CHAPTER 10

MICROWAVE RADIO EQUIPMENT

It is obvious from the preceding discussions that there is a very close interrelationship between the characteristics of the various items of the equipment to be used, and the engineering choices and performance parameters of the paths themselves.

Thus it is desirable, in fact almost essential, that the path survey engineers have enough advance knowledge of the frequency bands to be considered (often only one, but in some cases more than one), the kind of service, the number of channels (both present and future) to be accommodated by the system, the kind of performance and reliability criteria desired, and the pertinent parameters of the microwave equipment to be used (for example, transmitter output power, receiver noise figure and bandwidth, per channel deviation, et cetera), to allow an intelligent approach to the problem of path engineering. Many choices are involved in path selection, and choices made without a thorough knowledge of all the pertinent circumstances may not be the best ones.

Microwave systems can range from as little as 5 or 10 miles to distances as long as 4000 miles. Facility requirements can be relatively small, requiring structures and equipment for only a light route, or they may be very heavy, requiring multi-channel, heavy route layout with sophisticated switching. They can be constructed for nominally good service during certain limited hours of the day with considerable economy, or they can be built for a very high quality of service on a 24 hour a day, year-in and year-out basis.

Some systems are of a "through" type, with all or almost all of the channels going end-to-end, while others require multiple access, with dropping and inserting of channels at most, if not all, repeater points.

The two types of FM microwave equipment in common use are the IF heterodyne type and the baseband (remodulating) type. The IF heterodyne type, by eliminating demodulation and remodulation steps at repeaters, contributes the least amount of distortion, and is the preferred choice for systems handling exclusively, or almost exclusively, long-haul traffic, with little or no requirement for drop and insert along the route. The heterodyne type is also preferable for systems carrying color TV, if more than a few hops are involved. Equipment of the baseband type is widely used for short haul or for distributive systems. The great flexibility for drop and insert, plus maintenance advantages, are the determining factors. Heterodyne systems are inherently at a considerable disadvantage in such applications.

Apart from the choice between heterodyne or baseband equipments, primary considerations in the selection of the best radio equipment for a particular system include:

(a) characteristic of the end-to-end baseband facility (including bandwidth, frequency response, loading capability, noise figure, and noise performance); (b) the amount of radio gain available, as determined by transmitter power output and receiver noise characteristic; (c) operating frequency band, and required frequency spacing between radio channels, as determined by transmitter deviation, receiver selectivity and frequency stability; (d) primary power requirements and options available; (e) supervisory functions available, including order wire, alarms and controls; (f) equipment reliability, including availability of redundant versions such as frequency diversity, 1-for-N or 2-for-N multiline switching, hot standby, or hot standby at transmitters and space diversity at receivers; and (g) provisions for testing and maintenance.

With the rapidly changing nature of the state-of-the-art, and the continuing development of new equipments and upgrading of old ones, specific data on microwave equipment characteristics can become outdated in very short order. Consequently, the foregoing should be viewed in that light.

10.1 RECEIVER

The receiver consists basically of RF amplifiers, local oscillators, IF amplifiers, detectors, and video or audio amplifiers, along with gain and frequency control circuits. This chapter discusses the characteristics of terminal receivers used in microwave relay systems and the principles by which they operate.

At frequencies ranging from 600 - 13, 000 MHz a variety of differences in receiving conditions are found as compared to lower frequencies. The fluctuation noise of tubes and circuits in the receiver becomes greater than external noises, such as atmospheric disturbances and man-made interference. Therefore, receiver noise is one of the chief limitations of receiver sensitivity at such frequencies. The necessity of handling greater receiver bandwidths is also of great importance, both for determining the maximum rate at which information can be received and as a controlling factor of the total noise encountered. These greater bandwidths will increase the effective noise power originating in the resistors and tubes of the receiver as well as the received external noise.

10.1.1 Low Noise Preamplifier

There are few election tubes that will operate satisfactorily at 1 KMHz and above as RF amplifiers. Therefore, the use of RF pre-selection and a crystal mixer has been commonly used. The RF pre-selector is a tunable filter with low insertion loss to the operating frequency and high loss at undesired frequencies.

The limiting factor in the operation of RF amplifiers, especially at frequencies above 300 MHz, is the noise generated by the amplifier. Consequently, in certain applications (such as tropospheric systems) low noise devices such as parametric amplifiers have been used as RF amplifiers. Since the significant part of the noise generated in a cascaded operation is contributed by the first several stages, the use of such low noise devices is equivalent to improving receiver sensitivity thus permitting detection of extremely weak signals.

Probably the only significant parameters affecting RF-amplifier performance are noise figure and gain. Throughout this chapter all discussion relating to noise figure assumes that input noise is equal to KT. B and that this is the only noise contributed by the testing instrument or device - where the testing instrument output impedance is at room temerature (T_o = 290 K), K is Boltzman's constant (1.37×10^{-23}), and B is the bandwidth at 3 dB reference.

Although all parts of the receiver contribute noise to the output, the initial stages are the chief contributors. Unless the mixers and IF amplifiers have exceptionally high noise figures, the RF amplifier can be considered to be the determining factor. This is clearly indicated by the following expression for the total noise figure of cascade-connected multistage networks.

$$F_{1 \ldots n} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \ldots + \frac{F_n - 1}{G_1 G_2 \ldots G_{n-1}} \qquad (10\text{-}1)$$

where:

F_1 = noise factor of the first stage

F_2 = second stage, etc., and

G_1 = gain of first stage

G_2 = gain of second stage, etc.

Noise Figure = 10 log noise factor

Thus, it is evident from the expression, that after any high gain network the overall noise figure is not influenced greatly by additional networks even though their individual noise figures are relatively high. Consequently, for this measurement, it is sometimes assumed that the RF amplifier is the major contributor, and that the noise figure reflects the noise added by the amplifier alone. Such an approximation may not be valid especially where parametric amplifiers and masers are used, since their contributions to overall receiver noise are very small with respect to the mixer and the IF amplifier.

a. Parametric Amplifier. When the received signal is low, receiving system performance is most degraded by the addition of noise. Consequently, low noise preamplifiers are used for the first active electronic components in the receiving system. A low noise preamplifier chain may consist of a parametric amplifier and a tunnel diode amplifier. Figure 10-1 is the block diagram of a typical 4-stage parametric amplifier. The input is filtered to remove any transmit signals that have entered the receive channels. This filtered signal and a high frequency "pump" are applied to the amplifier stages (see figure 10-2). Three circulators and two loads are included to give stability to the amplifier by preventing reflections. Circulators are microwave devices in which signals that enter one port flow only to the adjacent clockwise port.

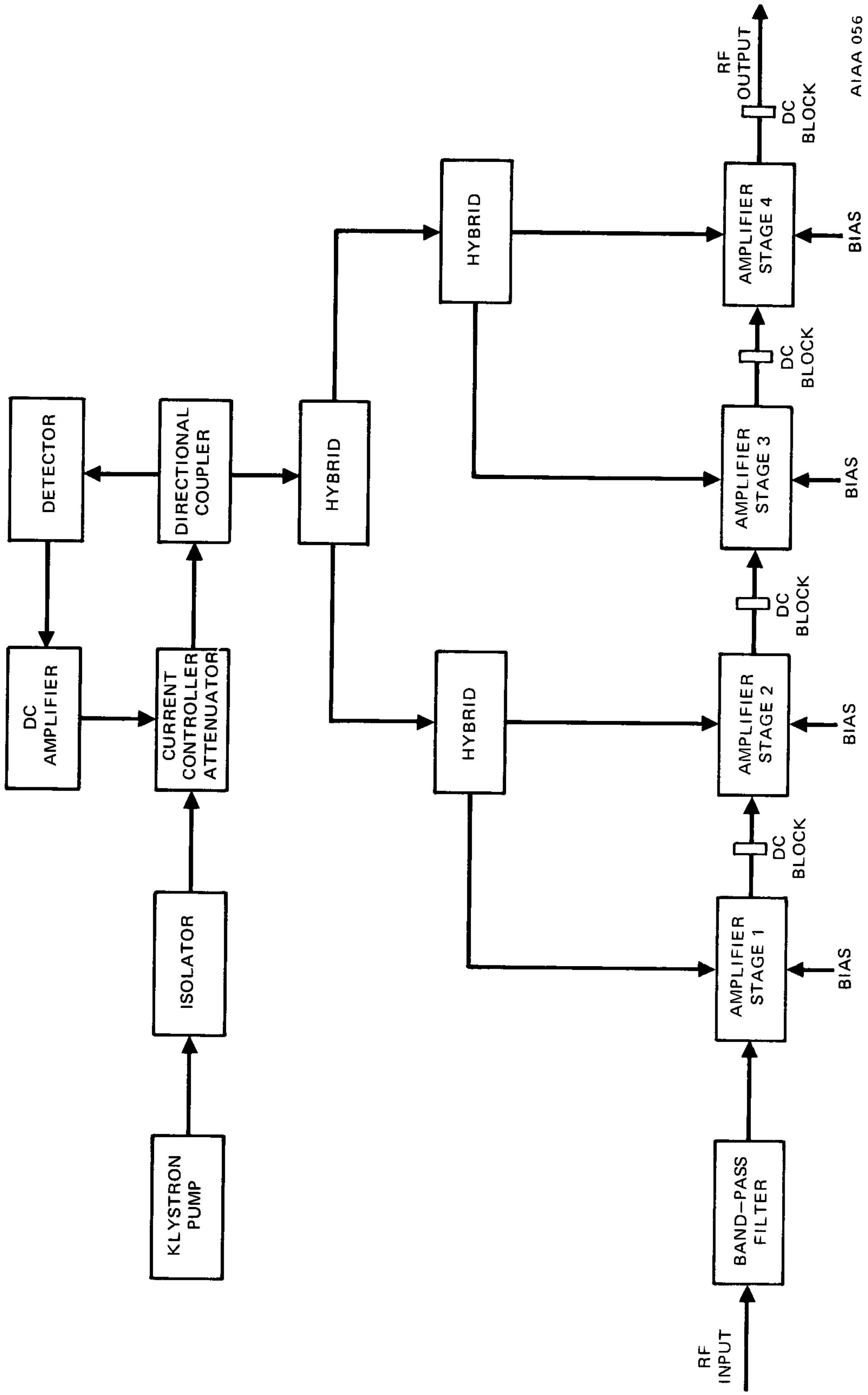

Figure 10-1. Parametric Amplifier, Block Diagram

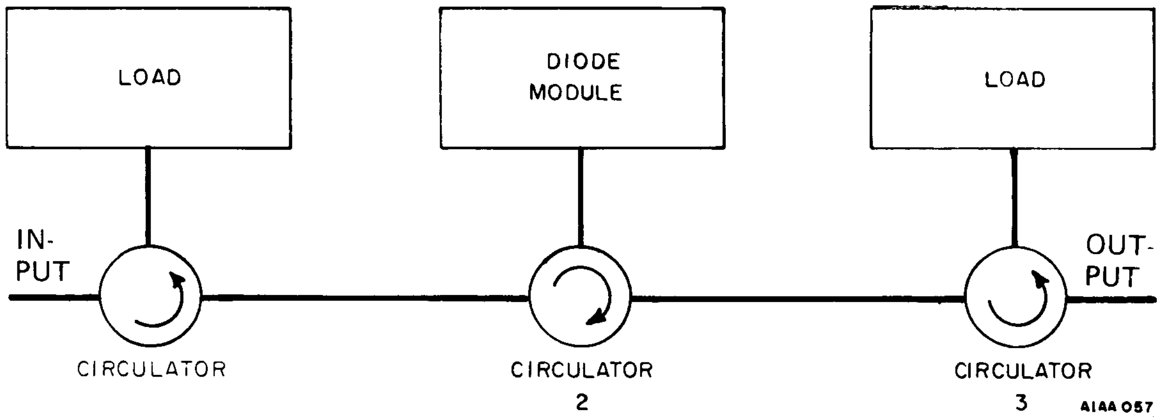

Figure 10-2. Amplifier - Stage Schematic

The device which amplifies is a varactor diode (see figure 10-3). The varactor diode is made part of two circuits, one resonant at the signal frequency and one resonant at the idler frequency (the difference between pump and signal frequencies). The junction capacitance of a varactor diode varies with the voltage across it. In a parametric amplifier, the capacitance is varied by application of the pump signal to the varactor diode. With the proper choice of pump level, pump frequency, and signal impedance matching, the diode presents a negative impedance to the incoming RF signal. This negative impedance causes the signal power reflected from the diode input impedance to be greater than the power incident on the impedance, thereby providing signal amplification.

