance from the source the measured value is naturally smaller and the effect of other sources in the environment begin to be felt. Though shielded enclosures for tests are not always demanded they are often used. The maximum distance is naturally restricted in such instances. The second problem concerns the measure of the noise itself (Reference 47). The instruments under discussion all have a moderately narrow RF-IF band; typically, bandwidths are in the order of 0.1-1 per cent of the tuned frequency. For the measurement of pure sinusoids all the instruments can be calibrated to give the same reading. For broad-band inputs the readings will depend on the detailed nature of the input, on the bandwidth, and on the function performed by the detector following the RF-IF amplifier. Three kinds of detectors will be found on standard instruments--namely, average of envelope detector, peak detector, and quasi-peak detector. Some instruments contain all three, and some omit the quasi-peak detector. The average of the envelope detector reads, as the name indicates, the average of the envelope of the output of the IF amplifier. The peak detector reads the peak value of the IF amplifier output. The quasi-peak detector schematically shown in Fig. 6.4, which reads less than the peak value is, in effect, defined by its charging and discharging time constants. In American standards the charging and discharging times (measured by applying a sine wave step input and determining the time to reach 63% of the final output, or by suddenly removing the input and determining the time required to fall 63% of the way to zero) are 1 millisecond and 600 milliseconds, respectively.

The response of such instruments to various inputs, assuming all of them have been calibrated to read the root-mean-square value, E, of an applied sine wave is shown in Table 6.1 (Reference 48). Indicated here is the reading obtained with an RMS detector, too, though instruments are not ordinarily equipped with such a device. The impulse input is a periodic pulse which occurs with a period not so high that successive pulses will overlap one another in the RF-IF amplifier. The quantity S_i is defined as the impulse strength and is twice the Fourier spectral intensity at the tuned frequency (see Section 4.1). For a pulse of duration τ ($\frac{1}{\tau}$ >> tuned frequency) and height A

$$S_i = 2A\tau \tag{6-1}$$

For the quasi-peak response to periodic impulses of the sort described the reading depends on the recurrence frequency of the pulses, on the bandwidth, and on the time constants. This dependence is shown in Fig. 6.5. The bandwidths, etc., in Table 6.1 are defined in Section 4.1.

6.3.1 Conducted Noise Measurements

Measurements of the noise level fed out of a piece of electrical apparatus on the power input line, are frequently carried out using the set-up of Fig. 6.2. The RF signal generator is replaced by an RFI measuring instrument. The line stabilization network prevents variability of the

Table 6.1

RESPONSE OF RIFI METER

Input	Detector Function			
	Peak	Average	Q.P. (1 ms-600 ms)	RMS
Sine Wave	E	E	E	E
Periodic Impulses (rate=f_p per sec.)	$\frac{S_i B_e}{\sqrt{2}}$	$\frac{S_i f_p}{\sqrt{2}}$	$\frac{S_i P(\alpha) B_e}{\sqrt{2}}$	$\frac{S_i \sqrt{f_p B_p}}{\sqrt{2}}$
Random Noise	--	$0.884 \sqrt{B_p N}$	$1.820 \sqrt{B_p N}$	$\sqrt{B_p N}$

N, B_e, and B_p are defined in Section 4.1.

power line impedance from affecting the readings. Alternative methods are suggested in some places involving capacitive connection to the line, particularly in instances where the power drawn by the equipment under test is high. Another procedure involves an inductive measurement of the noise current on the line using a clamp-on toroidal coil. In this latter method the RF impedances of the power source and the line do affect the readings. These methods can also be used on other interconnecting cables, such as control lines.

6.3.2 Radiated Noise Measurements

Measurements of noise field in the vicinity of an electrical device are made in a set-up essentially that of Fig. 6.3, with the antenna feeding a standard RFI measuring instrument. As in the case of susceptibility measurement the distance from the source depends on the particular specification as do the details of the setup. At low frequencies loops and rods are used; at higher frequencies dipoles and horns are used, as discussed in 6.2.2. Since the field, as pointed out

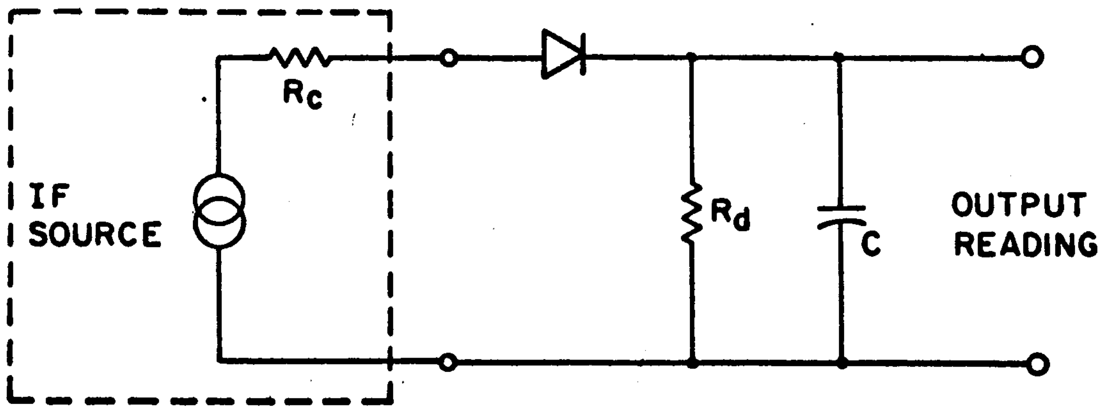


FIGURE 6-4

QUASI-PEAK CIRCUIT

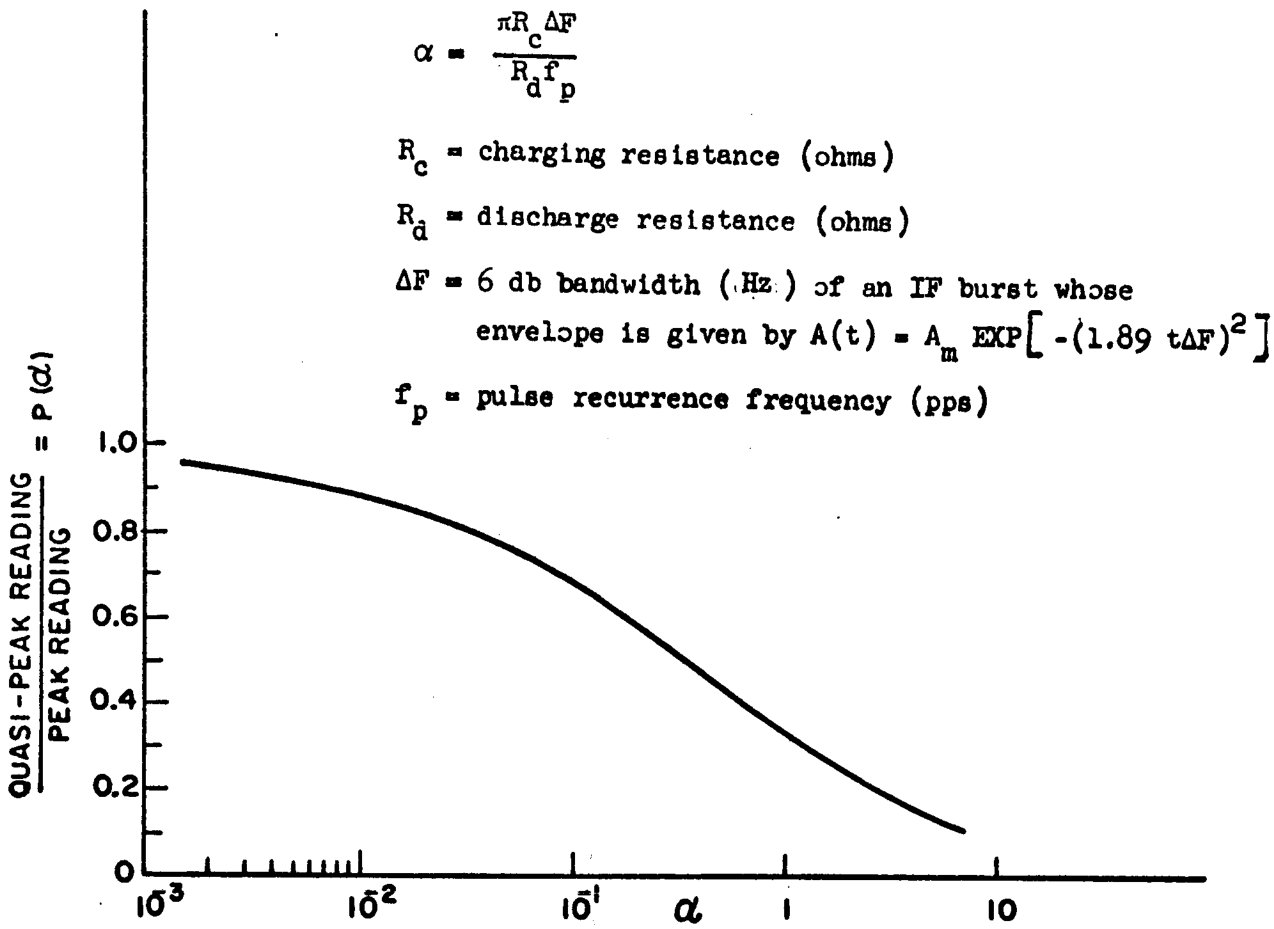


FIGURE 6-5

A RATIO OF QUASI-PEAK READINGS AS A FUNCTION OF RFI METER PARAMETERS before, depends on location it is common to find the requirement that the test sample be turned until the point of maximum emission is found.

6.4 SYSTEM TESTS

The objective of the component tests is to insure that the overall system in which the component is connected will not malfunction due to interference. Proper operation in a component test is however no absolute guarantee of proper operation of the system. An overall shake-down test is therefore advisable. Generalizations concerning systems tests are not too helpful and it will only be said here that the test should adequately simulate the normal operating conditions. Specifics concerning the system tests must be tailored to the system under study.

7.0 SITE SELECTION

The factors influencing site selection for ground stations are here discussed in brief. The material herein presented was assembled as part of an interference study program which was reported in Reference 42 (Section 2).

