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Bell System
TECHNICAL REFERENCE

TRANSMISSION PARAMETERS AFFECTING VOICEBAND DATA TRANSMISSION - DESCRIPTION OF PARAMETERS

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Bell System Data Communications

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Transmission Parameters Affecting Voiceband Data Transmission - Description of Parameters

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ENGINEERING MANAGER-ENGINEERING METHODS AND OBJECTIVES

NOTICE

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This Technical Reference supersedes the following Technical Reference (PUB 41008):

"Analog Parameters Affecting
Voiceband Data Transmission —
Description of Parameters —
October 1971"

PREFACE

Common carrier telephone channels, both switched and dedicated, now carry a variety of nonvoice signals. They consist of one or more tones, amplitude and/or phase modulated by either analog or digital information. Each signal format is impaired to varying degrees by the physical limitations, inevitable interferences, and design compromises imposed by nature and economics. Under increasing pressure to meet new service demands and improved quality, communications engineers have developed the methods and apparatus to measure, characterize, simulate, and control the transmission impairments to nonvoice telecommunications. This document collects the most up-to-date information available to describe these impairments. A companion Technical Reference, PUB 41009, describes currently accepted techniques for their measurements.

Transmission characteristics of a telecommunications channel are stratified as follows:

The Basic Channel – Linear Properties – Amplitude, Phase, Impedance Characteristics

Noise – Uncorrelated Interferences (of any origin)

Miscellaneous Impairments – Nonlinear Distortion, Unwanted AM, FM or PM, Rapid Gain or Phase Changes

Impairments that logically fall into these categories are considered in this order. Mathematical descriptions are used only where necessary to avoid awkward or lengthy verbiage. Where appropriate, we comment on the measuring techniques peculiar to the impairment under discussion, including where possible the advantages, weaknesses, ambiguities, and idiosyncrasies of each.

TABLE OF CONTENTS

LIST OF FIGURES

LIST OF FIGURES (Cont'd)

1. INTRODUCTION

1.1 Purpose

1.1.1 The purpose of this document is to provide an understanding of transmission characteristics of Bell System voice grade (nominally 300 to 3000 Hz) facilities and end-to-end connections.

1.1.2 It does not go into requirements or objectives but concentrates on defining transmission impairments. Tutorial and background information is provided where appropriate to point out subtleties of transmission parameters or apparatus. For example, considerable space is devoted to a discussion of envelope delay distortion since much confusion may exist in interpretation of results made with different measuring sets.

1.2 Composition of Transmission Channels Considered

1.2.1 The transmission channel may be composed of:

- two wires
- four wires (one pair for each direction of transmission)
- carrier facilities (equivalent to four wires)
- combinations of the above

In addition there may be other equipment such as amplifiers, attenuators, gain and/or phase equalizers, echo suppressors, switches, hybrid transformers (devices interconnecting 2-wire and 4-wire facilities), and signaling devices. All such equipment affects the analog signal characteristics of the channel.

1.3 Relative Effect of Analog Impairments

1.3.1 An attempt has been made in this Technical Reference to include all analog parameters which might adversely affect data transmission. Depending upon data modem design, many of these parameters may have little practical effect on particular data services. For some of the parameters, tests are made on every channel, whereas for others no test set is presently available. In still others, tests are made only where indicated, i.e., in trouble cases or on a non-routine basis. References 2 and 19 provide information on performance of many of these parameters on the switched telecommunications network; Reference 18 provides information on limits for many of these parameters for private line voiceband data channels.

2. THE BASIC CHANNEL — LINEAR PROPERTIES

Transmission Amplitude and Phase Characteristics

The amplitude and phase characteristics of a transmission medium are of fundamental importance. In contrast to a linear passive network, these quantities are not always readily measurable on a real transmission facility. Furthermore, these characteristics may be signal dependent due to equipment such as compandors. (Syllabic compandors used in analog facilities introduce loss during quiet intervals of speech to reduce noise and subjectively improve the channel for speech. Instantaneous compandors are used in digital facilities to provide an approximately constant signal-to-noise ratio, independent of the signal power over the design range of the facility.)[1] In telecommunications systems, the following three characteristics are typically required to define the basic channel: 1000 Hz loss, frequency response and phase characteristic.