The pump signal can be generated by a klystron oscillator or solid state source. A current-controlled attenuator in an automatic gain control (AGC) loop keeps the pump power to the amplification stages nearly constant despite changes in oscillator power. The pump power is sent through hybrids to all four amplifier stages. Each stage operates independently of the others. The frequency response of each stage is adjusted by setting the DC bias voltage to it. Blocking devices prevent the bias voltage from being sent along with the signal from one stage to the next. The amplifier stages are tuned separately and then together to give a combined gain of up to 30 dB across the receive bandwidth.

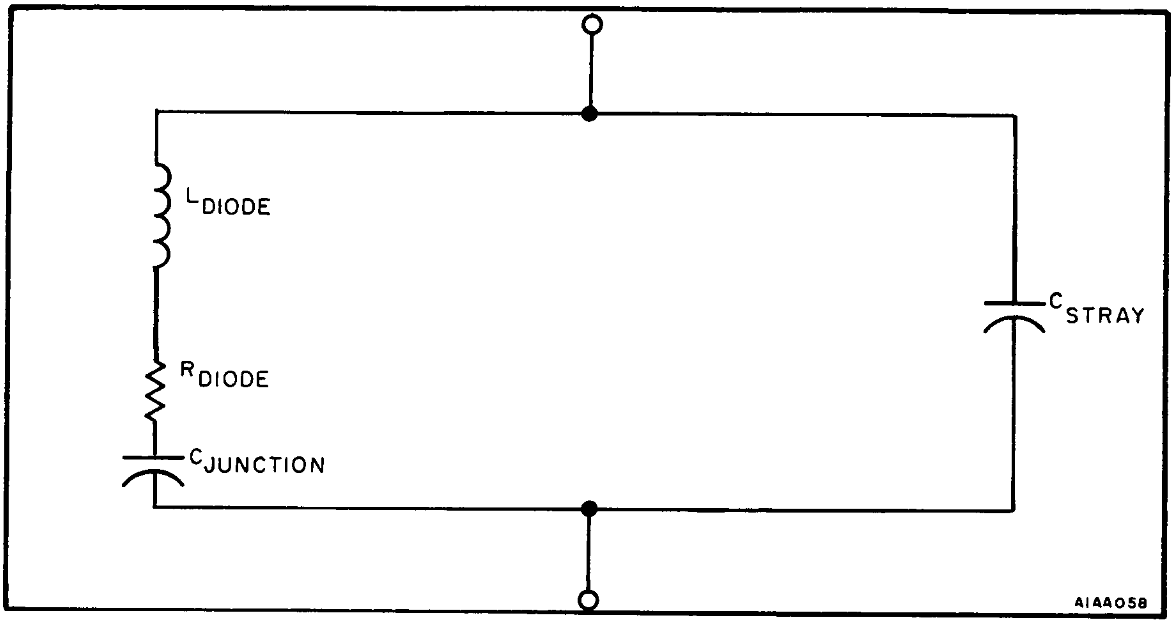

Figure 10-3. Varactor Diode Equivalent Circuit

To keep the parametric amplifier noise temperature as low as possible, some units are refrigerated. In very large systems, the physical temperature of the amplifying stages may be kept below 20K with a cryogenic refrigerating system consisting of helium compressor and expansion units. The expansion unit is insulated by a vacuum jacket. Metallic conductors transfer heat from the amplifier stages to the refrigerator. In smaller systems the more commonly used method is peltier cooling.

Electrical characteristics of a typical parametric amplifier are shown in table 10-1.

b. Tunnel Diode Amplifier. The parametric amplifier output is often further amplified by single-stage tunnel diode amplifiers. At microwave frequencies, a suitably-biased tunnel diode is basically a one-port device which can be used as a transmission-line termination. If the tunnel diode is used as a termination on a circulator port, the non-reciprocal properties of the circulator establish a one-way path along which signals may enter the circulator, undergo reflection at the tunnel diode port, and emerge from another circulator port (see figure 10-4). (The load shown on one port circulator is bypassed by the forward traveling signal, and is used only to absorb any reflected signals traveling in the reverse direction.)

Table 10-1. Typical Parametric Amplifier Characteristics

PARAMETER	CHARACTERISTICS
Gain	30 dB min
Instantaneous (0.5 dB) bandwidth	500 MHz min
Frequency band	3.7 to 4.2 GHz
Noise temperature without preselector filter	15 K max
Input or output SWR	1.3 max
Gain stability	$\pm$ 0.2 dB/minute $\pm$ 0.5 dB/12 hours $\pm$ 1.0 dB/week
Gain flatness	$\pm$ 0.2 dB/30 MHz
Amplitude response ripple	Less than 0.5 dB
Amplitude response slope	0.02 dB/MHz at any frequency in band
Dynamic range	Less than 1 dB compression for input signal from noise level up to -65 dBm
Spurious signals	60 dB below output level corresponding to -85 dBm input
Delay distortion	3 nanoseconds across any 200 MHz
	Slope of time delay shall not exceed 0.1 nanoseconds per MHz at any frequency across the band
Cool down time	About 4 hours
Phase Stability	$\pm$ 2°/24 hours
Intermodulation Products	More than 60 dB down for two carriers at -85 dBm each
Recovery from Incidental Cryogenic failures	30 minutes

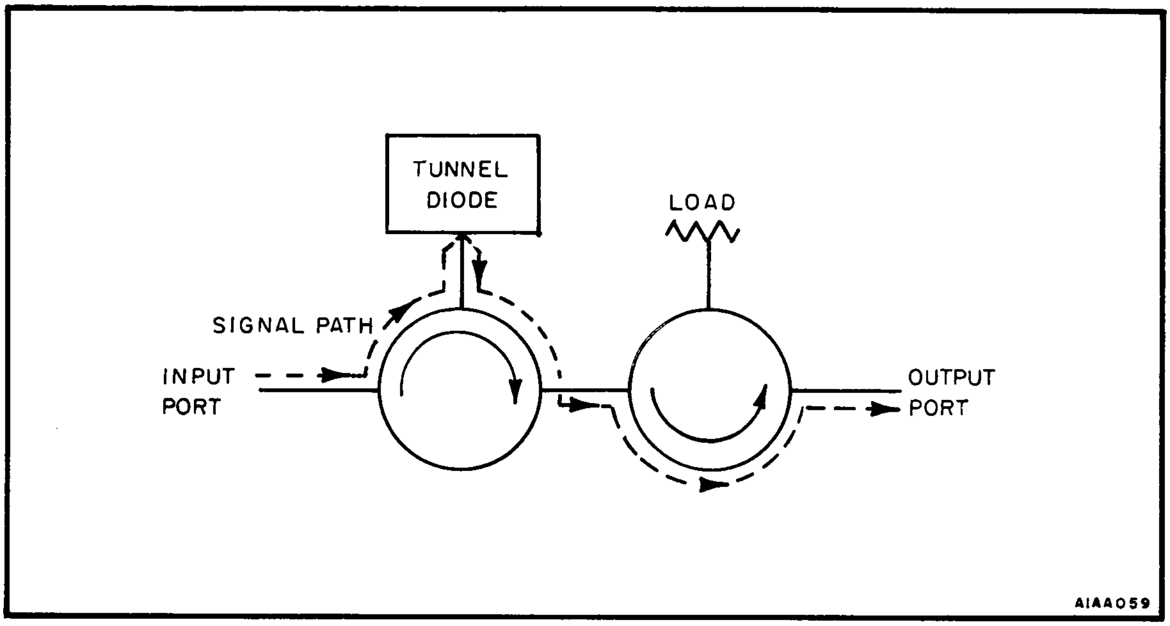

Figure 10-4. Tunnel Diode Amplifier Schematic

As with a parametric amplifier, a tunnel diode (see figure 10-5) has negative input impedance at its design frequencies and under its bias conditions. Reflection of signals from this negative impedance amplifies the signals. L_s and R_s are the series inductance and resistance, C_j is the junction capacitance, and $-R_n$ is the negative resistance of the tunnel diode. The input impedance, Z_{in}, consists of a real and an imaginary part, both of which are functions of frequency. According to semi-conductor physics, $-R_n$ and, to a lesser extent, C_j are both functions of the instantaneous voltage across the diode junction. Therefore, the input impedance depends on frequency, DC bias, and signal level. In practical amplifiers, resonant circuits and transformer sections are used with Z_{in} to produce an overall diode impedance. $Z_d = R_d + X_d$, which varies typically with frequency as shown in figure 10-6 for small-signal conditions, where the diode behavior is essentially linear.

Electrical characteristics of a typical tunnel diode amplifier used in a receiving system are listed in Table 10-2.

A limit to the tunnel diode amplifiers is that they become nonlinear above -40 dBm. Where amplification is necessary above -40 dBm, high peak-current tunnel diodes are used.

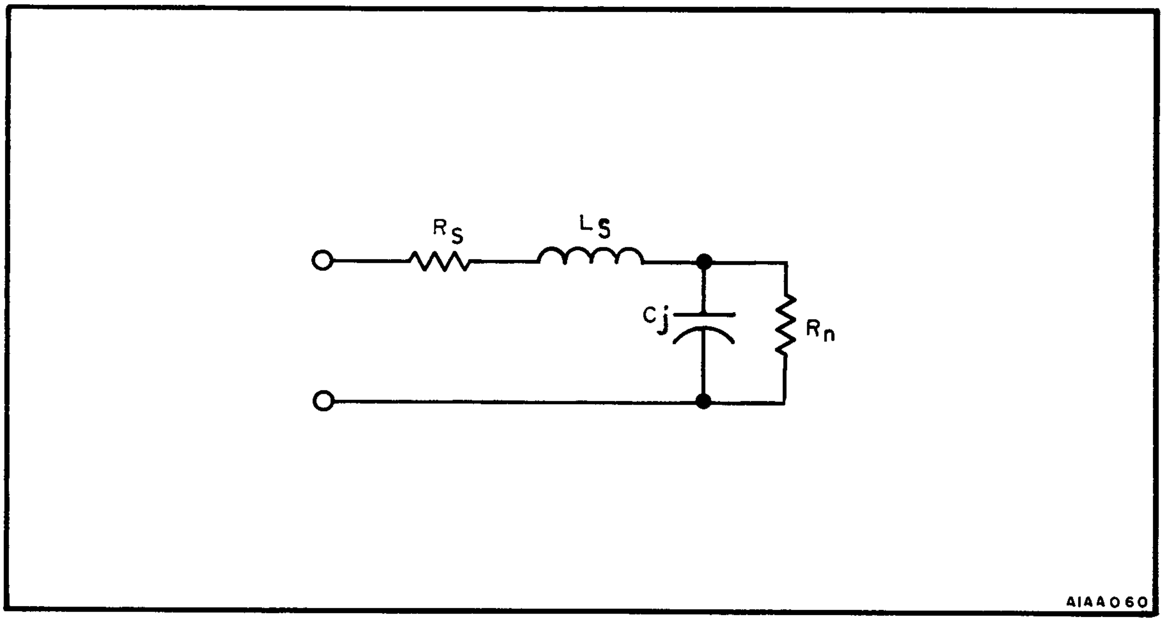

Figure 10-5. Tunnel Diode Equivalent Circuit

Table 10-2. Typical Tunnel Diode Amplifier Characteristics

PARAMETER	CHARACTERISTICS
Noise figure	5.0 dB max
Bandwidth, 0.5 dB	500 MHz min
Frequency band	3.7 to 4.2 GHz
Amplitude response ripple	Less than 0.5 dB
Amplitude response slope	0.02 dB/MHz max at any frequency in band
Gain	14 dB min
Gain stability	± 0.5 dB under all conditions
Burnout level	+ 10 dBm
Departure from linearity	0.01 dB up to -55 dBm input
Input match, SWR	1.2 max
Stability	Unconditionally stable with short or open at input or output at any phase
Delay slope	0.1 nanosec/MHz at any frequency in band

