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RADIO BROADBAND BEARER SYSTEM PERFORMANCE

1. INTRODUCTION

1.1 During the last financial year, Australians made almost 590 million trunk telephone calls and 8.4 million international telephone calls. They also made more than 37.4 million national Telex calls and 8.4 million international Telex calls. If we consider the number of telegrams sent, the television relays provided and the growing demand for data links, we can see that the national telecommunications network is a vital industry in Australia.

Broadband radio relay systems form a very significant part of the telecommunications network, as currently in Australia, there are more than 23,600 kilometres of broadband radio routes, providing 146,858 kilometres of one-way radio bearers. This network of radio systems requires a large amount of planning, provision and maintenance to ensure availability of high quality, reliable circuits.

1.2 When designing a radio broadband system, the design engineer is limited by various factors in getting the best system for the lowest cost. Most of the limitations are the result of the performance of currently available equipment: no equipment is ideal. Some limitations are not caused by equipment but by other factors such as interference from other sources. Planners must consider all of these factors when designing a system to conform to national and international standards.

At the completion of the installation of equipment for a system, acceptance tests are carried out to ensure that the equipment functions as expected and that the performance of each piece meets the required standards.

For proper maintenance of a radio broadband bearer, routine tests are performed on the bearer. The results of these tests may be compared with the acceptance tests to determine any variations. In the event of bearer degradation or failure, specific items of equipment may need to be tested to locate the fault.

1.3 The purpose of this paper is to examine how the performance of a radio bearer affects the quality of circuits provided by the system. An understanding of these factors will assist staff involved in planning, provision and maintenance of broadband radio systems to comprehend performance objectives.

This paper examines some quite complex aspects of radio broadband systems. To enable a complete understanding of this subject, the reader should be familiar with the principles explained in the paper, 'Radio Broadband Bearer System Principles'.

2. BROADBAND SYSTEM REQUIREMENTS

2.1 This section discusses the radio broadband bearer characteristics which must be optimised to ensure that all signals are transmitted to the output with stable gain, without significant alteration and with minimal addition of interference signals.

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The characteristics which affect these requirements are:

:: Amplitude linearity

:: Phase linearity

:: Amplitude - frequency characteristic

:: Phase - frequency characteristic

:: Noise performance

:: Gain Stability

:: Frequency stability

2.2 AMPLITUDE LINEARITY. Amplitude linearity is indicated by a characteristic relating input and output amplitudes of a system.

When non-linearity exists in a system, the output waveshape is different from that at the input; that is, distortion is produced (Fig. 1a).

For FDM broadband carrier systems, the important result of distortion is that extra components are produced in the band occupied by the carrier system. These components are harmonics of the lower frequency components of the signal and the intermodulation products that fall in the passband. Because of the large number of components in the carrier system frequency spectrum, intermodulation products are by far the more serious source of trouble. The intermodulation products appear over the complete baseband but, because of the large number of channels and the random nature of the signal, intelligible crosstalk is not produced. At the receiving end of the system, the products appear as a random noise signal in each telephone channel. Extremely good linearity is required to keep the noise in each channel below acceptable limits.

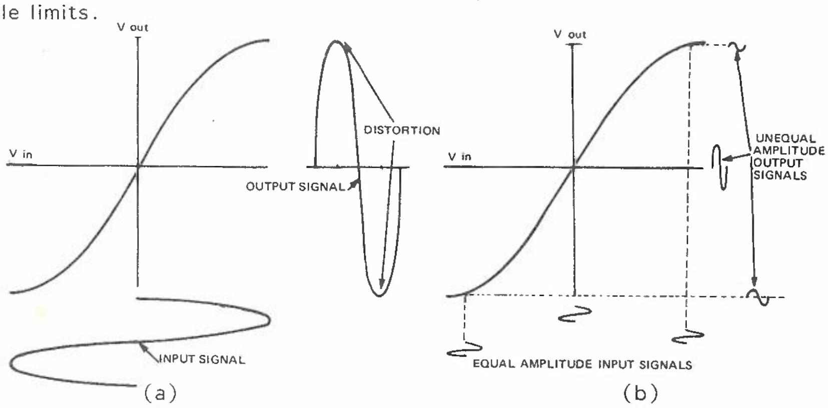


FIG. 1. EFFECTS OF AMPLITUDE NON-LINEARITY

As the position of a small amplitude signal on the input-output amplitude characteristic is varied, there should be no change in the amplitude of the output signal; that is, the gain of the signal should be constant over all of the characteristic. In a non-linear system, the small signal gain of the system varies between points on the characteristic (Fig. 1b). A measure of this variation indicates the 'differential gain' of the system.

Television bearers require that the shape of the output waveform should be identical with that at the input. While in a non-linear television bearer system additional frequencies are produced, more serious effects of the non-linearity are to vary the constrast of shades in the picture and to compress the height of the synchronising pulses. In a colour television system, the non-linearity may cause errors in the chrominance sub-carrier information, thus altering the colour saturation for high luminance colours. Sub-system distortions which give rise to system amplitude non-linearities are discussed more fully in Section 5 of this paper.

With Pulse Code Modulation (PCM) systems, good linearity is not a requirement since the signal is of a binary nature. In the receiving section of a PCM system, regenerating circuits are provided which effectively correct for any preceeding non-linearity.

2.3 PHASE LINEARITY. As the position of a small amplitude signal on the input-output amplitude characteristic is varied, no change in phase of the output signal should occur. However, in a system with phase non-linearity, phase changes do occur. When traffic is a TV signal, the variation in phase of small amplitude signals at different points in the amplitude characteristic is known as the 'differential phase' characteristic.

The hue of a reproduced colour at any instant depends on the phase of the sub-carrier relative to the phase of the colour burst. Errors in the differential phase characteristic cause variations in phase of the chrominance signal as the luminance level varies. This results in an incorrect phase difference between the chrominance signal and the colour burst, which is always at black level. In the NTSC system, the reproduced hue would be incorrect. In the PAL system, the technique of averaging the hue of adjacent lines allows reasonable phase errors to be tolerated. However, to ensure that the signal supplied to television broadcast transmitters meets the specified standards, bearer circuits are required to have good phase linearity. Though less easy to see in detail what effect this type of phase error has on TF traffic, it is clear that the baseband waveform will be distorted. In fact intermodulation noise results.

2.4 AMPLITUDE-FREQUENCY CHARACTERISTIC. The amplitude-frequency response of a bearer must be constant over at least the frequency range of the baseband signals. A reduction of the signal amplitude in a section of the frequency spectrum of an FDM broadband carrier system causes a reduction in the transmission equivalent for the group of channels occupying that band. For TDM carrier systems, the bearer amplitude-frequency response should have a smooth characteristic with neither peaks nor a sharp cut-off at the high frequency end of the range. A sharp cut-off introduces ringing associated with signal transitions. Also the characteristic should be free from ripples, as these introduce signals equivalent to echoes. Such degradations may cause errors to be introduced into the coded signal regenerated at the receiving end of the circuit.

In a television system, relative amplitude reductions at the high frequency end of the frequency range decrease the resolution in the reproduced picture, while peaks at high frequencies cause ringing. In addition, for colour video signal transmission, a reduction in amplitude of frequencies in the region of the colour subcarrier reduces the degree of colour saturation in the reproduced picture.

Frequencies below about 10 kHz are responsible for the vertical resolution of a television picture. These low frequencies represent background shading and the changes of shading in the vertical direction. If a completely white picture is scanned, the signal is a 50 Hz square-wave. If this low frequency square-wave is not amplified with its waveshape preserved, the reproduced picture will show a white screen, having a gradual change of intensity from top to bottom. When low-frequency response is particularly poor, the background in a normal reproduced scene is dull, with poor contrast. This is because the low video frequencies represent the main areas of the picture information.

To provide a TV signal with an adequate low frequency response over a radio bearer, the low frequency 3 dB point in the TV baseband amplitude-frequency characteristic must be lower than 30 Hz.

Errors in the response at intermediate frequencies in the baseband usually cause streaking in the picture. Degradation of television signals and the effects on picture reproduction are considered in detail in other Engineering Training publications.

2.5 PHASE-FREQUENCY CHARACTERISTIC. The phase-frequency characteristic of a bearer system must be given particular consideration to ensure accurate transmission of waveforms such as those used by TDM carrier systems and television. For FDM carrier systems, the phase-frequency characteristic normally is not so important. This is because the phase variations in each derived channel are usually only minor and the ear is not sensitive to these phase variations.

For the correct transmission of a pulse, however, all components of the spectrum must have the same transmission delay as delay errors cause the shape of the received pulse to be altered. To achieve a constant delay for all components, the phase shift produced by the system must be proportional to the frequency of the component. The phase-frequency characteristic may be examined by measuring the group delay of the system. Group delay (also known as envelope delay), see para 2.6 is a measure of the slope of the phase shift-frequency characteristic. A system with a phase shift proportional to frequency, gives a constant group delay. This is covered more fully in a later paragraph.

Considering the transmission of colour television signals, small changes in the phase-frequency characteristic, which affect both the colour burst and the chrominance signal during the line time, do not produce errors in the line of the reproduced colour picture. However, a difference in delay between the section of the frequency spectrum occupied by the chrominance signal and the lower frequency section occupied by the luminance signal only, causes misregistration of the colour and luminance signals.

2.6 NOMENCLATURE. Most existing radio broadband bearers employ frequency modulation. This means that varying base band voltages are converted to varying radio frequencies.

The distortions which occur in the radio frequency circuits, and which vary with frequency, then appear to vary with amplitude from an overall system point of view. For example, the bearer may employ receivers with an IF of 70MHz. When the base-band voltage is OV (no modulation) the value of the IF is 70MHz, and when the base-band voltage is ± 1V the value of the IF is 80MHz or 60MHz. If the RF circuitry, which comprises filters of various kinds, causes a variation in group delay between 60 and 80 MHz of say 10ns, then the positive peak of a 5MHz baseband sinewave of 2V p-p will be transmitted 10ns earlier (or later) than the negative peak. Relative to the negative peak, the phase of the positive peak will be shifted by:

$$10/200 \times 360 = 18^\circ$$

The extent of the distortion will clearly be less if the amplitude of the 5MHz sinewave is less.

Clearly, we have here a distortion which, from a system or end-to-end point of view, is a non-linearity i.e., a distortion which varies with amplitude. However, the cause of the distortion is not non-linearity.

This phenomenon has given rise to the use of words and phrases which, in other areas, might mean other things. In order to clarify the meanings of such words and phases, the following explanations are given:

:: LINEARITY means the departure from linearity of the amplitude-frequency, frequency-amplitude, and amplitude-frequency-amplitude conversion process. It is affected by the linearity of modems and by the response of IF and RF circuits. For the method of measurement see later paragraphs

:: DIFF. GAIN refers to the apparent linearity when the measuring signal is a fairly high frequency (generally greater than 1MHz). In these circumstances other distortions can interfere with the measurement which, though not strictly then linearity, is still of significance for the portion of the baseband near the measuring frequency

NOTE: The Diff. Gain of TV tests corresponds fairly closely to Linearity if a plain staircase, or a ramp or staircase modulated with 1MHz is used. It corresponds to Diff. Gain if a staircase or ramp modulated with colour sub-carrier is used

:: GROUP DELAY is the variation in propagation time with variation in radio frequency (or baseband amplitude since there is a linear relationship between transmitted frequency and baseband amplitude in an FM system). It is affected by the phase response of RF and IF circuits

:: DIFF. PHASE converted to a phase shift at the measurement frequency, is the apparent group delay, when the measuring frequency is fairly high (generally greater than 1MHz). Though not really a measure of group delay, this measurement is of significance to baseband frequencies near to the test frequency

NOTE: The Diff. Phase of TV tests corresponds with Diff. Phase as defined above, when the test frequency corresponds to the TV colour sub-carrier. The Group Delay (or envelope delay) of long line tests is quite different from the Group Delay we are concerned with. In the case of the long line tests we measure the change in propagation time as the baseband frequency varies. This type of distortion is normally of little significance for broadband radio bearers as the number of baseband circuits and filters is so low.

2.7 NOISE. Bearer systems are designed to meet specified signal-to-noise ratios in the basic traffic channel. For a telephony bearer, the specification is mainly concerned with the signal-to-noise ratio in each telephone channel. For a television bearer, the signal-to-noise ratio is specified in the luminance and chrominance channels.

Noise in a broadband bearer results from:

:: Thermal noise in the equipment, particularly in low level stages of amplifiers at repeaters and terminals

:: Interfering signals due to overshoots or other bearers

:: 'Hum' introducted into amplifier circuits from power supplies and because of longitudinal current in cables carrying video signals. Hum is usually not a problem for a telephony bearer

:: Intermodulation between frequencies in the spectrum because of system non-linearity and echoes

Details of the noise requirements of a system and the origin and distribution of noise are included in later sections.

2.8 GAIN STABILITY. Variations in amplifier gain or in modem conversion gain will change the overall system gain. This will have to be corrected by manual means in the case of TV, or in the gain correction devices used in TF systems. Echoes can produce short term variations in gain in the higher channels (i.e. the variation appears as a change in the amplitude/frequency response).

2.9 FREQUENCY STABILITY. All IF and RF distortions which are frequency sensitive (i.e. group delay, filter amplitude response and distortions coupled to these) are corrected by means of cancellation. In some cases, the cancellation takes place at a point distant from the distortion e.g. group delay may be mopped up at the end of a modem section. The stability of such a cancellation depends on the frequency stability of local oscillators, in the intervening transmitters and receivers.

2.10 SIGNAL LEVELS. The amplitude of the signal at any point in a circuit associated with an FDM broadband carrier system is normally expressed in dBr; indicating the level in dB relative to the reference point in the system. Levels quoted in dBr are the relative levels using a test tone for one channel of the system only. When more than one channel is present at a point in the system, as occurs in a multi-channel FDM carrier system, the power level at the point is greater than the power level for an individual channel. With test tones applied to more than one channel, the power level at a point where these channels are combined would increase. In general terms, the level would be proportional to N, the number of channels. However, with normal telephone conversations, all channels are not operating at maximum levels at all times because:

:: Speech volume continuously varies

:: Mostly, only one of each two conversants speaks at a time

:: The customer does not speak continuously : there are punctuation pauses

:: A system is usually designed so that there are a number of channels in reserve even during 'busy hour' conditions

The result of these factors is a reduction in the multichannel power level from the theoretical (0 + 10 log N) dBmO where the test tone level is 0 dBm. Other signals which increase the level at points common to all channels of a multichannel system are pilot and signalling tones.

During busy-hour periods the baseband spectrum of a multichannel microwave system contains so many signals that it takes on the appearance of white noise. Because of this, the busy-hour of a broadband system can be simulated by loading the system with white (random) noise of the proper amplitude and bandwidth.

The term 'white noise' is derived from white light which has a uniform distribution of energy across the visible spectrum. White noise is a random noise band where all frequencies of the band appear simultaneously and with the same energy density. Internationally accepted practice is to simulate the conventional loading level in a multi-channel baseband with white noise of mean power as given by the following equations:

Mean white noise power = $-15 + 10 \log N$ dBmO for $N \geq 240$ channels
Mean white noise power = $-1 + 4 \log N$ dBmO for $12 < N < 240$ channels

(where N = number of channels in the system).

Mean white noise power levels for multichannel systems of various capacities are shown in Table 1.

Number of Channels	Mean White Noise Power Level of System During 'Busy Hour' (dBm0)	Frequency Band Occupied (kHz)
60	6.1	60 - 300
120	7.3	60 - 552
240	8.8	60 - 1,052
300	9.8	60 - 1,300
600	12.8	60 - 2,540
900	14.6	312 - 4,028
		316 - 4,188
960	14.8	60 - 4,028
1,200	15.8	312 - 5,564
1,260	16.0	60 - 5,564
1,800	17.6	312 - 8,204
		316 - 8,204
2,700	19.3	312 - 12,388
		316 - 12,388

TABLE 1. F.D.M. BROADBAND CARRIER SYSTEM MEAN WHITE NOISE POWER LEVELS AND FREQUENCY BANDS

The values in Table 1 correspond to the mean power output of the multichannel carrier system during the busy hour relative to the single channel test tone level at any point in the system. As an example, the mean white noise power level of a 960 channel system during the busy hour is expected to be 14.8 dB above the single channel test tone level at the measuring point. The waveform representing the output of a multi-channel FDM carrier system normally has voltage peaks approximately 12 dB above its RMS value.

Broadband bearer circuits are tested using white noise signals, with bandwidths and levels corresponding to Table 1 to simulate the multichannel system during the busy hour.

Data signals, such as those originating from teletype machines, are often transmitted over the channels of multichannel carrier systems by firstly converting them into either AM or FM voice frequencies. Because data signals are not random in nature as are speech signals, they result in an almost constant power output on these channels.

With normal data signals, these levels cause the mean power output from broadband carrier systems to increase above the figures quoted in Table 1. This signal level increase must be allowed for when data signals occupy a significant section of the total frequency spectrum.

2.11 TESTING TECHNIQUES. The testing techniques used for broadband bearer system overall performance measurements are designed to critically examine the characteristics important to the traffic to be carried. Different tests are normally suitable for telephony and television links, though both types of tests can be applied to each type of bearer if suitable test points are available.

For telephony bearers, testing for adequate linearity is done by using a white noise test signal to simulate the normal baseband signal and then measuring the intermodulation noise produced in telephone channels at the bottom, centre and top of the baseband spectrum. At the same time, basic noise can be measured in a channel when the white noise is removed. The amplitude-frequency characteristic is normally measured at spot frequencies in the spectrum or by using swept-frequency techniques.

Since the quality of television pictures depends largely on the maintenance of the original waveform, testing of television bearers is normally carried out using waveform testing techniques. Linearity tests normally use sawtooth test signals and modulated sawtooth test signals. Both the amplitude-frequency and phase-frequency characteristics of the bearer are examined using sine squared pulse and bar test signals, 50 Hz square wave test signals and a square wave test signal of approximately 1 Hz. Basic noise and other interfering signals are normally measured using a CRO.

Testing techniques are not examined in this paper.

3. CHARACTERISTICS OF FM RADIO BEARERS.

3.1 FM SIGNAL DEVIATION. Radio broadband bearer systems use frequency modulation (FM). In an FM system the amplitude of the carrier signal is constant and the carrier frequency is shifted above and below the mean in proportion to the instantaneous amplitude of the modulating signal. Assuming a simple sine wave modulating signal, the frequency spectrum of an FM signal contains a number of sideband components on both sides of the carrier at frequency intervals equal to the modulating frequency.

Theoretically, an infinite number of pairs of sideband frequencies is produced but in practice, only those sideband components with an amplitude greater than 1% (-40 dB) of the unmodulated carrier amplitude are deemed to be significant. The number of significant sidebands depends on the amplitude and frequency of the modulating signal. For example, when the amplitude of a constant frequency modulating signal is increased, the carrier frequency deviation is increased and the number of significant sidebands increases (Fig. 2).

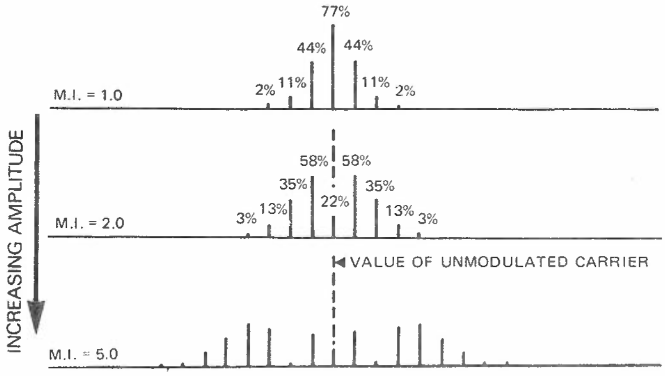


FIG. 2. FREQUENCY MODULATION SIDEBANDS-
EFFECTS OF INCREASING MODULATING SIGNAL AMPLITUDE

When the frequency of a constant amplitude modulating signal is increased, the deviation remains constant but the number of significant sidebands is reduced and the bandwidth is increased (Fig. 3).

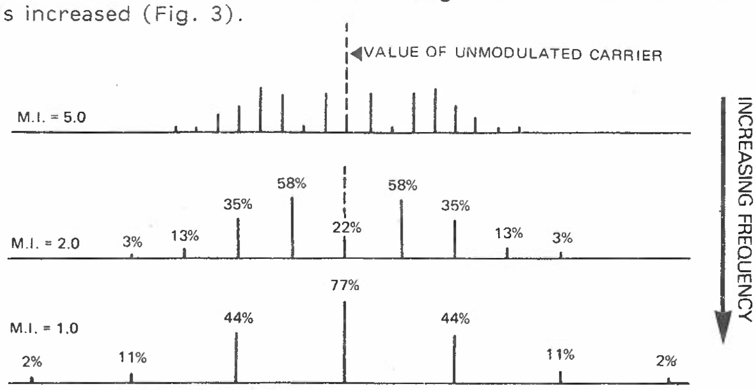


FIG. 3. FREQUENCY MODULATION SIDEBANDS-
EFFECTS OF INCREASING MODULATING FREQUENCY

The bandwidth required by a frequency modulated carrier depends on both the amplitude and frequency of the modulating signal, as does the number of significant sidebands, for all significant sidebands must be accommodated within the bandwidth. The number of significant sideband components is related to the ratio of carrier frequency deviation to the modulating frequency. This ratio is known as the Modulation Index:

$$\text{Modulation Index (M.I.)} = \frac{\Delta fc}{fm}$$

where Δfc = maximum carrier frequency deviation
fm = modulating frequency.

With a complex modulating signal containing many frequency components, the number of frequency components in the spectrum of the FM waveform is large. However, the frequency band occupied by the components does not increase. Normally, the bandwidth required for a particular deviation with a complex modulating signal is less than that for the same deviation with a single sine wave modulating signal. This is because the modulation index for each component of the complex wave is much lower than the modulation index for the single sine wave producing the same deviation. An analysis using a white noise modulating signal to simulate a typical telephony baseband signal indicates that the bandwidth for a frequency modulated complex signal is approximately:

$$\text{F M Signal Bandwidth (B)} \cong 2(\Delta F + fm) \quad \text{(Carson's rule)}$$

where ΔF = Peak deviation.
fm = Maximum modulating frequency.

3.2 DEVIATION SPECIFICATIONS. The basic definition of deviation refers to the peak shift of the carrier frequency at the peak amplitude of the modulating signal. Because the telephony baseband signal is of a random nature, it is more convenient to specify deviation in terms of an RMS value. For a sine wave modulating signal, the peak deviation is $\sqrt{2}$ times the RMS deviation. The deviation produced by a telephony baseband signal is quoted on a per channel basis as the RMS deviation produced by a test tone of 1 mW at the zero relative level point in the system; that is, OdBmO per channel. The standard deviations for systems of various capacities are listed in Table 2.