2.1 Loss

2.1.1 The 1000 Hz loss of a channel means the loss, expressed in dB, experienced by a 1000 Hz tone in traversing the transmission medium. The power of the received tone is measured on an averaging type of instrument, calibrated in dBm (power with respect to 1 mw). Due to mistracking in some compandored facilities, the loss may vary by as much as a dB as the test tone power level is changed.

2.2 Amplitude Characteristics

2.2.1 The amplitude characteristic of a network or system is commonly determined by simply measuring the loss of a single frequency tone as it is tuned across the bandwidth of interest. Let this measurement be referred to as the static frequency response. On telecommunications channels, due to level and frequency sensitive devices, and the presence of nonlinearities, a static measurement may not yield the same amplitude characterisitc as that experienced by a complex waveform. Differences of up to 2 dB have been observed on some compandored facilities when using other measuring techniques. These differences arise from the relative placement of filters with respect to level sensitive and nonlinear devices and generally are greatest near the band edges. In these cases, the frequency response is a function of the spectral content of the signal on the channel. This latter response is referred to as the dynamic frequency response. Comparison tests have shown that the dynamic response can be measured by using two tones. One tone, in the vicinity of 1000 Hz is referred to as a holding tone, and serves to activate the channel. The second, or measuring tone, is used to measure the frequency response. The holding tone is held 5 dB above the measuring tone and the response is measured using a frequency selective voltmeter. The composite test tone power should approximate the expected signal power that will appear on the channel when in normal use.

2.2.2 Even though differences (±2 dB) may be encountered between the static and dynamic testing methods, the static one has long been the method of measurement in the telecommunications industry. Because of its simplicity, it will probably continue to be used. The differences encountered, on analog compandored facilities, are generally in a direction such as to make the channel appear to have a narrower bandwidth than it does in the presence of a signal with a broad spectrum.

2.2.3 Frequency response is checked and controlled by the following simple test: the loss on a channel is measured at three different frequencies, 400 Hz, 1000 Hz, and 2800 Hz. The loss at 1000 Hz is then subtracted from the loss at 400 and 2800 Hz. These differential losses are then referred to as the slope at 400 or 2800 Hz. These two slopes are a measure of the frequency response of the channel under test.

2.3 Bandwidth

2.3.1 The bandwidth of a channel is determined from measurements made of the amplitude characteristic. It is defined as the band of frequencies within which the loss is no more than 10 dB greater than the loss at 1000 Hz. Experience has shown that the simple slope tests are sufficient to control the bandwidth on transmission facilities in use. Therefore, bandwidth measurements are seldom made.

2.4 Envelope Delay Distortion

2.4.1 It is difficult to measure the phase characteristic of a transmission system since a phase reference is hard to establish at the receiving end of the circuit and the channel may have a varying zero frequency phase intercept (see Section 5.2). Because of this, the derivative of phase, the envelope delay ($d\phi/df$), has been used as a measure of the phase linearity of circuits.[2,5] In practice, the true derivative cannot be measured either but can be approximated by measuring the difference in shift ($\Delta\phi$) experienced by the sidebands of a narrowband AM signal, and presenting the approximate derivative, $\Delta\phi/\Delta f$, on an instrument. The quantity Δf is twice the modulating frequency and is referred to as the aperture of the instrument. The difference between the quantity $\Delta\phi/\Delta f$ measured at some frequency, and that measured at some reference frequency is referred to as envelope delay distortion.

2.4.2 The phase characteristic of any system can be represented by a constant plus a linear term plus the Fourier expansion of the phase distortion. Thus $\phi(f) = A_O + af + \sum_i A_i \sin(2\pi f \tau_i)$ where $1/(\tau_i) = f_i$ is the period of the ith phase component in Hz. Measured envelope delay, D, is related to the true derivative of the phase by the expression $D = \frac{\Delta\phi}{\Delta f} = \frac{\sin x}{x}\frac{d\phi}{df}$; where $x = \frac{\Delta f \pi}{f_i}$. Note that when $f_i = \frac{\Delta f}{n}$; $n = 1, 2, 3, \ldots$, then $D = O$ so the instrument is blind to components equal in period to submultiples of the aperture of the set. As the component periods increase, $\frac{\sin x}{x}$ approaches 1 and D approximates the derivative more closely. Thus there exists a weighting of the phase derivative by the measuring instrument which is a function of the component period and the aperture of the set. Some weighting is desirable as shown below.