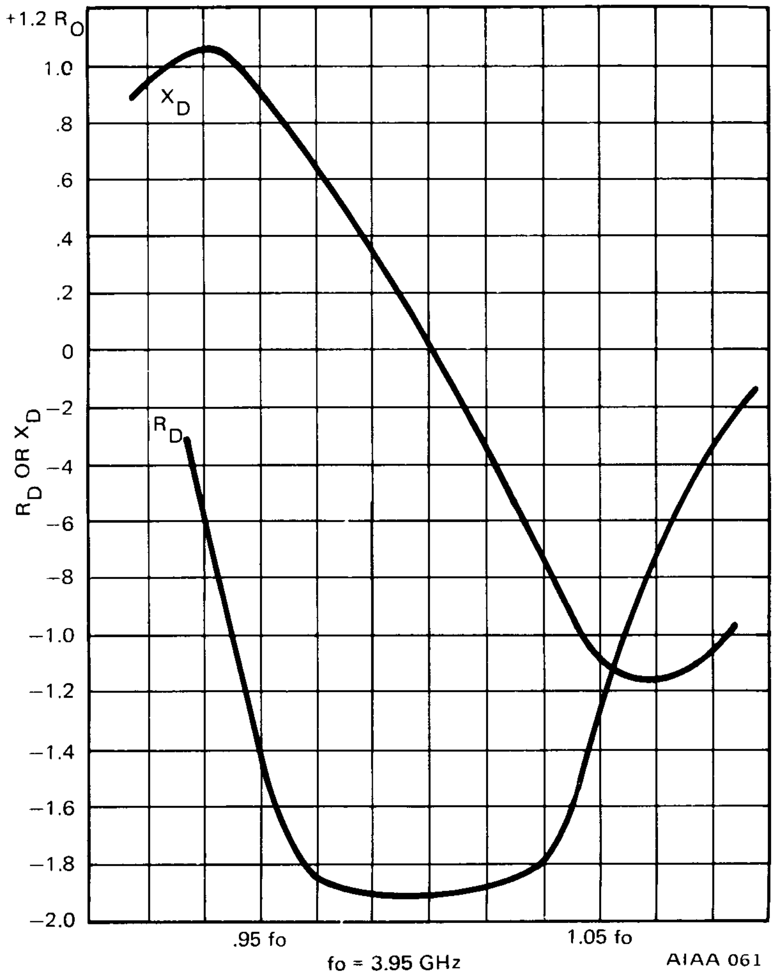

Figure 10-6. Tunnel Diode Impedance Versus Normalized Frequency

The basic difference between high-peak tunnel diodes and the others is that the area of the junction is larger. Although the negative resistance of high-current tunnel diodes is smaller than that of low-current tunnel diodes, causing lower gain, the large junction area lowers distortion and increases the power handling capability. A graph of output versus input for a high level tunnel diode amplifier is shown in figure 10-7. The graph also shows the magnitude of intermodulation products versus input for two carriers. The magnitude of the intermodulation products gives a measure of the nonlinearity of the tunnel diode amplifier.

10.1.2 Superheterodyne Receiver

The main type of receiver used at frequencies from 1 to 13 KMHz, as at lower frequencies, is the superheterodyne. Variations of the individual stages occur as the frequency is increased, but the general principles of operation remain the same.

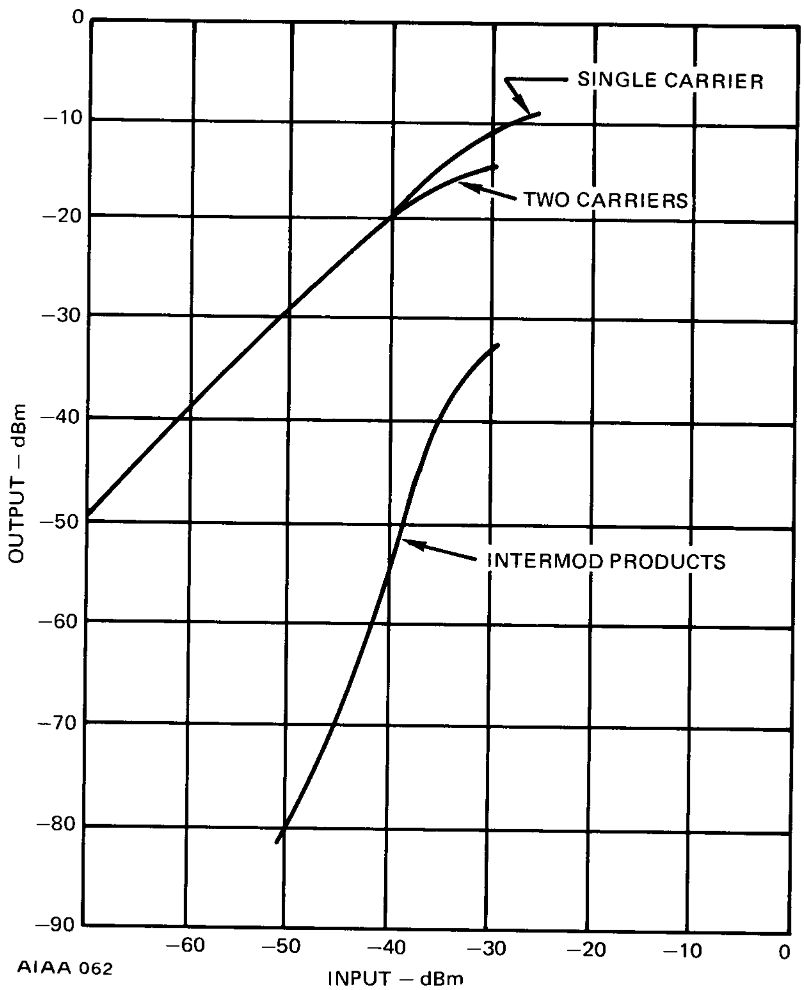

Figure 10-7. High Peak-Current Tunnel Diode Amplifier: Input Versus Output and Intermodulation Products

A superheterodyne receiver (see figure 10-8) operates by heterodyning, or mixing, the received RF signal with a locally generated RF voltage obtained from the local oscillator. These two voltages are combined in a non-linear device such as a crystal rectifier (mixer), producing, in addition to the original frequencies, the sum and difference frequencies. This process is identical with amplitude modulation as used in transmitters.

The difference frequency (lower sideband) is selected and amplified by a fixed tuned intermediate frequency (IF) amplifier system. The IF amplifier frequency remains the same in a given receiver regardless of the incoming signal frequency. This is accomplished by changing the local oscillator frequency so that the difference frequency between the desired signal and the local oscillator signal remains constant. Currently, the microwave receiver IF is specified to be 70 MHz by the Defense Communication Agency (DCA).

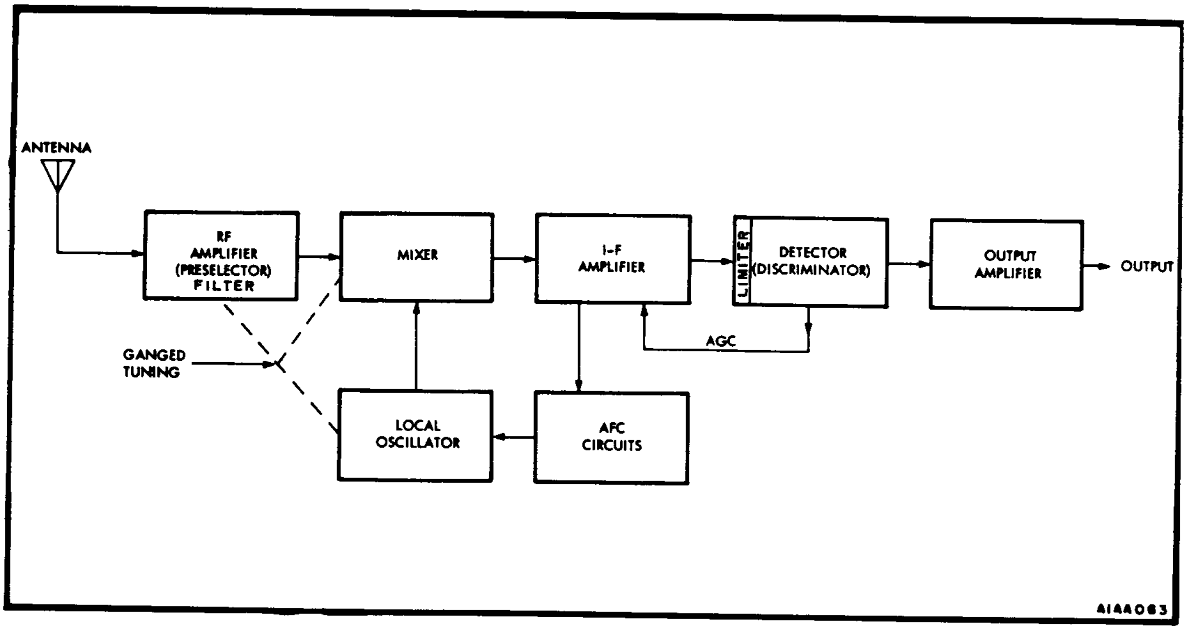

Figure 10-8. Superheterodyne Receiver, Block Diagram

After amplification, the IF signal is demodulated. The information and the carrier are separated and the signal containing the information is passed on to the output amplifier. Here the signal is amplified enough to be used as desired.

At some point in the frequency band about 1500 MHz, the use of an RF stage in a receiver results in a lower signal-to-noise ratio than without the stage. Klystrons produce amplification at these frequencies and are useful as transmitting amplifiers, but they are too noisy for use in receivers as RF amplifiers. Tube noise is important only when small signal levels are being handled. If the signal strength is of the same order as the noise produced in a stage, a large amount of distortion will result. However, if the signal level is of much larger amplitude than the noise, little if any effect will be produced. For this reason, noise produced in amplifiers is of importance only in low power level stages such as the initial stages of a receiver. Traveling wave tubes are useful as RF amplifiers in receivers and are made with as low a noise figure as the crystal mixers used for the first stages in receivers above 500 MHz. Receivers used for higher frequencies have as their first stages (1) the antenna, (2) a tuned circuit or preselector, and (3) a crystal mixer or traveling wave tube (TWT) for converting the signal to a lower intermediate frequency. The preselector is composed of one or more tuned circuits, which reject all frequencies except the desired frequency band. Coaxial cavities may be used for frequencies from about 500 MHz up.

a. Preselector. In general, a preselector is fed from a low-impedance source, the antenna, and is coupled to the higher impedance of a crystal mixer. The tuner must therefore act also as an impedance transformer to obtain maximum energy transfer. It must have a pass band that is at least as wide as the pass band of the IF amplifier and an off-frequency attenuation which produced the largest amount of image and harmonic rejection that can be obtained. If the desired attenuation characteristic cannot be obtained by a single-tuned circuit and still maintain the required pass band, double- or triple-tuned circuits must be used.

b. Local Oscillators. The local oscillator used in the receiver must operate at a frequency similar to the received frequency. The exact frequency of the oscillator is the desired signal frequency plus or minus the intermediate frequency. The local oscillator may operate either at a higher or a lower frequency than the received signal, since the difference frequency will remain the same. Klystron oscillators, usually of the reflex variety, are used as local oscillators almost exclusively above 3 KMHz. Additionally, solid-state oscillators are currently available for several frequency bands.

c. Mixers. The device used to produce the IF from the received signal and the local oscillator signal is the mixer (first detector). Any nonlinear device may be used for this purpose, although at frequencies above 500 MHz crystal mixers are appreciably superior to any triode mixers now available because of the small amount of noise generated. The conversion loss of type-1N26 crystals is low, even at 23 KMHz, with a maximum conversion loss varying from 5.5 to 7.5 dB in the frequency range from 3 to 16 KMHz. The maximum noise ratio (the amount of noise generated by the mixer compared to the noise produced in an equivalent impedance resistor) varies from 1.5 to 2.5 in the same frequency range. Silicon crystal mixers have the disadvantage that they are damaged by overloads.

d. IF Amplifiers. The difference frequency produced in the mixer may be any desired value. Selection of the frequency to be used in a given receiver will depend on several factors. A high IF is desirable to eliminate image response, but the selectivity and noise figure of an amplifier becomes worse with an increase in frequency, and ganged tuning of the local oscillator and preselector is thus a compromise between the desired image-rejection ration, on one hand, and the desired sensitivity, selectivity, and circuit simplicity on the other.

A method of obtaining both good image-rejection by using a high IF, and good selectivity and simplicity of circuits through the use of a low IF, is the double-conversion superheterodyne receiver. A high IF is first produced to obtain a good image-rejection characteristic. This frequency is further reduced by a second frequency conversion, bringing the IF down to a frequency that can be easily amplified and that provides good selectivity. The oscillator used to produce the second IF may be crystal controlled since it heterodynes with a fixed frequency. The disadvantage of this system is the additional local oscillator and mixer stages required.

e. Detectors. Some means of separating the carrier from the desired information must follow the IF amplifier. The second detector (demodulator) is used for this purpose.

Detectors used for FM are often called discriminators. A discriminator produces a DC voltage proportional to the frequency of an input signal. This may be accomplished in a number of ways. One method uses two tuned circuits, one tuned above the center frequency and one tuned below to obtain two IF voltages whose amplitudes depend directly on frequency. These voltages are rectified and combined so that zero voltage output is obtained at the center frequency. A difference voltage of the two IF voltages that is proportional to the frequency is obtained when the frequency of the IF signal is above or below the center frequency. A discriminator of this type is sensitive to amplitude variations in the input, so a clipper-limiter stage must be used preceding the discriminator to remove any AM. Another method of obtaining frequency discrimination is the use of the phase detector. This operates by comparing the phase relationships of two signals, one the IF signal and the other the IF signal phase-shifted in a resonant circuit tuned to the center frequency. The amount of phase shift will be proportional to the frequency of one signal. A limiter stage is also required for the phase detector.