7.1 FACTORS INFLUENCING THE SELECTION OF A GROUND STATION SITE

The selection of an optimum site for a ground station is bound by a number of constraints. These may be generally grouped into the following categories: (a) geographical region, (b) local topography and soil conditions, and (c) RF environment.

In order to ensure that communications between satellite and ground are the most reliable possible, it is essential to establish a ground station in "quiet" areas where the level of local ambient RF noise is as low as possible and where there are not nearby ground or airborne transmitters operating on the same or adjacent frequencies.

Operational requirements will largely dictate the selection of the geographical region in which the station is to be located. The topography criteria, outlined in the second category, will be briefly discussed in the following sub-section. However, most of this section is devoted to a detailed consideration of the RF environment criteria outlined in the last category.

7.2 LOCAL TOPOGRAPHY

For very sensitive receiving systems, the most suitable topography for a Data Acquisition Facility site is actually a very shallow "hollow", composed of a relatively flat plain and completely surrounded by distant hills. The hills will effectively shield the site from unwanted RF signals and noise being propagated along the ground. Unfortunately, these hills will slightly reduce the effective hemispheric coverage of the antennas. However, it is usually not possible to obtain much data from a satellite when it is positioned right on the horizon, owing to the large increase in antenna noise temperature which results when the receiving antenna is pointing at the horizon. Consequently, very little data will be really lost if only a fraction of the full 180^{o} coverage is obstructed by the surrounding hills. This shielding should not exceed about 10^{o} above the horizontal. This, then dictates the allowable height for the surrounding hills, depending in turn on their distance from the proposed site.

7.3 FACTORS THAT INFLUENCE THE RF ENVIRONMENT OF A SITE

These factors may be briefly summarized as follows:

(a) natural and man-made ambient noise background at a site;

(b) proximity of fixed and mobile transmitters operating on the same frequency or adjacent frequencies as the satellite to ground communications;

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

(d) the natural shielding which a site possesses (already discussed in 7.2).

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

7.3.1 Natural and Man-Made Ambient Noise Levels

At VHF/UHF frequencies, man-made noise generally represents by far the largest part of the overall background noise level at any particular site. Noise from natural sources is entirely of extra-terrestrial origin and is relatively insignificant. Figure 7-1 shows the relative noise levels from different sources which can be expected when using a half-wave dipole as the sensor.

7.3.2 Man-Made Noise

In a recent paper, Skomal (Reference 49) presents available data on the background noise level which is of man-made origin. Most of the sources of man-made noise are naturally concentrated in urban centers of population, especially those which are heavily industrialized. Such

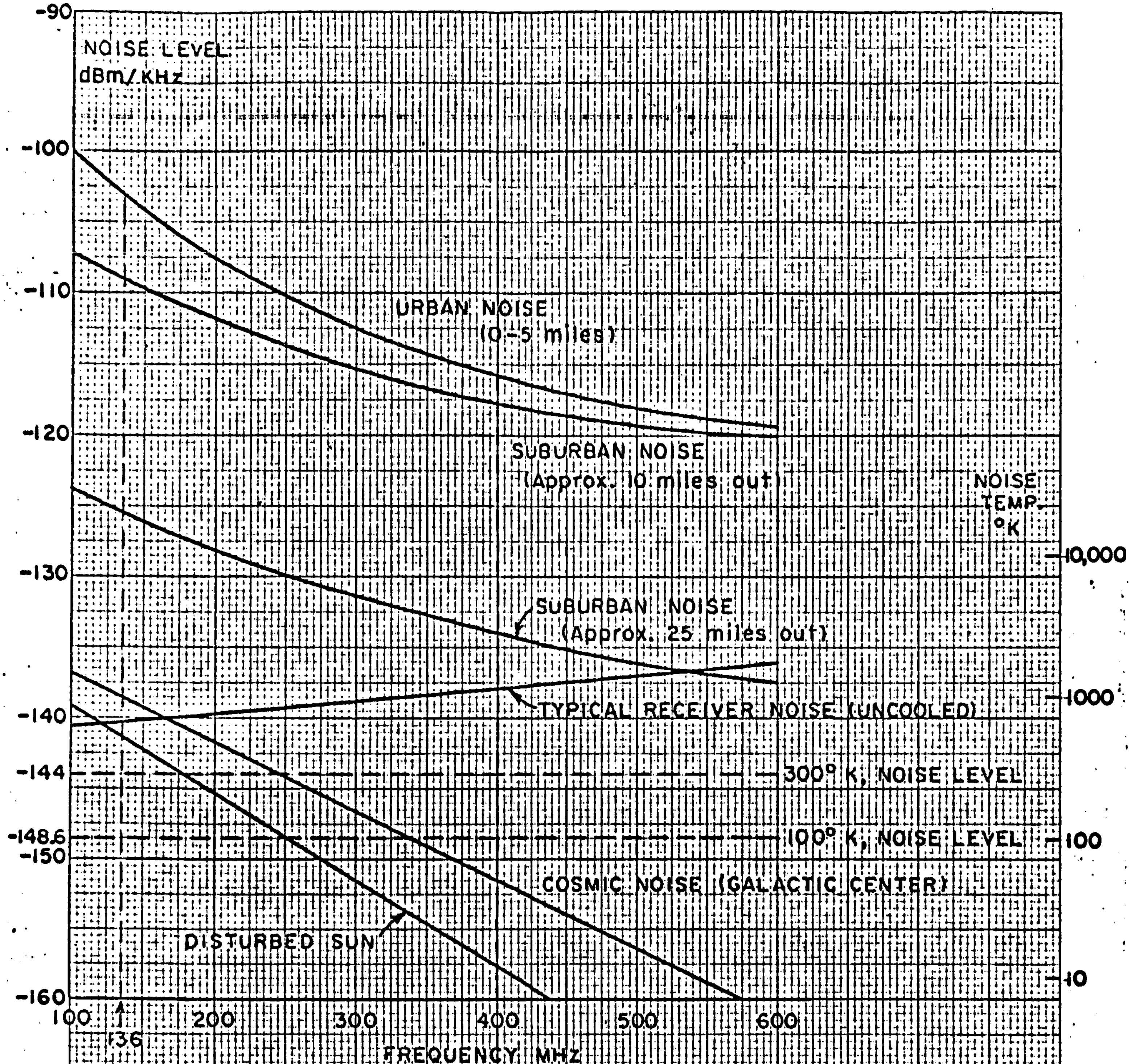


FIG. 7-1 RELATIVE NOISE LEVELS FOR A HALF-WAVE DIPOLE.

sources include automobile ignition systems, corona discharge from power lines, fluorescent lights, switching systems, motors, rectifiers, arc welders, etc. Figure 7-1 presents Skomal's data for the noise background in a typical city, and at points approximately 10 miles and 25 miles from the city center. These data were also obtained using a half-wave dipole. It will be noted that the noise levels within a 10 mile radius of the center are prohibitively high, thus making it undesirable to locate a sensitive receiving system within or very close to a city. Figure 7-2 shows the average fall-off of the noise levels with distance from the urban center.

Various authors have attempted to correlate the level of man-made noise existing in an urban environment with the population of the town or city. However, it has not been possible to obtain meaningful correlations owing to the general lack of sufficient data. Skomal has concluded that the data shown in Figs. 7-1 and 7-2 is applicable to any town or city with a population of 50,000 or greater. Based upon experimental evidence Skomal furthermore concludes that the major source of man-made noise at VHF-UHF frequencies is automotive ignition interference.

Automobiles are, of course, not confined to urban areas, so that low background levels of ignition noise can be expected in rural areas. Generally, these levels will be too low to be even detectable. However, ground stations should be located away from major highways to ensure that the receiving systems are not disturbed by ignition noise. This will be especially true of the DAF sites which utilize the more directional antennas.

Since there are obvious practical and economic advantages to locating a site in close proximity to towns or cities, the question which

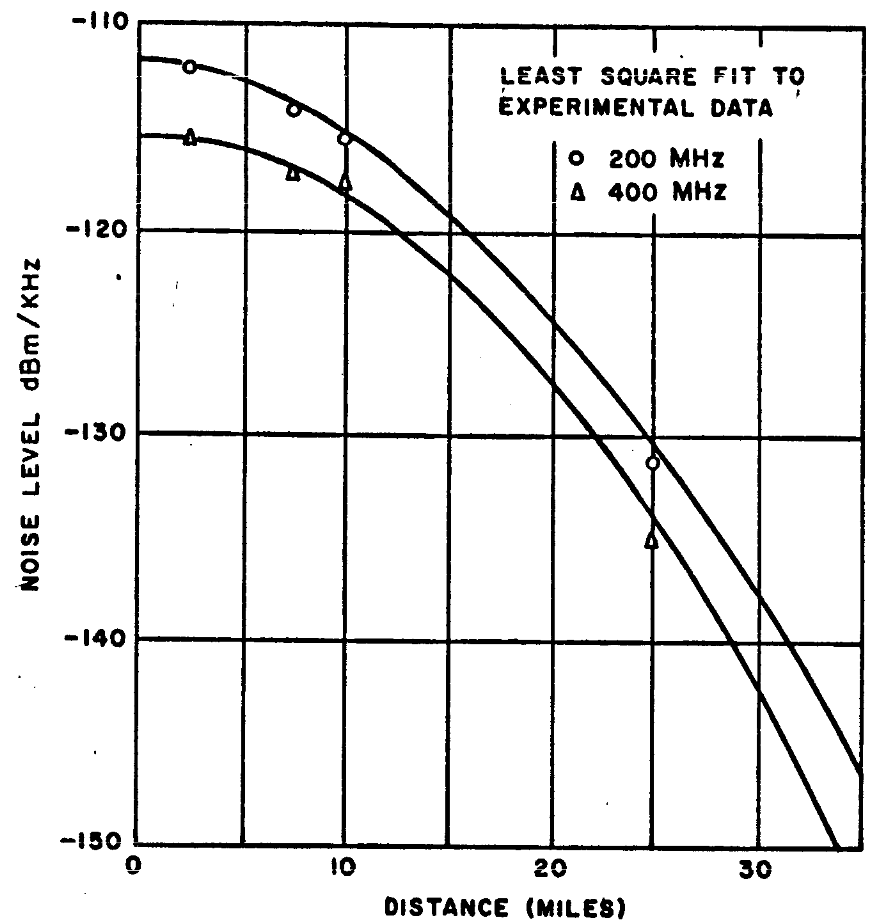


FIGURE 7-2

CHANGE OF MAN-MADE NOISE WITH DISTANCE
FROM URBAN CENTER

now remains is how close to a town can such a site be located and still continue to operate without interference? From the data already presented in this section, it may be safely concluded that stations of the Minitrack type can be located as close as 30 to 35 miles of a town of 50,000 population or greater. A site which effectively shields the receiving system from the urban noise, could actually be chosen even closer to the town. Sites for the Data Acquisition Facilities, which use the more sensitive receivers and more directional antennas, should be located further away to eliminate the possibility of serious interference when the antenna is pointing directly at the town; a distance of approximately 50-60 miles should be a good compromise. This distance will be modified by the height of the antenna above ground and the irregularity of the terrain.