2.4.3 Consider one component of the phase characteristic to be given by $\phi(f) = A \sin \frac{2\pi f}{f_O}$.: If A is small, as it usually is, this results in one leading and one lagging echo of the transmitted signal which will cause intersymbol interference.[1] The magnitude of the echoes will be given by A/2 and they will be displaced in time from the original signal by τ_O seconds ($\tau_O = 1/f_O$). The derivative of the above phase curve is $\frac{A2\pi}{f_O} \cos \frac{2\pi f}{f_O}$ in the delay domain. Note that the amplitude of the component in the delay domain is changed by the factor $2\pi/f_O$ and so echo magnitude is not immediately evident from the delay curve. With A held constant and f_O variable, Figure 1 demonstrates the delay amplitude, $A2\pi/f_O$ as a function of f_O, where contours indicate equally interfering delay component peaks (i.e., constant echo magnitude). The magnitude of the delay component, for echoes of equal magnitude, increases as f_O decreases. If a true phase derivative instrument could be built, small period components resulting in echoes equally as interfering as large period components would tend to completely obscure the delay plot. Thus some smoothing of small period components is desirable in order to keep the observed delay distortion irregularities in proper perspective.

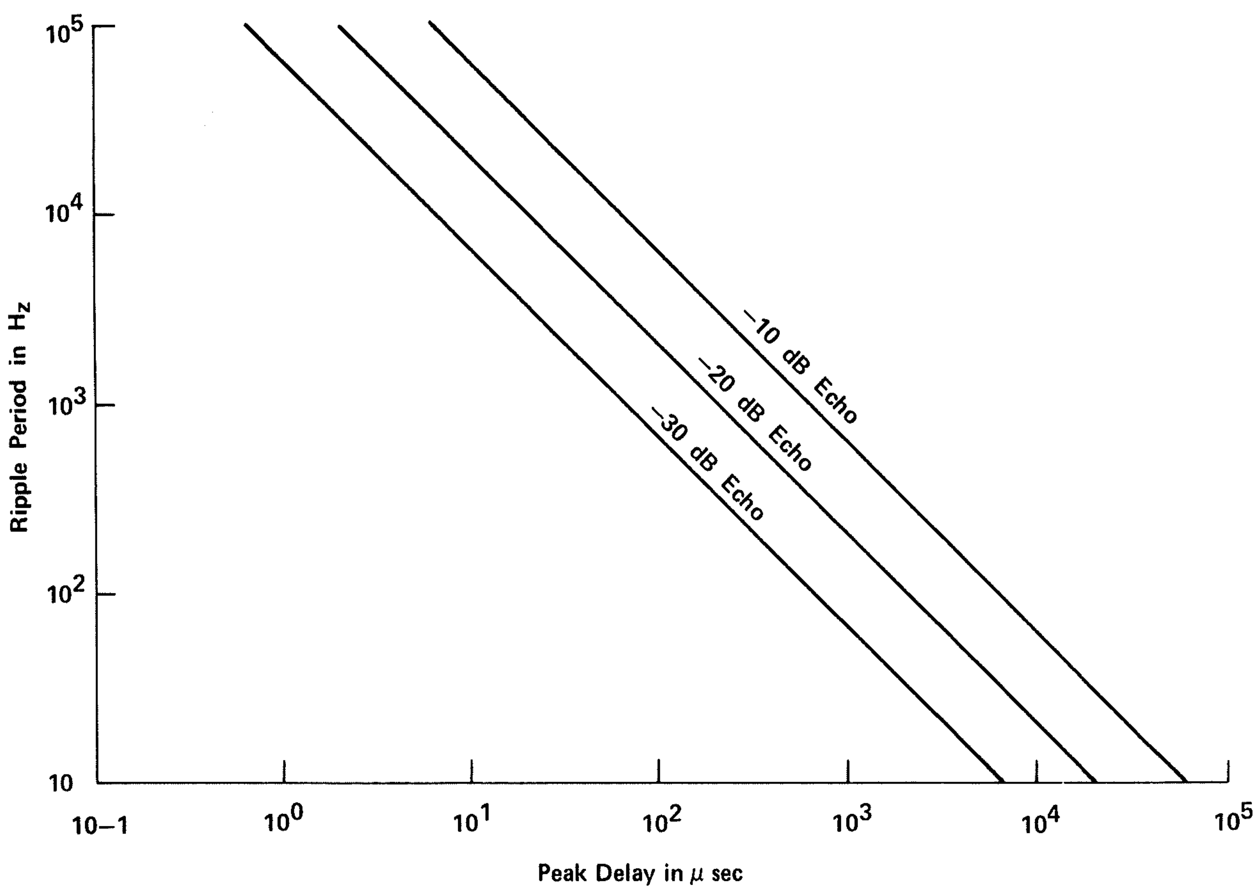


Fig.1 – Peak Delay for Single Sinusoidal Ripple With Constant Echo Magnitude

2.4.4 Several different apertures have been used in practice but further discussion will be limited to two. One is the current Bell System standard of 166-2/3 Hz and the other is the international standard (CCITT) of 83-1/3 Hz. Figure 2 shows how each of these apertures modifies the true derivative of the phase characteristic by the factor $\frac{\sin x}{x}$. The factor $\frac{\sin x}{x}$ is plotted as a function of f_0 for each of the two apertures. Echoes are usually of importance only where part or all of the transmission path consists of 2-wire facilities. The plots have been truncated at $f_0 = 20$ Hz because it is the practice to install echo suppressors on 2-wire connections when the round trip delay exceeds 45 milliseconds, and f_0 is the reciprocal of the round trip delay or echo time. Thus components finer than 20 Hz should be of no interest. One exception to this rule will occur if echo suppressors are disabled to permit simultaneous operation in both directions over a single 2-wire connection.[2] In practice, echoes with a delay time of about 45 milliseconds or greater are usually not a problem because of the extra loss they encounter in traversing the relatively long echo path.