To eliminate the need for a limiter stage, several types of discriminators have been developed that will respond to amplitude changes. The first of these is the ratio detector. Instead of the two rectified voltages of the ordinary discriminator being combined with opposite polarity, they are combined so as to add. The sum of the voltages is kept constant by a large-value capacitor. This eliminates any amplitude variations of the IF signal and makes the use of a limiter stage unnecessary. The ratio of the two voltages changes as the frequency input to the ratio detector changes and the output is taken from one of the rectifier loads.

f. AFC Systems. Receiver stability is even more important than transmitter stability. In contrast to the AFC systems used in transmitters, which tend to keep the operating frequency at an absolute value, receiver AFC systems usually compare the receiver's operating frequency to the received signal. This is done by using a discriminator in the IF channel to develop the error signal used to retune the oscillator. An FM receiver may use the same discriminator for signal detection and AFC provided the signal variation is averaged over a period of time so that the oscillator does not attempt to follow the modulation, but only slow frequency drifts. With the exception of the development of the error signal, AFC is obtained by the same methods as in transmitters and is discussed in that paragraph.

g. AGC Systems. Since received signal strength will vary over a wide range some means of automatic gain control (AGC) is required in the receivers. This is accomplished by rectifying the received carrier and averaging the voltage over several cycles of modulation. This voltage is used as bias for the IF amplifiers so that as the average signal strength increases, the gain of the IF amplifier is decreased. The bias is usually not applied to the first amplifier stage since the stage tends to become noisier with AGC applied. With some types of modulation, such as for TV, the rectified IF signal may be too large to make practical the use of the average value to control the receiver gain. Special techniques must be used to provide AGC in these cases.

h. Output Amplifiers. The detected modulation must be amplified for further use. The output amplifier must amplify the output frequency range, which in some equipment is up to 20 MHz wide, with uniform response. For pulse transmission, the bandwidth in MHz must be at least twice the reciprocal of the pulse duration in microseconds if the pulse is to approximate its original shape. Phase shift characteristics are as important as bandwidths. Since phase distortion will change the pulse shape by phase shifting the harmonics which make up the pulse difference angles.

10.1.3 FM Noise Threshold

FM threshold is defined as the point at which the received RF carrier signal peak, equals or exceeds noise peaks 99.999 percent of the time when the RF-RMS carrier level is 10 dB above noise threshold. The detector in an FM receiver is controlled by peak voltages. When noise peaks exceed signal peaks, the receiver follows noise peaks rather than the signal variations which are marked by noise. Peak noise is defined as the voltage level which is exceeded a certain percentage of the time, depending on the peak reference chosen. In a random noise distribution, the arbitrary ratio choice of noise peaks to noise RMS equal to 13 dB means that RF peak signal input equals or exceeds noise peak 99.999 percent of the time (figure 10-9). The ratio of peak to RMS for a sinusoidal signal is 3 dB. Therefore, the ratio of the RMS value of signal to RMS value of noise at threshold where noise peaks equal signal peaks is 10 dB.

Analytical Derivation of FM Threshold:

N_p = Peak Noise Voltage

RMS = Effective Noise Voltage at Noise Threshold

S_p = Peak Carrier Signal Voltage

S_{rms} = Effective Carrier Signal Voltage

From a statistical distribution of noise, the following statements can be made:

$N_p = N_{rms} + 13$ (Noise peaks are exceeded 0.001 percent of the time)

$S_p = S_{rms} + 3$

By definition at FM Threshold, the peak signal voltage shall be equal to or greater than the peak noise voltage 99.999 percent of the time.

$$Sp \geq Np$$

$$S_{rms} + 3 \geq N_{rms} + 13$$

$$S_{rms} \geq N_{rms} + 10$$

The expression arbitrarily defines FM Threshold. The RMS RF signal level, if 10 dB above noise threshold, means that noise peaks and noise bursts will exceed the peak signal level 0.001 percent or less of the time, or that the input signal voltage peaks will exceed noise peaks 99.999 percent of the time.

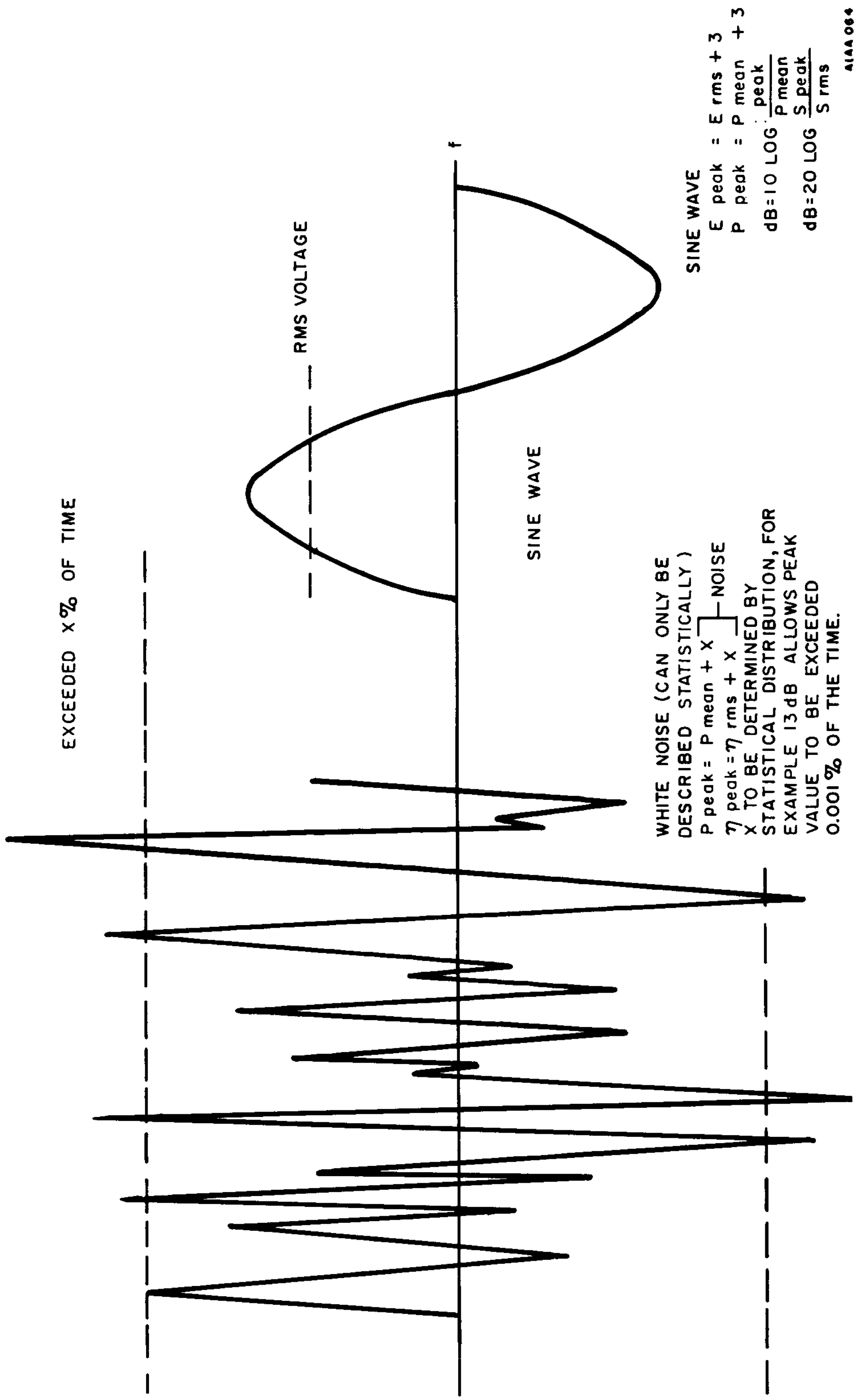

Figure 10-9. White Noise Versus Sine Wave

Noise Threshold - The RF input level at which signal power just equals the internally generated front end noise power or where RMS noise equals RMS signal.

10.1.4 Receiver Characteristics

To properly describe and evaluate overall FM receiver performance, it is necessary to analytically trace the curve of output signal-to-noise (S/N) ratio as a function of the input RF carrier, that is, S/N = f(C/N) and correlate this expression with experimental data. In many specifications, usually only one point is defined. This is insufficient to properly describe this function, with operation below FM threshold not discussed or described. To evaluate FM receiver performance the function must be described. For example, when a 3 dB threshold improvement is measured, what is a proper reference? This can be determined by use of this curve. In carefully analyzing experimental data and curves for many typical FM receivers, it appears reasonable to conclude that the curve [S/N = f(C/N] can be approximated by three asymptotic straight lines, each with a different slope. The curves can be defined analytically, and will be developed from noise threshold to receiver saturation.

Refer to figure 10-10 which shows the theoretical development of S/N = f(C/N).

a. Noise Threshold to FM Threshold

From Figure 10-10, Slope I, at noise threshold,

C/N axis,

$$\text{Noise Threshold} = 10 \log KTB + NF(dB) \qquad (10\text{-}2)$$

S/N axis,

$$(S/N)_1 = 0 \text{ (dB)} \qquad (10\text{-}3)$$

at FM threshold,

C/N axis,

$$\text{FM Threshold} = \frac{10 \log KTB + NF\ (dB) + 10\ dB}{\text{Noise Threshold}} \qquad (10\text{-}4)$$

S/N axis,

$$(S/N)_2 \text{ (Thermal)} = 10 + (10 \log \frac{B_{IF}}{2b} + 20 \log \frac{F}{FM} - L + P + W + \frac{C}{N}(dB)) \qquad (10\text{-}5)$$

where,

B_{IF} = Carson rule bandwidth = 2 F + 2 F_m, normally corresponds to IF bandwidth

F = Peak deviation

FM = Highest modulating frequency

b = Voice channel bandwidth = 3 KHZ

L = The multichannel loading factor

P = Pre/De-emphasis improvement

W = The weighting factor improvement or effective voice channel shaping for measurement purposes.

Equation 10-5 defines S/N ratio due to thermal noise in one voice channel for a multichannel FDM system. Equations 10-2 through 10-5 therefore, locate the points connected by straight line I between noise threshold and FM threshold. The slope of line I is,

$$\text{Slope} = \frac{(S/N)_2 - (S/N)_1}{\text{FM threshold - noise threshold}} \tag{10-6}$$

$$= \frac{(S/N)_2 - 0}{10 \text{ lot KTB} + \text{NF} + 10 - (10 \log \text{KTB} + \text{NF})}$$

$$\text{Slope} = \frac{(S/N)_2}{10}$$

b. <u>FM Threshold to Receiver Saturation</u>

From figure 10-10, Slope II,

$$\text{S/N (THERMAL)} = \text{C/N} + \boxed{10 \log \frac{B_{i-f}}{2b} + 20 \text{ Log } \frac{F}{Fm} - L + P + W} \tag{10-7}$$

The encircled portion of the equation can be described as a constant for a particular receiver once the variables are fixed. The equation can then be written as follows:

$$S/N = C/N + K$$

From this equation, it is apparent that a linear relationship exists between output S/N and input C/N. The coefficients of both are equal and the slope for curve II is equal exactly to one. It can also be stated that the slope is exactly one for all FM receivers when operating above FM threshold and up to receiver saturation; that is, assuming that the only variable is the received RF input (C/N) with the other variables held constant. In addition, it is possible to obtain a family of curves for one receiver by varying any one of the parameters B_{IF}, F, or Fm.