7.3.3 Fixed and Mobile Transmitters Operating on the Same or Adjacent Frequencies

In recent years, certain frequency bands have been allocated by the International Telecommunications Union for the transmission of telemetry data from scientific satellites. Those frequency bands, which are presently of concern in ground station operations,* are as follows:

(a) 136-138 MHz

(b) 400.04-402 MHz

(c) 1700-1710 MHz.

* Other frequencies are being used by NASA for such purposes as communication satellite down-links and down-links from deep space probes. However, such space craft are ordinarily not associated with the STADAN system and will therefore not be considered in this study.

Although the exclusive allocation of the above frequency bands to the space telemetry service has been realized within the United States, this is not the case throughout the rest of the world. In many nations of the world, the above frequency bands are, in effect, shared by both the space service and the fixed and mobile service (or other services). Consequently, low power transmitters continue to operate in these bands. The location of sites in close proximity to any of these legally operated transmitters, should be avoided.

The fact that the fixed and mobile services continue to operate in frequency bands adjacent to the three space research bands has also created the possibility of adjacent channel interference to ground station operations. A serious problem already exists in the United States due to adjacent channel interference from the aeronautical mobile service which operates at frequencies immediately below 136 MHz. Some sites are particularly susceptbile to interference from airborne transmitters, owing to the lack of path attenuation between transmitter and site. This problem is discussed further in Section 7.3.3.3.

7.3.3.1 I.T.U. Radio Regulations

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

The details of the three space research allocations which were made at the EARC, as well as the existing allocations in the adjacent bands above and below the three bands, are given in an appendix to this section (Reference 52).

7.3.3.2 Distribution of Fixed Transmitters Throughout the World

In order to determine suitable sites on a world-wide basis, it is advisable to try to determine beforehand some of the locations of the fixed low power transmitters that are being legally operated throughout the world in the space research bands and adjacent bands. If these locations are known, then the surrounding areas can be avoided during the preliminary site selection process.

Figures 7-3 to 7-8 are maps showing the locations of some of these fixed transmitters throughout the world. These locations and the data concerning the transmitters, have been obtained from the International Frequency List, compiled by the International Frequency Registration Board in Geneva (Reference 7) (note that these listings cannot be considered as complete). In the six maps which have been prepared, the so-called "aeronautical stations" or "aircraft stations," which are used for ground communications with aircraft, have been separated from the remaining fixed transmitters that are employed for communications with other fixed land stations or with mobile land or ship stations. The aviation service transmitters are represented by a cross (x) on the maps while the other transmitters are marked by dots. Interference is likely to be created, not so much by the ground transmitter itself, but by aircraft which are likely to be flying in the surrounding areas and which will be in radio communication with that ground station. Therefore, sites should be located, Norwegian Provinces: (See Note [illegible])

Agder
Hordaland
Horten
Kongsberg
Kongsvinger
Moss
Nordland
Oslo
Rogaland
Sandefjord
Sarpsburg
Troms

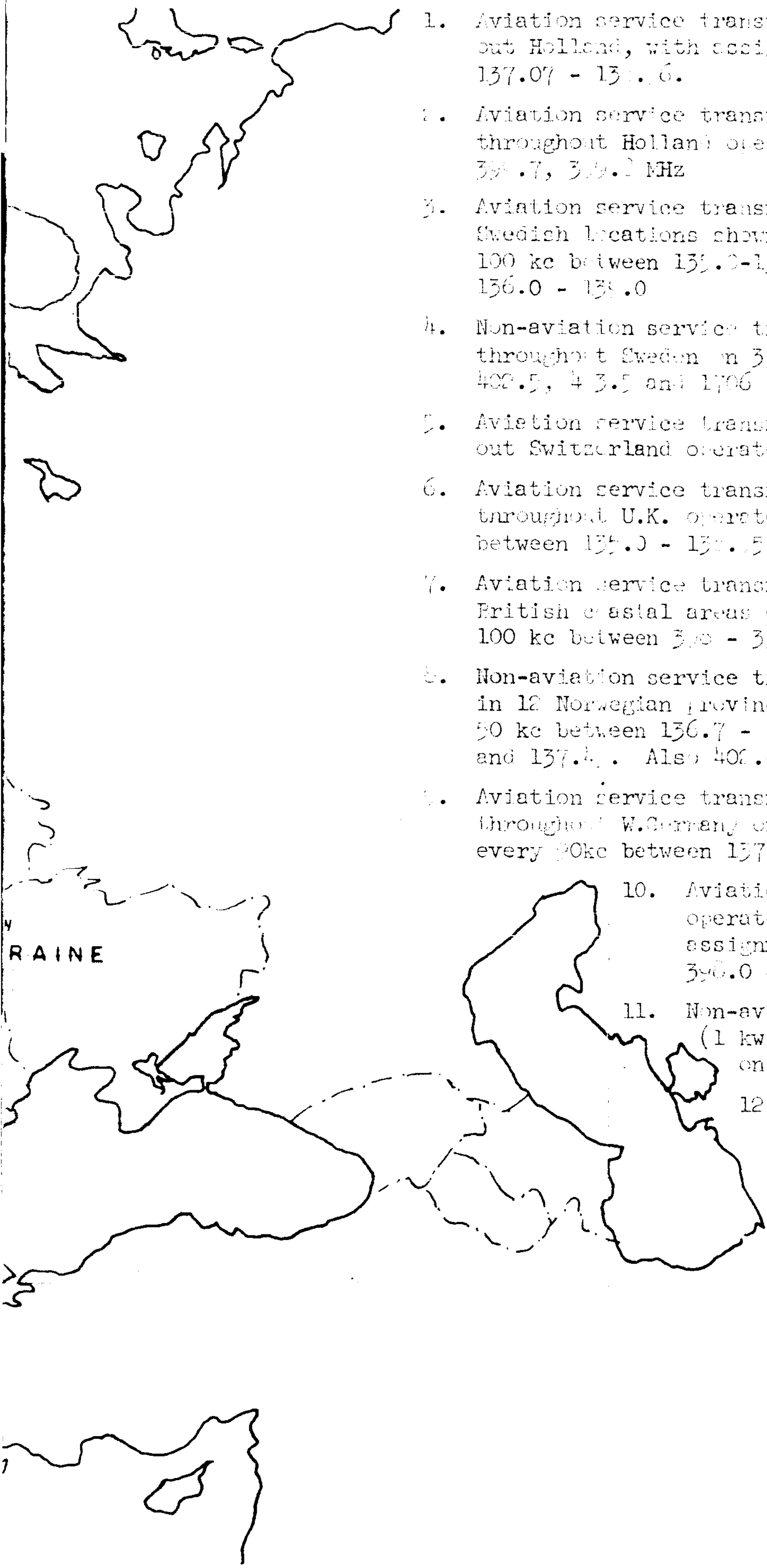
FIGURE 7-3

DISTRIBUTION OF FIXED TRANSMITTERS IN EUROPE

KEY

+ AERONAUTICAL SERVICE TRANSMITTER
o NON-AERO SERVICE TRANSMITTER
□ STADAN SITE

FOLDOUT FRAME



1. Aviation service transmitters(50w) located throughout Holland, with assignments every [illegible]0 kc between 137.07 - 13[illegible].[illegible]6.

2. Aviation service transmitters(100w) located throughout Holland operating on 3[illegible].7, 3[illegible].5, 399.2, 39[illegible].7, 3[illegible].[illegible] MHz

3. Aviation service transmitters(50w) operate at 18 Swedish locations shown, with assignments every 100 kc between 135.0-136.0 and every 50 kHz between 136.0 - 13[illegible].0

4. Non-aviation service transmitters (1 kw) operate throughout Sweden on 3[illegible].5, 39[illegible].5, 400.5, 401.5, 402.5, 4[illegible]3.5 and 1706 (50w)

5. Aviation service transmitters (20w) located throughout Switzerland operate on 135.1 MHz

6. Aviation service transmitters (100w) located throughout U.K. operate on assignments every 90 kc between 135.0 - 13[illegible].5

7. Aviation service transmitters (200w) located in British coastal areas operate on assignments every 100 kc between 3[illegible] - 399.9

8. Non-aviation service transmitters (15-20w) operate in 12 Norwegian provinces on assignments every 50 kc between 136.7 - 137.1 and on 137.2, 137.4 and 137.[illegible]. Also 40[illegible].0