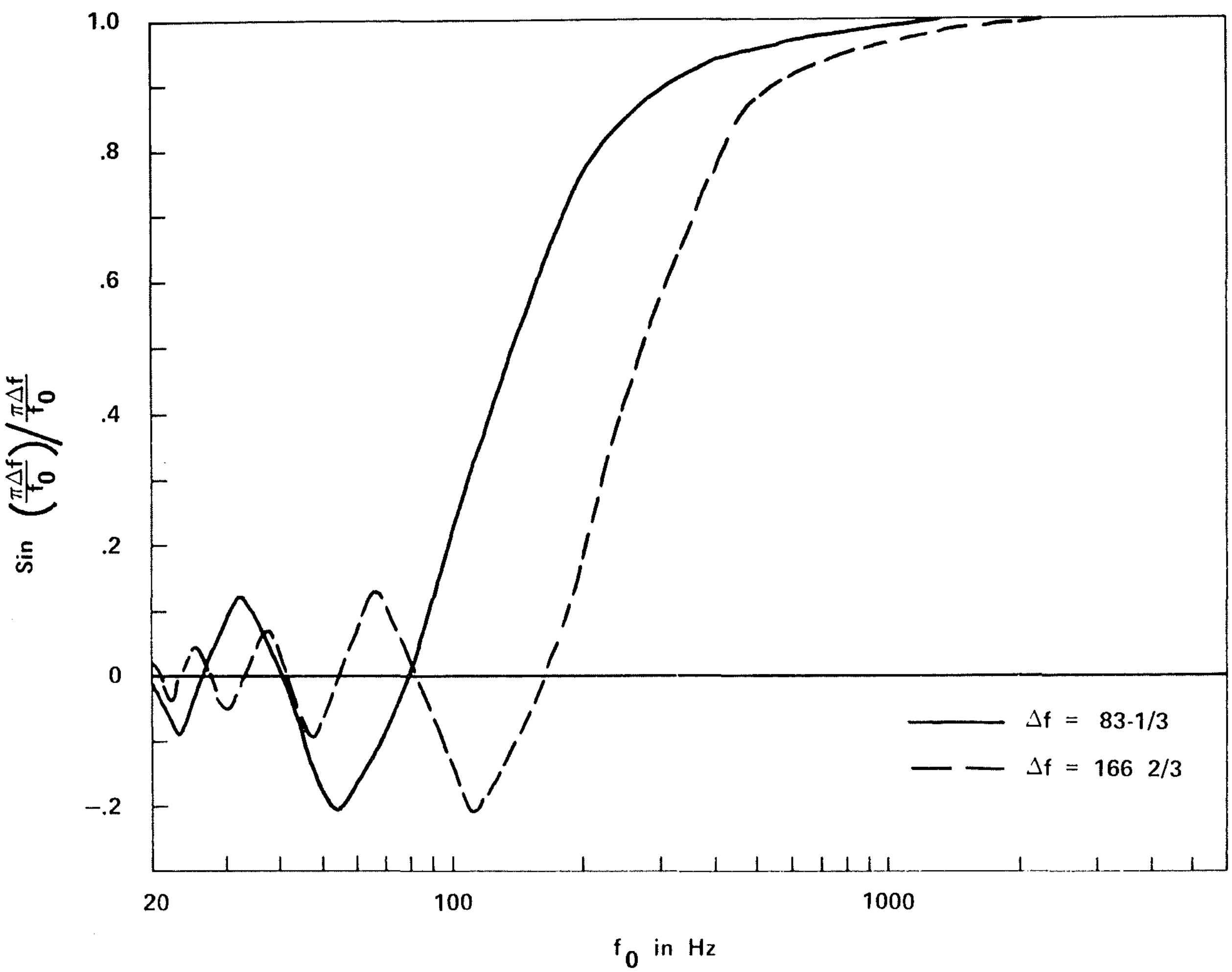


Fig. 2 — Response of Two Envelope Delay Test Sets to Ripples in the Phase Characteristic of a Transmission Line

2.4.5 From Figure 2 it can be seen that the Bell System sets have eight blind spots (zero crossings) and come within 90 percent of the true delay for ripples at 500 Hz and higher. The international standard aperture has four blind spots and comes within 90 percent of the true delay at about 300 Hz and above.

2.4.6 The magnitude of the intersymbol interference, due to a single sinusoidal component of the phase characteristic, is given by $A_i/2$. The peak possible intersymbol interference is given by $\Sigma|A_i|$, where the A_i are the coefficients of the Fourier series expansion of the phase characteristic across the band of interest. In theory, this could occur for a given phase characteristic for a particular signal sequence such that, at one instant of time, all the echoes generated by the preceding signals added in phase. Because of the importance of the quantities A_i, it is of interest to examine possible errors in the extraction of the A_i from delay measurements. This can be done by examining how the A_i are modified by a perfect envelope delay measuring set and by a realizable envelope delay measuring set.

2.4.7 A typical example of the usual manner of specifying delay requirements[18] is shown in the following table:

Table 1

Permissible Envelope Delay Distortion Limits	Within Frequency (Hz) From	To
500 microseconds	1150	2300
900 microseconds	1000	2500
1750 microseconds	800	2700