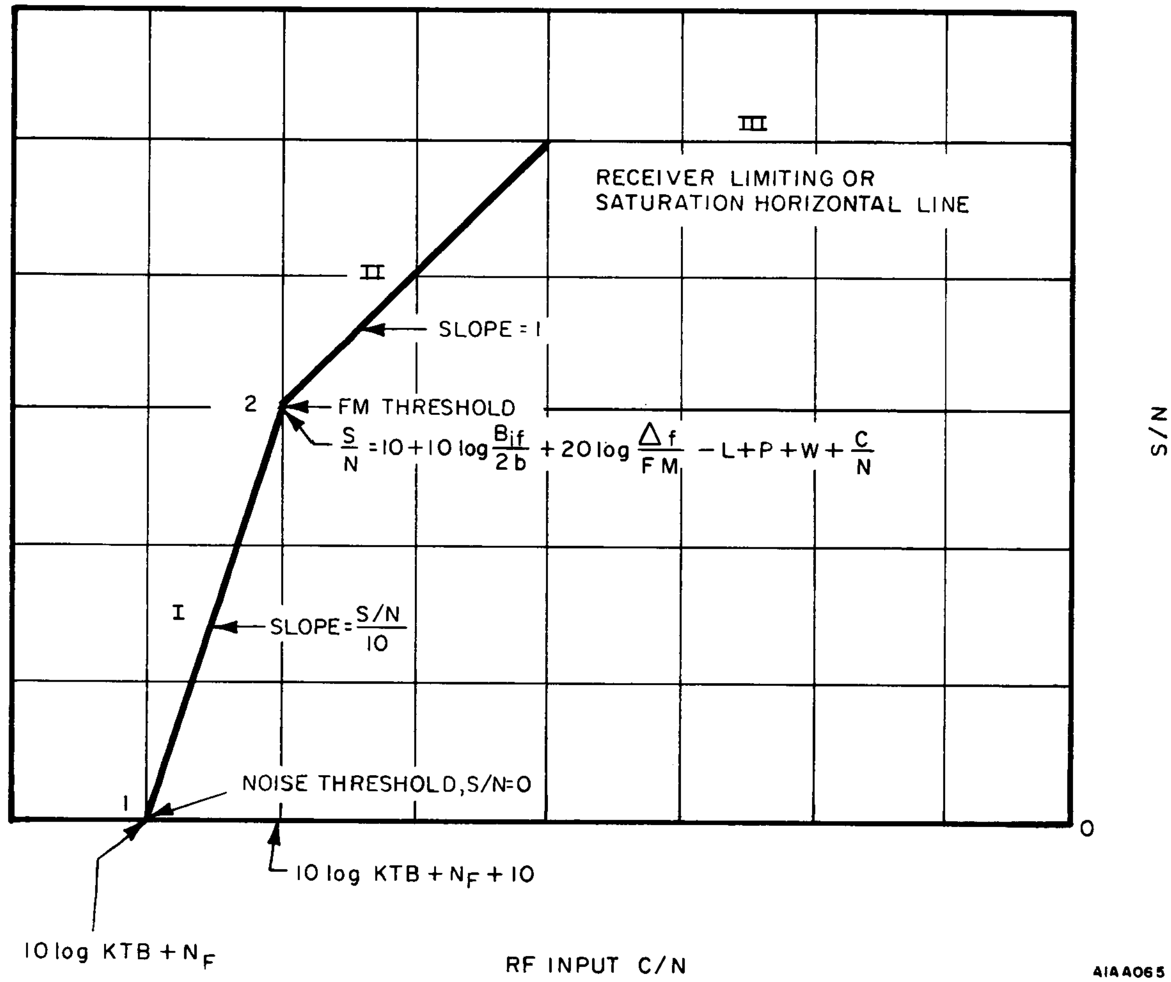

Figure 10-10. Analytically Developed Characteristic Curve of S/N Versus C/N for FM Receivers

c. <u>Receiver Saturation</u>

The output S/N ratio in one voice channel will keep increasing for a corresponding increase in received RF level, since this decreases the affects of thermal noise (see figures 10-11 through 10-14). The increase in RF input signal level increases the affects of AGC which then decreases receiver gain and thus the output thermal noise contribution will decrease, since output thermal noise is equal to KTBG. This process will continue until the received RF carrier level drives the RF amplifier beyond AGC control. At this juncture, an increase in received RF level does not decrease the affects of thermal noise and, consequently, the S/N ratio will be held virtually constant. It should be emphasized that the equation for output S/N is for a single FDM voice channel (3.1 kHz. This equation cannot be used for other applications, such as wideband signals, TV, PCM, radar video, etc. In practice, the receiver RF amplifier does not saturate completely but does saturate in a gradual asymptotic fashion; therefore,

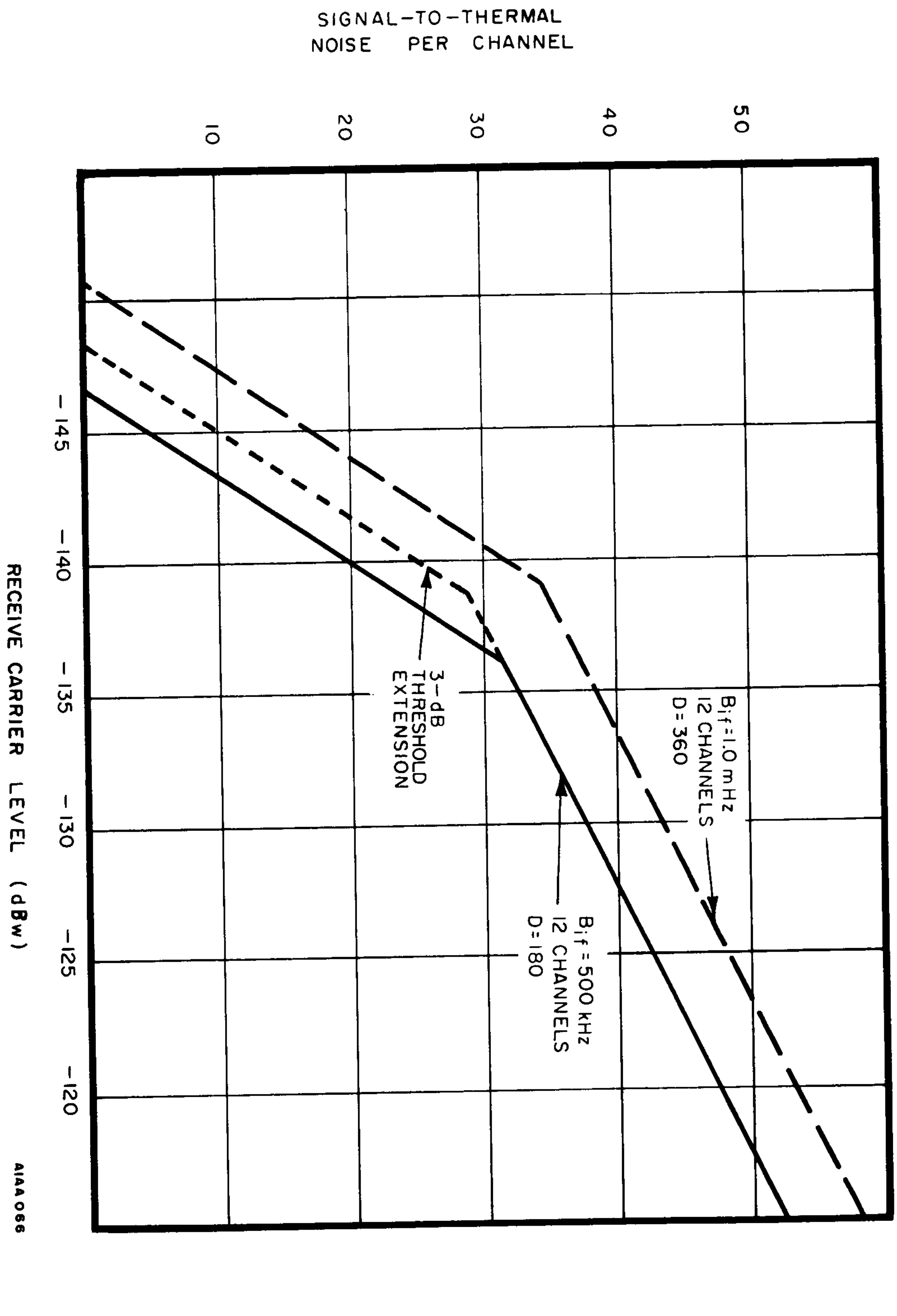

Figure 10-11. Receiver Comparison Using S/N to C/N Application

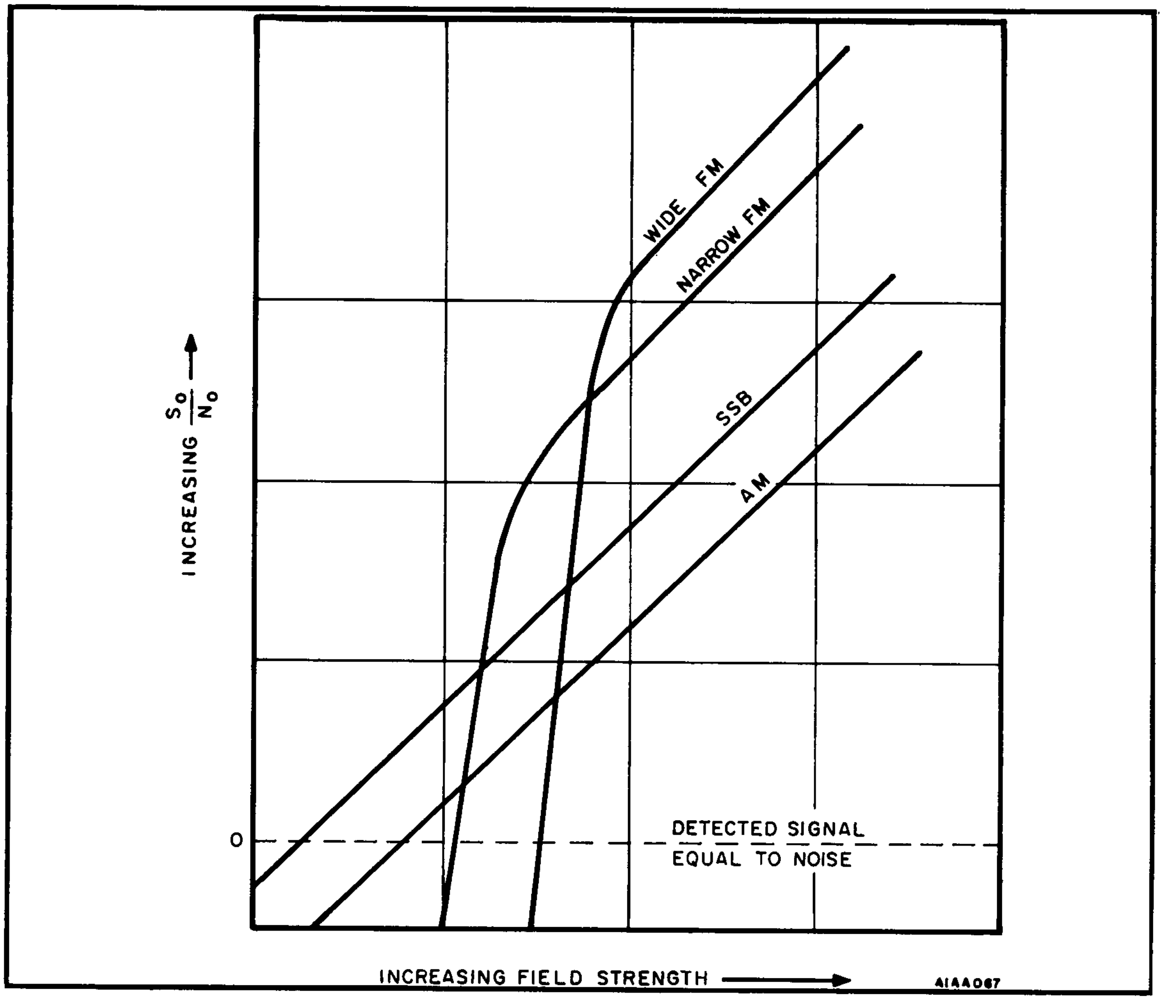

Figure 10-12. Relative S/N Ratio of AM and FM Systems as a Function of Field Intensity

the S/N at saturation is not a horizontal line. However, from a practical point of view, describing it in this manner will be reasonably accurate for most engineering considerations. The point at which receiver saturation occurs cannot be determined without having receiver design data available.