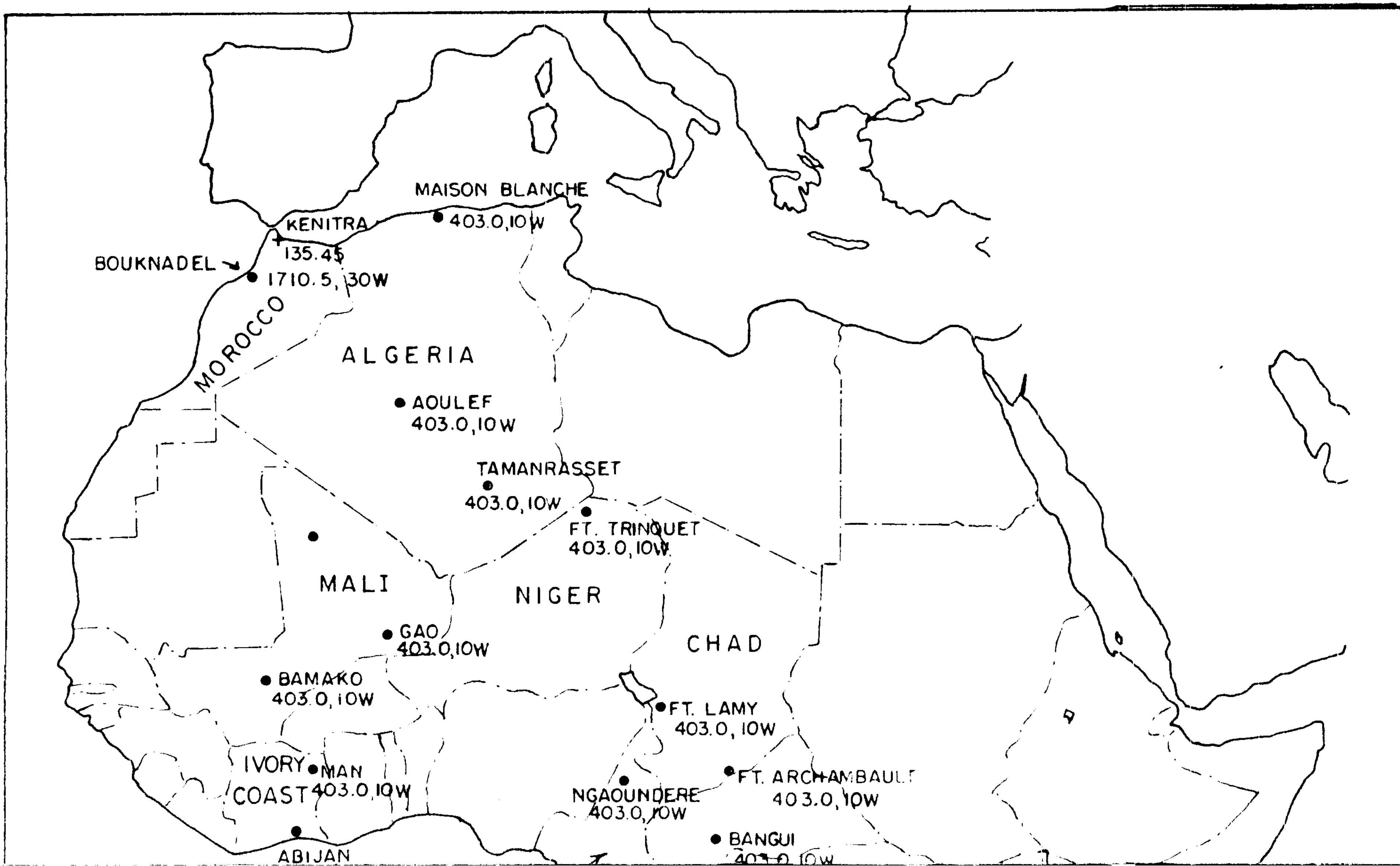
9. Aviation service transmitters (100w) operate throughout W.Germany on 135.[illegible] and on assignments every [illegible]0kc between 137.07 - 138.96

10. Aviation service transmitters(200w) operate throughout W.Germany on assignments every 100kc between 390.0 - 399.9

11. Non-aviation service transmitters (1 kw) operate throughout W.Germany on 1705.0 Mcs

12. Non-aviation service transmitters (5w) operate throughout Czechoslavakia on 1690.0 and 1702.5

FOLDOUT FRAME



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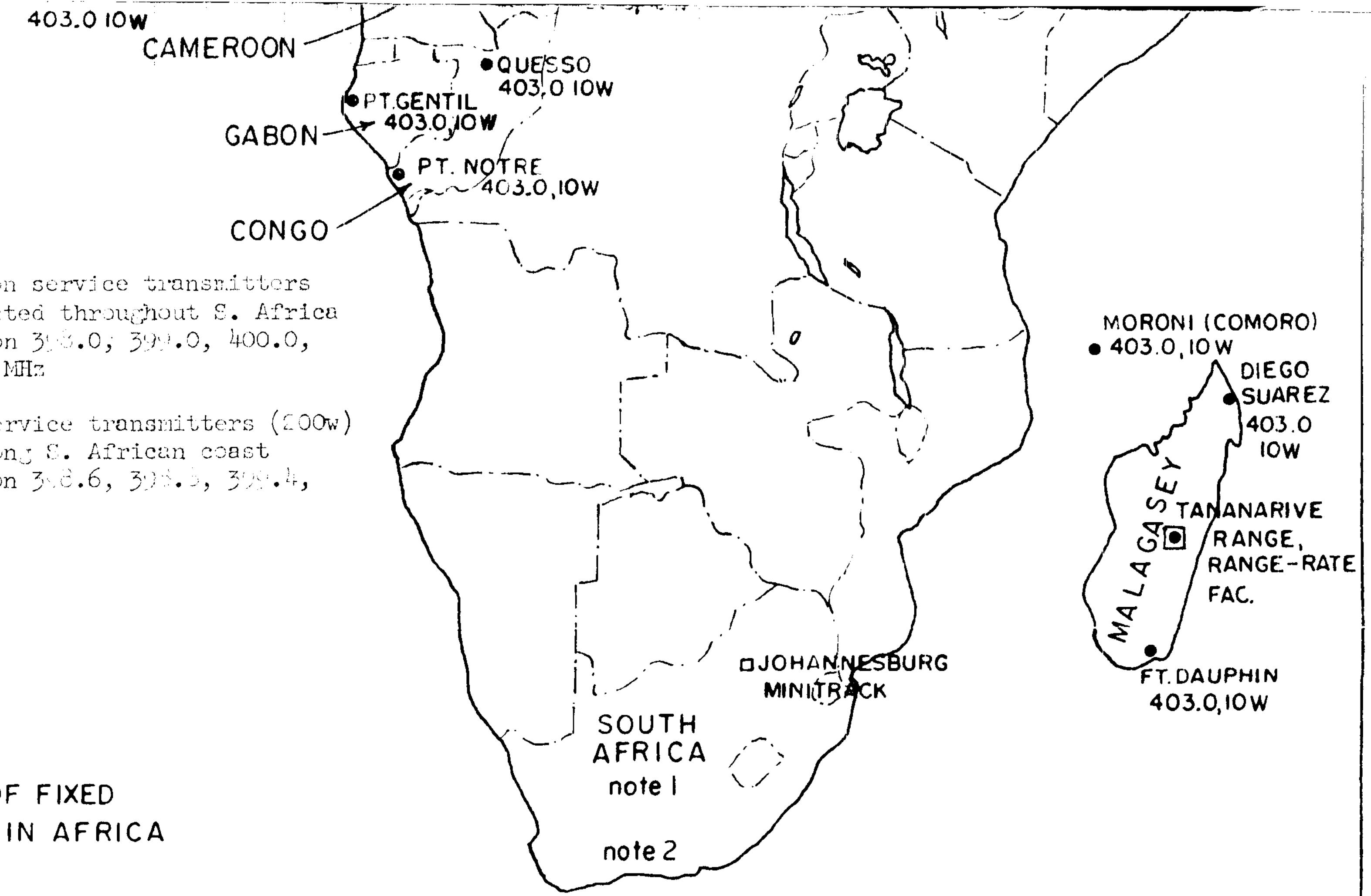


Note 1. Non-aviation service transmitters (200w) located throughout S. Africa operating on 3[illegible].0, 39[illegible].0, 400.0, 401.0, 402 MHz

Note 2. Aviation service transmitters (200w) located along S. African coast operating on 3[illegible].6, 39[illegible], 39[illegible].4, 39[illegible]. MHz