In addition, listener echoes[2] should not exceed a power of -12 to -18 dB with respect to the received signal. This means, for example, that a distortion consisting of a single isolated echo should be no larger than one-fourth of the main signal. Such echoes are proportional to the peak-to-peak phase deviations on the channel. However, the peak-to-peak envelope delay distortion is proportional to the product of the amplitude of an echo and the time displacement between the echo and the main received signal $1/f_i$. (Peak-to-peak envelope delay is proportional to A_i/f_i). Because of this, echoes having long delay paths tend to cause very large envelope delay distortion even though they may cause very little difficulty to data transmission. This effect is shown by curve I of Figure 3. As the curve shows, 12 dB listener echoes that are delayed more than a few milliseconds cause peak-to-peak envelope delay excursions that greatly exceed the 500-microsecond requirement, for the example shown, of Table 1. If their delay were on the order of 10 to 20 milliseconds, such echoes could completely obscure a plot of envelope delay distortion even though they would not be significantly interfering. Fortunately, the aperture error determined by the modulating frequency of a delay measuring set will tend to offset this effect. Curve II of Figure 3 shows the response of delay sets having an aperture of 166-2/3 Hz to a 12 dB listener echo. The response of the delay set, while having peaks of 1000 microseconds, remains in the neighborhood of 500 microseconds peak-to-peak.