Number of Channels	RMS Deviation per Channel (kHz) for 1 mW at a point of Zero Relative Level.
60 to 120	50, 100 or 200 (alternatives)
120 to 960	200
960 to 1,260	140 or 200 (alternatives)
1,260 to 2,700	140

TABLE 2. TELEPHONY BEARER DEVIATIONS

3.3 TELEPHONY BEARER BANDWIDTH. A knowledge of the baseband signal and the deviation specifications allows the bandwidth required for the FM signal to be calculated. Assume a 960 channel system adjusted for 200 kHz RMS deviation per channel. From Table 1, the baseband signal level during busy hour is 14.8 dB above the single channel test tone level. Since deviation is proportional to signal voltage, the RMS deviation for the total baseband signal is greater than for a single channel in proportion to the voltage ratio corresponding to 14.8 dB.

That is:

RMS deviation for 960 channel baseband signal $= 200 \times \text{antilog}\ \frac{14.8}{20}$ kHz

$= 200 \times 5.5$ kHz

$= 1.1$ MHz

The peak value of a telephony baseband signal is approximately 12 dB above its RMS value. This means that the peak deviation exceeds the RMS deviation in proportion to the voltage ratio corresponding to 12 dB. That is:

Peak deviation for 960 channel baseband signal $= 1.1 \times \text{antilog}\ \frac{12}{20}$ MHz

$= 1.1 \times 4$ MHz

$= 4.4$ MHz.

The maximum baseband frequency for a 960 channel system is 4.028 MHz.

Using the formula in para. 3.1:

FM Signal Bandwidth (B) $\cong 2(\Delta f + fm)$

B (960 channels) $= 2\ (4.4 + 4.028)$ MHz

$= 16.856$ MHz

The bandwidths required for systems carrying between 60 and 2,700 telephony channels are shown in Fig. 4. The curves are designated to show the RMS frequency deviation per telephony channel.

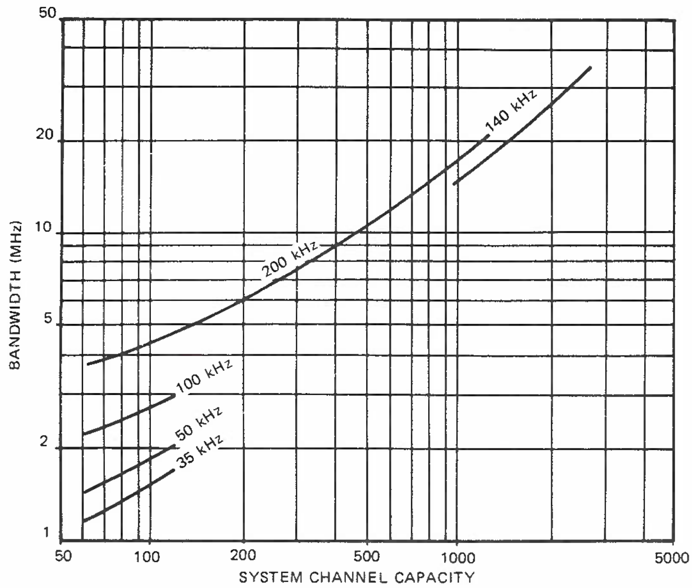


FIG. 4. BANDWIDTH REQUIREMENTS OF FM TELEPHONY BEARERS

3.4 TELEVISION BEARER BANDWIDTH. For a television bearer, the deviation is specified as 8 MHz peak-to-peak for 1V peak-to-peak modulating signal at the reference point in the system. The resultant TV bearer bandwidth is approximately the same as that of a 960 channel telephony bearer.

The sound channel for a television bearer is normally provided by using a 7.5 MHz subcarrier. The subcarrier is frequency modulated by the sound programme information and then frequency division multiplexed with the television video signal to form a combined baseband signal which frequency modulates the RF carrier. The primary deviation of the 7.5 MHz sound programme subcarrier is 140 kHz RMS for a +16 dBm signal at the reference point in the system. The peak deviation for this test signal is, therefore, 200 kHz. For a 7.5 MHz subcarrier sound channel, the maximum audio frequency is 15 kHz. The deviation ratio (maximum deviation ÷ maximum modulating frequency) is 13.3 which results in approximately 17 pairs of significant sideband frequencies. This means that the bandwidth required by the modulated subcarrier is approximately 510 kHz.

The secondary deviation of the RF carrier by the unmodulated 7.5 MHz subcarrier is 300 kHz RMS. The resultant frequency modulated signal is accommodated within the RF bandwidth available for the television bearer.

3.5 AMPLITUDE-FREQUENCY RESPONSE OF FM TRANSMISSION EQUIPMENT. It is often thought that, because an FM receiver includes a limiter to remove amplitude variations, the amplitude-frequency response of the RF and IF circuits of an FM system is not particularly important. This appears to be justified because the amplitude-frequency response of an IF amplifier operated in the limiting condition appears to depend only on the response of the circuits following the limiter. However, the amplitude-frequency response of the RF and IF circuits is important for two reasons:

:: The relative amplitudes of the carrier and sideband components depend on the amplitude-frequency characteristics of all the circuits; subsequent amplitude limiting cannot correct the FM sidebands.

:: Variations from the normal indicate that amplitude variations of the modulated carrier envelope have occurred. Consequently, any subsequent conversion of the AM to FM (such as occurs in non-linear reactive components) will result in distortion. This aspect is examined in para. 5.7.

3.6 PHASE-FREQUENCY RESPONSE OF FM TRANSMISSION EQUIPMENT. Complex signals are formed by the algebraic sum of a series of discrete frequency sine waves. For undistorted transmission of such a signal, the amplitude ratios and relative phase relationships of the component sine waves must remain unchanged. Also the transmission time delay for all component frequencies must be the same; that is, the phase shift must be proportional to frequency over the operating frequency band. Lines A and B in Fig. 5 represent ideal characteristics with each giving a different transmission time delay or phase delay. The phase delay at any frequency can be calculated from the phase-frequency characteristic:

$$\text{Phase Delay} = \frac{\text{Phase shift}}{\text{Frequency}} = \frac{\varnothing}{f} \text{ seconds}$$

where $\varnothing$ is in radians, f is in radians/second OR $\varnothing$ is in parts of a cycle, i.e. $\frac{\text{degrees}}{360}$

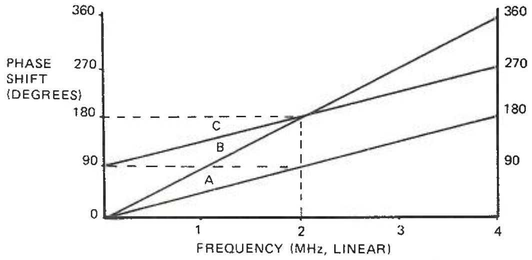
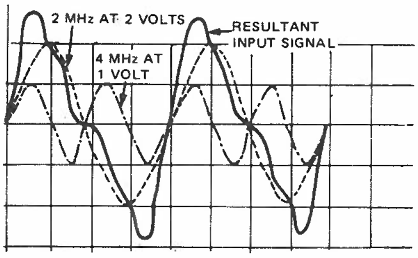


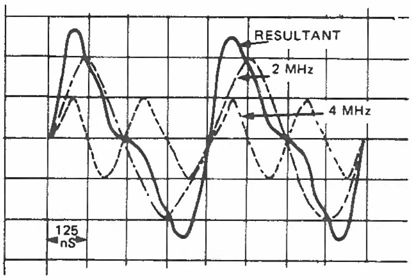
FIG. 5. PHASE-FREQUENCY CHARACTERISTICS

To examine the effect of phase delay on a signal, let us compare three systems (A, B and C) with flat amplitude-frequency characteristics but with phase-frequency characteristics as shown in Fig. 5. Assume that the input signal to the systems is a complex signal consisting of two frequencies, 2 MHz with an amplitude of 2 volts and 4 MHz with an amplitude of 1 volt. The resultant input signal to the three systems is shown in Fig. 6a.



In System A, the phase shift at 2 MHz is 90° from Fig. 5, and the phase delay is 0.125 μs or 125 nanoseconds (ns) by calculation. The phase shift at 4 MHz is 180° and the phase delay is again, 125 ns.

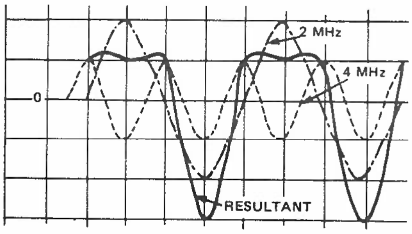
(a)



The resultant output signal is shown in Fig. 6b. This shows that the output signal is a replica of the input signal, but is delayed by 125 ns from the input signal.

In system B, the phase shift at 2 MHz and 4 MHz is 180° and 360° respectively and the phase delay at both these frequencies is 250 ns. The output signal (not shown) would be a replica of the input but delayed by 250 ns.

(b)



In system C, the phase shift at 2 MHz is 180° and the phase delay is 250 ns. At 4 MHz, the phase shift is 270° and the phase delay is 187.5 ns. This means that the two frequencies are delayed by different times and the output waveform (Fig. 6c) is not the same as the input waveform.

(c)

FIG. 6. PHASE DELAY EFFECTS

3.7 GROUP DELAY. In practice, measurement of the slope at numerous points on the phase-frequency characteristic gives a sensitive indication of its linearity. The slope of the phase-frequency characteristic at any point is the GROUP DELAY (or envelope delay) of the system at that frequency. Group delay is defined as the derivative of the phase-frequency response:

$$\text{Group Delay } (\tau g) = \frac{\text{small change in phase}}{\text{small change in angular frequency}} = \frac{d\emptyset}{df}$$

For each of the characteristics shown in Fig. 5, the slope is constant at all frequencies so the group delay for each is constant over the frequency range. The group delay for A is 125 ns, for B it is 250 ns and for C the group delay is 125 ns.

3.8 COMPARISION OF PHASE DELAY AND GROUP DELAY. A phase-frequency characteristic which differs from the ideal is illustrated in Fig. 7a. For convenience, abrupt changes in the slope of the characteristic are shown, though more gradual changes would be introduced by normal equipment. At frequencies below f1 the phase shift is proportional to frequency and this section of the characteristic is ideal.

Both the phase delay and the group delay for frequencies up to f1 are constant as indicated by Figs. 7b and c. Between f1 and f2, the phase shift characteristic is not proportional to the frequency only but has a constant phase shift added and the phase delay decreases. However, over this frequency range the slope of the phase shift-frequency characteristic is constant but lower than that below f1 and the group delay between f1 and f2 has a constant but low value. The mathematic expressions are as follows:

:: For the frequency range 0 to f1

Phase shift = af (proportional to freq.)

Phase delay = $\frac{\text{phase shift}}{\text{freq.}} = \frac{af}{f}$ = a (constant with frequency)

Group delay = $\frac{d}{af}$ (phase shift) = a (constant with frequency)

:: For the frequency range f1 to f2

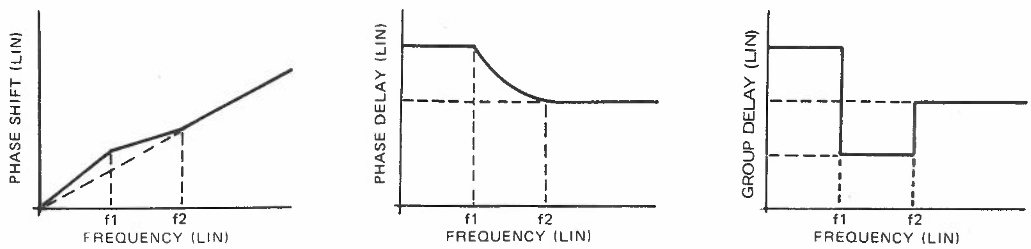
Phase shift = af + c (proportional to frequency with a constant phase shift added)

Phase delay = $\frac{af + c}{f} = a + \frac{c}{f}$ (inversely proportional to frequency with a constant phase delay added)

Group delay = $\frac{d}{af}$ (af + c) = a + 0 (constant with frequency)

where f is the angular frequency in radians/second. a and c are constants

Above f2 the characteristic in Fig. 7a is again proportional to frequency and would be ideal if frequency components were restricted to this section. The constant slope of this section results in constant intermediate values for both phase delay and group delay.



(a) Phase Shift (b) Phase Delay (c) Group Delay

FIG. 7. COMPARISON OF PHASE SHIFT, PHASE DELAY AND GROUP DELAY

A phase shift proportional to frequency, giving a constant phase delay, is a necessity for pulse circuits such as those used for television video signals. However, for the transmission of a modulated carrier signal, it is necessary only that the group delay is constant over the frequency range. A circuit with a phase-frequency characteristic as in Fig. 7a, which has a constant group delay between f1 and f2, can be used to transmit a modulated carrier whose frequency spectrum is restricted to the frequency range between f1 and f2. Even though the phase shift is not directly proportional to frequency and the phase delay is not constant over this range, no degradation of the signal information impressed on the carrier will be produced.

3.9 PRACTICAL GROUP DELAY CHARACTERISTICS. RF and IF circuits of transmitters and receivers normally have a phase-frequency characteristic considerably different from that shown in Fig. 7a. The general form of the characteristic of tuned amplifiers is illustrated by the full line curve in Fig. 8b which represents the phase-frequency characteristic of an amplifier with a single tuned circuit. The amplitude-frequency response for the same circuit is shown in Fig. 8a.

Relative to the zero reference phase shift at the resonant frequency of the tuned circuit, the output voltage leads and lags the input voltage at lower and higher frequencies respectively. The group delay-frequency characteristic corresponding to Fig. 8b is illustrated by Fig. 8c. It shows an approximately constant group delay for a limited frequency range on either side of the resonant frequency. For a single resonant circuit, the maximum variation in phase is ±90° but greater numbers of resonant circuits will produce proportionally greater amounts of maximum phase shift and also inflections in the characteristic. The broken line curves in Fig. 8 illustrate the characteristics of a circuit including an overcoupled double-tuned resonant circuit. By suitable design, bandpass amplifiers with a constant group delay over the passband can be produced and these will give distortionless transmission of a modulated carrier.

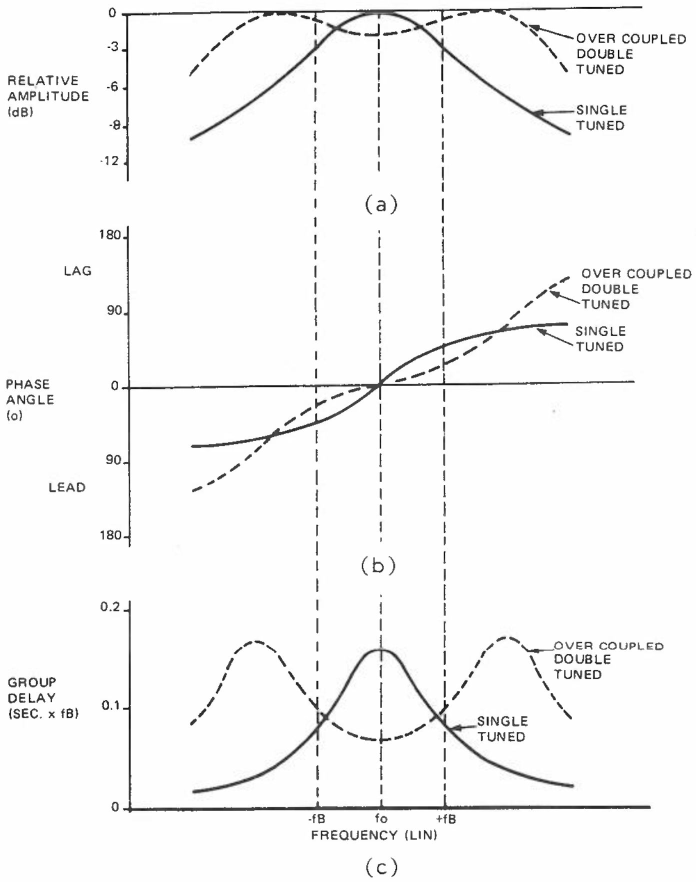


FIG. 8. CHARACTERISTICS OF TUNED AMPLIFIER

3.10 EFFECT OF CONSTANT GROUP DELAY. The effect of a phase-frequency characteristic which is not proportional to frequency but gives a constant group delay, is most readily shown by using an amplitude modulated signal as an example. The same conclusions apply to frequency and phase modulated signals. Assume that a circuit has a phase-frequency characteristic as illustrated by Fig. 9a. This has a phase shift that is proportional to the frequency difference from the centre frequency where the phase shift is zero, and gives a constant group delay-frequency characteristic over the frequency range of interest (Fig. 9b). The magnitude of the group delay is:

$$\tau_g = \frac{d\emptyset}{df} \text{ secs}$$

$$= \frac{18 \times 10^6}{360 \times 10^3} = 50 \ \mu s$$

Consider that the input signal to the circuit is a 10 kHz carrier frequency which is 100% amplitude modulated by a 1 kHz sine wave. The AM signal can be thought of as consisting of the carrier frequency (10 kHz) and two sideband frequencies at 9 kHz and 11 kHz, each with amplitudes equal to half the carrier amplitude (Fig. 9c). The input AM signal has a sine wave shaped envelope as shown in Fig. 9d.

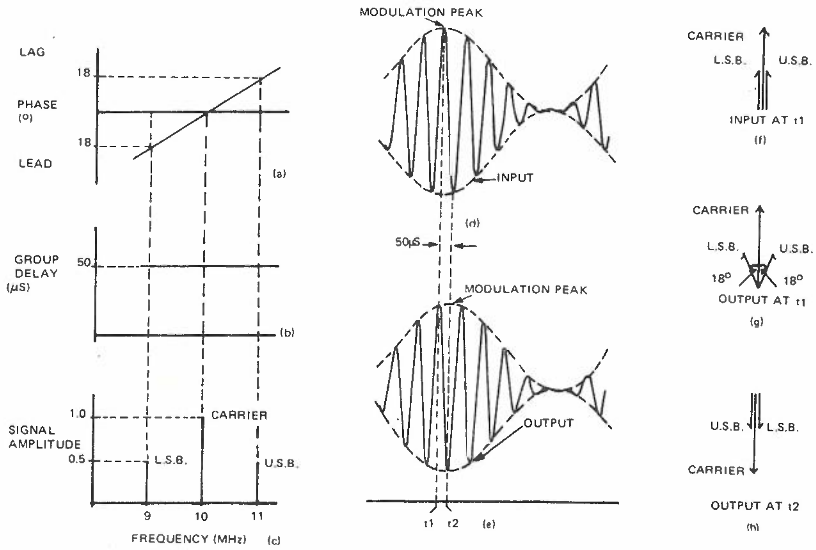


FIG. 9. EFFECT OF CONSTANT GROUP DELAY - FREQUENCY CHARACTERISTIC

The frequency components of the signal can be represented by phasors. At the peak of the envelope at time t1 the phasors will be in phase as illustrated by Fig. 9f. At the output at this same time the carrier signal has not changed its phase but the lower sideband frequency has advanced by 18° and the upper sideband frequency has retarded by 18° as shown in Fig. 9g. The addition of the phasors now gives less than the maximum envelope amplitude at time t1.

Since the group delay is 18° of the modulation cycle (50 μs), it is expected that the modulation envelope will be delayed by this amount. To prove that this is so, we will examine the phasors which represent frequency components at time t2 which is 50 μs after t1.

In 50 μs the 10 kHz carrier has advanced by:

$$\frac{50 \times 10^4 \times 360}{10^6} = 180 \text{ degrees,}$$

the lower sideband has advanced by:

$$\frac{50 \times 9 \times 10^3 \times 360}{10^6} = 162 \text{ degrees.}$$

and the upper sideband has advanced by:

$$\frac{50 \times 11 \times 10^3 \times 360}{10^6} = 198 \text{ degrees.}$$

The phasors for the frequency components of the output signal at time t2 are shown in Fig. 9h. Notice that all the phasors are again in phase, though the resulting maximum envelope amplitude now occurs with the carrier in a different phase.

In summary, the modulation envelope is delayed in accordance with the group delay and the carrier cycles are delayed in accordance with the phase delay at the carrier frequency. (In this example the delay at the carrier frequency is zero). The resultant envelope is shown in Fig. 9e. The important observation is that, with a constant group delay, the shape of the modulation envelope containing the signal information has not changed. The magnitude of the group delay, and the resultant modulation envelope delay, is not important. For an FM signal the resultant deviation is delayed in accordance with the group delay and the mean carrier frequency in accordance with the phase delay at this frequency but, with a constant group delay at all frequencies in the spectrum, no distortion of the deviation is produced.

3.11 EFFECT OF FM EQUIPMENT ON BASEBAND FREQUENCY RESPONSE. In a system where the deviation ratio (maximum deviation ' maximum modulating frequency) is low, as in a radio broadband bearer system, the amplitude-frequency and phase-frequency characteristics of the FM transmission circuits affect the amplitude-frequency and phase-frequency characteristics of the derived baseband circuit in a complex manner. These characteristics are particularly important for a television bearer.

The results obtained from an FM system with amplitude and phase errors depend on whether the characteristic's errors are symmetrical on both sides of the carrier frequency or skew-symmetrical giving a tilt in the characteristic over the passband so that upper and lower sidebands are affected in opposite ways. An idea of the effects can be obtained with some simple examples. Consider a system in which only the first sideband components are signficant. When the amplitude-frequency response of the FM transmission equipment reduces the amplitudes of these sideband components by equal amounts on each side of the carrier frequency, the resultant waveform has its deviation reduced accordingly. The signal obtained after demodulation is also reduced in amplitude. This results in a baseband signal amplitude-frequency response with approximately the same shape as the amplitude-frequency response of the FM equipment on either side of the carrier frequency. When the FM equipment amplitude-frequency response has a tilt over the passband, with one sideband being reduced and the other sideband being increased, the baseband amplitude-frequency response is not greatly affected. Typically, amplitude-frequency response variations over the bandwidth required (i.e. about 17 MHz for a 960 channel bearer) are less than ±0.2 dB.

If the group delay varies in a symmetrical manner over the passband, giving a typical practical characteristic of parabolic shape, the sideband frequencies for a particular modulating frequency are affected equally, but the delay of the modulation information varies with frequency. This results in a baseband group delay-frequency characteristic with approximately the same form as the FM equipment characteristic on each side of the carrier. The baseband group delay-frequency characteristic is not greatly affected by FM equipment with a group delay having a linear tilt over the passband. However, this form of group delay characteristic does affect the amplitude-frequency response of the baseband circuit.

In any practical case, the magnitude of these effects is likely to be quite small as the uncorrected group delay is not likely to exceed a few nanoseconds. Group delay errors also produce system non-linearity which results in intermodulation. This is examined in Section 5.

4. RECEIVER NOISE.

4.1 NOISE IN AN FM SYSTEM. The noise performance of a radio broadband bearer is a very important characteristic and is a major design factor, particularly when the bearer is to be used for telephony. To obtain an understanding of the origin of noise and the effects of different parameters and sections of equipment on system noise, some of the design factors of radio broadband bearer systems are examined in the following paragraphs. An understanding of these factors is necessary if measurements on bearer systems are to be correctly interpreted.