FIGURE 7-4
DISTRIBUTION OF FIXED TRANSMITTERS IN AFRICA

KEY

- + AERONAUTICAL SERVICE TRANSMITTER
- ● NON-AERONAUTICAL SERVICE TRANSMITTER
- □ STADAN SITE

FOLDOUT FRAME

137-2

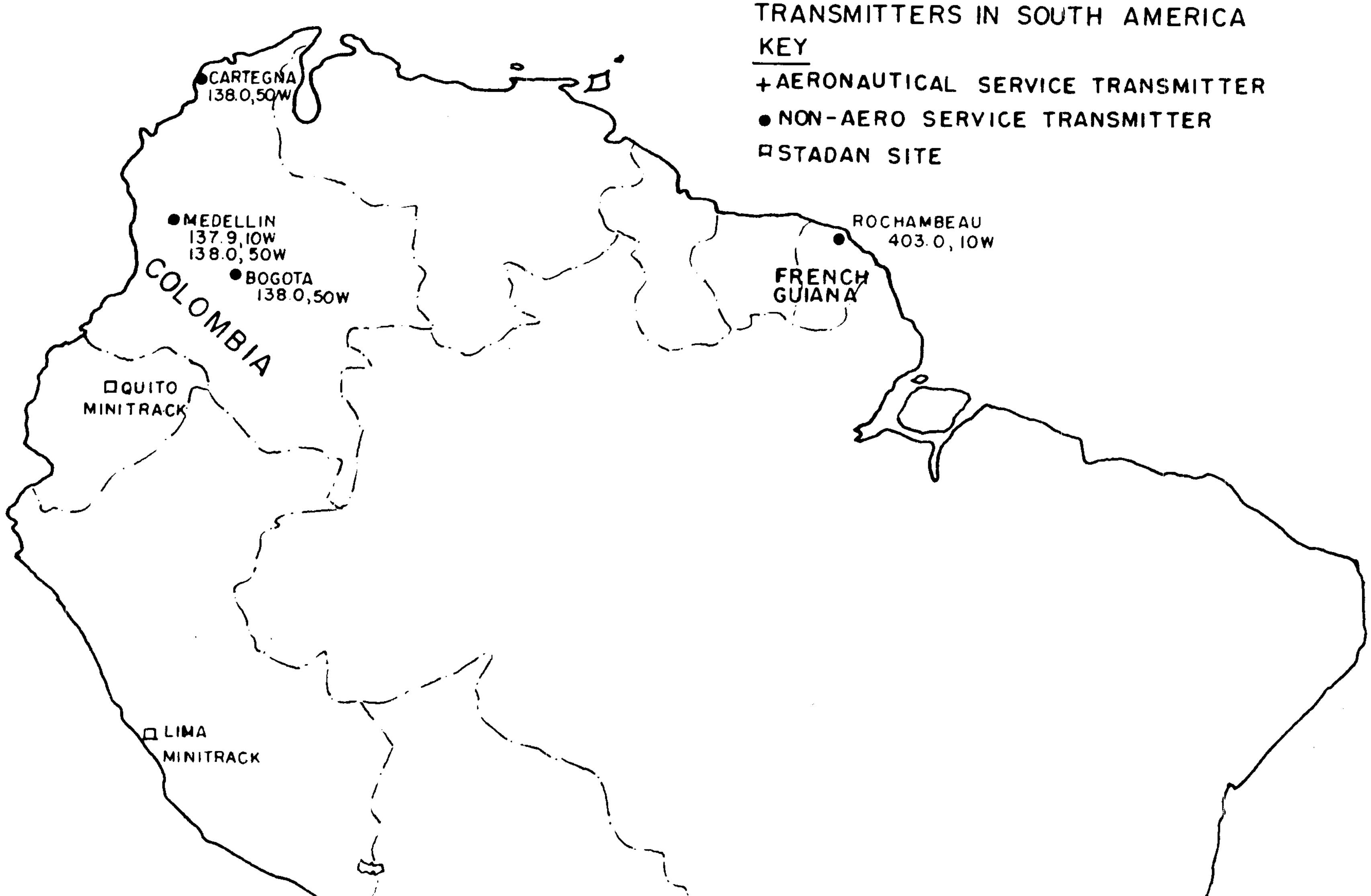
FIGURE 7-5 DISTRIBUTION OF FIXED TRANSMITTERS IN SOUTH AMERICA

KEY

+ AERONAUTICAL SERVICE TRANSMITTER

● NON-AERO SERVICE TRANSMITTER

□ STADAN SITE



FOLDOUT FRAME

138-1

FOLDOUT FRAME

1. Aviation service transmitters (10w) operate at locations shown on some or all of the following frequencies: 136.11, 136.23, 136.71 137.19, 137.31, 137.43, 137.55, 137.67

2. Non-aviation service transmitters (60w or less) operate in Buenos Aires on 136.0, 136.34, 136.47, 136.59, 136.[illegible]3, 136.[illegible]5, 137.07 and 399.25 and 1705.0

138-2

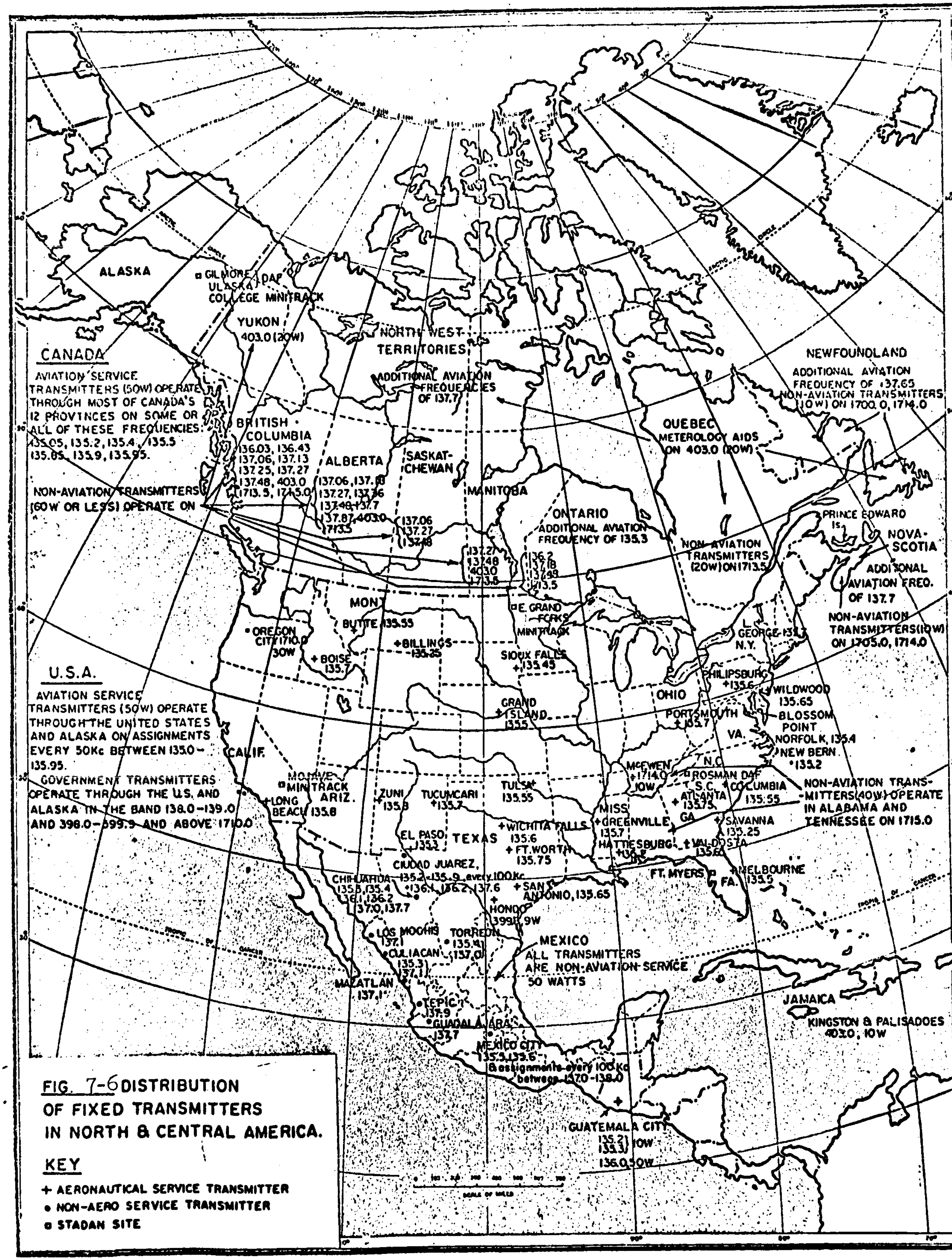


FIG. 7-6 DISTRIBUTION OF FIXED TRANSMITTERS IN NORTH & CENTRAL AMERICA.