10.1.5 Threshold Extension

In a system with a perfectly linear input-output transfer characteristic, the noise bandwidth would be the bandwidth of the narrowest filter. Most receiving systems become non-linear below the point identified as the "threshold." It is therefore convenient to use two bandwidths when establishing system performance. The first is the noise bandwidth which is either the pre-detection bandwidth or the bandwidth inherent in the detection system itself, whichever is smaller; and the second is the base bandwidth

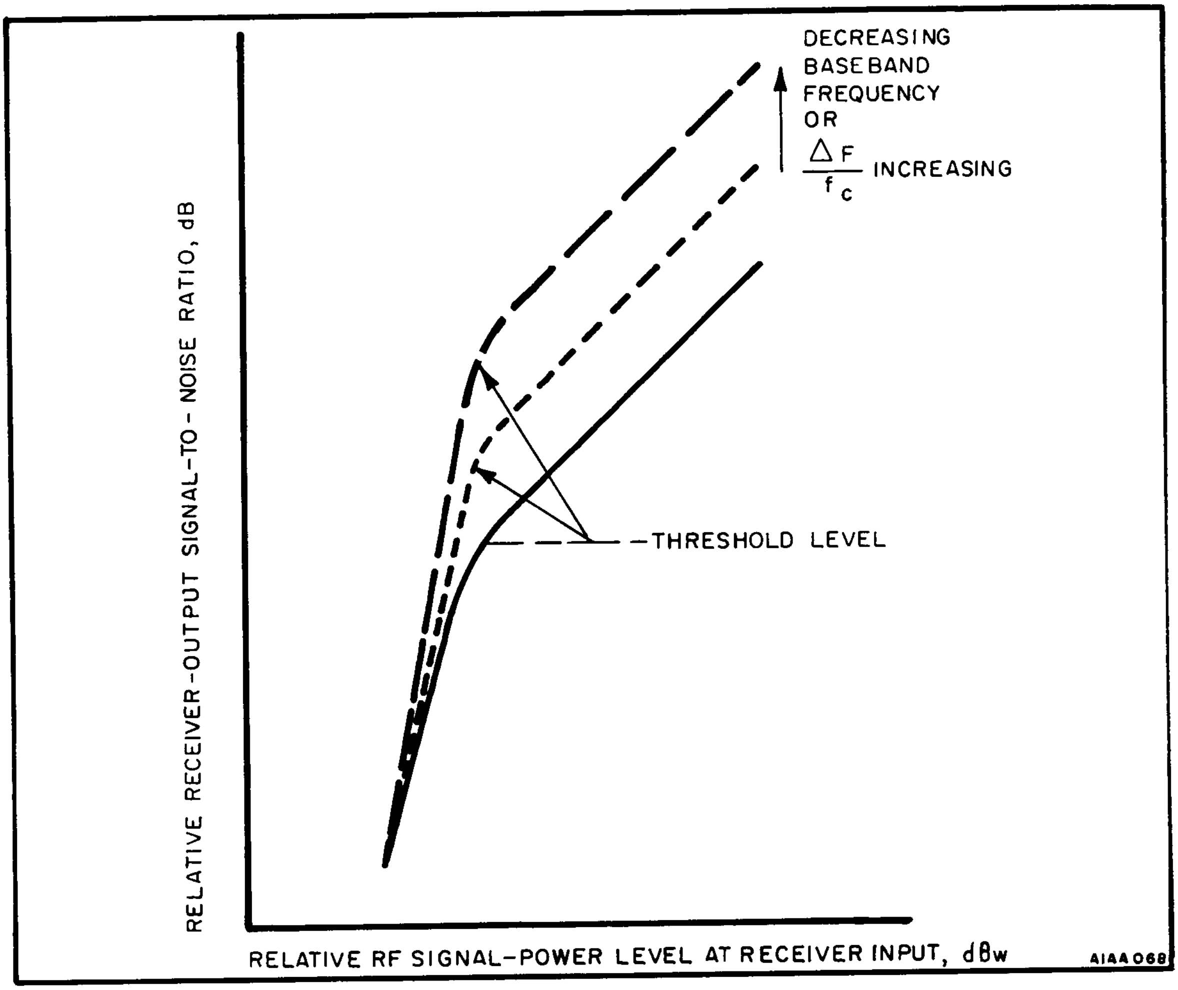

Figure 10-13. FM Receiver Characteristic Curves, Without Pre-emphasis

which is the effective bandwidth of the post-detection and instrument filters. To identify signal powers in each of these bandwidths, it is conventional to use the term "carrier" for the total signal power in the noise bandwidth and the term "signal" for the demodulated signal power in the base bandwidth. In the case of multiplex telephone, carrier means the total received signal while signal means the demodulated intelligence in one channel only, usually the highest. The conventional plots of receiver performance show the S/N ratio measured in the base bandwidth as the ordinate, and the C/N power ratio measured in the noise bandwidth as the abscissa. These plots are meaningless unless the noise bandwidth is precisely known. As this is seldom the case, it is much more meaningful to plot carrier power at the input terminals of the first amplifier as the abscissa. It is difficult to assign an exact number to the base bandwidth because of the effect of shaping networks, filters, and instrument characteristics. It is more meaningful to identify the frequency response of the entire post-detection circuit and also the actual measuring instrument.

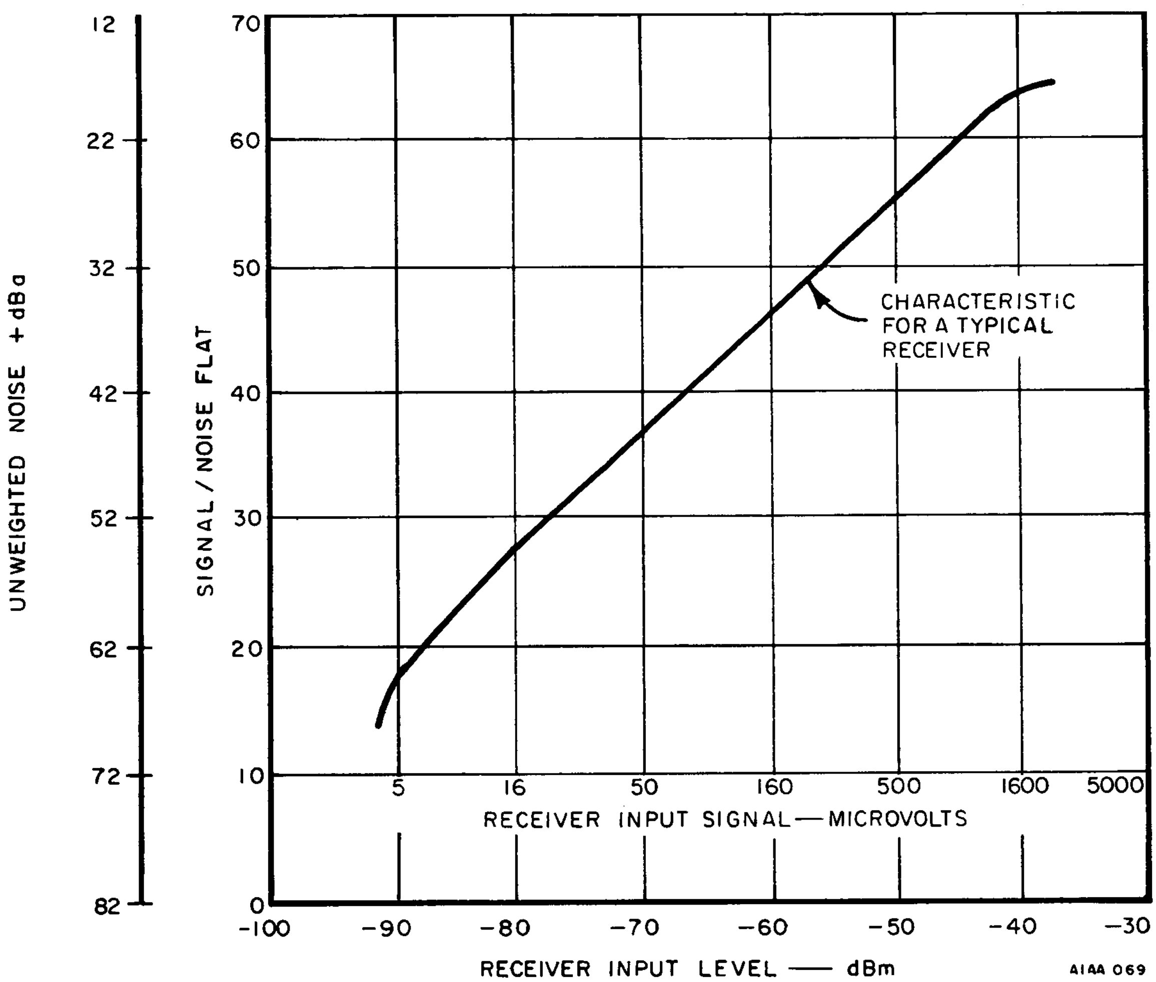

Figure 10-14. Thermal Noise Characteristics Showing Noise as a Function of Receiver Input Power

For design, the noise bandwidth must be known. When measuring a complete receiver using FM feedback, the noise bandwidth is not a simple measurement. The noise bandwidth of typical receivers can be estimated from Table 10-3.

Selection of noise bandwidths for a receiving system depends on the receiver threshold, which dictates how wide the bandwidth can be. The amount of information that must pass through the system dictates how narrow the bandwidth can be.

Threshold is the minimum C/N ratio that yields an FM improvement which is not significantly deteriorated from the value predicted by the usual small-noise S/N formulas. For comparison, the threshold is picked as that carrier power at which the output S/N ratio has dropped 1 dB below, i.e., departed from, the linear relationship between S/N ratio and carrier power.

Table 10-3. Noise Bandwidth of Receivers

TYPE OF RECEIVER	NOISE BANDWIDTH
Conventional	Same as 3 dB bandwidth of IF amplifier
F-M Feedback	1.57 times the 3-dB bandwidth of the filter in the feedback loop if it is a simple single-pole network; or 1.25 times the 3-dB bandwidth of the filter in the feedback loop if it is a double-tuned circuit.
Phase Lock Feedback	3.24 times the 3 dB bandwidth of the closed loop, provided that the damping factor of the loop is 0.707 and provided that the input noise is white, additive gaussian.

If the noise bandwidth is reduced below the base bandwidth by a simple filter, severe distortion results in the output unless the carrier deviation is reduced accordingly at the transmitter so that the base bandwidth is no longer wider than the noise bandwidth. Reducing the deviation, in turn, greatly reduces the advantages gained by using frequency modulation. A brief review of the basic fundamentals of frequency modulation shows that when modulation is applied, a theoretically infinite number of sidebands, spaced symmetrically above and below the carrier, are generated. In actual practice, only the significant sidebands, i.e., those containing significant power, are considered. The bandwidth required to convey this information by a frequency modulated carrier depends upon the amplitude and frequency of the modulating signal. This leads to the formation of the modulation index (m) which is defined as

$$m = \frac{F}{f_m} \tag{10-8}$$

where

F = carrier deviation

f_m = modulating frequency.

The modulation index, when applied to mathematical tables derived from Bessel functions, provides the number of significant sidebands that will be generated for the particular index, and therefore the required bandwidth. As shown in Table 10-4, a large index results in a large number of sidebands and a large bandwidth; a small index results in a smaller number of sidebands and a corresponding reduction in bandwidth requirements.

In a conventional FM receiver, the IF and demodulator bandwidth is made to be $2f_m(m + 1)$. A list of required bandwidths for several modulation indices is given in Table 10-4. To lower the threshold by decreasing receiver bandwidth, the modulation index must be reduced to ensure an undistorted output signal. This process is called threshold extension.

Table 10-4. Modified Bessel Chart

MODULATION INDEX (m)	NO. OF SIGNIFICANT SIDEBAND PAIRS	BANDWIDTH (bw) (f = MODULATING FREQUENCY)
0.01	1	2f
0.1	1	2f
0.4	1	2f
0.5	2	4f
1.0	3	6f
2.0	4	8f
3.0	6	12f
4.0	7	14f
5.0	8	16f
Note: When m>3, a reasonable approximation of occupancy is bw = 2f (1 + m)		

The term threshold extension is also used to define the performance of an FM feedback receiver with reference to a conventional receiver with a noise bandwidth equal to $2f_m(m + 1)$. The extension is given in terms of the reduced carrier power, referred to the carrier power at threshold in a conventional receiver, expressed in dB. Since designers have claimed both reduced noise bandwidth and reduced carrier-to-noise ratio in this bandwidth, the value of threshold extension must be given by:

$$TE = 10 \log \frac{\text{carrier power to reach threshold in conventional receiver}}{\text{carrier power to reach threshold in feedback receiver}} \quad (10\text{-}9)$$

$$= 10 \log d/N_c = 10 \log C/N_t + 10 \log B_c/B_t$$

where TE = threshold extension in dB

C/N_c = C/N ratio measured on conventional receiver with noise bandwidth = B_c

C/N_t = C/N ratio measured on feedback receiver with noise bandwidth = B_t

B_c = noise bandwidth of conventional receiver

B_t = noise bandwidth of feedback receiver

One wideband FM system characteristic facilitates reducing effective bandwidth: although the incoming signal may occupy any position in the passband, the transmitted baseband information is restricted in the rate at which it may move from one passband position to another. This signal-modulation restraint is used to obtain effective bandwidth reduction in all currently used FM-feedback threshold-extension systems.

10.1.6 Noise Figure

Noise figure in dB (noise factor) is defined as:

$$N_f = \frac{(C/N)_{in}}{(C/N)_{out}} = \frac{C_{in}/N_{in}}{\frac{C_{in}G}{N_{out}}} = \frac{\text{NOISE OUT}}{\text{NOISE IN X G}} \qquad (10\text{-}10)$$

$$N_f = \frac{\text{NOISE IN X G} + N_x}{\text{NOISE IN X G}} = \frac{kT_\circ BG + N_x}{kT_\circ BG}$$

$$N_f = 1 + \frac{N_x}{kT_\circ BG} = 1 + \frac{T_e}{T_o} \text{ [in dB]}$$

In a perfect device, the noise output and the noise input are equal, assuming unity gain. An actual amplifier has a finite noise contribution and that is why the noise figure >1.