Noise at the output of a radio telephony bearer can be divided into three broad categories:

:: BASIC NOISE is that which is present at all times, even when the system is idle. Thermal noise at the input of each receiver and that introduced by the equipment, particularly low signal level stages, forms the main component of basic noise. Additional sources of basic noise are spurious outputs from transmitter and receiver local oscillators and random variations of local oscillator frequencies. This latter source may vary with the amount of traffic that the interfering systems are carrying

:: INTERFERENCE may arise from within a system as a result of finite front-to-back ratios of antennas, from overshoot, or may come from other systems. The level of interference will vary due to atmospheric conditions varying propagation factors

:: INTERMODULATION NOISE. This type of noise is produced by intermodulation between component frequencies of the signal. Intermodulation noise results from system non-linearities such as non-linearity of the modulator and demodulator characteristics of an FM system, errors in the amplitude and phase characteristics of the FM transmission equipment and reflections in antenna feeders and in propagation paths

In a television bearer, noise and interference results from:

:: Basic noise, as in a telephony bearer

:: 'Hum' introduced from power supplies

:: Interfering signals from antenna feeder or propagation path reflections which, in an FM system, usually cause waveform degradations. Such echoes have negligible effect on TV systems

4.2 ADDITION OF NOISE CONTRIBUTIONS. When noise is random in nature, as is the case with thermal noise, the total noise produced is found by adding the individual noise powers. If noise originates from similar sources so that individual contributions tend to be in phase, the noise is added on a voltage basis. Although some tendency towards voltage addition occurs in a radio broadband bearer, most of the noise contributions, including thermal and intermodulation noise, are sufficiently unrelated that they can be considered as random and may be added on a power basis. This approach is normally used in all but the most detailed examinations for it allows the noise power developed by each item of equipment at each station on the route to be added. This enables the total noise power for a system to be estimated to see if it satisfies the design requirements for the signal-to-noise ratio of the channels. Alternately, the total noise corresponding to the required output signal-to-noise ratio can be distributed to allocate a proportion to each item of equipment, keeping in mind the performance that can be practicably and economically achieved at that time.

4.3 RECEIVER INPUT SIGNAL. The effect of noise introduced by the early stages of each receiver in a bearer system depends on the signal strength of the received carrier and the effective noise power of the receiver input. The factors affecting the receiver input signal power are:

:: Transmitter power output

:: Propagation path attenuation

:: Fading

:: Antenna gains

:: Feeder attenuation

:: Attenuation in RF multiplexing equipment and filters

These factors are discussed in the subsequent paragraphs.

4.4 TRANSMITTER POWER OUTPUT. The transmitter output power of a radio broadband system depends on the system design. If no output amplifier stage is included, as may be the case on a short path, the maximum power output of current equipment typically varies between 2 W at 2 GHz and 300 mW at 7GHz. When a travelling wave tube amplifier is included, the maximum power output is typically 20 W to 2 W over the same frequency range.

To simplify calculations of signal-to-noise ratio, the power output is often quoted in dB relative to 1 W (dBW):

$$P_T = 10 \log W_T$$

where P_T = Transmitter power output in dBW

W_T = Transmitter power output in watts

A typical 4 GHz transmitter with a travelling wave tube output amplifier has a power output of 5 W. The power output in dBW is:

$$\begin{aligned} P_T &= 10 \log 5 \\ &= +7 \text{ dBW} \end{aligned}$$

This power output is then fed to the antenna to be radiated.

4.5 ANTENNA GAIN. The gain of an antenna system is often specified relative to an isotropic antenna. Antenna systems for use in the upper UHF and SHF bands usually obtain high gain and directivity by using large area reflecting and focusing surfaces.

The gain of such an antenna system is:

$$G = \eta \frac{4 \pi A}{\lambda^2}$$

where G = Gain relative to isotropic antenna expressed as a ratio

A = Area of reflecting surface

λ = Wavelength of signal

η = Aperture efficiency

The aperture efficiency (η) is a factor dependent on how evenly the field from the radiating source is distributed over the reflecting surface. For a parabolic antenna, η is normally 55% and for a horn reflecting antenna, typically 67%. This means that the gain of a 3 metre diameter paraboloid reflector antenna at 4 GHz

($\lambda = \frac{300}{4,000} = 0.75$ metres) is:

$$\begin{aligned} G &= \eta \frac{4\pi A}{\lambda^2} \\ &= \frac{0.55 \times 4 \times \pi \times 1.5 \times 1.5 \times \pi}{0.75 \times 0.75} \\ &= 8685 \end{aligned}$$

The gain relative to an isotropic antenna (in dBi) = 10 log 8685
= 39.4 dBi

4.6 PROPAGATION PATH ATTENUATION. The propagation characteristics of signals in the UHF and SHF bands are examined in other papers but path attenuation is discussed here to obtain an understanding of the magnitude of the received signal and its effect on system noise.

The reference to which all variations in path attenuation (or path loss) are compared is known as the FREE SPACE LOSS, which is the loss that would be present if the path were located in space, remote from the earth and its effects. For a line-of-sight radio system with path clearance greater than 0.6 of the first Fresnel zone radius and no major path reflections, the path attenuation under no-fade conditions is approximately equal to the free space loss.

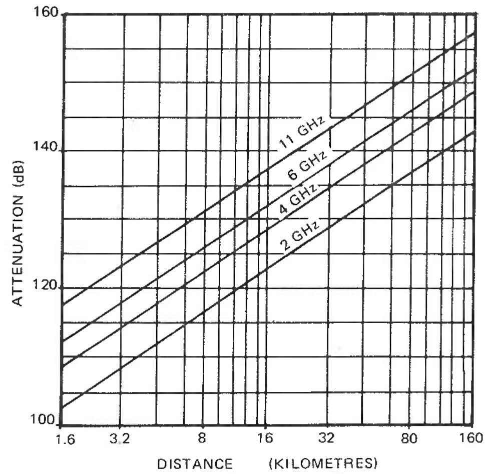


FIG. 10. FREE-SPACE ATTENUATION BETWEEN ISOTROPIC ANTENNAE.

Calculations of path attenuation assume a comparison with an ISOTROPIC antenna. An isotropic antenna is a theoretical, point-source antenna, radiating equally in all directions. With such an antenna, the transmitted power at a distance is reduced as if it were distributed over the area of a sphere with a radius equal to the distance from the transmitting antenna. The ratio of transmitted power to received power is given by:

$$\frac{P_T}{P_R} = \left(\frac{4\pi D}{\lambda}\right)^2$$

$$= \left(\frac{4\pi Df}{V}\right)^2$$

where

P_T = Transmitted power

P_R = Received power

D = Distance between transmitter and receiver

λ = Wavelength of signal

V = Velocity of propagation

f = Frequency in Hertz

From these formulae we can see that the path attenuation ratio, P_T/P_R, is proportional to the square of the distance and to the square of the operating frequency.

When path attenuation is required in dB, the formula can be expressed in logarithmic form as:

$$FSL_i = 32.4 + 20 \log f + 20 \log D$$

where FSL_i = Free space loss between isotropic antennae (in dB)

f = Frequency in MHz

D = Distance in km

It can be shown from the formula that the free space loss between two isotropic antennae is 32.4 dB for the first kilometre when the frequency is 1 MHz and increases by 6 dB each time either the frequency or the distance is doubled. For example, the path attenuation between two isotropic antennae for a 4 GHz signal over a 40 km path is:

$$FSL_i = 32.4 + 20 \log 4000 + 20 \log 40$$

$$= 32.4 + 72.04 + 32.04$$

$$\cong 136.5 \text{ dB}$$

Fig. 10 shows the variation in free space loss between isotropic antennae for several frequency bands used for radio broadband systems.

4.7 FADING. When a signal is received over two or more propagation paths, fading can be produced by the signals arriving with relative phase differences. At some instants the signals may be in-phase and the resultant signal will have a greater than free-space level. At other instants, the signals may be in opposite phase and the resultant level is less than the free-space level. The result is that the signal at the receiver varies in level and this variation is called 'fading'. Fading is discussed more fully in other Engineering Training publications, but this paragraph briefly examines the magnitude of variation in received signal level that is produced by fading so that its effect on the resulting noise in a bearer system can be estimated.

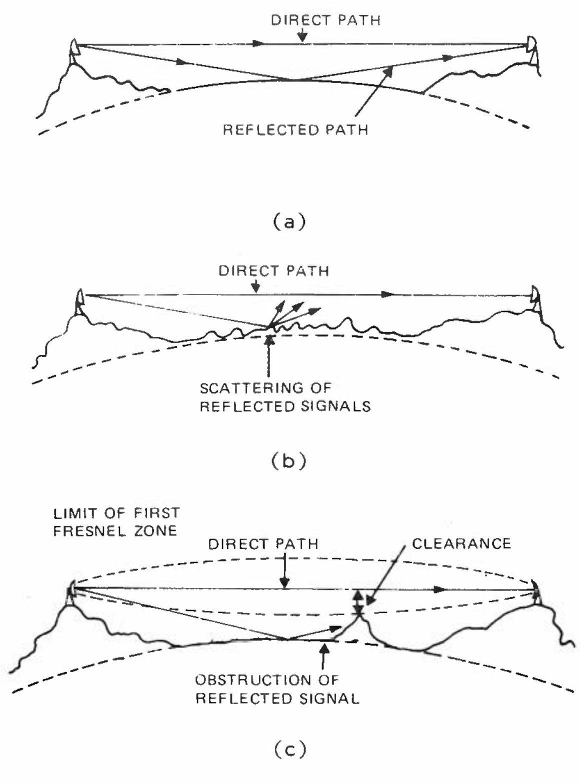


FIG. 11. PROPAGATION PATHS

A possible source of fading is produced by the simultaneous reception of the direct signal and a signal reflected from the earth's surface (Fig. 11a). When the area between the transmitter and receiver is flat country or water, a large amplitude reflected signal is received and large amplitude variations of the resultant signal occur with small changes in path propagation conditions. Such a path normally produces slow fading that follows a regular pattern; for example, variations in humidity, temperature, time of day or tide height. Variations in signal strength of 40 dB can be expected. When this type of path cannot be avoided, space diversity reception is normally used and this almost eliminates fading caused by ground reflections. Where possible, a path with strong mid-path reflections is avoided. This can be done by selecting a path with rough terrain between stations to scatter the reflections or with a mountain range running across the path so that mountains intercept major ground reflected signals. These paths are illustrated in Figs. 11b and 11c.

Fading may also occur as a result of changes in the refractive index of the atmosphere in various ways.

If the variations are stratified we may experience a bending of the signal path toward the earth so that the signal power is intercepted before it reaches the antenna, or it may be refracted away from the earth.

Stratification can also lead to 'ducts' which may cause focussing of the signal leading to a fade (if the signal is not focussed on the receiving antenna) or to a great increase in level if it is focussed at the receiving antenna. Ducts may also trap the signal in a region different from that in which the receiving antenna is located, again leading to fading.

Multipath signals are often produced by reflection, refraction and ducting, resulting from rapid variations in the refractive index of the atmosphere with height. Many signals appear at the receiving antenna with random amplitudes and phases and cause severe fading. The fading is very rapid and may last for only a fraction of a second.

Severe fading during a bad fading period tends to follow a statistically predictable pattern and this type of fading is often referred to as 'Rayleigh' fading. Severe multipath fading is most likely to occur during the summer months at night, when the air is still and when the propagation path is long. Using statistical information it is possible to predict the percentage of time that a fade of a certain magnitude will occur. As an example, for 99% of the time the received signal power is expected to exceed a level which may be 18 dB below the median power (that is, the received signal power for 50% of the time). This allows calculations to predict the noise during all but a small percentage of time to see if the short term noise requirements are satisfied.

Because of the random nature of fading, it is not usual for a large magnitude fade to occur on all hops of a multihop radio bearer system at the one time. To estimate the mean noise in any hour for a long system, it is often considered that 20% of the path is suffering Rayleigh fading and the rest only mild fading. An allowance for fading is added to the propagation path loss. Typically it is considered that the mean received power on all hops is reduced by 4 dB below the no-fade (free space) level. The likelihood of simultaneous fades occurring on many paths reduces as the length of the period being examined is reduced. When the noise during a small percentage of the time is being predicted, calculations are often made by assuming that one hop is suffering from a severe fade and that no fading is occurring on the other paths.

4.8 MISCELLANEOUS LOSSES. Waveguide is normally used for the feeder between the transmitting and receiving equipment and the antenna. At the frequencies used for radio broadband bearers, the attenuation of coaxial cables would be too high. The attenuation in waveguides depends on the frequency, material used and the cross-sectional shape. Practical waveguides may be rectangular, circular or elliptical. The attenuation at particular frequencies for copper waveguide of each type, with sizes suitable for the frequency bands used for bearer systems, is shown in Table 3. A system operating at 4 GHz typically has a 60 metre length of elliptical waveguide between the equipment and the antenna. From the table, this waveguide has an attenuation of approximately 1.7 dB.

FREQ. (GHz)	RECTANGULAR		CIRCULAR		ELLIPTICAL	
	Dimensions (internal-mm)	Atten. (dB/100m)	Diameter (ID-mm)	Atten. (dB/100m)	Axes (overall-mm)	Atten. (dB/100m)
2	109.22 x 54.61	1.08			144.3 x 77.5	1.14
4	58.17 x 29.09	2.78		1.6	73.7 x 47.2	2.59
6	40.39 x 20.19	4.98		1.3	49.8 x 32.3	4.95
	34.84 x 15.8	6.46	71.42			
7.5	34.84 x 15.8	5.77	(Siemens)	1.2	48.4 x 28.4	4.75
	28.5 x 12.62	8.913				
11	22.86 x 10.16	12.39			29.5 x 20.3	11.05
	19.05 x 9.53	14.8			(Andrew)	

TABLE 3. WAVEGUIDE ATTENUATION

Because of its low attenuation, circular waveguide is occasionally used for the mast section of the feeder to an antenna when the waveguide run is long. In these cases, rectangular or elliptical waveguide is used for the feed between the equipment and the bottom of the mast and a waveguide transducer is required. Each transducer typically has a loss of 0.1 dB. However, coaxial cable is most commonly used in the 2GHz band, whilst elliptical waveguide is most commonly used in the higher frequency bands. Additional losses in the feeder system between the equipment and the antenna are produced by the RF multiplexing equipment which allows several bearer systems to use a common antenna. The multiplexing equipment uses circulators, isolators and filters for combining and separating each RF channel. The loss depends on the number of systems sharing the one antenna. A loss of 2 dB is typical for a system with three RF channels.

4.9 RECEIVER INPUT NOISE POWER. In any equipment, noise is generated because of the random motion of thermally agitated free electrons in the resistance of the components. This noise depends on the temperature and resistance of the components and the bandwidth of the circuit. When the resistance is connected to a matched load, maximum transfer of the noise power occurs. In a matched circuit, the noise power due to the noise generated in the original resistance, and dissipated in the load resistance is calculated as follows:

Noise Power in Watts (W_N) = KTB

where - K = Boltzmann's constant,

(1.38×10^{-23} joules/° Kelvin),

T = Temperature in degrees Kelvin,

B = Bandwidth in Hz

Thermal noise is introduced by all items of equipment in a radio bearer system. However the most important sources of noise are those at the input to each receiver and generated in their source impedances. These noise sources are significant because of the low carrier level at these points. Each receiver amplifies both the carrier and the noise and, if no additional noise were added by the equipment, the same carrier-to-noise ratio would be maintained throughout the particular receiver.

For a line of sight broadband bearer system, the effective temperature of the antenna system, which is pointing horizontally at the lower atmosphere and the earth, is approximately equal to the ambient temperature. This is normally taken at 290°K (17°C, 62.6°F). A typical broadband radio receiver has an effective noise bandwidth (normally taken at the -3 dB points on the amplitude-frequency characteristic) of 40 MHz. (Compare this 'receiver noise bandwidth' with the ±0.2 dB requirement for the bandwidths in paras. 3.3 and 3.4).

In this example, the noise power input to the receiver is:

$$W_N = KTB$$

$$W_N = 1.38 \times 10^{-23} \times 290 \times 40 \times 10^6 \text{ watts}$$

which in logarithmic units is -128 dBW

As stated earlier, noise has a random nature, and when several noise sources influence a circuit, the total noise power is the sum of the individual noise powers. For this reason it is often convenient to express noise in linear units (for example, pW). However, for some purposes logarithmic units (for example, dB relative to 1 watt - dBW) are preferred and the noise power is calculated by either of the following methods:

Noise power in dBW (P_N) = 10 log W_N where W_N = Noise power in watts.

or P_N = 10 log KTB

Using the values in the preceding example:

$P_N = 10 \log W_N$	OR	$P_N = 10 \log KTB$
$= 10 \log (0.16 \times 10^{-12})$		$= 10 \times (\log k + \log T + \log B)$
$= 10 \times 13.2041$		$= 10 \times (23.1399 + 2.4624 + 7.6021)$
$= 10 \times (-12.7959)$		$= 10 \times (-12.7956)$
$= -128$ dBW		$= -128$ dBW

4.10 NOISE FIGURE. In practice, receivers are not ideal. Thermal noise is generated internally and this is added to the source noise. The noise at the receiver output is then greater than it would be at the output of an ideal noiseless receiver with the same gain, bandwidth and source noise at the input. The noise generated in a receiver can be specified by quoting the receiver noise figure. Noise figure is the ratio of the actual noise output of a receiver to the noise output of a similar but noiseless receiver, with the input correctly terminated in each case in a resistance at a standard ambient temperature of 290°K. Noise figure is normally expressed in dB.

$$\text{Noise figure } (F_N) = 10 \log \frac{\text{Actual receiver noise output}}{\text{Ideal receiver noise output}} \text{ dB}$$

The noise power at the output of an ideal receiver is the input noise power in the receiver bandwidth multiplied by the receiver power gain. In a practical receiver, the output noise power includes noise produced in the receiver. Consequently, the noise figure for a practical receiver is found from:

$$\text{Noise figure } (F_N) = 10 \log \frac{GW_N + W_M}{GW_N} \text{ dB}$$

Where W_N = Noise power (in Watts) in the receiver bandwidth due to source at 290°K

W_M = Noise power (in Watts) generated in the receiver

G = Matched power gain of the receiver

A small variation of the source temperature from the standard has little practical effect. For example, a 20°K change in temperature changes the noise figure by only 0.3 dB.

A large proportion of the noise generated in a receiver is developed in the receiver input stage which is a diode mixer in the majority of broadband radio receivers. The noise figure of a receiver increases slightly with frequency and depends on the receiver design. Typical values for receivers using diode mixers are between 6 dB and 9 dB over the frequency range from 2 GHz to 7 GHz.

The noise figure of a receiver can be improved by including a low-noise RF amplifier before the mixer. Such an amplifier is now practicable and economical and takes the form of either a negative resistance amplifier using a tunnel diode or a parametric amplifier using a varactor diode. A receiver with this arrangement may have a noise figure between 4 dB and 7 dB.

Since the noise generated in any resistance is dependent on its temperature, the noise figure of a receiver can be improved by cooling the early stages of the receiver. This technique is not economical for direct line-of-sight systems but is used for receivers of satellite repeater links.

When the noise figure of a receiver is known, it is convenient to consider that the receiver is ideal and that the source noise at the receiver input is increased by a factor equal to the noise figure. If the source noise is -128 dBW, as in the preceding example, and the receiver noise figure is 10 dB, the practical receiver is equivalent to an ideal receiver with the same gain and bandwidth but with an equivalent source noise (P_E) of -118 dBW.

4.11 RECEIVER INPUT SIGNAL POWER. During design of a bearer system, the input signal power to each receiver is calculated. As a typical example, the signal level for an average hop can be determined by using the figures for gain and loss quoted as examples in the preceding paragraphs. The figures under no-fade conditions are summarised in Table 4. The power levels at each point in the system have been determined from the gains and losses.

Item	Gain (+) or Loss (-) in dB	Power Level in dBW
Transmitter power output (5W) (P_T)	-	+7
Transmitter RF multiplexing loss	2.0 (-)	+5
Transmitter feeder loss (60 m elliptical waveguide)	1.7 (-)	+3.3
Transmitter antenna gain (3 m paraboloid reflector)	39.4 (+)	+42.7 *
Propagation attenuation (40 km free space)	136.5 (-)	-93.8
Receiver antenna gain (3 m paraboloid reflector)	39.4 (+)	-54.4
Receiver feeder loss (60 m rectangular waveguide)	1.7 (-)	-56.1
Receiver RF multiplexing loss	2.0 (-)	-58.1
Total loss between transmitter output and receiver input (L)	65.1 dB(-)	
Receiver signal level input		-58.1 dBW

*The power shown following the transmitter antenna gain is called the 'equivalent isotropically radiated power' (EIRP).

TABLE 4. TYPICAL BEARER SYSTEM LEVELS (4 GHz)

Table 4 shows that under no-fade conditions the signal level at the receiver input is -58.1 dBW and the total loss between the transmitter output and the receiver input is 65.1 dB.

When Rayleigh fading is experienced on the propagation path, the mean signal level is typically reduced by 4 dB. Assuming a fading allowance of 4 dB, the mean signal level at the receiver is expected to be -62.1 dBW. A graphical indication of the gains, losses and levels shown in Table 4 is illustrated by Fig. 12.

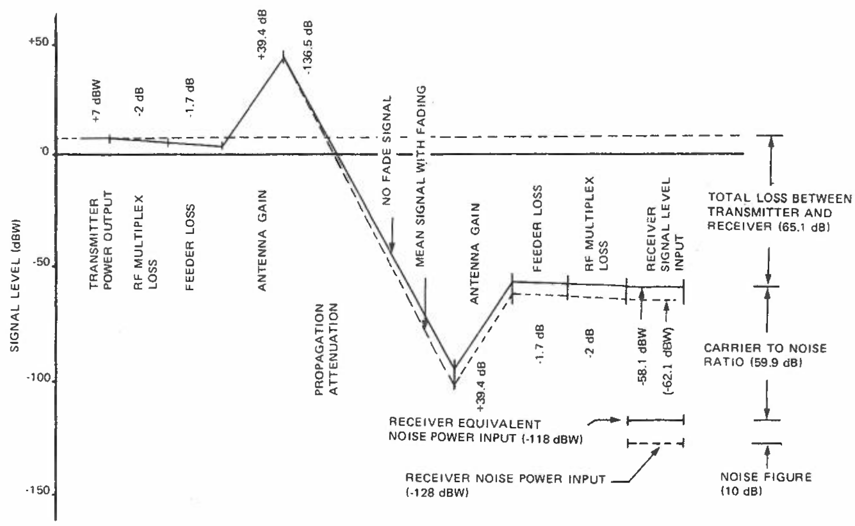


FIG. 12. TYPICAL BEARER SYSTEM CARRIER AND NOISE LEVELS

4.12 CARRIER-TO-NOISE RATIO. A knowledge of the received signal level, the source noise level at the receiver input and the receiver noise figure, allows the ratio of the carrier level to the noise level to be calculated. In the example given in para. 4.9, the source noise in the receiver bandwidth is -128 dBW, and the noise figure is 10 dB, so the receiver is equivalent to an ideal receiver with a noise input of -118 dBW. Therefore, with a received signal level of -58.1 dBW, the carrier-to-noise ratio (C/N) expressed in dB is:

$$\frac{C}{N} = P_T - L - P_E$$

$$= P_R - P_E$$

$$= -58.1 - (-118) \text{ dB}$$

$$= 59.9 \text{ dB}$$

where

P_T = Transmitter power output in dBW

L = Total loss (dB) between transmitter output and receiver input

P_E = Receiver equivalent source noise in dBW

P_R = Receiver input signal level (dBW)

Under fading conditions, the mean carrier-to-noise ratio is decreased by 4 dB to 55.9 dB. The receiver carrier signal and noise levels are shown in Fig. 12. The difference between these levels represents the carrier-to-noise ratio.