KEY

+ AERONAUTICAL SERVICE TRANSMITTER
• NON-AERO SERVICE TRANSMITTER
▫ STADAN SITE

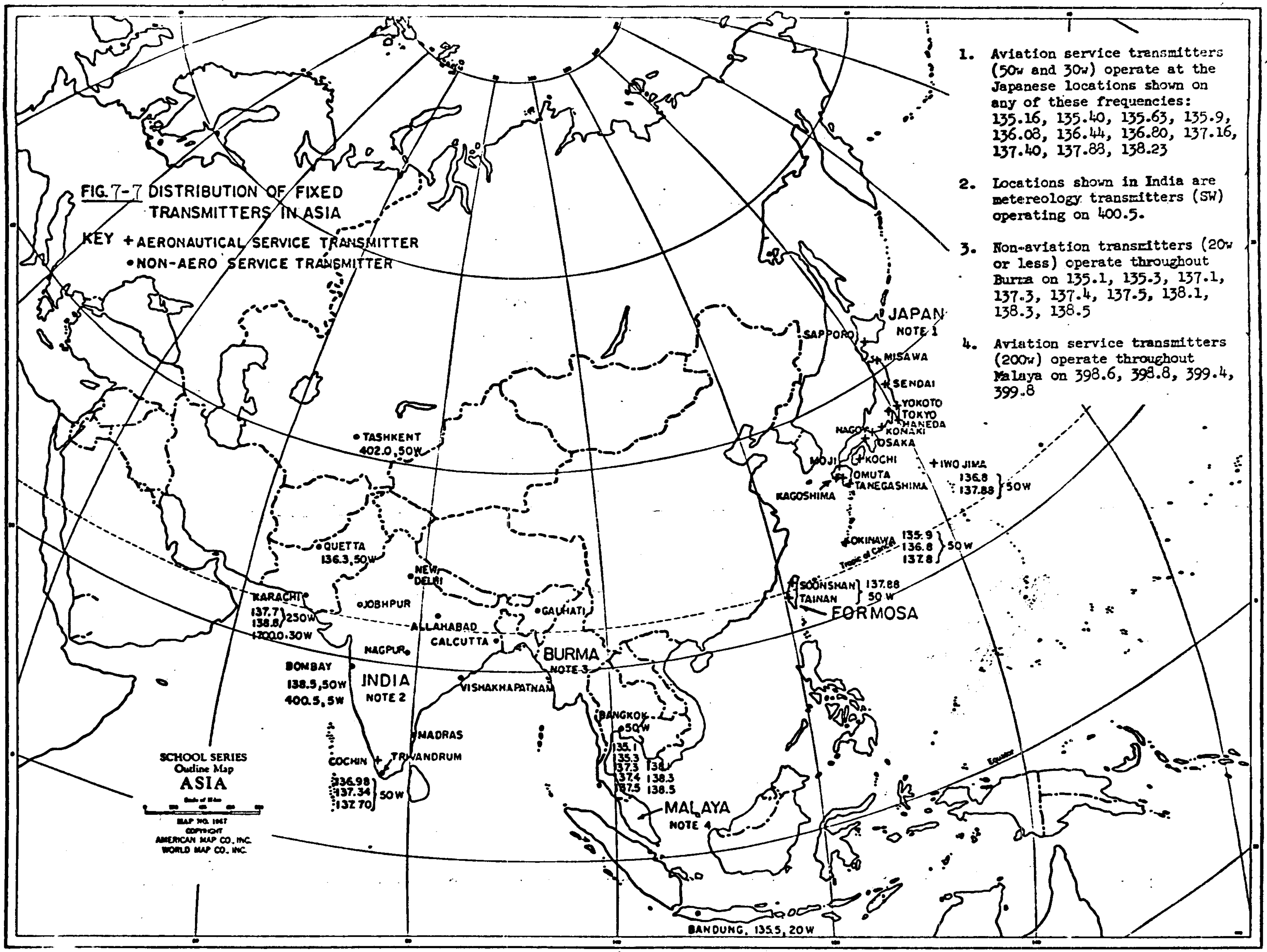


FIG. 7-7 DISTRIBUTION OF FIXED TRANSMITTERS IN ASIA
KEY + AERONAUTICAL SERVICE TRANSMITTER
• NON-AERO SERVICE TRANSMITTER
1. Aviation service transmitters (50w and 30w) operate at the Japanese locations shown on any of these frequencies: 135.16, 135.40, 135.63, 135.9, 136.08, 136.44, 136.80, 137.16, 137.40, 137.88, 138.23
2. Locations shown in India are metereology transmitters (SW) operating on 400.5.
3. Non-aviation transmitters (20w or less) operate throughout Burma on 135.1, 135.3, 137.1, 137.3, 137.4, 137.5, 138.1, 138.3, 138.5
4. Aviation service transmitters (200w) operate throughout Malaya on 398.6, 398.8, 399.4, 399.8
JAPAN
NOTE 1
SAPPORO
MISAWA
SENDAI
YOKOTO
TOKYO
HANEDA
KONAKI
OSAKA
KOCHI
MOJI
OMUTA
TANEGASHIMA
KAGOSHIMA
IWO JIMA
136.8
137.88
50W
OKINAWA
135.9
136.8
137.8
50W
SOONSHAN
TAINAN
137.88
50 W
FORMOSA
TASHKENT
402.0, 50W
QUETTA
136.3, 50W
NEW DELHI
JOBHPUR
ALLAHABAD
CALCUTTA
GAUHATI
KARACHI
137.7
138.8
1200.0
250W
30W
NAGPUR
BOMBAY
138.5, 50W
400.5, 5W
INDIA
NOTE 2
VISHAKHAPATNAM
MADRAS
TRIVANDRUM
COCHIN
136.98
137.34
137.70
50W
BURMA
NOTE 3
BANGKOK
50W
135.1
135.3
137.3
137.4
137.5
138.3
138.5
MALAYA
NOTE 4
BANDUNG, 135.5, 20W
SCHOOL SERIES
Outline Map
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MAP NO. 1047
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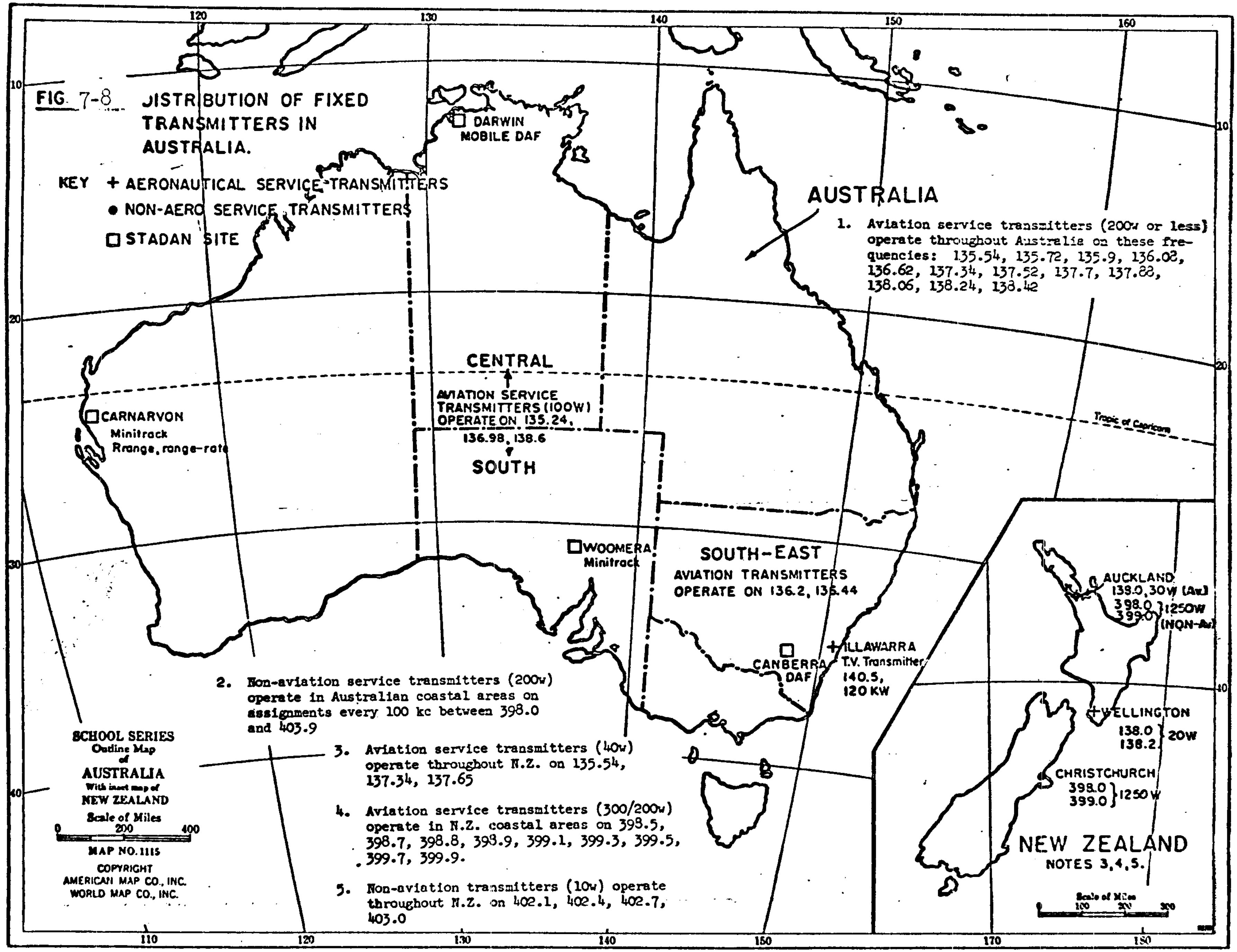


FIG. 7-8 DISTRIBUTION OF FIXED TRANSMITTERS IN AUSTRALIA.

in general, as far away from the aeronautical ground stations as possible. The proximity of a site to the other fixed transmitters will depend on the power being radiated and on whether the transmitter is operating on a frequency assignment within one of the three space research bands or on an adjacent frequency assignment. It will also depend on the type of terrain which exists in the region where the transmitter is located.

The data obtained from the Frequency Lists only provides an approximate picture of the overall RF environment as it exists throughout the world today. More detailed information on any particular region of the world can be obtained by contacting the telecommunications administration responsible for frequency assignments within the country of interest. Within the United States, the agencies responsible for telecommunication matters are the Federal Communications Commission (FCC) and the Interdepartmental Radio Advisory Committee (IRAC).

7.3.3.3 Airborne Transmitters in the United States

Of the many potential sources of interference to station operations considered so far (natural and man-made noise, fixed transmitters, mobile transmitters, etc.) the most serious threat comes from airborne transmitters.

The lack of sufficient path attenuation between the airborne transmitter and the site represents the chief cause of this interference problem, owing to the consequent high power levels arriving at the site. The expression for the power received at a 136 MHz telemetry receiver input is given by

$$P_r = \frac{P\lambda^2}{8\pi^2\ 1600^2} \cdot \frac{G}{R^2} \text{ watts} \qquad (7.3\text{-}1)$$

where P = the output of the airborne transmitter in watts

λ = the wavelength in meters

G = receiving antenna power gain (above isotropic)

R = distance in miles between transmitter and site.

This expression has been plotted as a function of distance in Fig. 7-9 for an aircraft transmitter operating in the range of 135.0 MHz to 135.95 MHz with a 25-watt output, and an antenna gain of 3 db (hemispheric coverage). Curves for three different values of G, the receiving antenna gain, corresponding to three of the different types of receiving antennas used at STADAN sites, are shown.

It can be seen from Fig. 7-9 that the power levels received from aircraft transmitters are prohibitively high. The signal level received from an aircraft 100 miles away may be as much as 60 db greater than the corresponding level being received from a satellite, so that the aircraft signal will, very likely, completely destroy the data arriving from the satellite. Even if this aircraft is flying in a region outside the main lobe of the STADAN receiving antenna, there will still be sufficient unwanted energy entering through the antenna side lobes or back lobes to exceed, by 30-40 db, the level of the satellite signal.

Whether or not interference is actually created depends on a number of factors, including the general air traffic density in the area, the height at which these aircraft fly, and the amount of time these aircraft are in communication with the ground. From Fig. 7-9 it is evident that an aircraft flying at 40,000 ft. will be "visible" to the ground station from as far away as 250 miles, assuming smooth earth conditions. This distance obviously depends also on the height of the