G = Power Gain

T = Source Temperature assume 290°K

B = Bandwidth at 3 dB reference (in FM receivers, this is I_f bandwidth)

N_x = Noise power contributed by device that is being measured = K T_e BG

N_{in} = Input Noise power

N_{out} = Output Noise power from device

T_e = Effective input noise temperature, a measure of internal noise sources of device

The noise figure of a device is a function of temperature of the noise generator, and may be defined for any source temperature. Usual custom is to define noise figure in terms of standard temperature $T_\circ$ = 290K, and noise figure can be determined at different source temperatures T, by using the equation:

$$N_{fs} = 1 + \frac{N_x}{GKT_sB} \quad \text{or} \quad N_{fs} = \frac{T_o}{T_s}(N_f - 1) + 1 \quad \text{or} \qquad (10\text{-}11)$$

$$N_{fs} = \frac{T_o}{T_s}\left(\frac{N_x}{GKT_\circ B}\right) + 1.$$

Noise figure of a complete receiver

$$N_f = 10 \log n_f = n_{F_1} + \frac{n_{F_2} - 1}{G_1} + \frac{n_{F_3} - 1}{G_1 G_2} \cdots \frac{n_{F_n} - 1}{G_1 G_2 \cdots G_{n-1}} \qquad (10\text{-}12)$$

where

N_f = Voice Figure

G_1 = Gain of first stage

G_2 = Gain of second stage

F_1 = Noise factor of first stage, etc.

It is evident that the system noise figure is affected primarily by the first stage gain and noise figure, with each subsequent stage having a decreasing effect. When using a high gain, low noise, front end, the system noise figure is approximately equal to the noise figure of the front end (tunnel diode amplifier or parametric amplifier).

10.1.7 Emphasis

Pre-emphasis is an FM practice of increasing the amplitude of the higher frequency components of the baseband, according to a pre-determined plan (a pre-emphasis curve). An examination of the problems in FM which give rise to the necessity of using pre-emphasis is probably the best means of arriving at an understanding of this practice and the reasons for it. The fundamental problem necessitating the use of pre-emphasis lies in the output characteristics of an FM discriminator. If uniform white noise (constant amplitude with respect to frequency) is placed at the input of an FM discriminator, the output will be a finite band with linearly increasing amplitude with increasing frequency. Discriminator input and characteristics are shown in figures 10-15 and 10-16. The ramp-like output resulting from a white noise input of finite bandwidth is illustrated in figure 10-17. The problem results from the fact that white noise is applied at the discriminator along with the IF signal output. After detection, the higher frequency baseband components are degraded by the ramp-like noise output. The result is a decreasing S/N ratio with increasing baseband frequency. Since optimum performance requires constant S/N ratio across the baseband spectrum, this effect must be corrected. Pre-emphasis at the transmitter modulator is used to accomplish this purpose.

Since increasing peak frequency deviation (ΔF) will increase baseband amplitude, pre-emphasis is accomplished by increasing the peak frequency deviation in accordance with a curve design to effect compensation for the ramp-like noise at the discriminator output. A pre-emphasis curve is shown in figure 10-18. Note that the curve is in terms of deviation ratio (D) rather than ΔF, despite our previous statement that ΔF is the parameter which is varied for pre-emphasis. This apparent inconsistency is resolved when we examine the definition of D and note that in any given baseband configuration, the modulating frequency (F_m) is essentially constant, being taken by convention as either the middle or upper baseband subcarrier frequency. Drawing out pre-emphasis curve in terms of the deviation ratio (D) extends the usefulness of the curve by making it applicable to a variety of different baseband configurations with different subcarrier frequencies.

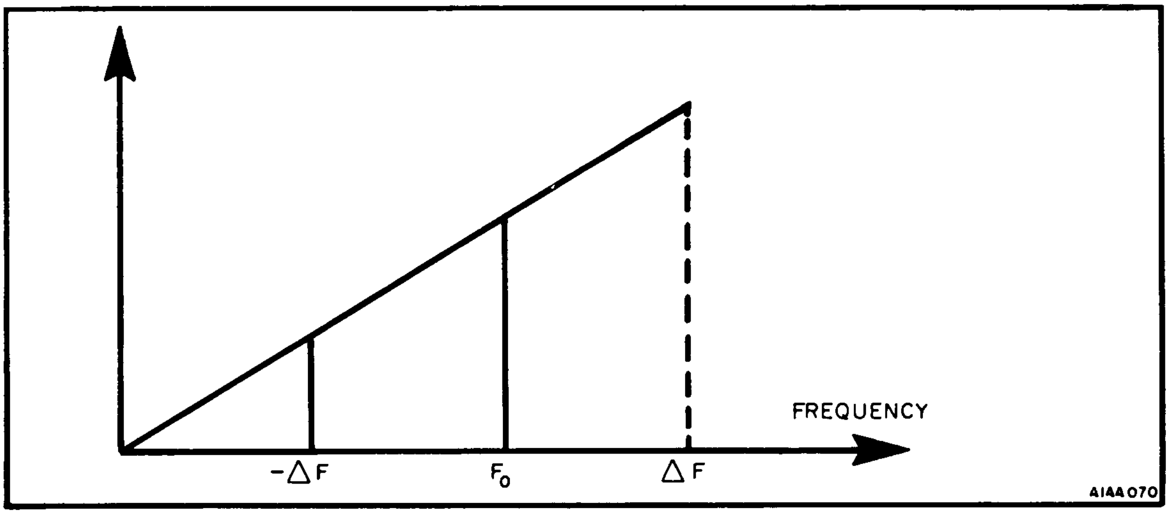

Figure 10-15. Discriminator Characteristics

Once pre-emphasis has been allowed to accomplish its purpose of preserving a constant S/N across the baseband, it is then necessary to restore the detected baseband to its original amplitude configuration. This restoration (de-emphasis) will affect both noise and the signal, preserving the original S/N ratio. The de-emphasizing process can be performed by a single passive network having an attenuation characteristic (de-emphasis curve) inverse to the pre-emphasis curve. Pre-emphasis and de-emphasis are also used for individual FDM voice channels. This is necessary because of the spectral amplitude distribution of speech which shows a high amplitude level concentration at the lower frequency end of the voice channel. This signal distribution, when applied to a baseband, can result in masking the higher frequency low-amplitude components by white noise present before modulation, and in the receiver after detection. The amplitude of the higher frequency components of the voice channels increased to compensate for this condition. De-emphasis is used at the demultiplexer to restore the proper amplitude configuration.

10.1.8 Selectivity

The selectivity of a receiver is its ability to differentiate between the desired signal and signals at other frequencies. Selectivity is usually defined as the ratio of the sensitivity for desired signals to the sensitivity for undesired signals, expressed in decibels. The sensitivity of the receiver for a range of frequencies about the desired frequency is often plotted as a selectivity curve whose shape is primarily determined by the response of the IF amplifiers.