4.13 CHANNEL SIGNAL-TO-NOISE RATIO. The noise at the input to a receiver and that introduced by the receiver appears in the channels being provided by the bearer. For a television bearer it is the complete baseband noise that is of interest and for a telephony bearer it is the resultant noise that appears in each telephony channel. The channel signal-to-noise ratio characteristics of the FM system and the difference between the RF bandwidth and the channel bandwidth must be taken into account when determining channel signal-to-noise ratio. These factors are examined in Section 6.

5. EFFECTS OF NON-LINEARITY.

5.1 INTERMODULATION NOISE. When non-linearity is present in a bearer system, harmonic and intermodulation products of the desired baseband signal are produced. The distribution of these products over the baseband depends on the form and origin of the non-linearity. Assume that the non-linearity of the characteristic is such that second harmonic and second order intermodulation products are produced. With two input frequencies (fa and fb), the output would also include 2fa, 2fb, fa + fb and fa - fb. Depending on the frequencies of the signals, some of the resulting products could be outside the passband of the system and would not produce interference. For example, the second harmonic components (2fa and 2fb) would be in the passband only if fa and fb were less than half of the maximum baseband frequency. Also, for some combinations of frequencies, fa + fb would be above the baseband and fa - fb would be below the baseband.

When the non-linearity is such that third harmonic and third order intermodulation products are produced, many resultant frequencies, each involving three factors, are possible. Some examples are 3fa, 3fb, 3fc (third harmonics), 2fa + fb, 2fa - fb, fa + fb + fc, fa + fb - fc, fa - fb - fc. Many frequencies involving these and other combinations will fall in the baseband spectrum. With higher order non-linearities, the possibility of combinations producing frequencies in the baseband becomes even more complex.

So far we have considered only a few discrete signal frequencies. However, when the baseband signal has components evenly distributed over the complete baseband, similar to white noise, the number of possible combinations giving interfering signals in the baseband is almost limitless.

For a multichannel FDM telephony system, the multitude of products produced by non-linearity results in unintelligible crosstalk between channels and this appears as random noise interference in each telephony channel. Intermodulation products normally account for the majority of noise produced, and we will refer to the noise due to non-linearity as intermodulation noise.

5.2 TELEVISION SYSTEM NON-LINEARITY. It was stated in Section 2.2 that for a television system, non-linearity results in waveform errors. The effect of the non-linearity is readily interpreted from the shape of the video waveform. Assume that a video signal, with a picture signal containing equal amplitude steps, is transmitted through a non-linear circuit with a transfer characteristic as shown in Fig. 13.

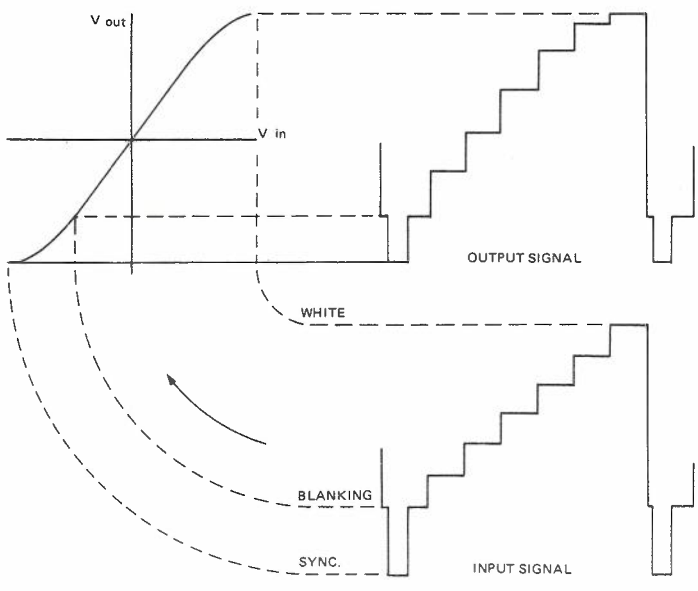


FIG. 13. EFFECT OF NON-LINEAR SYSTEM ON MONOCHROME TELEVISION SIGNAL

The output waveform indicates non-linearity in the picture signal region because the equal amplitude steps are not maintained. Using such a system, the range of shades in the reproduced picture would be compressed compared to what they should be in the white region. Reasonable non-linearity is not objectionable in a reproduced monochrome picture. The output signal in Fig. 13 also indicates non-linearity in the sync. region because of a change in the picture-to-sync. ratio. In practice considerable variation in the sync. pulse amplitude can be tolerated.

As the position of a small amplitude signal on the input-output amplitude characteristic is varied, there should be no change in the amplitude of the output signal; that is, the gain of the signal should be constant over all of the characteristic. In a non-linear system, the small signal gain of the system varies between points on the characteristic. A measure of this variation indicates the 'differential gain' of the system. This characteristic is important for colour television signals since the position of the chrominance subcarrier signal on the characteristic depends on the magnitude of the luminance component. Fig. 14 shows the change in a saturated colour bar video signal when it is transmitted through a non-linear circuit. As well as the compression of the luminance signal, particularly in the white region, and the compression of the sync. pulse amplitude, the amplitude of the chrominance subcarrier is also incorrect. In the example of Fig. 14, the amplitude of the subcarrier is reduced at high values of luminance signal, and this causes a reduction in the colour saturation for high luminance colours.

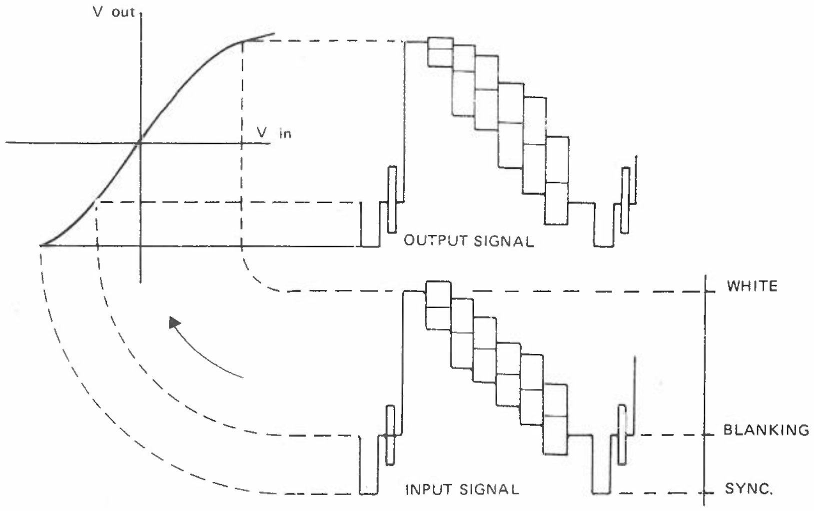


FIG. 14. EFFECT OF NON-LINEAR SYSTEM ON COLOUR TELEVISION SIGNAL

5.3 SOURCES OF NON-LINEARITY. The main sources of non-linearity in an FM bearer system are:

:: Non-linearity of the output-amplitude/input-amplitude characteristic of baseband equipment, mainly baseband amplifiers

:: Non-linearity of the signal-amplitude/output-frequency characteristic of modulators and the reciprocal characteristic of demodulators of the FM system

:: Variation of the amplitude-frequency characteristic of IF and RF equipment from the ideal flat characteristic over the passband

:: Variation of the phase-frequency characteristic of IF and RF equipment from the ideal so that the group delay is not constant

:: Reflections in feeders and propagation paths producing echo signals. Such echo signals vary the amplitude-frequency and phase-frequency characteristics of the system, and can be examined by using these characteristics or by considering the echo signals as co-channel interfering signals

5.4 BASEBAND EQUIPMENT NON-LINEARITY. The most likely source of non-linearity in baseband equipment is the baseband amplifiers. However, the intermodulation noise introduced by baseband amplifiers is small. This is because baseband amplifiers can have much better linearity than other sections of the system by using feedback techniques.

Switching of baseband signals is normally carried out by using relays rather than by electronic switches such as diodes. Relays are used because it is difficult to obtain adequate linearity with electronic switches.

5.5 MODULATOR AND DEMODULATOR NON-LINEARITY. The ideal characteristic for the frequency modulator of a broadband bearer system is one in which the output frequency deviation on either side of the carrier frequency is directly proportional to the amplitude of the broadband signal. Such a characteristic is illustrated by Fig. 15a.

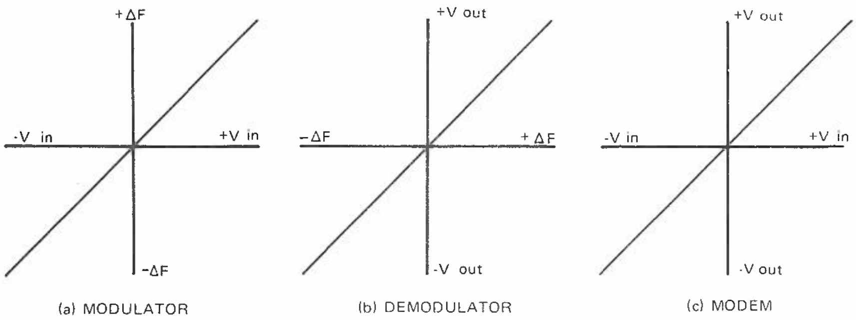


FIG. 15. FREQUENCY MODEM CHARACTERISTICS

The demodulator should also have a linear characteristic giving an output voltage proportional to the deviation of the carrier frequency (Fig. 15b). Since each system used to provide a baseband to baseband bearer circuit includes both a modulator and a demodulator, it is convenient to think of the modulator and demodulator as a pair, operating back-to-back. This allows the non-linearity due to the modem equipment to be treated as one item although modem test equipment can accurately measure the linearity of each unit. An ideal circuit has a linear characteristic which relates modulator input voltage to demodulator output voltage. (see Fig. 15c).

Two forms of modem overall characteristics showing non-linearity are illustrated in Fig. 16. The characteristic (Fig. 16a) is asymmetrical and results in second order products being produced (the linearity error depends on the square of the input voltage). The symmetrical but non-linear characteristic (Fig. 16b) produces third order products (the linearity error depends on the cube of the input voltage). Typical equipment characteristics are a combination of these types of characteristic in various proportions and may also include higher order non-linearities. Non-linearity may be distributed between the modulator and the demodulator in any proportion, but in practice, each is adjusted so that its characteristic is as near as possible to ideal. Test equipment that indicates the relative slopes of the individual and overall characteristics is used for these adjustments.

The slope of the overall characteristic at various points over the amplitude range indicates the gain of the system. The variation of the slope (that is, the variation in gain) over the operating range is a measure of the linearity of the system and is expressed as a percentage of the maximum gain. The measurement of linearity is carried out by measuring the change in amplitude of a small amplitude, high frequency (usually 250-500 kHz) signal as its position on the characteristic is varied or swept. The frequency of the small amplitude signal is referred to as the 'test frequency' or 'search frequency'.

Figs. 16a and 16b each show a small amplitude signal applied to three points on the characteristic. The output signal amplitude varies for the different points on the characteristic and depends on its slope at that point. To obtain the amplitude sweep in practical test equipment, the small amplitude, high frequency signal is superimposed on a low frequency, large amplitude signal. For telephony systems this is usually a low frequency sine wave, often 50 Hz. For television systems, a line rate sawtooth waveform is used as the amplitude sweep waveform. The linearity of the modem characteristics in Figs. 16a and 16b is illustrated in Figs. 16c and 16d respectively. Fig. 16a has a slope that is lowest for the most negative input voltages and increases with increase in input voltage. The gain varies in a linear manner over the operating range (Fig. 16c). Fig. 16b has a slope that is high at each end of the characteristic and lowest when the input voltage is zero. This results in a parabolic gain characteristic (Fig. 16d). The linearity for each is the maximum gain variation over the operating range expressed as a percentage of the maximum gain.

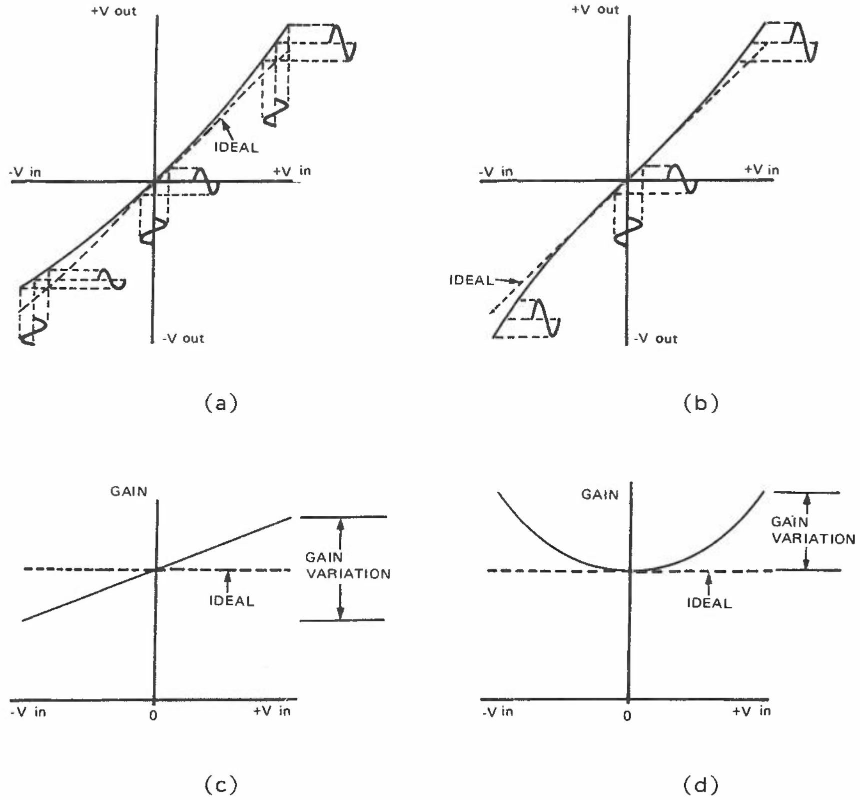


FIG. 16. MODEM OVERALL CHARACTERISTICS

If the test frequency is a high frequency (above 1 MHz), the result of the test is the 'differential gain'.

The value of the slope of the modulator and demodulator characteristics at various points over the operating range is known as the modulator or demodulator derivative respectively. Though the term does not exactly suit the characteristic, the variation in the derivative, relative to the maximum derivative is often called the differential gain because of the similarity of modem tests to the overall differential gain tests.

The noise produced by modem non-linearity depends on the magnitude of the gain variation over the amplitude range (that is the differential gain) and on the shape of the gain characteristic. It can be proved that, when a particular signal-to-noise ratio is to be achieved for the modem of a broadband bearer system, the equipment can have larger parabolic gain errors than linear tilt errors over the operating range.

5.6 NOISE DUE TO AMPLITUDE-FREQUENCY CHARACTERISTIC. The standard deviations and bandwidths required for FM bearer systems of various capacities were discussed in para. 3.2. If the RF and IF equipment bandwidth is less than the specified requirements or the deviation is greater than the standard, significant sideband components of the FM signal are either reduced in amplitude or perhaps almost eliminated. Sideband limiting in an FM system produces an effect similar to the clipping of peaks of the modulating signal information. The gross non-linearity resulting from insufficient bandwidth or excessive deviation produces large amounts of intermodulation noise in the derived telephony channels. An FM system effectively has an overload point and, for a given bandwidth, a small increase in deviation above the overload value causes a large increase in the intermodulation noise.

5.7 FREQUENCY-TO-AMPLITUDE CONVERSION. The amplitude-frequency response of the RF and IF equipment of an FM bearer over the desired passband is controlled mainly by the IF circuit and should be constant. In practice, small deviations from the ideal occur. This causes the FM signal to vary in amplitude as the carrier frequency varies with modulation. The effect is illustrated by Fig. 17 and is called frequency-to-amplitude conversion. Assume that a circuit has an amplitude-frequency response which includes a tilt over the passband as shown in Fig. 17. When an FM signal is applied, the carrier frequency deviates within the passband of the circuit. The tilt of the characteristic causes the amplitude of the output to vary in accordance with the frequency variation of the input. Therefore the output carrier signal is amplitude modulated in addition to being frequency modulated.

Other more complex passband shapes, for example, parabolic and double humped amplitude-frequency characteristics, produce more complex variations in the shape of the output envelope. The envelope amplitude variations produced by errors in the amplitude-frequency response are removed by limiter stages in the FM equipment and, if the limiters are ideal, no amplitude variations remain to cause degradation of the traffic signal. Some intermodulation products can be produced if the limiters are not ideal.

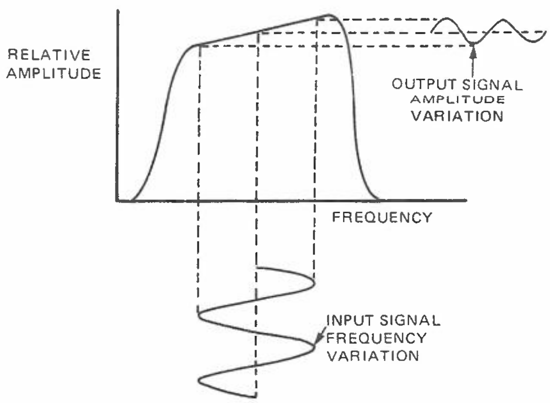


FIG. 17. DEGRADED AMPLITUDE-FREQUENCY CHARACTERISTIC

In addition to producing amplitude variations, the amplitude-frequency response causes phase variations to be produced, even when the phase-frequency characteristic is ideal. This results from the effect of the amplitude-frequency response on the various sideband components. The phase variations can represent non-linearities resulting in intermodulation distortion. A small linear tilt over the passband does not produce any intermodulation distortion, and a strictly parabolic passband characteristic results in a negligible amount of distortion. However, practical circuits have more complex passband shapes with sharp cut-offs towards the edges of the band and often have double humped characteristics. These types of characteristics produce significant intermodulation products. A symmetrical characteristic produces less intermodulation noise than does an asymmetrical characteristic.

To achieve low intermodulation distortion, the amplitude-frequency response of the IF and RF equipment typically has a symmetrical parabolic or double humped characteristic with a maximum of ±0.2 dB amplitude variation over the bandwidth required. The particular shape of the response is determined by the equipment designer because, in some equipment, the parabolic response of one section is used to compensate for the double humped response of other equipment.

As explained earlier, an FM signal may become amplitude modulated as it passes through the several stages preceeding the system limiters. As a result of the amplitude modulation, the FM signal may be subjected to some amplitude to phase modulation conversion in subsequent stages. This is possible because of the inherent non-linear voltage dependent capacitance between the terminals of the semiconductor devices used in these stages. The AM to PhM conversion can be produced in transmit and receive mixers, RF and IF amplifiers and limiters. It is a further possible source of intermodulation distortion dependent on the amplitude-frequency characteristic of the RF and IF equipment.

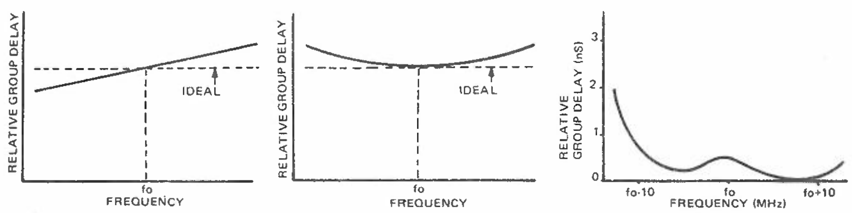
5.8 NOISE DUE TO EQUIPMENT GROUP DELAY-FREQUENCY CHARACTERISTIC.

Errors in the group delay-frequency characteristic are very important and in practice are the source of a significant part of the intermodulation noise of a radio broadband bearer.

To transmit the modulated signal without distortion, IF and RF sections of the equipment require a constant group delay over the passband as any variation from the ideal causes phase modulation of the FM signal. After detection of the signal in the FM detector, the phase modulation results in intermodulation distortion.

Group delay-frequency characteristics showing basic forms of errors are illustrated in Fig. 18. In Fig. 18a the group delay has a linear tilt over the passband. A parabolic group delay-frequency characteristic is illustrated in Fig. 18b. Practical equipment has characteristics which are a combination of these and higher order characteristics. The main sources of group delay errors in bearer equipment are the bandpass filters of RF sections and the bandpass characteristics produced by the coupling circuits or filters of IF amplifiers. To reduce group delay errors in a system, each repeater and terminal normally includes a group delay equaliser to compensate for the errors in that section of the equipment. Further equalisers are usually provided to compensate for errors accumulated over each section between demodulating stations.

Note: A group delay equaliser is a device with an equal and opposite group delay characteristic to that of the equipment it is to equalise.



(a) (b) (c)

FIG. 18. GROUP DELAY CHARACTERISTICS

The amount of noise produced by group delay errors depends on the magnitude of the errors and also on the shape of the group delay-frequency characteristic. It can be proved that a particular magnitude of tilt in the characteristic over the passband produces considerably more noise than does a parabolic group delay with the same maximum error. It is desirable therefore, that bearers be equalised to give a symmetrical group delay-frequency characteristic. After equalisation, the group delay measured between two IF points of a system is approximately constant over the centre of the passband and increases rapidly towards the edges of the band. A typical characteristic is illustrated by Fig. 18c.

The group delay-frequency characteristic is normally specified for a slightly narrower band than the amplitude-frequency characteristic.

A typical specification for a 960 channel system requires:

:: Linear group delay variation (tilt) < 1 ns at ±7 MHz

:: Parabolic group delay variation (symmetrical) < 3 ns at ±7 MHz

5.9 ECHO SIGNALS. When a transmission line is not terminated in its characteristic impedance, not all of the incident power is accepted by the load: some is reflected. The amplitude of the reflected signal depends on the reflection co-efficient. Also, when the generator feeding the line does not have an impedance equal to the line characteristic impedance, a further reflection occurs, again reduced in amplitude dependent on the reflection co-efficient. The double reflection produces an echo signal at the output which is delayed by a time equal to twice the propagation time along the line. The echo amplitude depends on the reflection co-efficients at each end of the line and also on the attenuation of the transmission line.