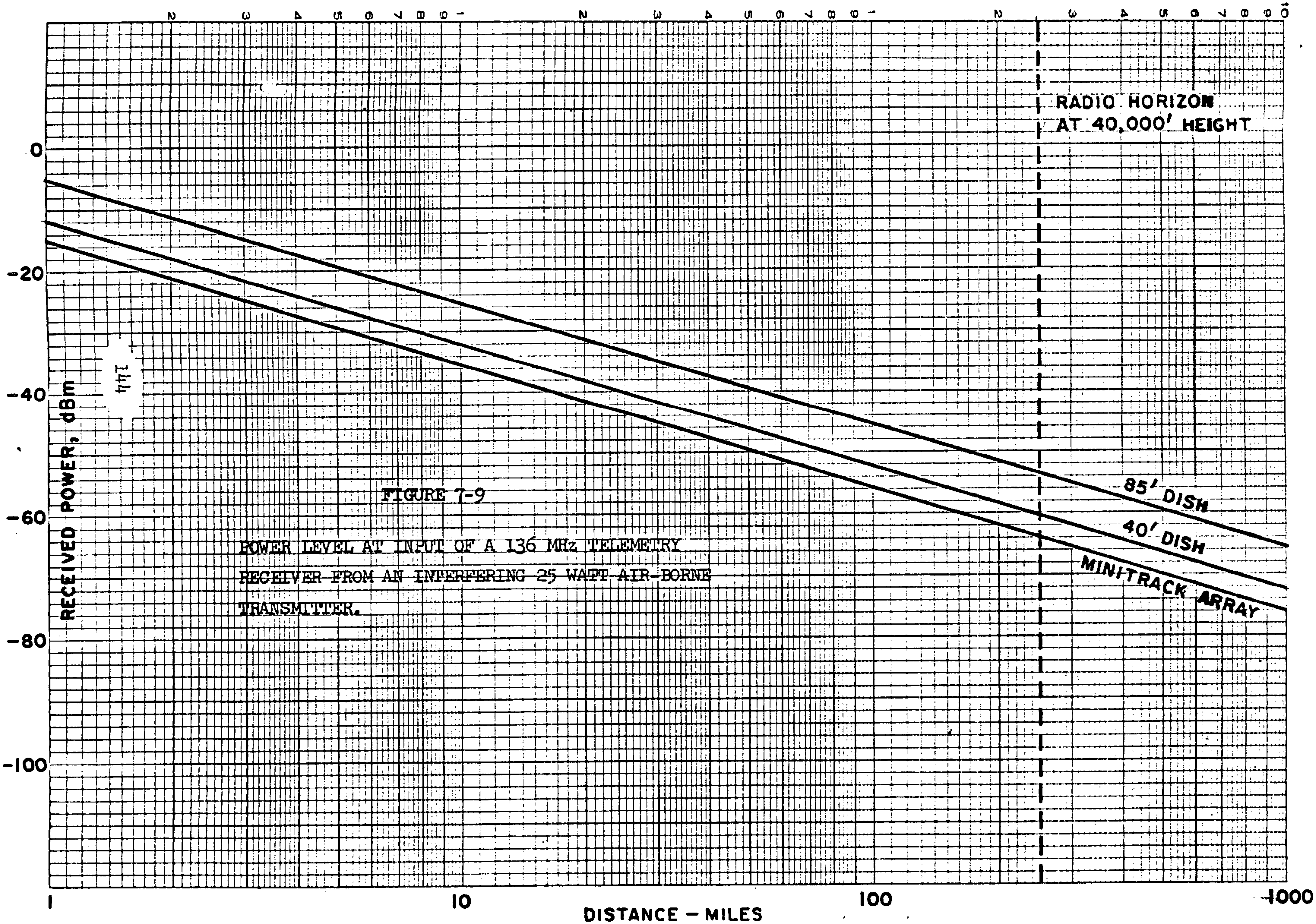


FIGURE 7-9

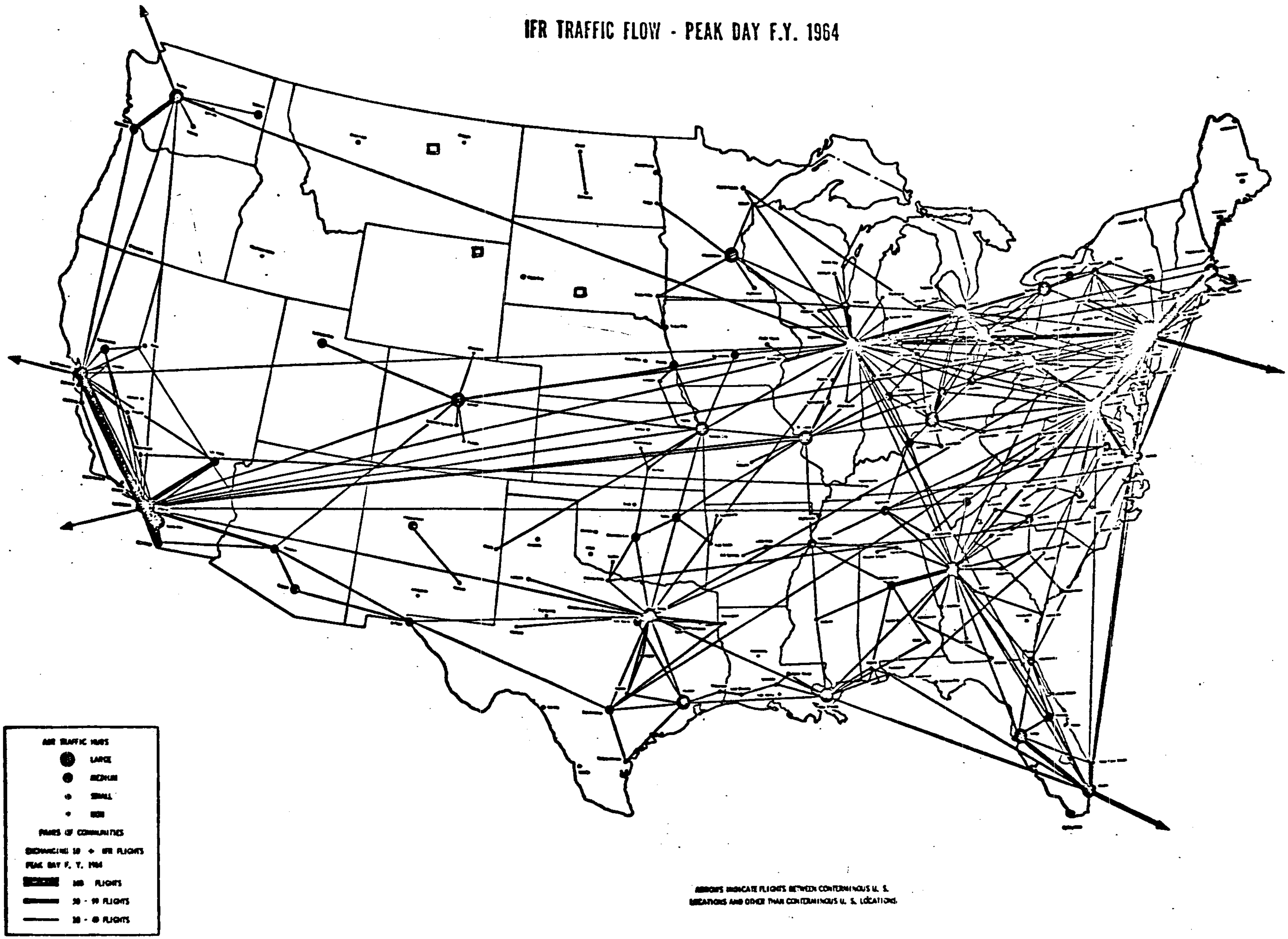
POWER LEVEL AT INPUT OF A 136 MHz TELEMETRY RECEIVER FROM AN INTERFERING 25 WATT AIR-BORNE TRANSMITTER.

STADAN receiving antennas above ground; an aircraft will be "visible" to the 85 ft. parabolic dish antenna, from much farther away than it will be to the Minitrack array, since the dish antenna is mounted about 100 ft. above the ground. Consequently, the higher an aircraft is flying, the greater is the probability of interference. Also, we see that the Data Acquisition Facilities, which employ the directional dish antennas, are more susceptible to aircraft interference than are the Minitrack stations. The Rosman, N.C., and Blossom Point, Md. STADAN sites have experienced more aircraft interference since they are located in regions which contain a very high air traffic density. This situation is well illustrated by Fig. 7-10 (Reference 55) which represents a map of the air traffic density which existed in the peak or busiest day of the fiscal year 1964. This map gives the traffic flow for aircraft which were flying under "Instrument Flight Rules" (IFR). A considerable number of aircraft (mostly private and business planes) are usually also in the air at any time, operating under "Visual Flight Rules" (VFR). These apparently were not included in the survey for which Fig. 7-10 applies.

7.3.4 VHF Propagation Characteristics

At the frequencies presently used in the STADAN system, the ground or surface wave which predominates at the lower frequencies is rapidly attenuated, so that propagation is restricted to the "space-wave" consisting of a direct wave and a ground reflected wave. For relatively short distances between the transmitting and receiving antennas (20 miles or less), "line-of-sight" conditions are said to exist and, to a good approximation, the earth may be considered to be flat. In this region, the direct wave and reflected wave create interference effects, so that

IFR TRAFFIC FLOW - PEAK DAY F.Y. 1964



□ POSSIBLE LOCATION FOR NEW STADAN SITES

FIG 7-10 AIR TRAFFIC FLOW WITHIN THE UNITED STATES.

the resulting change of field strength with distance contains many deep nulls, assuming a perfectly smooth reflecting earth (see Fig. 7-11). In practice, the earth has a rough surface and is not a perfect conductor. Consequently, the ground reflected wave is both attenuated and altered in phase relative to the wave reflected from the smooth earth. The curve of Fig. 7-11 (Reference 59) will therefore be modified to that shown by the dotted curve for actual rough earth conditions. The form of both curves in Fig. 7-11 is dependent on the height of the transmitting and receiving antennas and on the frequency used; increasing these parameters will greatly increase the number of nulls.

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

Diffraction of a wave by a rough surface is difficult to evaluate theoretically since it depends very much on the type of terrain

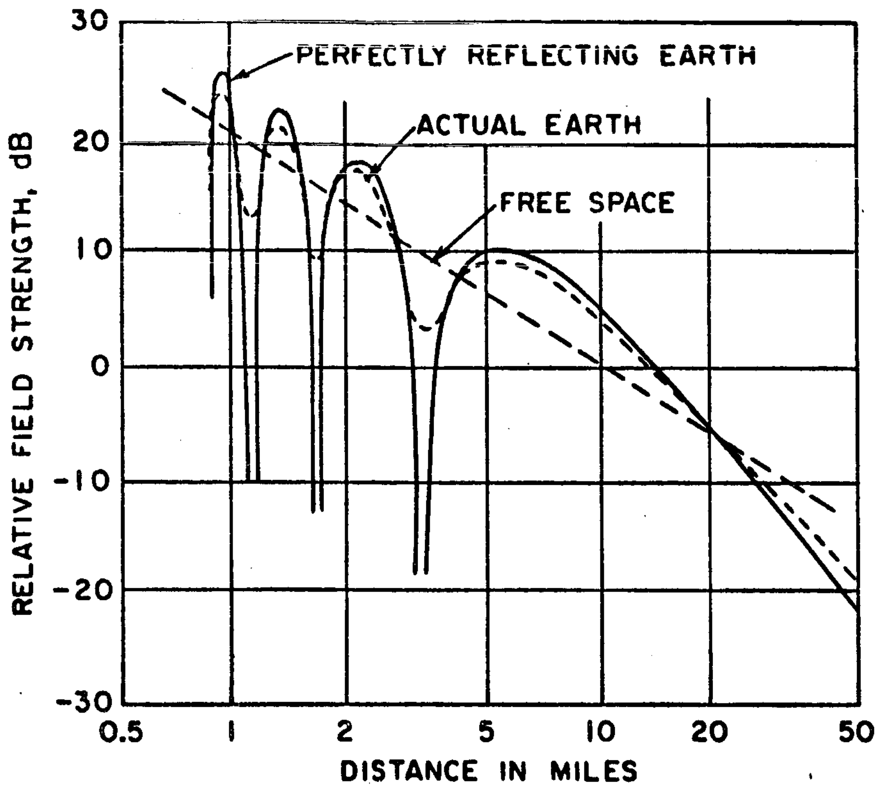


FIGURE 7-11 FIELD STRENGTH AS A FUNCTION OF DISTANCE, FLAT EARTH

Variation of field strength as a function of distance for hypothetical but nevertheless typical conditions of space-wave propagation. These curves assume a flat earth and horizontally polarized waves.

being considered, including its roughness and its conductivity. Some work has been done on knife edge diffraction and the results of this are occasionally applicable in the case where a single large mountain or hill happens to be located between transmitting and receiving antenna. However, this is not commonly found in practice.