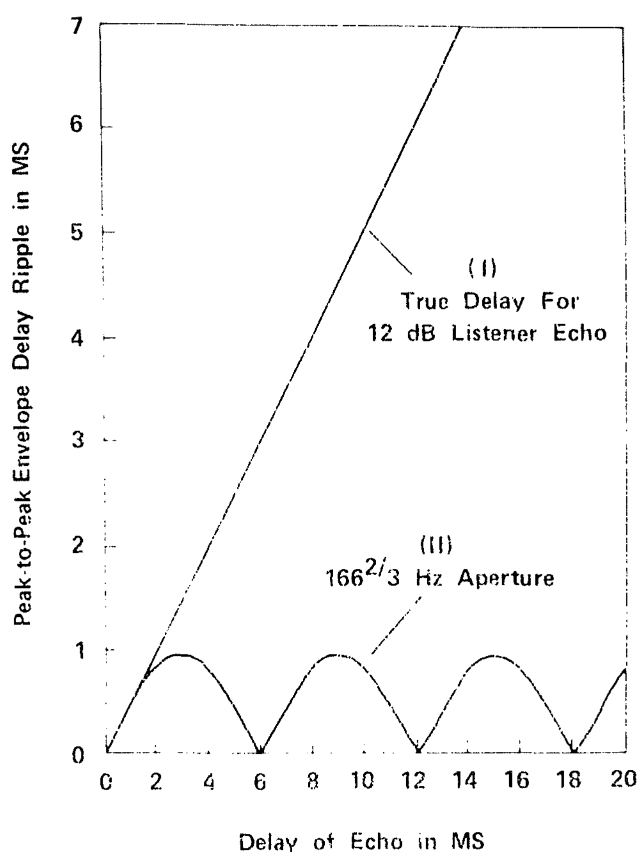


Fig.3 — Response of Bell System Envelope Delay Test Set to Echoes as Compared to True Delay

2.5 Return Loss (Impedance Control)

2.5.1 Impedance mismatches in a circuit give rise to reflections of signal energy called echoes. A single reflection causes energy to return to the transmitter (talker echo). If the single reflection is again reflected at an impedance mismatch, signal energy will arrive at the receiver some time after the original signal (listener echo). Echoes due to multiple reflections also occur, but they are usually insignificant by comparison with those due to single or double reflections.

2.5.2 Echoes generally cause no problems on 4-wire channels. However, on 2-wire channels, and on channels containing both 2-wire and 4-wire sections, echoes may be interfering to both voice and data communications. Echoes are controlled by matching impedances and by controlling loss in the echo path.

2.5.3 Figure 4 illustrates the use of a hybrid transformer which joins 2-wire and 4-wire facilities. The impedance of the balancing network ideally is the same as the impedance looking into the 2-wire facility. If this condition is met, signal power from the receive side of the 4-wire facility will divide equally between the balancing network and the 2-wire facility and no signal power will "return" on the transmit side of the 4-wire facility. In practice, the match is never perfect. Moreover, the match is different at different frequencies. Some portion of the received signal power is thus returned on the transmit side of the 4-wire facility.

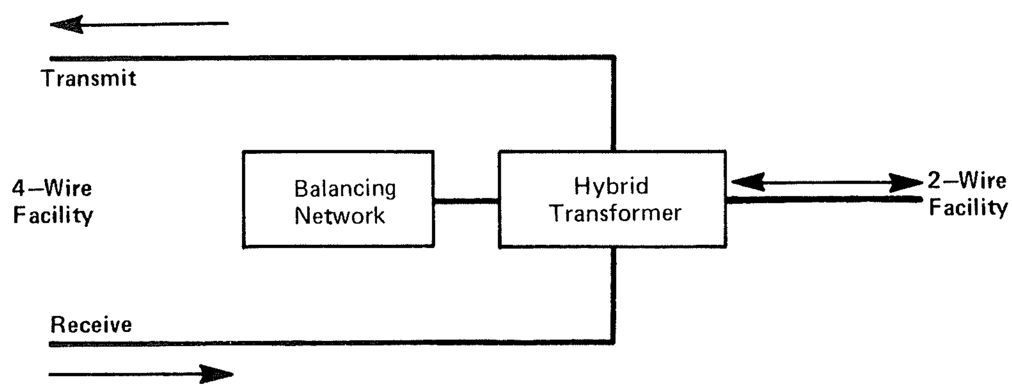


Fig. 4 — Typical Interface Between 2-Wire and 4-Wire Facilities

2.5.4 Various tests are used to determine channel quality with respect to echoes. These include return loss, echo return loss, singing return loss, singing point, and singing margin.