The most likely sources of troublesome transmission line echoes in a radio broadband bearer are the feeders between the transmitters and receivers and their associated antennae. These feeders are normally in the form of a waveguide. In a typical example the waveguide run is 60 metres. If the waveguide group velocity is two thirds of the free space velocity, the echo time delay is:

$$\text{Delay} = t = \frac{2 \times 60 \times 3}{300 \times 2}\ \mu s = 0.6\ \mu s.$$

The phase relationship between the echo signal and the original signal depends on the time delay introduced by the line and the frequency of the signal.

A typical requirement for the return loss of a waveguide antenna feeder used with 960 channel bearer system is 26 dB. The amount of energy reflected because of mismatch can be expressed by the reflection co-efficient (ρ), which is equal to the ratio of the reflected voltage to the incident voltage. The relationship between return loss and reflection co-efficient is:

$$\text{Return loss (dB)} = 20 \log \frac{1}{\rho}$$

When a waveguide antenna feeder has a return loss of 26 dB, this corresponds to a reflection co-efficient of approximately 0.05. Assume that the source feeding the waveguide has the same specifications. If the waveguide run has an attenuation of 1.7 dB (voltage ratio output to input of approximately 0.82) the echo signal at the output has an amplitude less than the original output signal amplitude by 26 + 1.7 + 26 + 1.7 = 55.4 dB. Therefore, the ratio (ρ) of the reflected or echo signal amplitude to the original or direct signal amplitude is:

$$\frac{\text{Echo signal amplitude}}{\text{Direct signal amplitude}} = \rho = \frac{1}{\text{Antilog } \frac{55.4}{20}} = \frac{1}{589} \cong 0.0017$$

In practical antenna feeders, small reflections are produced at bends and joints as well as from the terminations.

The transmitting and receiving equipment connected to an antenna feeder includes multi-section filters for selection of the required frequency band and for rejection of undesired signals. These filters introduce a significant delay to the signal modulation envelope because of their group delay characteristic. The delay depends on the filter bandwidth and the number of sections and is typically 25 ns. Therefore, if a filter is not correctly terminated at both the input and the output, the reflections produced cause an echo signal delayed by approximately 50 ns.

In some installations, only a short cable run is required for the intermediate frequency connections between the transmitter-receiver and modulator-demodulator racks of a bearer. In other installations, these cables can have lengths up to about 30 metres, and again assuming that the IF cable group velocity is 2/3 that of free space, this would give echo delays of approximately 0.3 μs.

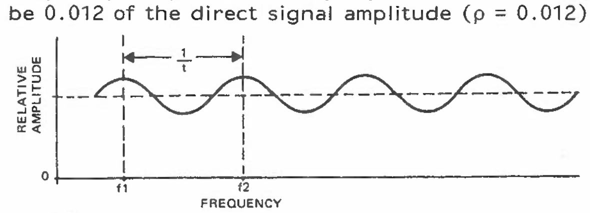
To prevent echoes, IF cables must be correctly terminated. Typically, a return loss better than 24 dB is required of the input and output impedances of the IF equipment of a 960 channel system.

We see, therefore, that typical transmission lines usually have echo signals with low amplitudes relative to the desired signal and the echo delay times are often significant compared with the time for one cycle of the maximum baseband frequency. Echo signals can also be produced by reflections in propagation paths and by multiple propagation paths between transmitters and receivers. These echoes are responsible for the fading of the RF signal. The time delays experienced by reflected signals in most cases are very short. For a typical path, a signal arriving via a ground reflection is delayed less than 1 ns. An echo signal varies in both amplitude and phase depending on propagation conditions and it is possible for the echo signal amplitude to be comparable with that of the direct signal. Longer propagation echoes are possible from reflections close to an antenna but directional antennae usually make the amplitude of such echoes negligible.

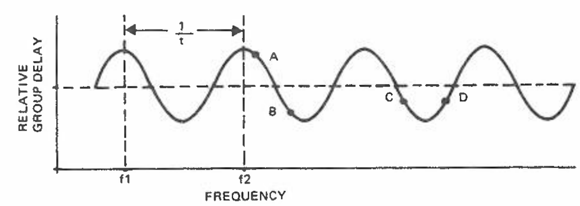
5.10 EFFECT OF ECHOES ON BEARER CHARACTERISTICS. The magnitudes of echoes on a transmission line can be determined by return loss measurements. In addition, the presence of echoes somewhere in a bearer system is indicated by the way in which the amplitude-frequency and group delay-frequency characteristics of the system are modified. Assume that the input signal to an incorrectly terminated transmission line is a sine wave with a frequency that is varied over the frequency range of interest. At some frequencies the delay produced by the transmission line is such that the echo signal at the output is in phase with the direct or original signal. The resultant output signal amplitude is the sum of the two signal amplitudes. At other frequencies the delay causes the echo to be in inverse phase to the direct signal and the amplitude of the output signal is the difference between the two signal amplitudes. That is, the output amplitude is dependent on frequency, with the maximum and minimum amplitudes being the sum and difference respectively of the direct and echo signals. When the ratio (ρ) of the echo signal amplitude to direct signal amplitude is relatively small, the output amplitude varies approximately sinusoidally with frequency. The echo has caused a ripple in the amplitude-frequency response of the transmission line and this characteristic is illustrated in Fig. 19a. When the line has the characteristics of the example in para. 5.9, the variation in amplitude is ±0.0017 of the average amplitude. The ratio of the maximum amplitude to the average amplitude in dB is:

$$\begin{aligned} 20 \log 1.0017 &= 20 \times 0.0007 \\ &\cong 0.015 \text{ dB} \end{aligned}$$

That is, the amplitude-frequency response ripple has a variation of only ±0.015 dB. For the amplitude-frequency response to vary by ±0.1 dB, the amplitude of a single echo would have to be 0.012 of the direct signal amplitude (ρ = 0.012).



(a) Amplitude - Frequency Characteristic



(b) Group Delay-Frequency Characteristic

FIG. 19. EFFECT OF ECHOES ON AMPLITUDE-FREQUENCY AND GROUP DELAY - FREQUENCY CHARACTERISTICS

Accompanying the variation in the amplitude-frequency characteristic is a variation in the group delay-frequency characteristic for the line. When ρ is small, the variation in the characteristic is sinusoidal (Fig. 19b). The peak magnitude of the group delay variation from the average equals the product of ρ and the echo delay. When ρ is 0.0017 and the delay is 0.6 µs (para. 5.9) the group delay variation is

$$\pm 0.0017 \times 0.6$$

$$= \pm 1.02 \text{ ns}$$

Typical test equipment is capable of displaying amplitude-frequency characteristics with a sensitivity of 0.1 dB per centimetre division and group delay-frequency characteristics with a sensitivity of 1 ns per centimetre division. The preceding typical values indicate that a more sensitive indication of the presence of degrading echoes is obtained from a group delay-frequency characteristic.

Let us examine the relationship between the echo delay and the frequency spacing of the ripples in the amplitude-frequency and group delay-frequency characteristics. Assume that the echo delay (t) causes the echo to be in phase with the direct signal at a frequency (f1) and that the delay corresponds to n cycles of f1.

$$\text{Time for one cycle of f1} = \frac{1}{f1}$$

$$\therefore \quad \text{Delay} = t = \frac{n}{f1}$$

$$\therefore \quad f1 = \frac{n}{t}$$

At the next higher frequency that the direct and echo signals are in phase (f2), the delay corresponds to n + 1 cycles of f2.

$$\therefore \quad t = \frac{n+1}{f2}$$

$$f2 = \frac{n+1}{t}$$

These frequencies correspond to adjacent peaks on the amplitude-frequency characteristic (Fig. 19a) and the spacing of the peaks or the pitch of the ripple is:

$$f2 - f1 = \frac{n+1}{t} - \frac{n}{t}$$

$$= \frac{1}{t}$$

From the above, we see that the pitch of the ripples in the characteristic is inversely proportional to the delay time. In a typical example with an echo delay of 0.6 µs, the spacing of the ripples in the characteristic is 1.67 MHz. When the group delay is being measured over a range of ±9 MHz using swept frequency techniques, the display could appear as shown in Fig. 20a.

A change in the centre frequency of the sweep range would alter the phase of the display. For shorter delays the number of ripples in the display for a particular frequency range decreases. If the echo delay is very short, only a part of a cycle of the sine wave response is displayed. For example, assume that the echo delay is 18.5 ns. Such a short delay could be produced by a return echo in 1.85 metres of waveguide and could originate from a reflection from a faulty waveguide joint or bend close to the equipment racks. With an echo delay of 18.5 ns, the ripples in the group delay characteristic are 1000 ÷ 18.5 = 54 MHz apart. This means that the ±9 MHz display is one third of a sine wave cycle and could appear as a tilt as shown in Fig. 20b when the swept range is covering section AB of the response in Fig. 19b. A change in the swept range could produce an approximately parabolic characteristic (Fig. 20c) corresponding to section CD of Fig. 19b.

The relationship between echo delay and the ripples in the swept responses of a system allows the time delay of an objectionable echo to be determined. With this information it is possible to predict the likely origin of an echo. In a practical system involving many hops, the feeder lengths vary in a random manner. When all feeders meet the same return loss specifications, the regular pattern of ripples in the characteristic tends to disappear. However, when one section is degraded, the echo produced, results in a distinguishable ripple which is superimposed on the random display corresponding to the remainder of the sections.

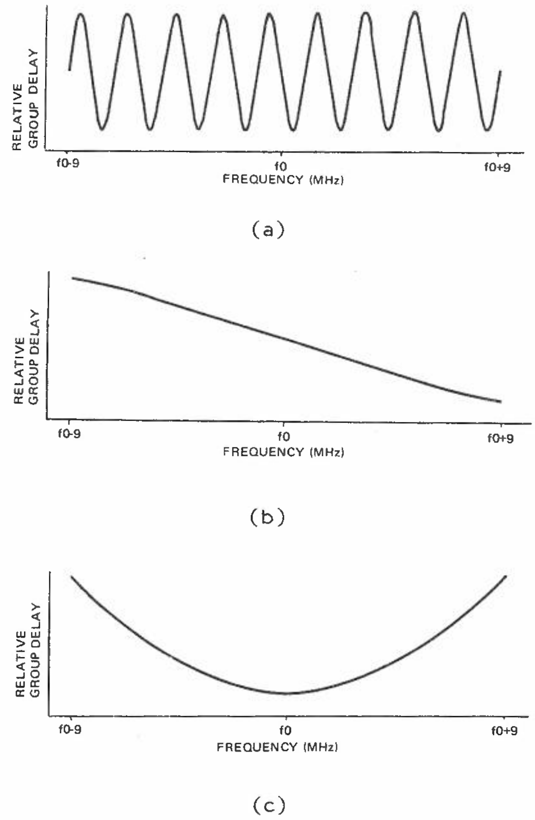


FIG. 20. TYPICAL GROUP DELAY - FREQUENCY CHARACTERISTICS

5.11 NOISE DUE TO ECHOES. Echo signals cause variations in amplitude-frequency and group delay-frequency characteristics and these variations result in intermodulation noise in the channels of a telephony broadband system. The magnitude of the intermodulation noise depends directly on ρ for all values of delay. When the delay is long (greater than the reciprocal of the maximum baseband frequency) the noise is at its maximum and is independent of the magnitude of the delay. The echo behaves as if it is unrelated to the direct signal and the delay is not important. A delay that is the reciprocal of the maximum baseband frequency gives a characteristic with a ripple spacing equal to the maximum baseband frequency. The delays introduced by the terminations of practical antenna feeders are in the long delay category. For example, assume a 960 channel bearer with a maximum baseband frequency of approximately 4 MHz. The reciprocal of the maximum baseband frequency corresponds to a delay of 0.25 μs. This echo delay is produced by a waveguide of approximately 25 metres.

For short delays the noise produced is dependent on the delay and the echo phasing, as well as being dependent on the echo amplitude. The intermodulation noise level decreases rapidly as the delay is reduced. In the phase condition that gives maximum noise, the noise reduces by approximately 12 dB each time the delay is halved. Because of this, the very short delay echoes produced by propagation conditions do not normally produce significant noise.

Reduction of signal amplitudes produced by the echo allows the receiver input thermal noise to predominate and, when this fading is severe, switching to the protection bearer of a diversity system normally occurs before echo intermodulation noise becomes objectionable.

The noise produced by echoes tends to be distributed fairly evenly over the upper two-thirds of the baseband spectrum, with lower noise levels in channels at the lower end of the baseband. When emphasis networks are included, the highest frequency channels are improved by a few dB and low frequency channels are degraded by a similar amount. This results in a noise distribution that is within a few dB over the upper 80% of the baseband spectrum, with a broad peak of approximately 3 dB in the centre baseband channel (compared with the high frequency channels). Pre-emphasis and de-emphasis of the baseband signal is examined in para 6.13.

5.12 EFFECT OF ECHO ON TELEVISION SIGNALS. When a television signal is transmitted via an AM system, an echo produces a ghost of the desired image in the reproduced picture. An echo in an FM system has a considerably different effect. Assume that a step transition is to be transmitted via an FM bearer. Before the transition, the frequency transmitted is constant at f1 (Fig. 21a).

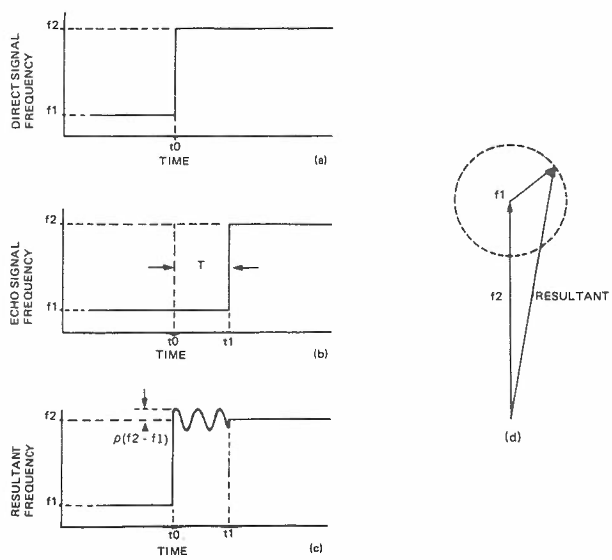


FIG. 21. EFFECT OF ECHO ON TELEVISION SIGNALS

After the transition, the transmitted frequency is again constant and equal to f2. Assume that the output from a circuit includes the original or direct signal and also an echo signal with a delay equal to T. The echo signal has a reduced amplitude but the frequencies before and after the step in the modulating signal amplitude are the same as for the direct signal as shown in Fig. 21b.

Before time t0 and after time t1, both the direct and echo signals have the same frequency. The resultant combined signal (Fig. 21c) is therefore unchanged in frequency at these times. The signal amplitude depends on the phase relationship between the direct and echo signals. Between times t0 and t1, however, two different frequencies are present and the lower amplitude echo signal behaves as an interfering signal interacting with the desired direct signal. The two signals are represented by the vector diagram in Fig. 21d. If the vector for f2 is considered as stationary, the vector for f1 rotates and describes a circular locus. The resultant is the vector sum of these two components and it follows the locus of f1. The resultant varies in magnitude and phase at a frequency corresponding to the difference between f1 and f2. The phase variation of the resultant is significant in an FM system as it corresponds to an equivalent frequency variation of the FM signal.

If the echo amplitude is small when compared with the direct signal amplitude, the interfering frequency variation is sinusoidal. The peak deviation resulting from the interference equals the ratio of the echo to direct signal amplitudes (ρ) times the frequency change produced by the step. A corresponding amplitude degradation is produced after demodulation. The phase relationship of the interference depends on the relative phase of f1 and f2. The interference commences abruptly when the direct signal arrives and terminates abruptly when the echo signal arrives. Notice that this is different from ringing that is often encountered in video circuits.

A considerably more complex situation exists when the television emphasis networks are included. The resultant interference varies in frequency for a time dependent on the equivalent time constant of the emphasis networks and continues after the arrival of the echo signal.

When pre-emphasis is not included, a black to white transition using the standard deviation corresponds to a frequency shift of approximately 5 MHz. Smaller amplitude steps produce lower frequency shifts. This means that all possible interference frequencies are within the passband of the television baseband. With a typical echo delay of 0.6 μs (para. 5.9), 3 cycles of a 5 MHz interference would be reproduced. When the delay is less than approximately 0.1 μs, the duration of the interference produced results in products outside the television passband. Hence short delay echoes produced by propagation conditions do not normally give significant waveform degradation. For these echoes, it is more important to consider the effect of fading of the RF signal level.

When a feeder system meets the return loss specifications assumed in para. 5.9, the peak-to-peak amplitude of the interference is negligible, being 0.36% of the step amplitude. However, echo interference could become significant should a fault introduce excessive feeder reflections. Since feeders for all of the stations on a bearer route tend to have random lengths, the interference of several echoes increases on a power basis, that is, when the number of echoes of the same amplitude is doubled, the amplitude of the resulting interference is increased by $\sqrt{2}$ times.

6. CHANNEL SIGNAL-TO-NOISE RATIO.

6.1 CHANNEL NOISE PRODUCED BY AN FM SYSTEM. When noise is superimposed on a carrier signal, it causes both amplitude and phase variations of the carrier. In an FM system, amplitude variations of the envelope are removed by the limiter, but phase variations still remain. The frequency deviation, produced from phase modulation of the carrier by noise, results in noise being present at the baseband output of the receiver. When the carrier level is significantly greater than the noise level, the signal-to-noise ratio at the receiver baseband output is proportional to the carrier-to-noise ratio and is therefore proportional to the receiver input signal level. In an FM system, when an interfering signal (noise in this case) becomes greater in amplitude than the desired carrier amplitude, the interfering signal overrides the required signal. The output of the FM detector is almost completely 'information' from the interfering signal and the modulation of the desired signal is suppressed. This causes the signal-to-noise ratio to decrease to zero.

6.2 SIGNAL-TO-NOISE/RECEIVER INPUT LEVEL CHARACTERISTIC. The variation of channel signal-to-noise ratio (S/N) with receiver input carrier signal level is shown in Fig. 22. The curve shows a linear decrease in channel S/N as the receiver carrier level is decreased, until the abrupt fall in S/N as the THRESHOLD LEVEL is approached. The theoretical threshold level of an FM receiver is the level at which the peak amplitude of the interfering noise equals the peak amplitude of the carrier signal. The peak amplitude of white noise is approximately 12 dB above its RMS value, and the peak amplitude of a sine wave is 3 dB above its RMS value. This means that the theoretical threshold level is the level at which the equivalent noise power at the receiver input is 9 dB below the carrier level; that is, the carrier-to-noise ratio (C/N) is 9 dB.

In the example in para. 4.10, the equivalent receiver input noise (source thermal noise plus noise figure) is -118 dBW. This results in a theoretical threshold level of -109 dBW. The full S/N improvement that can be achieved by a particular FM system is not available until the carrier level is above the practical threshold at the lower end of the linear section of the characteristic in Fig. 22.

The practical threshold is normally arbitrarily chosen as 3 dB above the theoretical threshold at a C/N of 12 dB. The low S/N below the practical threshold makes the bearer unusable. To prevent high noise levels which could simulate pilot signals, muting circuits are included to disable the receiver under low input signal level conditions. Muting circuits are normally adjusted to operate at a receiver carrier level slightly below the practical threshold.

6.3 SIGNAL-TO-NOISE RATIO LIMITING. Above the practical threshold, the S/N due to the equivalent receiver input noise level is proportional to the carrier level. At high carrier levels, S/N is limited by other factors. When no baseband signal is present, the limit is determined by basic noise components that are not dependent on receiver input level. When the bearer is carrying traffic, S/N is limited at a lower value by intermodulation noise. Fig. 22 illustrates the limiting of S/N characteristics of the receiver at high signal levels for mean busy hour traffic. With reduced traffic, less intermodulation noise is produced and the S/N limit set by intermodulation noise increases towards the constant basic noise limit. Systems are usually designed so that the mean receiver input carrier level falls on the section of the curve where, during busy hour, the intermodulation noise is a significant factor. In the carrier signal level range between the free space signal level (no-fade value) and the practical threshold level, the system operates successfully, but with a poorer than normal channel signal-to-noise ratio. This range of levels allows for fading and is known as the FADE MARGIN of the system. The fade margin of each receiver is normally arranged to be at least 40 dB. In a typical system, the bearer is switched and replaced by a protection bearer when a carrier signal level fade of about 25 to 35 dB occurs.

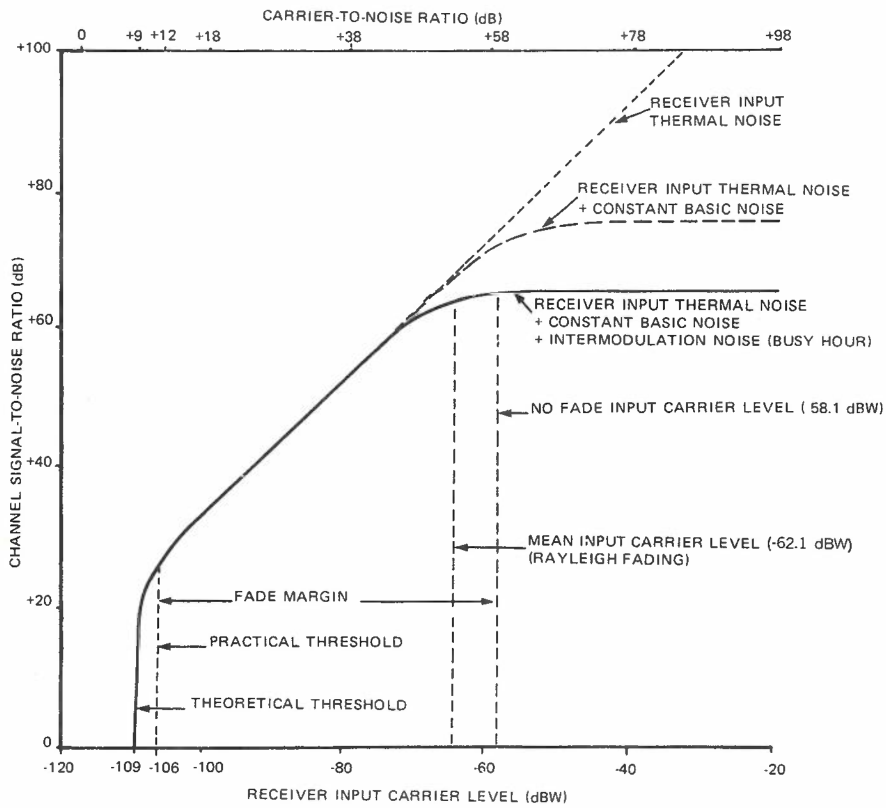


FIG. 22. VARIATION OF SIGNAL-TO-NOISE RATIO WITH CARRIER LEVEL

6.4 EFFECT OF VARYING THE RECEIVER BANDWIDTH. Because receiver input noise depends on receiver noise bandwidth (para. 4.9), the threshold level would be decreased if the bandwidth were decreased. Though the receiver threshold would be extended (Fig. 23a), this would not change S/N above threshold since the channel bandwidth is unchanged. However, for a given channel capacity and deviation, the reduction of receiver bandwidth could introduce non-linearity and increase intermodulation noise. This would result in a poorer S/N at high receiver signal levels. These conditions are illustrated by the characteristic in Fig. 23a.