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

Figure 7-12 (Reference 60) shows the change of both theoretical and measured field intensities with distance at a frequency of about 50 MHz. The curve for the measured data represents an average curve, since in practice, the experimental data varies over wide limits. It can be seen that the measured curve falls roughly in between the smooth earth curve and that predicted for knife edge diffraction. The experimental curve probably falls close to that which can be predicted by rough surface diffraction theory.

7.4 APPENDIX

PARTIAL REVISION OF THE RADIO REGULATIONS, GENEVA, 1959, ADOPTED BY THE EXTRAORDINARY ADMINISTRATIVE RADIO CONFERENCE CONVENED IN GENEVA, Oct. 63

Notes

1. The data in this appendix applies only to the frequency bands which have been allocated to the space research service. The adjacent band

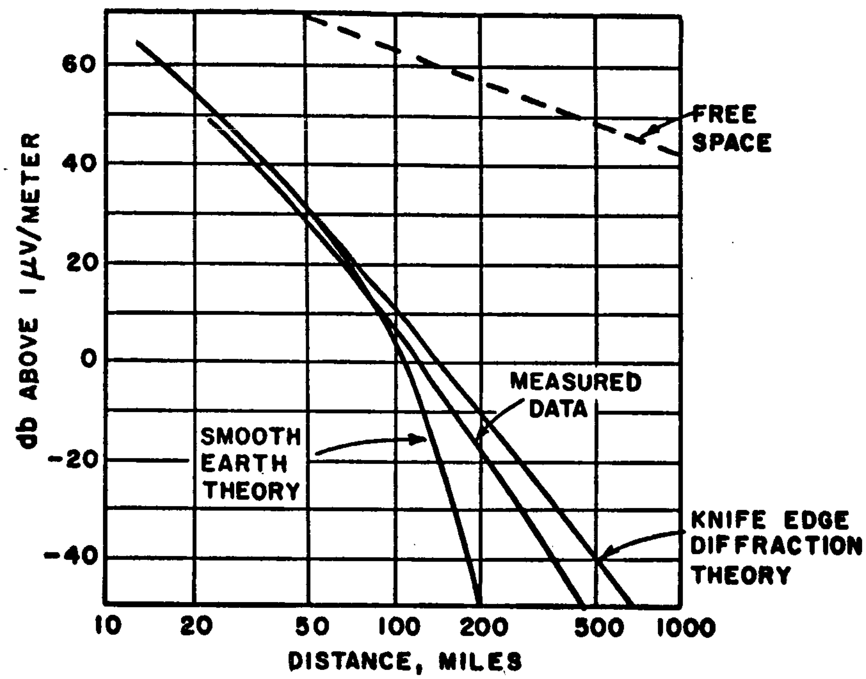


FIGURE 7-12

MEASURED AND THEORETICAL FIELD INTENSITIES AT 50 MHz
ASSUMING A RADIATED POWER OF 1 kW AND AN ANTENNA HEIGHT OF 500 FEET

allocations are also included.

2. The allocations made, differ according to the different parts of the world. For purposes of frequency allocations, the ITU has divided the world up into three different regions.

3. Services, the names of which are printed in capital letters (e.g., FIXED), represent "primary" services. If only one primary service is allocated to a particular frequency band, then that allocation is an <u>exclusive</u> one. If more than one primary service has been allocated to a particular band, then the band is <u>shared</u> by these services.

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

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

132-136 MHz

Region 1	Region 2	Region 3
Western Europe, Middle-East, North and East Africa: AERONAUTICAL MOBILE Western, Central, and Southern Africa: FIXED, MOBILE	FIXED, MOBILE (AERO. MOBILE is exclusive to the U.S.A.)	All countries (except Australia and New Zealand): (Eventually to be exclusively allocated to AERO. MOBILE) FIXED, MOBILE Australia and New Zealand: AERO.MOBILE

The application of space communication techniques for use in the Aero.Mobile service is permitted in this band. This is to be limited to satellite relay stations belonging to the Aero.Mobile service.

136-137 MHz

Region 1	Region 2	Region 3
FIXED, MOBILE, SPACE RESEARCH (Telemetry and Tracking)	SPACE RESEARCH This band also allocated to FIXED, MOBILE until 1 Jan. 1969 (indefinite in Cuba)	FIXED, MOBILE, SPACE RESEARCH

137-138 MHz

SPACE RESEARCH (Tracking and Telemetry)
SPACE, OPERATIONAL (Tracking and Telemetry)
METEOROLOGICAL-SATELLITE

Norway, Switzerland, Turkey: This band also allocated to Fixed, Mobile (excl. Aviation) until 1 Jan, 1969. U.S.S.R. Bloc, Algeria, Morrocco, U.A.R., Lebanon: This band also allocated to Aero Mobile indefinitely. All other countries in Western Europe, Middle East, North and East Africa: This band also	This band also allocated to FIXED, MOBILE until 1 Jan. 1969 (indefinite in Cuba).	This band also allocated to FIXED, MOBILE until 1 Jan. 1969 (indefinite in Malayasia, Pakistan, and Phillippines) Australia: This band also allocated to BROADCASTING (TV)

allocated to Aero Mobile service until 1 Jan 1969. Central and Western Africa: This band also allocated to FIXED, MOBILE indefinitely.		

138-143.6 MHz

Region 1	Region 2	Region 3
Austria, Denmark, Greece, Netherlands, Norway, Portugal, Sweden, Switzerland, Turkey, U.K., and Western Germany: This band will be allocated in the future to the Fixed and Mobile (excluding Aeronautical) service. Western, Central, and South Africa: FIXED, MOBILE AERONAUTICAL MOBILE	FIXED, MOBILE Permitted Service: Radiolocation (In the U.S.A., this band is reserved for government use)	All countries (except New Zealand) FIXED, MOBILE New Zealand: AERO. MOBILE Australia: This band also allocated to TV Broadcasting Taiwan: This band also allocated to Radiolocation

335.4-399.9 MHz

Region 1	Region 2	Region 3
	FIXED, MOBILE (Reserved for government use in the U.S.A.)	

399.9-400.05MHz

RADIONAVIGATION-SATELLITE

Stations operating in the Fixed and Mobile Services may continue to operate in this band until 1 Jan. 1969, except for the following countries where the band is shared with the Fixed and Mobile Service indefinitely.

Bulgaria, Greece, Hungary, Lebanon, Morrocco, U.A.R., Yugoslavia.	Cuba	Iran, Kuwait

400.05-401 MHz

Region 1	Region 2	Region 3
METEOROLOGICAL AIDS METEOROLOGICAL-SATELLITE(Maintenance Telemetering) SPACE RESEARCH(Telemetering and Tracking		
U.S.S.R. bloc, Greece, U.A.R.: This band also allocated indefinitely to Fixed and Mobile Service U.K.: This band also allocated on a secondary basis to Radiolocation		

401-402 MHz

Region 1	Region 2	Region 3
METEOROLOGICAL AIDS SPACE (Telemetering and Tracking) Secondary Services: Fixed Mobile (excluding Aviation)		
U.S.S.R. bloc, Greece, Norway, Sweden, Switzerland, Turkey: This		Iran: This band also allocated on a primary basis to Fixed and Mobile

band also allocated on a primary basis to Fixed and Mobile (excl. Aviation) service. France: Meteorological Aids is the only primary allocation. U.K.: Secondary allocation to Radiolocation		(excl. Aviation) service. Australia: The Space Service operates in this band on a secondary basis only.

402-406 MHz

METEOROLOGICAL AIDS

Secondary Services: Fixed
Mobile (excluding Aviation)

All of the exceptions listed in the band 401-402 MHz also apply in this band (omit Australia).

1690-1700 MHz

Region 1	Region 2	Region 3
METEOROLOGICAL AIDS METEOROLOGICAL-SATELLITE		
Secondary Service: Fixed Mobile (excl. Aviation) U.S.S.R. bloc, Algeria, Lebanon, Morrocco, U.A.R.: This band is also allocated on a primary basis to the Fixed and Mobile (excl. Aviation) service. Austria, Finland: Meteorological-Aids is the only primary allocation.	Cuba: Primary allocation also to Fixed and Mobile (excl. Aviation) service.	Pakistan, Kuwait: Primary allocation also to Fixed and Mobile (excl. Aviation) service. Australia, Indonesia, New Zealand: Secondary allocation to Fixed and Mobile (excl. Aviation) service.

1700-1710 MHz

Region 1	Region 2	Region 3
FIXED SPACE RESEARCH (Telemetering and Tracking) Secondary Service: Mobile	SPACE RESEARCH (Telemetering and Tracking) Cuba: Primary allocation also to Fixed and Mobile Service.	FIXED MOBILE SPACE RESEARCH (Telemetering and Tracking)

1710-1770 MHz

Region 1	Region 2	Region 3
FIXED Secondary Service: Mobile Switzerland: Primary allocation to both Fixed and Mobile (excl. Aviation) services.	FIXED, MOBILE (Allocated for government use in the U.S.A.)	FIXED MOBILE

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