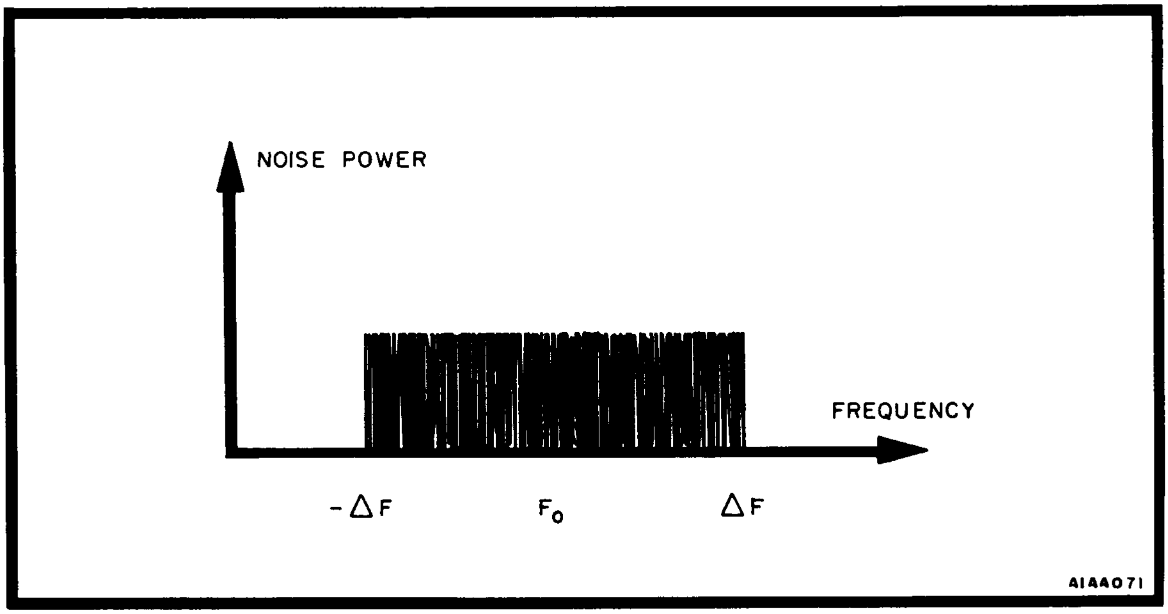

Figure 10-16. Limited Noise Input to Discriminator

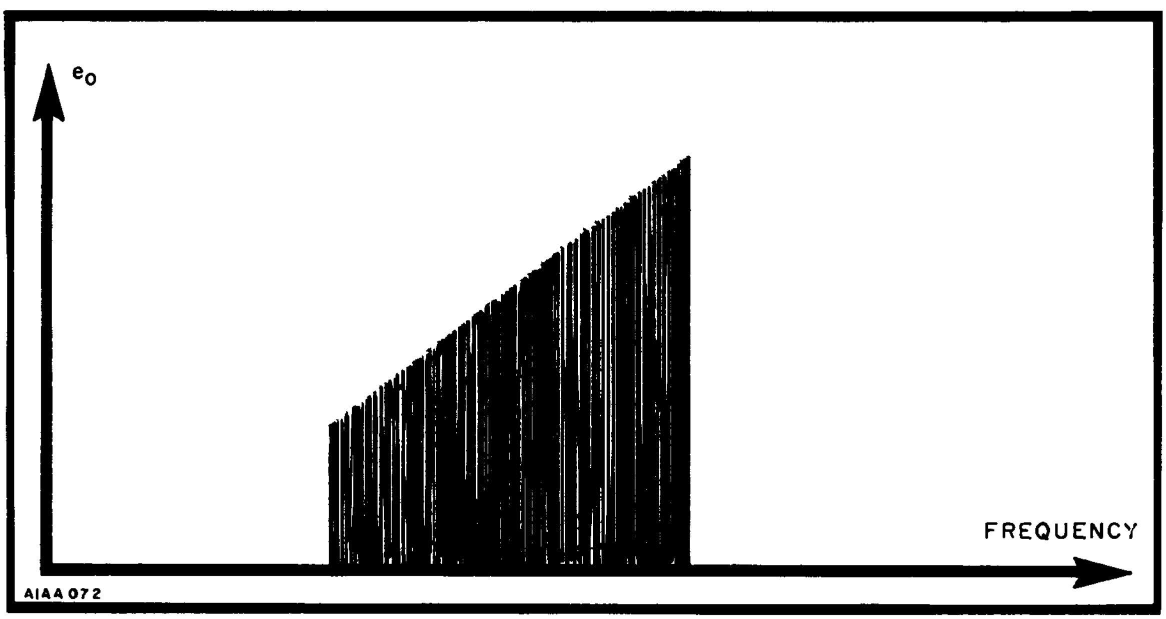

Figure 10-17. Discriminator Output Characteristics, Triangular Noise

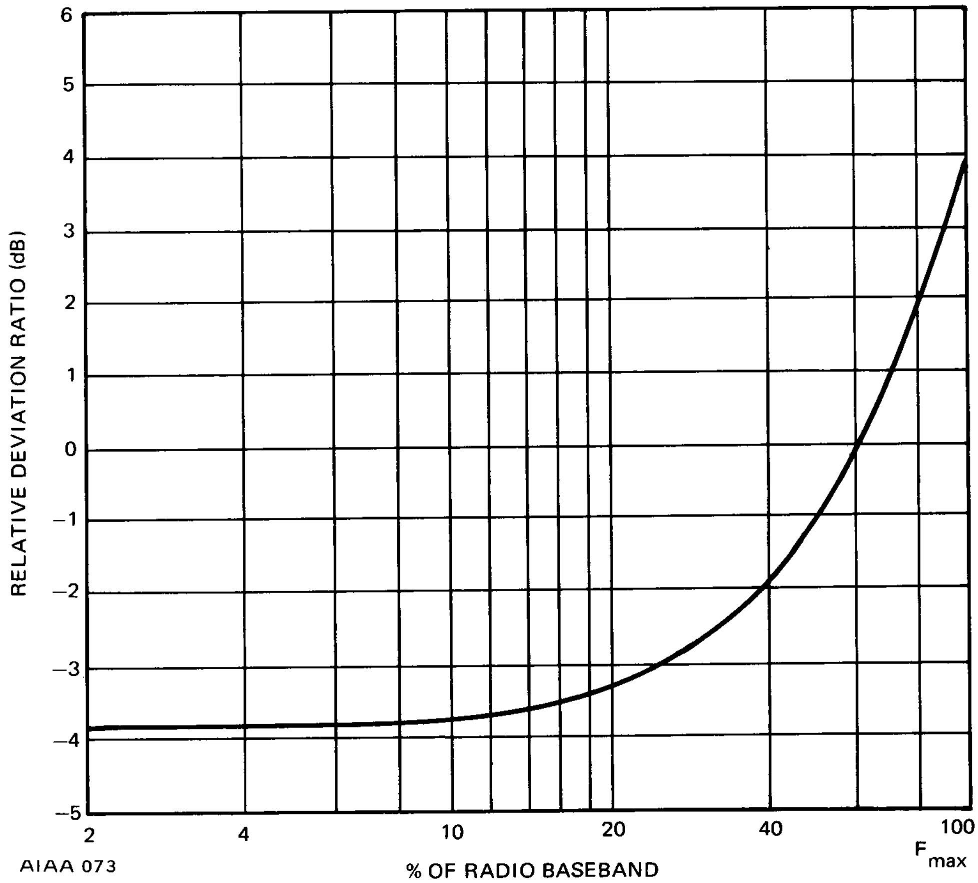

Figure 10-18. Normalized Pre- and De-emphasis Curve

10.1.9 Image Rejection and Spurious Response

The image ratio at the receiver is a measure of the rejection of that frequency which, when mixed with the local oscillator signal, produces the same IF as the desired signal. If the local oscillator is operated below the signal frequency, the image frequency will be the same amount below the local oscillator frequency that the signal frequency is above. The image rejection must take place prior to the mixer in the RF amplifier or preselector stages. Presently the DCA specifies that the image and out-of-band frequencies shall be rejected by at least 60 dB.

10.1.10 Frequency Stability

Frequency tolerance of the receiver shall be 1 part in 10^6 per day ($\pm$ 0.0001%) or better.

10.2 TRANSMITTER

The transmitter consists basically of an oscillator, modulator, and amplifier, with the complexity of these units depending on frequency, power output required, and type of modulation used. This section discusses oscillators as they apply to microwaves, methods of frequency stabilization, microwave modulators, and power amplifiers used in microwave transmitters.

10.2.1 Oscillators

As microwave frequencies are approached the triode coaxial oscillators begin to fall off in efficiency and again new systems must be resorted to in order to produce oscillations at these frequencies. Three general types of oscillators have been developed for use in these ranges: the klystron, the magnetron, and the traveling wave tube. While in the coaxial triode oscillators the problem of interelectrode transit time was encountered, these new types of oscillators take advantage of the finite speed of the electron.

a. Klystron. Two main categories of klystron oscillators exist: the reflex klystron and the 2-cavity klystron. The reflex klystron has low efficiency (about 1 percent) and generates relatively low power. It is primarily used in receivers, local oscillators, and test equipment. Two-cavity klystrons are much more efficient (from 20 to 40 percent) and can be designed to generate high power levels. Most of the limitations of conventional negative grid tubes do not exist in klystrons. The cathode and anode are outside the RF field and therefore may be made as large as desired. The cathode to anode spacing is of the order of one inch so that extremely high voltages may be used without danger of internal arcing. The only limiting factor in the amount of power which may be produced by a klystron is the loss in the dielectrics making up the windows between the output cavity and the load.

In a 2-cavity klystron, a stream of electrons from and electron gun passes through a resonant cavity called a buncher. This cavity is the input cavity and contains an RF field corresponding to the signal input. In the case of an oscillator, the input is fed back with the proper phase relationship from the output or catcher cavity. The buncher either accelerates or retards the electrons in the stream depending on the portion of the RF cycle. Following the buncher there is the drift space where the electron beam is unaffected except by the uniform accelerating force of the anode voltage. In this space the electrons form into bunches, the retarded electrons falling back and the accelerated electrons moving forward to the next bunch. When the electrons, now bunched, reach the catcher, they set up a varying electric field from which energy may be taken. The electrons themselves continue on to the anode, or collector, where their remaining kinetic energy is dissipated as heat.

A modified version of the 2-cavity klystron is the reflex klystron oscillator. The operating principles remain nearly the same. In this klystron the coupling between the input and the output is accomplished by the electron beam itself. In fact, the two resonating cavities are replaced by one cavity which functions both as buncher and catcher. The electrons are produced and accelerated by an electron gun as before. Then they pass through the cavity for the first time, being velocity modulated as in the buncher of a 2-cavity klystron. The electrons travel into the drift-space but instead of being further accelerated, they are in a uniform retarding field produced by the negative repeller plate. The electron beam slows to a stop and then reverses direction being accelerated back toward the cavity. As the bunched electrons pass through the cavity for the second time, they give up part of their energy to the cavity and are then stopped by the cavity, which also functions as the collector. The frequency of operation can be changed to a limited extent by changing the repeller voltage, thus changing the transit time in the drift stage.

The reflex klystron is less efficient than the 2-cavity klystron because a single resonator performs both functions of bunching and catching. On the other hand, the single resonator tuning and the ease of electrical tuning by varying the repeller voltage makes it better for use when only small amounts of power are required. Another advantage of the reflex klystron is its greater stability as compared to a 2-cavity klystron when used as a master oscillator.

b. Magnetron. To aid understanding of the principle of operation in a traveling-wave magnetron, the most commonly used, consider the movement of an electron in magnetic and electric fields. An electron moving at right angles to a magnetic field will be acted upon by a force perpendicular to both its direction of motion and the magnetic field. This force does not change the velocity of the electron but causes it to move in a circular path, the radius of the path being determined by the magnetic field strength and the velocity of the electron. An electron moving parallel to an electric field will be either accelerated, taking energy from the field, or retarded, giving energy to the electric field.

The magnetron is basically a fixed-frequency device, but certain of the newer types may be frequency modulated by changing the potentials on certain elements. Anode power is normally applied to a magnetron in very short pulses of very high amplitude. Voltages of 40 kilovolts and currents of 100 amperes are not unusual in pulsed magnetron service. It is possible to produce a peak power of 2.5 megawatts at 3000 MHz with an efficiency as high as 50 percent. At frequencies of 25,000 MHz, more than 50 kW may be obtained, but the efficiency will fall to about 25 percent.

The traveling wave tube (TWT) connected as an oscillator is essentially an amplifier which uses an electron beam and an RF wave traveling together in such a way that the wave accepts energy from the electron beam. In some ways it is similar to the linear magnetron discussed previously. The TWT consists of a helical coil inside a conductor. It may be considered as a coaxial cable with a helical inner conductor. In operation, the RF wave to be amplified travels along the helical coil which greatly reduces the velocity of the wave. This slower velocity causes the RF wave to travel at the same speed as an electron beam centered in the helical coil, which enables the RF wave to accept energy from the beam.

Assuming the electron velocity and the wave propagation velocity are the same, the electrons in the beam will be retarded or accelerated by the electric field. This will cause bunching to occur, with the electron bunches forming in alternate points of zero longitudinal electric field. In producing these bunches, as many electrons are retarded as are accelerated and no net transfer of energy is made in either direction. Since this would produce no amplification, some means must be found of obtaining a transfer of energy from the electron beam to the electric field. This can be done by a slight increase in the velocity of the electron beam. The electron bunches are now at a retarding point of the electric field and the electrons are retarded for a longer period of time than they are accelerated. This will produce a transfer of energy to the wave, and therefore the wave is amplified. A necessary addition to the TWT is some means of preventing the electron beam from spreading. This is done by using a longitudinal magnetic field. As long as the electron beam moves parallel to the magnetic field it has no effect on the electrons. When the electron strays from a parallel path, however, the magnetic field forces the electron back into the beam. By coupling the output to the input in the proper phase relationship, oscillation may be produced. The TWT can be used over a great range of frequencies with high gain, at a cost of low power output and efficiency.

10.2.2 Frequency Stability

An ideal oscillator would be one in which the frequency could be easily adjusted and, once set, remain at that frequency regardless of temperature, output load, or voltage input. At low frequencies these conditions are relatively easy to approach, but as operating frequencies are increased, stability of operating frequency becomes more difficult to obtain. Even by attaining the same percentage of stability, which in itself is hard to do, serious frequency shifts may occur at microwave frequencies. A frequency shift of 0.01 percent at 1 MHz is only 100 cycles which presents no problems, but this same percentage shift at 10,000 MHz is equal to 1 MHz which is enough to interfere with satisfactory operation.

Three primary factors affect the operating frequency of an oscillator. There are, first, geometric factors in which the effective inductance and capacitance are changed directly through mechanical motion; second, pulling factors in which reactance is coupled into the oscillatory circuit from the load; and third, pushing factors in which reactance is introduced by changes in input conditions, such as voltage, current, or magnetic field.

There are three means of insuring stable operating frequencies. One is by the use of frequency stabilizers that tend to maintain a constant frequency of oscillations; another is by automatic frequency control (AFC) systems that mechanically or electronically retune the oscillator when it shifts from a reference frequency; the third is the use of synthesizing circuits, where crystal control is necessary to maintain the required frequency stability.

As required by DCA standards, equipment design shall be such that the center frequency of the radiated signal from any transmitter which generates its own RF signal internally shall be maintained to within 150 kHz of the assigned frequency.

10.2.3 Emphasis

Both pre-emphasis and de-emphasis are defined and discussed in paragraph 10.1.7.

10.2.4 RF Extraneous and Spurious Outputs

The average power of any extraneous or spurious emissions in the $f_o \pm 5\% f_o$ band. MIL-STD-461 and Figure 14, Appendix C of MIL-STD-188C specify the out-of-band emission limits in terms of absolute power levels versus the transmitted power of the fundamental. Measurements shall be made taking full advantage of transmission line and antenna filtering characteristics. (For RF leakage, other undesired-emanation measurements and permissible limits refer to Military Standards 461 and 462.)

10.2.5 Deviation-Mod Index

In FM, the varied parameter of the carrier, which carries the amplitude of the modulating wave, is its instantaneous frequency, but the maximum deviation of frequency from its assigned value is limited arbitrarily and independently of the modulation. Thus, frequency allocations for FM broadcast are for a 200-kHz channel. A guard-band of 25 kHz is used at each side of the channel, leaving a 150-kHz bandwidth, or plus and minus 75 kHz from the carrier resting frequency. Other allocations for FM services may limit the total band to 50 kHz, or 10 kHz. A function called "modulation index" is the ratio of the maximum frequency difference between the modulated and the unmodulated carrier, or deviation frequency to the modulation frequency. It is sometimes referred to as the "deviation ratio." The degree of modulation in an FM system is usually defined as the ratio of the frequency deviation to the maximum frequency deviation allowable, or the ratio of frequency deviation to the maximum deviation of which the system is capable. Degree of modulation in a frequency modulation system, therefore, is not a property of the signal itself.

10.2.6 Power Amplifiers

The modulated signal may be passed to an amplifier to increase the amplitude of the outgoing signal. The same limitations of conventional circuits at microwave frequencies that applied to oscillators apply as well to amplifiers. In the microwave region no amplification is possible with conventional vacuum tubes and circuits; either the oscillator itself must supply enough power, or specially designed amplifiers must be used.

a. Klystron Amplifier. The klystron amplifier may be a 2-cavity klystron as used for an oscillator or a special amplifier klystron used for high power, which is known as a cascade or 3-cavity klystron. This tube is effectively two klystrons connected in cascade within the same envelope, with the catcher for the first section functioning as the buncher for the second section. The signal to be amplified is applied to the first cavity and the power output is taken from the third cavity. The second cavity is energized by the bunched electron beam and is not supplied with external RF driving power. These tubes are capable of power gains up to 30 dB, efficiencies of 30 to 40 percent, with a power output of 12.5 kW.

b. Traveling Wave Tube Amplifiers. Traveling wave tube amplifiers are the second type of amplifier tube that may be used at microwave frequencies. The tubes have inherent regenerative feedback due to wave reflections in the tube. When designed for amplifier service, some means of attenuating the reflected wave must be provided. However, they are capable of large amplifications with a wide passband and high efficiency.