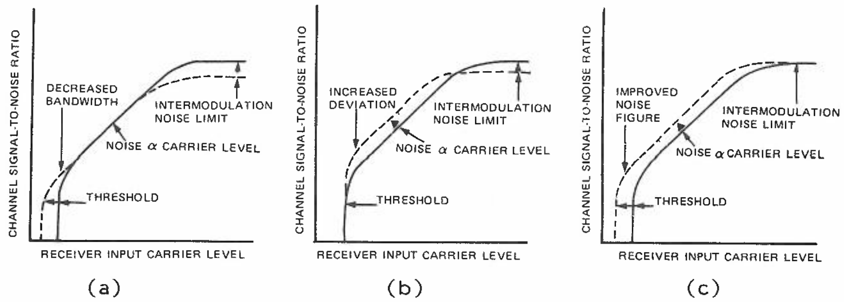


FIG. 23. FACTORS AFFECTING SIGNAL-TO-NOISE RATIO CHARACTERISTICS

6.5 EFFECT OF VARYING THE DEVIATION. The S/N at the output of an FM system depends on the ratio of system deviation to the deviation resulting from the phase modulation produced by noise. Increasing the reference level deviation specified for a system would improve S/N at low carrier signal levels. However, this would not alter the threshold level. With the bandwidth of the system unchanged, including that of modulators and demodulators, an increase in deviation above the normal operating deviation would normally cause the intermodulation noise to increase more rapidly than the deviation increase.

Therefore, the S/N limit produced by intermodulation would normally deteriorate. The resultant S/N variation with receiver input carrier signal level is shown in Fig. 23b. These factors were taken into account when the deviation specifications (para. 3.2) were determined.

6.6 EFFECT OF VARYING THE NOISE FIGURE. If the noise figure of a receiver is improved while other factors are maintained, the equivalent receiver input noise level is decreased and the threshold level is similarly decreased. The equivalent input noise improvement gives a corresponding improvement in C/N and also the resulting channel S/N. The S/N limiting produced by intermodulation noise is not affected. Fig. 23c illustrates this condition.

6.7 EFFECT OF CHANNEL BANDWIDTH. To determine the noise in a 3.1 kHz telephone channel at the output of a receiver, it is necessary to take into account the difference between the bandwidth of the receiver RF and IF circuits and the bandwidth of the output channel. This is because only that part of the total noise power which has a bandwidth equal to the channel bandwidth can appear in the output channel. Since receiver noise is evenly distributed over the frequency spectrum of the receiver bandwidth, the noise in the output channel is reduced in proportion to the ratio of the receiver noise bandwidth to the channel bandwidth.

6.8 BASEBAND DISTRIBUTION OF RECEIVER NOISE. The frequency deviation produced by phase modulation, in addition to being proportional to the phase change, is also proportional to the rate of change of the modulating signal. Since the rate of change of a sine wave signal of given amplitude is proportional to the frequency of the signal, the frequency deviation produced by phase modulation is proportional to the frequency of the modulating signal. Therefore:

$$\Delta F \propto \Delta\phi . f_M$$

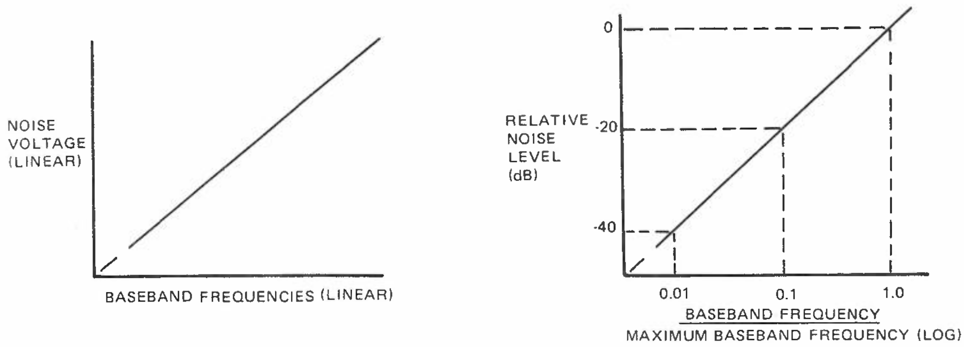
Where

ΔF = frequency deviation

$\Delta\phi$ = phase deviation

f_M = modulating frequency

Noise produced by thermal agitation in the receiver source resistance and in the early stages of the receiver has a frequency spectrum with components evenly distributed over the receiver passband. After demodulation of the noisy carrier signal, the resultant noise voltage at the baseband output is proportional to frequency (Fig. 24a); it is low at the low frequency end of the baseband and increases linearly towards the high frequency end. This results in a noise power at the baseband output that is proportional to the square of the baseband frequency; that is, the noise increases by 20 dB when the baseband frequency is increased by a factor of 10 (Fig. 24b).



(a) (b)

FIG. 24. BASEBAND OUTPUT NOISE DISTRIBUTION

6.9 NOISE IN A TELEPHONY CHANNEL. Factors that affect noise have been examined in preceding paragraphs. Consideration of these factors enables us to determine the noise in a telephony channel which results from receiver source and internal thermal noise. For a bearer operating with a carrier signal level greater than the threshold level, S/N depends directly on receiver C/N. The S/N is better than C/N in proportion to the ratio of the receiver noise bandwidth to channel bandwidth. The use of an FM system further modifies the S/N in accordance with the ratio of the system per channel deviation to the baseband frequency corresponding to the centre frequency of the output channel. As each channel of a broadband system occupies only a small proportion of the total baseband, the noise can be assumed to be distributed approximately evenly over the channel, even though the distribution over the baseband is triangular. The resultant S/N due to the receiver noise, expressed in logarithmic form, is:

$$\text{Channel signal-to-noise ratio} = \frac{C}{N}\ (\text{in dB}) + 10 \log \frac{B}{b} + 20 \log \frac{\Delta f\ (\text{RMS})}{f_M}\ \text{dB}$$

or alternatively

$$\text{Channel S/N} = \frac{C}{N}\ (\text{in dB}) + 10 \log B - 10 \log b + 20 \log \Delta F\ (\text{RMS}) - 20 \log f_M\ \text{dB}$$

where -

$\frac{C}{N}$ = Carrier-to-Noise ratio in dB

B = Receiver IF bandwidth

b = Channel bandwidth (3.1 kHz)

ΔF (RMS) = RMS Channel deviation

f_M = Baseband frequency corresponding to the channel centre frequency

It would appear from the formula that channel S/N depends on the receiver bandwidth (B). However, this is not so since C/N includes the receiver bandwidth in the denominator. An increase in receiver bandwidth increases the receiver equivalent input noise power and decreases C/N, but this is cancelled by the corresponding increase in bandwidth ratio (B/b). The noise due to the receiver equivalent input noise depends only on the channel bandwidth.

Assume that a system has the following characteristics:

Carrier-to-noise ratio	(C/N)	=	59.9 dB (para. 4.12 - no fade condition)
Receiver noise bandwidth	(B)	=	40 MHz (para. 4.9)
Channel bandwidth	(b)	=	3.1 kHz
Channel deviation	(ΔF RMS)	=	200 kHz RMS for standard test tone level
Channel centre frequency	(f_M)	=	3 886 kHz

Then, for the no-fade condition:

Channel signal (standard test tone level) - to-noise ratio

$$= 59.9 + 10 \log 40\,000 - 10 \log 3.1 + 20 \log 200 - 20 \log 3\,886 \text{ dB}$$

$$= 59.9 + 46 - 4.9 + 46 - 71.8 \text{ dB}$$

$$= 75.2 \text{ dB}$$

When a 4 dB fade allowance is included, the mean C/N is decreased by 4 dB and a corresponding reduction in channel S/N is produced. Therefore, channel S/N under mean carrier level conditions is 71.2 dB.

By assuming other fade allowances, S/N for any fade condition can be determined. This is done when it is required to calculate the minimum channel S/N that is expected for all but the short periods of time during deep fades. The details of short time signal-to-noise ratios are not examined further in this paper.

Noise in a telephony system is normally referred to the output transmission level reference point and quoted relative to a 0 dBm test signal at that point. Therefore, with a 71.2 dB mean S/N, the mean receiver output noise level is -71.2 dBmO. The noise at the reference point is often quoted in linear power units. It is usually given in picowatts at the output transmission level reference point (pWO). In this example, the channel noise level is:

$$\text{Mean noise level at reference point} = \text{Antilog } \frac{-71.2}{10} \text{ mWO}$$

$$= \text{Antilog } -7.12 \text{ mWO}$$

$$= \text{Antilog } \bar{8}.88 \text{ mWO}$$

$$= \text{Antilog } 1.88 \text{ pWO}$$

$$= 75.88 \text{ pWO}$$

6.10 THERMAL NOISE VARIATION IN DIFFERENT CHANNELS. The baseband frequency of 3 886 kHz used in the previous example is a standard test fequency near the upper frequency limit of a 960 channel system baseband. The same calculations for a baseband frequency of 70 kHz (a standard test frequency near the lower frequency limit of a 960 channel system baseband) gives a mean S/N increase to 106.1 dB. This represents a mean noise level at the output transmission level reference point of 0.02 pWO. The two values of S/N indicate that the receiver input thermal noise has its greatest influence on the channels at the high frequency end of the baseband spectrum.

The dotted curves in Fig. 25 show the signal-to-receiver input thermal noise ratio (S/N_{th}) for variations in receiver input carrier signal level, for channels centred on three baseband frequencies. The frequencies correspond to standard test channels of a 960 channel bearer system. The curves show that the S/N_{th} of the 3 886 kHz channel is about 34.9 dB (corresponding to 106.1 - 71.2 dB) poorer than that of the 70 kHz channel. The intermediate 2438 kHz channel has an intermediate value of S/N_{th} in proportion to its relative baseband frequency.

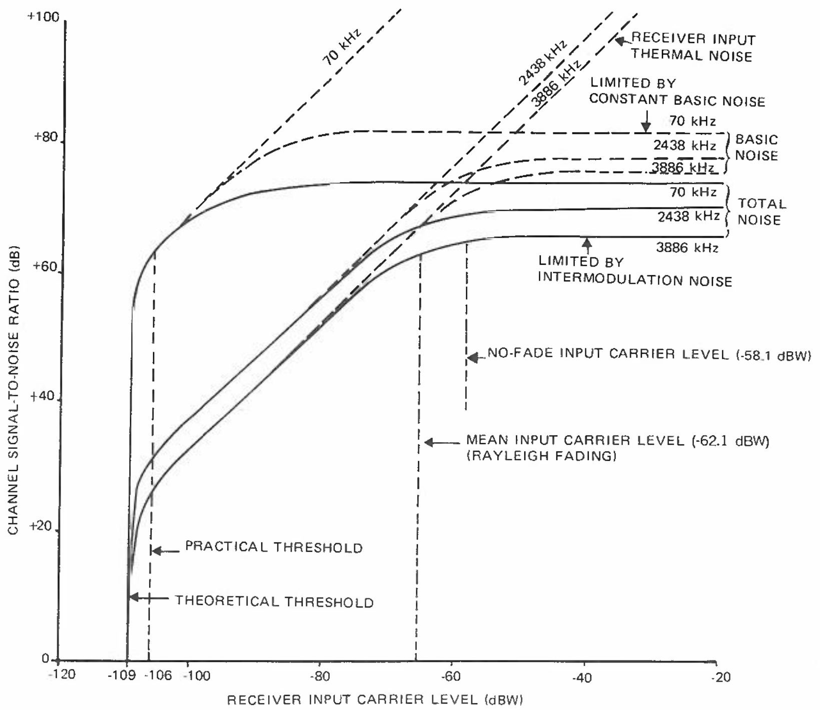


FIG. 25. SIGNAL-TO-NOISE RATIO FOR BASEBAND CHANNELS

In practice, at high receiver input signal levels when basic and intermodulation noise become significant in comparison with the receiver input thermal noise, the S/N difference between high and low frequency channels is much less than is suggested by their S/N_{th} differences. The broken line curves in Fig. 25 indicate that constant basic noise limits the S/N in all channels to values that are not greatly different from one another. Similarly, when intermodulation noise is included (full line curve, Fig. 25), the S/N limit in all channels does not vary over a wide range. Because the low frequency channel has the best S/N_{th}, the S/N limit in this channel is reached at a lower receiver input signal than for the high frequency channel. When pre-emphasis (para. 6.13) is taken into account, the S/N_{th} difference between high and low frequency channels is further reduced. The high frequency channel is improved by approximately 4 dB, the centre channel is not greatly affected, but the low frequency channel is degraded by approximately 4 dB.

6.11 S/N CHARACTERISTICS FOR MULTI-HOP SYSTEMS. The curves in Fig. 25 represent the characteristics of a single hop system. Characteristics with a similar appearance are produced by a multi-hop system when considering the received signal level for one hop, with the other hops having a constant level received signal. This type of characteristic occurs when one hop is fading but other hops are operating at normal levels. With a multi-hop system characteristic, the slope of the section between receiver threshold and the knee (Fig. 25) is the result of the input thermal noise of the receiver suffering a fade.

In this section, S/N is proportional to receiver input signal level. At receiver input levels above the knee, the S/N is limited to a value dependent on intermodulation noise, constant basic noise and the input thermal noise of the remaining receivers of the system. For a single hop system characteristic, the knee in the characteristic occurs at a receiver input carrier level where basic noise equals intermodulation noise. For a multi-hop system, this condition is approached when intermodulation noise is the controlling factor in every hop except the one assumed to be fading. When basic noise is the controlling factor for the noise introduced by each no-fade hop, there is negligible difference between the basic and total signal-to-noise ratios on the 'limited' section of the characteristic at high receiver input signal levels.

6.12 WEIGHTED NOISE MEASUREMENTS. When measuring noise, it is common to use a psophometric weighting network to modify the frequency response of measuring equipment so that the reading obtained gives an indication of the subjective annoyance of the noise or interference, to the telephone subscriber. Interfering signals at the high and low frequency ends of the channels are less annoying to the human ear than those around 1 kHz.

When noise is distributed evenly over the channel bandwidth, as is approximately true for the channels of a broadband system, it is not necessary to use a psophometric weighting network with the measuring equipment. This is because the weighting network for a telephony channel gives an improvement in channel S/N of 2.5 dB. To obtain a psophometrically weighted output noise level, 2.5 dB is subtracted from the level in dBmO and the result is designated dBmOp. For example a noise level of -69.2 dBmO when psophometrically weighted, is -71.7 dBmOp. This noise level in linear units is 67.6 pWOp, with the suffix 'p' indicating that the figure is a psophometrically weighted value.

6.13 TELEPHONY PRE-EMPHASIS. In an attempt to equalise the S/N of high and low frequency channels of a broadband system, it is normal to use pre-emphasis. The high baseband frequencies are increased in level relative to the low baseband frequencies at the transmitting end of the system and reduced again, along with the noise, at the receiving end with a complementary de-emphasis network. This gives an improvement in the high frequency S/N. The amount of pre-emphasis is a compromise. Pre-emphasis increases the intermodulation noise which causes additional noise to appear in low frequency channels.

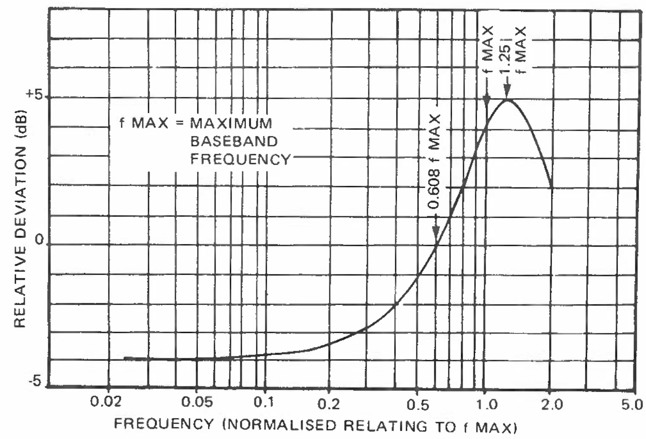


FIG. 26. TELEPHONY PRE-EMPHASIS CHARACTERISTIC

The standard CCIR pre-emphasis characteristic for a telephony bearer is shown in Fig. 26. This characteristic applies to broadband bearers of any channel capacity. The characteristic gives a 4 dB increase in the level of the maximum baseband frequency and a 4 dB decrease at the minimum baseband frequency. The deviations for channels at these frequencies are modified by the same amounts.

At a frequency 0.608 of the maximum baseband frequency, no level or deviation change occurs and this frequency is often used for test purposes so that adjustments of deviation can be made without disconnecting the pre-emphasis network. The pre-emphasis characteristic is such that the RMS deviation of the total baseband signal is the same with and without pre-emphasis and the bandwidth requirements of the system are not changed.

6.14 SPECTRAL DISTRIBUTION OF NOISE DUE TO ERRORS IN BEARER CHARACTERISTICS. The noise power due to errors in the amplitude-frquency response characteristics is distributed over the baseband approximately in proportion to the cube of the baseband frequency and is most noticeable in high frequency telephone channels, particularly in high capacity bearers. The noise distribution is modified by the de-emphasis network but it is still only the high frequency channels that are affected significantly. This means that faults which produce amplitude-frequency response errors, mainly affect the high frequency channels of a broadband carrier system. Changes in the amplitude-frequency response would invariably be accompanied by changes in the phase-frequency characteristic. Small changes in the latter normally produce relatively large amounts of intermodulation noise and usually give the first indication of changes in the system characteristics.

With telephony traffic, the noise generated by modem non-linearity is not evenly distributed over the baseband. Typically, for one modulator-demodulator pair, channels at the low frequency end of the baseband have a S/N due to modem non-linearity that is 3 to 6 dB poorer than the high frequency channels. When a fault or maladjustment condition causes a modulator-demodulator pair to become excessively non-linear, the additional intermodulation noise produced degrades the normal signal-to-noise ratios of the channels. This is particularly so when the modem noise, with its peculiar spectral distribution, becomes comparable with noise originating in other sections of the equipment. In a typical multi-hop system operating with a S/N that is greater for the low frequency channels than for the high frequency channels, a gradual modem degradation shows up first in the low frequency channels, but when the signal-to-noise ratios for all channels are normally very similar, the degradation appears in all channels. Should the modem linearity be so bad that intermodulation noise is the major noise contribution to the system, all channel signal-to-noise ratios become poor, with the low frequency channels a few dB worse than the high frequency channels. When the de-emphasis network is taken into account, the low frequency channel signal-to-noise ratios are further degraded by a few dB and the high frequency channel signal-to-noise ratios improved by a few dB. The changes are not necessarily the ±4 dB of the emphasis networks, because pre-emphasis of the input signal can slightly change the magnitude and distribution of the intermodulation noise produced.

The intermodulation noise resulting from linear and parabolic group delay errors has a baseband spectral distribution that is approximately proportional to frequency. The S/N is worst in the highest frequency channel and about 4 to 6 dB better in the channels in the centre of the baseband. Group delay errors have negligible effect on the signal-to-noise ratios of the lowest frequency channels. The high frequency channel S/N is improved by the de-emphasis network. If noise due to group delay errors is the predominant noise in the system, the S/N improvement produced by de-emphasis makes the S/N in the channels in the upper two thirds of the baseband spectrum approximately constant. Errors in the group delay characteristics that produce curvatures of a higher order than parabolic, affect mainly the high frequency end of the baseband spectrum, even when de-emphasis is taken into account. For this reason, fault conditions which cause group delay errors are characterised by an increase in the noise level in test channels in the centre and upper sections of the baseband spectrum, with perhaps a slightly greater effect in the high frequency test channel for some types of errors.

6.15 NOISE IN A TELEVISION CHANNEL. In a similar manner to telephony channel noise, the noise in a television channel which results from receiver thermal noise is dependent on receiver S/N, the ratio of receiver noise bandwidth to television channel bandwidth, and the system deviation relative to the modulating signal frequency. However, for television, additional factors must be taken into account.

The wide bandwidth of the television channel means that the triangular noise distribution over the band must be considered. Also, the deviation for television is quoted as a peak-to-peak value for the peak-to-peak composite video signal. However, S/N is specified using the peak-to-peak picture signal only (not including sync. pulses) as the reference. A convenient way of expressing S/N resulting from receiver thermal noise is:

Luminance Channel

$$\text{Signal-to-noise ratio} = 20 \log \frac{\text{peak-to-peak picture signal (in volts)}}{\text{RMS noise (in volts)}}$$

$$= \frac{C}{N} + 10 \log B + 20 \log \Delta f\,(p\text{-}p) - 30 \log f_V + 1.7 \text{ dB}$$

where - $\frac{C}{N}$ = Carrier-to-noise ratio in dB

B = Receiver noise bandwidth in same units as used when calculating carrier-to-noise ratio (C/N)

Δf (p-p) = Peak-to-peak deviation for a peak-to-peak composite video signal

f_V = Maximum video frequency

Note: f(p-p) and f_V must be in the same units.

For example assume that a television bearer has the following characteristics:

Carrier-to-noise ratio	=	59.9 dB (para. 4.12, no fade condition)
Receiver noise bandwidth	=	40 MHz (para. 4.9)
Peak-to-peak deviation	=	8 MHz
Maximum video frequency	=	5 MHz

Then:

T.V. S/N = 59.9 + 16.0 + 18.1 - 21.0 + 1.7

= 74.7 dB

The mean signal-to-noise ratio, assuming a 4 dB fade allowance, is 70.7 dB. Because of the triangular distribution of noise over the video signal bandwidth, most of the high amplitude noise appears at high video frequencies which produce the fine picture detail. The noise is displayed on the TV screen as a random pattern of small dots (snow). The subjective effect of this form of noise is that it causes less interference than a similar amplitude low frequency interfering signal. So that the value of S/N gives an indication of the annoyance of the interference, a weighting network is included in the measuring system. If the noise distribution over the bandwidth is triangular, the weighting network gives a S/N improvement, for a 625 line, 5 MHz television system, of 16.3 dB. This means that the system in the preceding example has a weighted mean S/N of 87 dB.

6.16 TELEVISION PRE-EMPHASIS. Pre-emphasis is used for television bearer systems. It gives a small improvement in high frequency noise, but low frequency noise, particularly line and field rate interference, is increased. A more important reason for including pre-emphasis is that it makes the asymmetrical video waveform closer to being symmetrical. This allows the same bearers to be used for either telephony or television and allows the carrier frequency stability to be controlled by an AFC system which senses the mean carrier frequency. Without pre-emphasis, a television link must have the black level maintained at a fixed frequency, effectively transmitting DC, if the standard deviation is to be used and the frequency components are to be maintained in the available bandwidth.

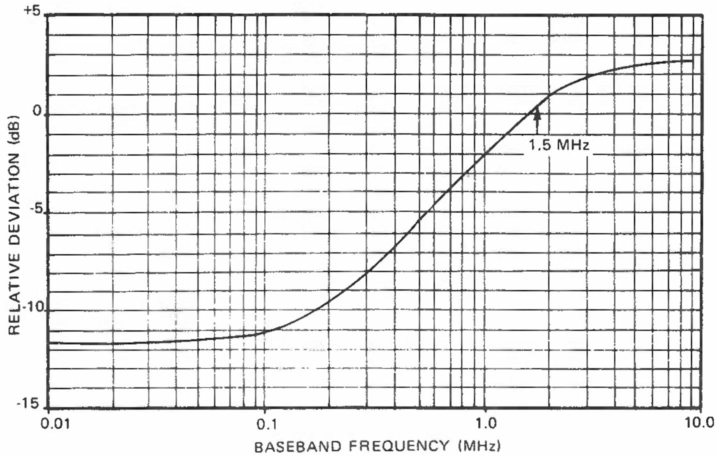
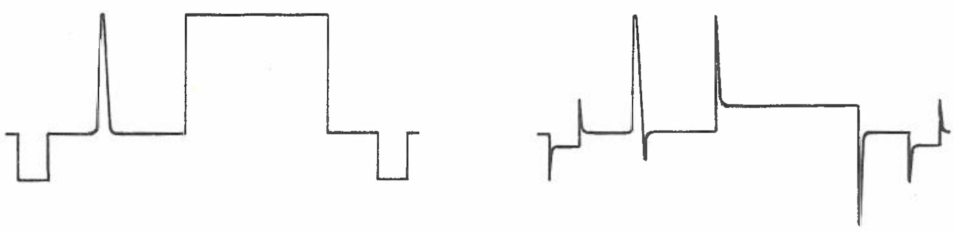


FIG. 27. TELEVISION PRE-EMPHASIS CHARACTERISTIC

The standard pre-emphasis characteristic (CCIR) for the transmitting terminal of a 625 line television bearer system is included in Fig. 27. It shows an increase of approximately 2.5 dB at high video frequencies and a decrease of about 11.5 dB for the low frequency end including the line frequency. At a frequency of 1.512 MHz, the pre-emphasis does not affect the deviation. A 1V peak-to-peak sine wave of this frequency can be used to adjust the system deviation to the standard 8 MHz peak-to-peak, without disconnecting the pre-emphasis network. The receiving end of the bearer includes a de-emphasis network with a characteristic that is the complement of that shown in Fig. 27.

6.17 EFFECT OF PRE-EMPHASIS ON TELEVISION WAVEFORM. Fig. 28 shows the effect of a television pre-emphasis network on a 2T pulse and bar test waveform. The output waveform includes large amplitude overshoots following transitions. Notice that the amplitude of the major section of the bar of the output waveform is approximately one quarter of the amplitude of that section of the input waveform. This corresponds to the low frequency amplitude reduction of almost 12 dB because of the pre-emphasis characteristic.



(a) Input (b) Output

FIG. 28. EFFECT OF TELEVISION PRE-EMPHASIS NETWORK

6.18 SYSTEM VALUE. In the preceding calculations to determine the thermal noise expected in a broadband bearer circuit, many of the factors are the same for each hop of a system where the same radio equipment is used at each station. The main variables in the calculations are path length and therefore the propagation loss, waveguide length and type and therefore its attenuation and, sometimes, the type of antenna used at each station. These factors are the main losses and gains between the transmitter output and the receiver input. To simplify calculations for each hop of a system, all the remaining factors are combined to indicate a figure of merit for the system known as the system value (V_S).

Because noise depends on the type of channel provided (telephony or television) and, for a telephony channel, also on its baseband frequency allocation, the system value varies for these different conditions. The main difference between the system value and S/N formulae is that C/N in the latter formula, which represents the ratio of receiver input signal power to receiver input noise, is replaced in the system value formula by the ratio of transmitter power to receiver input noise. Hence, for a telephony system:

System value (v_S) = $P_T - P_E + 10 \log B - 10 \log b + 20 \log \Delta f\ (\text{RMS}) - 20 \log f_M$

where -
- P_T = Transmitter power output in dBW
- P_E = Receiver equivalent source noise in dBW
- B = Receiver noise bandwidth
- b = Channel bandwidth
- Δf (RMS) = RMS channel deviation
- f_M = Baseband frequency corresponding to the channel centre frequency

For a system with characteristics as assumed in paras. 4.11 and 6.9:

$$V_S = 7 - (-118) + 46.0 - 4.9 + 46.0 - 71.8 \text{ dB}$$
$$= 140.3 \text{ dB}$$

To determine the channel S/N, it is necessary only to subtract the total loss between the transmitter and receiver from the system value. Thus:

Channel S/N $= V_S - L$
$= 140.3 - 65.1$
$= 75.2$ dB

where - V_S = System Value,
L = Total loss between transmitter output and receiver input

This agrees with the no-fade value determined in para. 6.9.

In a similar manner, a system value can be determined for a television bearer.

6.19 TELEVISION SOUND CHANNEL. When the thermal noise in the subcarrier sound channel of a television bearer is to be determined by calculations, it is necessary to take into account both the primary and the secondary deviations and bandwidths. The development of the formula is not included in this paper, but a simple formula applying to a subcarrier sound channel with standard specifications is included. The standard subcarrier sound channel has the following specifications:

Subcarrier frequency	= 7.5 MHz
Maximum audio frequency	= 15 kHz
Subcarrier deviation by audio frequency	= 140 kHz (RMS)
RF deviation by subcarrier	= 300 kHz (RMS)

For these specifications:

Sound channel signal-to-noise ratio $= \frac{C}{N} + 10 \log B - 15.55$ dB

For a system with:

C/N = 59.9 dB (para. 4.12, no-fade condition)

B = 40 MHz (para. 4.9)

Then:

Sound channel S/N = 59.9 + 46.0 - 15.55 dB

= 90.35 dB

The CCIR specifies the characteristic of a weighting network for measurement of noise in a programme channel. Whilst this weighting network varies the noise distribution over the channel bandwidth, it does not significantly change the signal-to-noise ratio resulting from receiver thermal noise.

7. CHANNEL NOISE DUE TO LOADING

7.1 EFFECT OF CHANNEL SIGNAL LEVEL ON NOISE. The manner in which the channel S/N, including both basic and intermodulation noise, varies with the carrier signal level is shown by the curves in Figs. 21 and 24. These curves show how a system behaves under fade conditions. Values for these curves cannot conveniently be obtained after a system has been put into service, and also the curves, particularly for multihop systems, are not the most useful for maintenance purposes for determining the source of noise degradation in the system. More useful information about a system can be obtained by plotting curves indicating the variation of the channel S/N with variation in channel signal level. In practice, equivalent curves are plotted from the results of measurements using a variable level white noise test signal.

For example, assume that the channel signal level in all channels is well below the normal signal level at the channel combining point in the system. This means that the amplitude of the total baseband signal is lower than normal and, with the modulators and demodulators set to operate with standard levels and deviations, the output channel signal level is also correspondingly lower than normal. For a given carrier signal level input to the receivers of the system, a particular basic noise power appears in the output channels. The reduced channel signal level results in a low S/N due to basic noise. The low input signal levels means that overload in the baseband equipment is unlikely, only a small section of each of the modulator and demodulator characteristics is used, the peak deviation is low and not likely to exceed the system bandwidth, and the group delay variation over the frequency band required is likely to be small. For these reasons very little intermodulation noise is generated under low channel signal level conditions. The effect of channel signal level on channel signal-to-noise ratio is shown in Fig. 29.

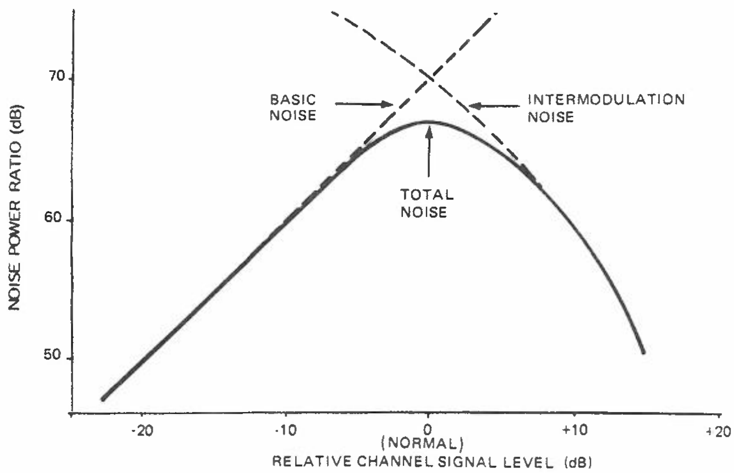


FIG. 29. EFFECT OF CHANNEL SIGNAL LEVEL ON NOISE

With the basic noise power unchanged, the channel S/N improves proportionally as the channel signal level is increased. However, the increase in the channel signal level gives an increase in the intermodulation noise produced by system non-linearities that depends directly on channel signal level and deviation. This includes all sources of intermodulation noise except that produced by long delay echoes. Above the normal channel signal level, and approaching the overload point of the system, the intermodulation noise increases rapidly and the signal-to-intermodulation noise ratio decreases correspondingly (Fig. 29).

The channel signal-to-total noise ratio peaks at the channel signal level where the signal-to-basic noise equals the signal-to-intermodulation noise ratio. To the left of the peak, S/N is controlled mainly by the basic noise and to the right of the peak, intermodulation noise is dominant. For optimum S/N, the peak in the characteristic occurs close to the normal channel signal level. This condition results from a balance between system standards, system design and equipment design.

7.2 EFFECT OF CARRIER SIGNAL LEVEL. The S/N of a bearer varies for different channels in the baseband and also with changes in the receiver input carrier signal level. To explain the effect of these parameters on S/N with varying channel signal level, we will assume that a single hop system has a S/N variation for changes in carrier signal level as shown in Fig. 30. The effect of pre-emphasis is included in the characteristics.

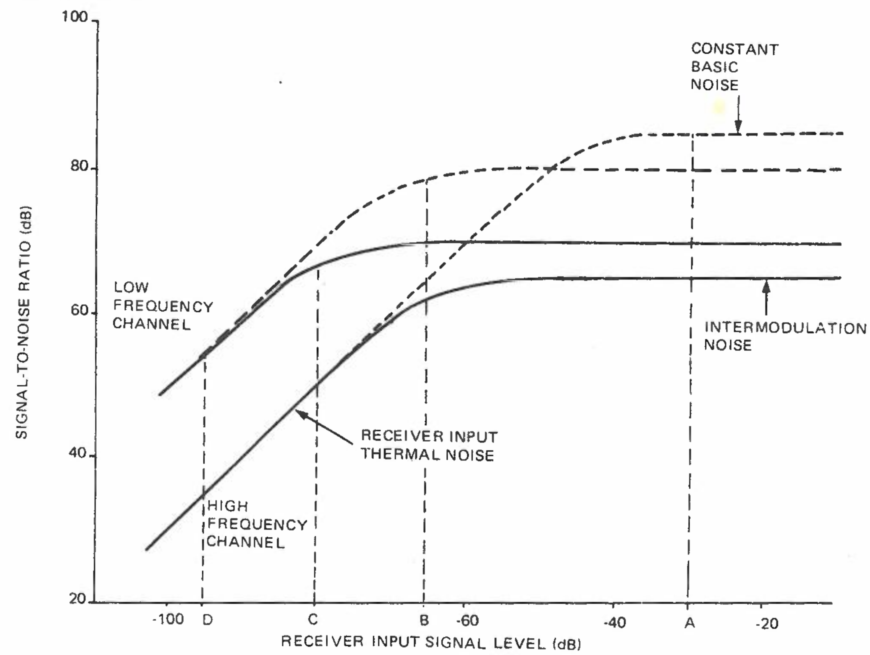


FIG. 30. SIGNAL TO NOISE - RECEIVER SIGNAL LEVEL CHARACTERISTIC

At low receiver input signal levels where receiver input thermal noise is the controlling factor, the S/N is much greater for the low frequency channel than the high frequency channel. At high receiver input signal levels, the constant basic noise is slightly better in the high frequency channel than the low frequency channel and the intermodulation noise for mean busy hour traffic is slightly poorer in the high frequency channel than the low frequency channel.

The curves in Fig. 31 show the variation in S/N with changes in channel signal level for a system operating at successively lower receiver input signal levels.

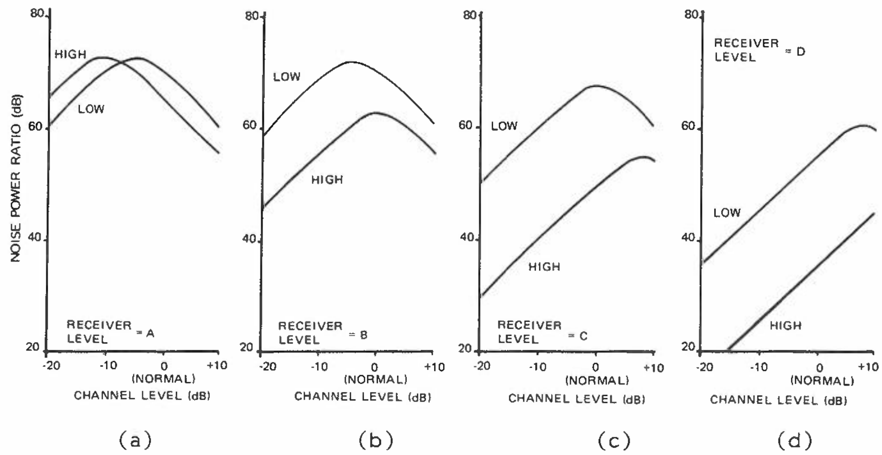


FIG. 31. SIGNAL TO NOISE RATIO-CHANNEL SIGNAL LEVEL CHARACTERISTIC

The main points determining the form of the characteristics are now examined. Fig. 31a shows the characteristic corresponding to a receiver input signal level 'A' in Fig. 30. At this level, with mean busy hour traffic, intermodulation noise predominates and the S/N peak (Fig. 31a) occurs at a lower than normal (0 dB) channel signal level. To the left of the peak, S/N decreases with reduction in the channel signal level. In this region basic noise is the controlling factor. At level 'A', (Fig. 30), the signal-to-basic noise ratio in the high channel is slightly better than in the low channel. This is also shown by the section of the curve in Fig. 31a to the left of the S/N peak. To the right of the peak, where intermodulation noise is the controlling factor, the S/N decreases with increase in channel signal level.

With the receiver input signal at level 'A' (Fig. 30) and normal channel signal level, corresponding to mean busy hour traffic, the S/N due to intermodulation is slightly poorer in the high frequency channel than in the low frequency channel. This condition is also shown in Fig. 31a.

The S/N (Fig. 31a) is shown as being degraded by intermodulation in the same proportions in both high and low frequency channels but this is not necessarily so. The characteristic in this region depends on the source of the intermodulation noise and its frequency distribution. A more detailed analysis, taking into account the changing channel signal level, allows the values of S/N in Fig. 31 to be determined from Fig. 30 and these are indicated by the scales in Fig. 31.

Fig. 31b illustrates the channel signal level characteristic for a receiver input signal level equal to 'B' in Fig. 30. At this level, basic and intermodulation noise contribute equally to the S/N in the high frequency channel and the curve for this channel (Fig. 31b) peaks at the normal channel signal level. Intermodulation noise still predominates in the low frequency channel and its peak is to the left of the normal channel signal level.

The signal-to-basic noise ratio for the low frequency channel is slightly reduced from that at level 'A' because the receiver input thermal noise has become comparable with the constant basic noise (Fig. 30). In the high frequency channel, receiver input thermal noise has taken over control of the signal-to-basic noise ratio which is considerably reduced from that at level 'A'. These changes may be seen by comparing the left sections of the curves shown in Figs. 31a and 31b. Since intermodulation noise is not dependent on the receiver input signal level, a change in this level does not affect the sections of the curves in Fig. 31 where intermodulation noise is the controlling factor, and so the extreme right sections of Figs. 31a and 31b are the same.

When the receiver input signal level is reduced to 'C' (Fig. 30), receiver input thermal noise controls the basic noise in both low and high frequency channels and the signal-to-basic noise ratios are lower than at level 'B'. The left sections of the characteristics for level 'C' (Fig. 31c) illustrate this condition. With the intermodulation noise controlled section of the characteristics unchanged, the peak signal-to-noise ratios shift towards the higher channel signal levels. The low frequency channel now has a peak S/N when the channel signal level is normal.

For a further reduction in the receiver input signal level to 'D', (Fig. 30) the signal-to-noise ratios at normal channel signal levels in both channels are controlled by basic noise originating from thermal noise in the receiver input. Also, the signal-to-noise ratios are less than at level 'C'. This is confirmed by the characteristics for this level in Fig. 31d where the normal channel signal level is to the left of the S/N peaks and is less than in Fig. 31c.

7.3 MULTI-HOP SYSTEM. The examination of the variation of S/N in para. 7.2 used a single hop system as an example. For a multi-hop system, the channel signal level characteristics are influenced by parameter changes in the same way. However, the relationship between the channel signal level and the receiver input signal level characteristics cannot conveniently be obtained because, as explained in para. 6.11, the knee in the receiver input signal level characteristic for a multi-hop system does not necessarily occur when the basic noise equals the intermodulation noise.

7.4 SUMMARY. We will now summarise, in general terms which apply to both single and multi-hop systems, the changes in the channel signal level characteristic (Fig. 31) produced by changes in the receiver input signal level. The characteristic to the left of the S/N peak is dependent on basic noise. When the signal level input to all receivers is high, constant basic noise predominates and normally both the high and low frequency channels have similar signal-to-noise ratios. As the input signal level to a receiver is reduced, receiver input thermal noise takes control and the signal-to-noise ratios for a particular channel signal level are reduced. The high frequency channel is affected first. When receiver input thermal noise controls the basic noise on both channels, the high frequency channel has a lower S/N than the low frequency channel. The difference depends mainly on the relative baseband frequencies but is modified by pre-emphasis. Intermodulation noise controls the characteristic to the right of the S/N peak and this section is not changed by changes in receiver input signal level. Because of this the S/N peak moves towards the high channel signal levels when the signal-to-basic noise ratio falls as a receiver input signal level falls. This is accompanied by a decrease in S/N at the peak and at the normal channel signal level.

Where possible, systems are normally designed with mean receiver input signal levels so that with mean busy hour traffic, the peak S/N in the worst channel occurs slightly below the normal channel signal level. Optimum conditions are then closely maintained for small fades.

7.5 TRAFFIC DENSITY. During times when the traffic is light the total baseband amplitude is below normal even though the channel signal levels are normal. This reduces the intermodulation noise and improves S/N for high channel signal levels. The peak in the S/N characteristics is shifted towards a higher channel signal level as shown in Fig. 32. This condition is sometimes found when a bearer is used with fewer channels than it was designed for.

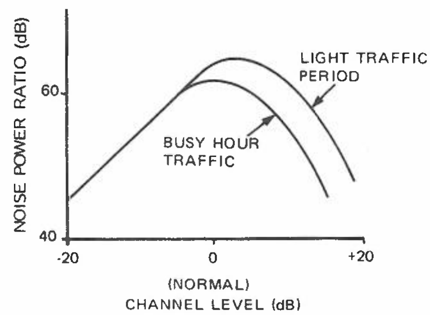


FIG. 32. EFFECT OF TRAFFIC DENSITY ON SIGNAL-TO-NOISE RATIO

8. INTERFERENCE NOISE

8.1 Thermal noise and intermodulation noise have been considered in the previous sections, but radio-relay systems are also subject to additional noise interference. This additional noise, which depends on the distribution of the radio-frequency modulated spectrum, can arise from many different sources. Some of these sources such as receiver selectivity, equipment screening, and antenna directivity, are under the control of the equipment designer. Other sources such as unwanted couplings between channels due to choice of frequencies, site considerations, foreground reflections, overshoot signals, or inadequate angular separation between crossing routes are under the control of the system planner. In addition, the system planner has the responsibility of deciding the acceptable level of interference, bearing in mind both the economic and engineering aspects of the problem. No radio-relay system can be completely free from interference and all practical installations are a compromise between what is desirable and what is economically achievable. Various types and sources of interference noise are considered below.

8.2 TYPES OF INTERFERENCE. Generally, the types of radio frequency interference may be classified by one of the following terms:

:: CO-CHANNEL INTERFERENCE. This term refers to interference from a source, modulated or otherwise, having a frequency the same as or close to that of the wanted carrier. When the interference is caused by the beat between two relatively high level carrier components having the same frequency or a frequency separation which falls within the baseband of the wanted signal, the predominant interference is single-tone in character. When the carrier frequency separation falls outside the baseband range of the wanted signal, and when the interference is from a dispersed signal, the character of the interference resembles that of random noise

:: ADJACENT CHANNEL INTERFERENCE. This term refers to interference due to the presence of one or more radio-frequency carrier, modulated or otherwise, immediately adjacent in frequency to that of the wanted modulated carrier. The term adjacent is normally taken as indicating the adjacent channel in the frequency channelling plan which has been adopted

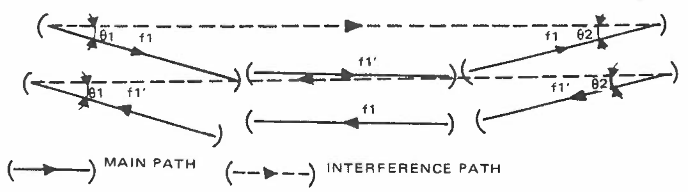
:: DIRECT ADJACENT CHANNEL INTERFERENCE. As in the case above, this term refers to interference due to the presence of adjacent carriers. The mechanism of the interference, however, requires the interfering carrier to be modulated, the modulation apparently being transferred to the wanted carrier

:: OTHER FORMS OF INTERFERENCE. This term refers to interference which can arise from external sources, or from unwanted couplings within the radio equipment such as, for example, the image response of the receiver

8.3 CO-CHANNEL INTERFERENCE. Co-channel interference arises when there is excessive coupling between signal sources operating on the same, or adjacent frequencies. Because the number of operating frequencies available for use on a radio route is limited, it is current practice to transmit a common carrier frequency (F1) from every alternate repeater section of a radio-relay system. The remaining stations of the system transmit a second carrier frequency (F1') which differs from the first by an amount which depends on the particular frequency plan employed (see Figs. 33 and 34). Such a frequency arrangement is known as the 'two frequency plan'. This arrangement can result in overreach interference where station A transmitting on frequency F1 illuminates the antenna of station D, and vice versa (see Fig. 33).

Protection against the above type of interference is normally provided by careful site selection, and by ensuring that the antenna side-lobe suppression given by the angles d_1 and d_2 are sufficient to reduce the coupling to insignificance. In difficult cases it may be possible to reinforce the measures mentioned by arranging that the polarization of the wanted and unwanted signals are orthogonal, but this is not always practicable if, for example, a spur route branches off from one of the stations and it is desirable to provide cross-polar discrimination between the main route and the spur route.

An alternative method of reducing interference is to employ an interleaved frequency plan over the section that is subject to interference. In this context, an interleaved frequency plan is one in which all carrier frequencies are changed in the same sense by an amount equal to half the adjacent channel frequency spacing.



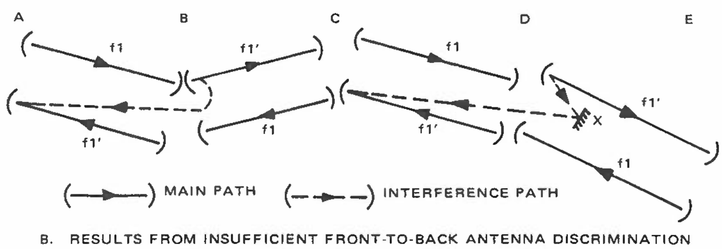
Note: The ratio (wanted signal/interfering signal) depends on the gain of the antennae at angles ϕ_1 and Φ_2 relative to the gain in beam and the difference of propagation losses between main and overreach paths.

FIG. 33. OVERSHOOT INTERFERENCE

The advantage to be gained by this arrangement depends on the baseband width of the channels concerned relative to their carrier spacing and also their relative carrier levels. It is generally desirable to use a carrier frequency spacing of not less than three times the highest baseband frequency in order that the first-order sidebands of one channel do not overlap the second-order sidebands of the other channel. Closer spacings in terms of multiples of the highest baseband frequency may be used with care, depending on the relative levels of the wanted and unwanted carriers and the degree of interference that is acceptable.

It can be shown that for a 960 telephone channel system, a carrier frequency shift of 14.5 MHz could result in a reduction in the level of interference by over 30 dB compared with the co-channel case. For systems of lower capacity the improvement may be even greater, but for systems of greater capacity and the same frequency shift, the possible improvement reduces and finally becomes zero for an 1800-channel system.

A further interference mechanism can be seen in Fig. 34 where radiation at frequency F1' from the back of the antenna at station B is received at station A. Even when the front-to-back discrimination of the antenna is sufficient to reduce this interference to an acceptable level, reflections from nearby objects can cause coupling, as shown at station D (Fig. 34). The reduction of interference from such a coupling must rely either on antennae of improved directivity or on the use of a more suitable site, or possibly on the use of an interleaved frequency plan. A front-to-back antenna discrimination of 65 dB, or better, at each station is necessary in order to limit co-channel interference to an acceptable level.



B. RESULTS FROM INSUFFICIENT FRONT-TO-BACK ANTENNA DISCRIMINATION
D. RESULTS FROM A REFLECTION FROM NEARBY TREES, HILL OR BUILDING, ETC. AT POINT X.

FIG. 34. ADJACENT PATH INTERFERENCE

Fig. 35 shows an example of interference from a spur or crossing which is similar to one of the interference mechanisms shown in Fig. 34 except that in the case of Fig. 35 the angles involved are less, and it is the front-to-side polar discrimination of the antennae which controls the level of interference. If, on the main route, signals of the same frequency arriving from opposite directions are co-polar, the spur route can be cross-polarized with respect to both directions. If however, the main route signals are cross-polar, the spur route may be cross-polarized with respect to the direction having the least angular separation.

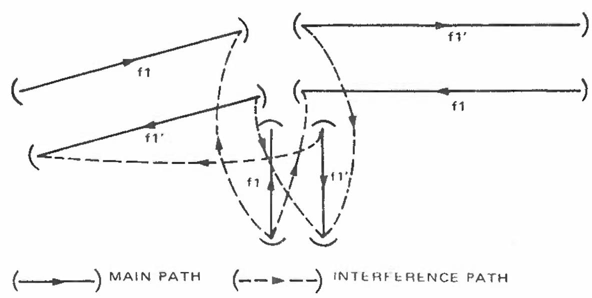


FIG. 35. SPUR ROUTE INTERFERENCE

8.4 ADJACENT CHANNEL INTERFERENCE. Adjacent channel inteference can arise when (for a given baseband capacity and adjacent channel separation) suppression of the adjacent carrier and overlapping unwanted sidebands is inadequate. The reduction to an acceptable level of interference from this source is dependent upon several factors. Firstly, adequate separation of the adjacent channel sidebands must be provided, and this is taken into account in the radio-frequency channel arrangements recommended by the International Radio Consultative Committee (CCIR). These RF channel arrangements have been adopted by Telecom Australia. Secondly, both RF and IF selectivity must be provided to reduce the level of unwanted adjacent sidebands. In providing such filtering the equipment designer must take into account the possibility of introducing distortion into the wanted signal path and of overloading the receiver mixer. Thirdly, cross-polar discrimination between adjacent channels is provided to supplement the selectivity given by the filtering. A cross-polar discrimination of some 25 to 30 dB is both practicable and necessary.

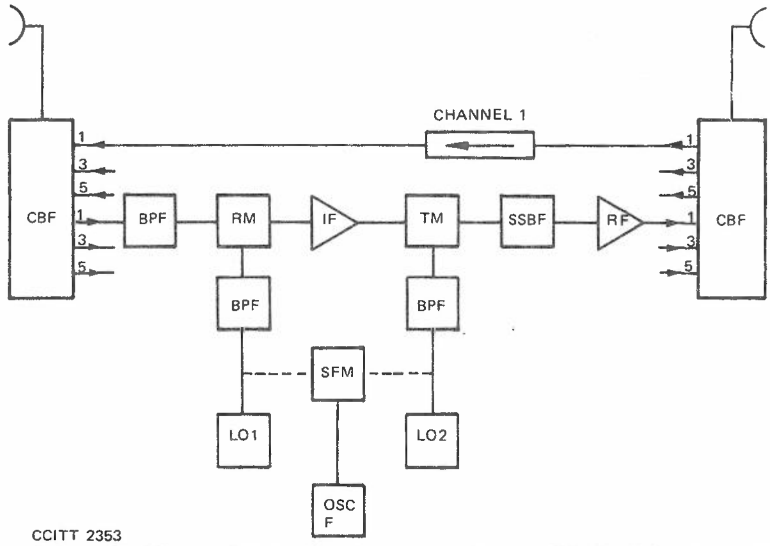
8.5 DIRECT ADJACENT CHANNEL INTERFERENCE. Direct adjacent channel interference arises when a modulated unwanted carrier is immediately adjacent in frequency to that of the wanted unmodulated carrier and, as in the case (para. 8.4) above, the suppression of the unwanted carrier and its overlapping sidebands is inadequate. The modulation of the unwanted carrier is apparently transferred to the wanted carrier. At the discriminator, the unwanted modulation appears at the output, along with the wanted signal. The effect is characterised by the fact that intelligible crosstalk is produced, and that the level of interference varies by 2 dB for each 1 dB variation in the wanted-to-unwanted carrier ratio. The problem of reducing interference from this source to an acceptable level is again one for the equipment designer who must ensure that, together with the cross-polar discrimination between adjacent channels, adequate selectivity is provided prior to limiting.

8.6 OTHER SOURCES OF INTERFERENCE. Interference can arise from within the radio equipment itself, or from sources external to the radio system. The latter sources include communication-satellite systems, which share the same frequency bands as line-of-sight radio-relay systems, radar installations, whose harmonics and/or spurious radiations fall within the radio-relay system frequency bands and radio and television broadcasting stations, whose radiated frequencies (or their harmonics) fall either into the IF, RF or baseband range of radio-relay systems.

To facilitate frequency sharing between communication-satellite systems and terrestrial radio-relay systems, the CCIR has evolved a number of Reports and Recommendations. Some of the latter place constraints on such characteristics as EIRP and power flux density, so that levels of interference can be limited to an agreed value.

Interference from such sources as radar installations and broadcasting stations can be minimized by careful consideration of matters such as the choice of radio-communication station sites, directions of shoot, screening of antennae and equipment rooms, in the selection of frequencies to be assigned, and any other aspect pertinent to the particular conditions involved.

Interference can occur within a system as a result of deficiencies in the equipment itself. Adequate selectivity prior to the low-level (receive) mixer (RM in Fig. 36) is necessary to densensitize the receiver in all regions but that of the channel carrier frequency and, in addition, both the local oscillator output (L02 in Fig. 36) and the unwanted sideband at a point beyond the medium-level (transmit) mixer (TM in Fig. 36) should be considerably attenuated. Further, the oscillator feed to the low-level mixer, whether derived directly from L01 or from OSC F via the shift-frequency mixer (SFM), can appear as an interfering signal in other receivers sharing the same feeder installation (particularly if harmonics of the basic crystal oscillator frequency are present in the feed) unless steps are taken to prevent this.



CCITT 2353

CBF = CHANNEL BRANCHING FILTERS	TM = MEDIUM LEVEL (TRANSMIT) MIXER
BPF = BANDPASS FILTER	LO = LOCAL OSCILLATOR
SFM = SHIFT FREQUENCY MIXER	SSBF = SINGLE-SIDE BAND FILTER
IF = INTERMEDIATE - FREQUENCY AMPLIFIER	RF = RADIO-FREQUENCY AMPLIFIER
RM = RECEIVE MIXER	OSCF = LOCAL OSCILLATOR FOR SFM.

Note: The block marked channel 1 is a replica of the repeater shown in more detail for the opposite direction of transmission. Channels 3 and 5 will comprise similar repeaters operating on different radio frequencies.

FIG. 36. OUTLINE OF A TYPICAL REPEATER FOR SEVERAL TWO-WAY RADIO CHANNELS

The design and maintenance of equipment must also ensure that the outputs of all local oscillators are of high spectral purity, i.e. as free as possible from spurious signals, random noise and both long-term and short-term frequency changes. Spurious signals in local oscillators can arise from imperfections of the crystal itself, inadequate filtering of undesired crystal harmonics or self-oscillation and general instability in varactor multiplier chains.

A further interference mechanism exists on systems which employ common antennae and feeders to transmit and receive. All waveguide feeders exhibit amplitude non-linearity to some extent, mainly resulting from imperfections at joints. Such non-linear elements result in intermodulation between the outputs of two or more transmitters, and the resulting unwanted products may fall close to receive frequencies either in the same frequency band or in a different frequency band from that of the originating transmit channels.

Careful attention to all feeder connections, tuning screws etc. is necessary.

9. SYSTEM NOISE DISTRIBUTION

9.1 HYPOTHETICAL REFERENCE CIRCUITS. A hypothetical reference circuit is used as a basis for determining the transmission characteristics of equipment in the planning stage and for calculating expected noise in the design of new systems. The hypothetical reference circuit is an artificial model of a long-distance circuit or chain of circuits having a defined length and a defined number of pairs of translating equipment for groups, supergroups, etc. The CCIR specifies hypothetical reference circuits for various types of services including terrestrial radiocommunication systems for both telephony and television. These circuits are used by equipment designers and system planners as guides in planning radio systems.

9.2 REFERENCE CIRCUIT FOR TELEPHONY. The CCIR recommends a hypothetical reference circuit for use in FDM broadband radio systems. This is a complete circuit (between an audio frequency terminal at each end) established over a hypothetical broadband radio system of 2500 km length. Such a circuit is shown in Fig. 37.

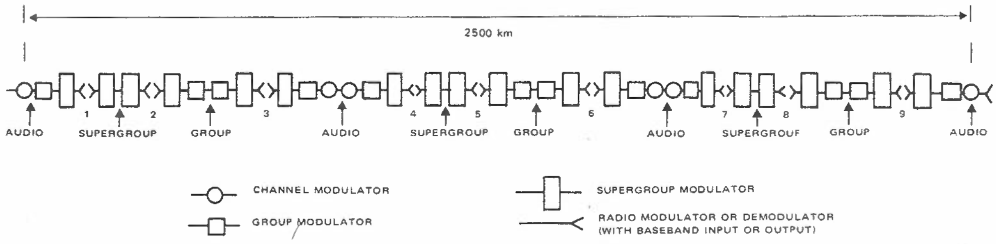
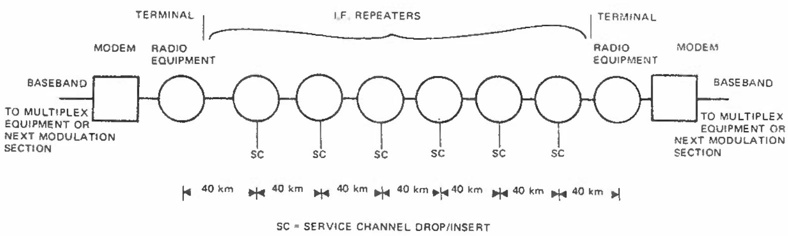


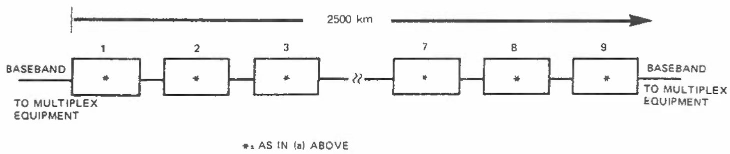
FIG. 37. CCIR HYPOTHETICAL REFERENCE CIRCUIT
(RADIO BROADBAND TELEPHONY BEARER)

The hypothetical reference circuit comprises a number of modulations and demodulations of the channels, groups and supergroups. The circuit includes 9 sets of radio modulators and demodulators for each direction of transmission and these divide the system into 9 'homogeneous' sections of equal length. A homogeneous section is a section without branching or modulation of any supergroup, group or channel established over the system.

9.3 AUSTRALIAN TELEPHONY REFERENCE CIRCUIT. Telecom Australia has established a reference circuit for broadband radio systems which has a slightly different form from the CCIR hypothetical reference circuit. The performance specifications for radio systems used by Telecom Australia for broadband telephony relate to the following circuit configurations:



(a) Reference Modem Section



(b) National Reference Telephony Circuit (NRC)

FIG. 38. AUSTRALIAN TELEPHONY REFERENCE CIRCUITS

:: REFERENCE MODEM SECTION. This section comprises two terminals and six I.F. repeaters (Fig. 38a). Each of the seven hops is nominally 40 km long and each terminal and repeater includes service channel insert and dropping facilities

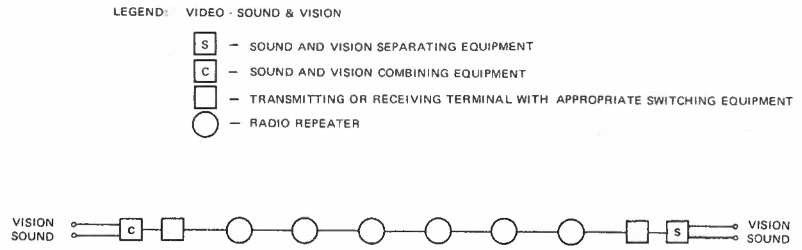
:: NATIONAL REFERENCE TELEPHONY CIRCUIT (NRC). This circuit comprises nine reference modem sections connected in series, with each modem section as described above. The NRC is nominally 2500 km in length (Fig. 38b)

9.4 TELEVISION REFERENCE CIRCUIT. In a similar manner to reference circuits for telephony, Telecom Australia specifies reference circuits for a radio bearer carrying a television signal. The performance specifications for radio bearers used for television relate to the following configurations:

:: REFERENCE VIDEO MODEM SECTION. This section comprises two terminals and six IF repeaters (Fig. 39a). Each of the seven hops is nominally 40 km long and each terminal includes sound-vision combining and separating equipment

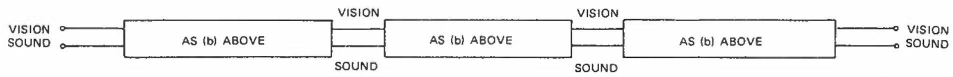
:: NATIONAL REFERENCE VIDEO CIRCUIT SUB-SECTION. This sub-section comprises three modem sections connected in series at baseband, with sound-vision combining and separating equipment at the end terminals only (Fig. 39b)

:: NATIONAL REFERENCE VIDEO CIRCUIT (RADIO RELAY). This circuit comprises three sub-sections, each as described above, connected in series separately for the vision and sound channels (Fig. 39c)



(a) Reference Video Modem Section

(b) National Reference Video Circuit Sub-Section



(c) National Reference Video Circuit (Radio Relay)

FIG. 39. REFERENCE VIDEO CIRCUIT CONFIGURATIONS

9.5 NOISE IN THE HYPOTHETICAL REFERENCE CIRCUITS. The CCIR recommends limits for the noise allowable on circuits similar in composition to the hypothetical reference circuit or to sections of it. Recommendations are also provided for the allowable noise performance of circuits which are not similar to the reference circuit.

Noise recommendations for telephony and television circuits are as follows:

:: TELEPHONY. For telephony, the noise recommendations comprise

a long term target 3L pWOp, (L is circuit length in km)
not to be exceeded for more than 20% of the time

a short term target 47500 pWOp, not to be exceeded for more than

$\frac{L}{2500}$ x 0.1% of the time

a very short term target, 10^6 pWOp not be exceeded for more than

$\frac{L}{2500}$ x 0.01% of the time

In addition, 3L pWOp (when averaged for one hour) is not to be exceeded for 5% of all hours. Margins are allowable when the circuit is not similar to a hypothetical reference circuit.

:: TELEVISION. In the 2500 km hypothetical reference circuit for television transmission, the signal-to weighted noise ratio should fall below the following values:

luminance channel, 56dB for more than 20% of any month, and
44 dB for more than 0.1% of any month

chrominance channel, 50 dB for more than 20% of any month, and
38 dB for more than 0.1% of any month.

Waveform targets are also recommended and certain laws of addition
are included

9.6 NOISE IN AUSTRALIAN REFERENCE CIRCUITS. Apart from very short links, the noise criteria applied to Australian circuits follow the CCIR recommendations. The Australian TV hypothetical circuits may be different from the CCIR circuits, but the noise (and waveform) criteria are such that when a National Reference Video Circuit (NRVC) is formed, the same criteria apply for the overall circuit as for a CCIR Hypothetical Reference Circuit.

9.7 EQUIPMENT NOISE ALLOCATION. The following list gives another survey of the types of noise that can appear in radio communications:

:: Load independent noise (basic noise):

thermal receiver noise
basic noise of RF equipment
basic noise of modem equipment

:: Load dependent noise:

intermodulation noise of RF equipment (including reflections)
intermodulation noise of modem equipment

:: Noise due to other radio channels of the same system or satellite radio (interference noise)

To estimate the total value of noise over a radio link and to determine if this noise is within the specification for the link, the contributions shown above must be added, using the methods for adding noise contributions which were outlined in para. 4.2.

The receiver noise in a radio-link can be calculated relatively simply from the equipment data : transmitting power, noise figure, and phase deviation in conjunction with the total hop attenuation, not however, the load-dependent intermodulation noise.

For the values of these contributions which can be attained with careful development, good manufacture and maintenance, experimental and measured values must be resorted to. Similar conditions hold concerning the load-independent basic noise of the equipment. All these values can be taken from specifications and detailed equipment descriptions. Knowledge of the specific values of a certain radio system is needed in order to calculate the absolute magnitude of the attainable S/N ratio for a certain system.

9.8 DISTRIBUTION OF NOISE TO VARIOUS EQUIPMENT. In assessing the noise for the end of a long radio link, the noise contributions of the devices listed in the previous paragraph must be known by calculation (in the case of the thermal receiver noise) or information of the equipment maker (in the case of the other noise components). A suitable selection of the frequencies and decoupling means must ensure that the mutual interference with use of several RF channels will not exceed permissible values. Isolation must be attained, for instance, by sufficient sidelobe attenuation of the antennae, suitable selection of the polarization, terrain structure, artificial shielding and selectivity in the devices. When planning a radiocommunication system, designers often assume, in calculations of noise performance, that interference noise will contribute about 5 pWOp per hop.

Number of modulation sections		(1)	1	
Number of hops		(2)	7	
Measuring channel		(3)	3886	kHz
Basic Noise	Thermal receiver noise	(4) (5)	-71.5 * 70.8	dBmOp pWOp
	Basic noise of RF equipment	(6)	110 *	pWOp
	Basic noise of modem equipment	(7)	60 *	pWOp
Thermal noise under no-fade condition (sum of 5 + 6 + 7)		(8) (9)	240.8 -66.2	pWOp dBmOp
Intermodulation Noise	RF and IF equipment	(10)	224 *	pWOp
	Modem	(11)	55 *	pWOp
Total intermodulation noise (sum of 10 + 11)		(12) (13)	279 -65.5	pWOp dBmOp
Interference Noise		(14)	35 *	pWOp
Total noise under no-fade condition (sum of 8 + 12 + 14)		(15) (16)	554.8 -62.6	pWOp dBmOp
Permissible total noise		(17) (18)	708 -61.5	pWOp dBmOp
Permissible rise in receiver thermal noise (17-15)		(19)	153.2	pWOp
Permissible thermal receiver noise (sum of 19+5)		(20) (21)	224 -66.5	pWOp dBmOp
Average fading margin (21 - 4)		(22)	5	dB

*Assumed values to show procedure of calculation.

TABLE 5 TYPICAL EQUIPMENT NOISE DISTRIBUTION

Table 5 shows how the noise contributions of the various components are added to give the total permissable noise in one modulation section. The table enables calculation of the noise appearing in the top measuring channel of a 960 channel system under busy hour conditions.

It should be noted that a specification of special data which strongly depends on the equipment would be too comprehensive and specific : only approximate values are given here for general understanding.

From the assumed noise contributions shown in Table 5, the total noise under no-fade conditions (line 15) is 554.8 pWOp or -62.6 dBmOp. This is equivalent to 1.98 pW/km over a reference modem section.

When a receiver is suffering from fading, the receiver input carrier level is reduced. The receiver's automatic gain control (agc) enables the faded carrier to be amplified back to a suitable level. However, in doing so, the receiver also amplifies its own thermal noise. In the output channels of the receiver, the effect of fading is an increase in noise for each channel.

When a receiver signal fade is greater than a predetermined value (known as the fade margin) it is usual to mute the repeater by removing the IF from the following transmitter and supplying, instead, the output of a 70MHz oscillator. This ensures that the affected RF channel becomes very quiet instead of very noisy.

Note: The fade margin of a receiver is the difference (in dB) between the unfaded received signal level, and that signal level at which the C/N = 12dB. In broadband systems, the fade margin is usually 40 to 50 dB on each hop.

END