2.5.5 Return loss is the ratio in dB of the power of a single frequency signal placed on the receive side of the 4-wire facility (Figure 4) to the resulting power at that frequency appearing on the transmit side.

2.5.6 Echo return loss is a weighted average of the power returned at all frequencies between 500 and 2500 Hz and then expressed in dB.

2.5.7 Singing return loss is also a weighted average of the reflected power at all frequencies in a frequency band expressed in dB. There is a low frequency test covering the 200 to 500 Hz band, and a high frequency test covering the 2500 to 3200 Hz band.

2.5.8 A singing point test is made by placing a variable gain amplifier between the two sides of the 4-wire facility and increasing the gain until self-sustained oscillation, or "singing," occurs at some frequency. The singing condition is detected by audio monitoring of the circuit. The amount of gain in dB required to cause singing is called the singing point.

2.5.9 A singing margin test is similar to the singing point test. In this case, the channel is terminated in its normal impedances.

A variable gain amplifier is inserted on one side of the 4-wire facility and the gain increased until singing occurs. The amount of gain in dB required to cause singing is called the singing margin.

2.5.10 The return loss, echo return loss, singing return loss, and singing point tests can all be made on a 2-wire facility by incorporating the hybrid in the test set and making measurements as described above. The balancing network in this case is commonly 600 or 900 ohms resistance in series with 2.16 microfarads capacitance. Since these networks are compromises, they represent only the nominal impedance of the facility being tested. Provision is therefore made to connect different impedances to the test set to more precisely balance the 2-wire facility.

2.6 Peak-to-Average Ratio (P/AR)

2.6.1 Due to the difficulties in evaluating the resultant intersymbol interference from an envelope delay measurement, it is difficult to establish requirements in the delay domain which satisfactorily characterize the channel quality. To help overcome this problem, the Bell System has developed a technique to measure channel dispersion (spreading in time of signal amplitude) due to transmission imperfections. The concept is simple: generate a pulse train with spectral content shaped to be representative of a data modulated voice-band signal with spectral components chosen at the generator to give rise to a high peak-to-average ratio (signal peak to full wave rectified average, abbreviated P/AR) of the signal. As such a signal traverses a dispersive medium, the P/AR will deteriorate. Then by measuring the P/AR at the receive end, a simple measure of the dispersion is obtained. If the prime source of dispersion is phase nonlinearity, as it is in telecommunication channels, then a quick measure of this impairment is possible.

2.6.2 The test set used to measure P/AR also responds to other channel impairments but to a lesser extent than phase nonlinearity. Frequency response and nonlinearities are the second and third most important degradations followed by C-notched noise, and incidental FM.

2.6.3 The P/AR test signal has a spectrum centered at about 1650 Hz and has 12 dB points at about 1000 and 2400 Hz. The detector responds approximately according to the relation:

$$P/AR = 100 \left[2 \frac{E_p}{E_{FWA}} - 1 \right]$$

where E_p is the normalized peak and E_{FWA} is the normalized full wave rectified average of the envelope.

2.6.4 In general, P/AR ratings of about 50 or higher indicate that intersymbol interference for medium speed (2400 bits per second) data transmission will be acceptable. For details on the P/AR signal, effects of various impairments and correlations with data transmission performance, the reader is referred to References 7, 8, 9 and 14.

2.6.5 The P/AR ratings of a system may be calculated from the gain and phase or gain and delay characteristics of the channel if it is desired to know the rating precisely.[7] The P/AR test is used as a quick method of evaluating telecommunications channels for nonvoice transmission. It provides a rapid means of sorting between channels having acceptable and unacceptable phase characteristics. The accuracy of the meter is typically ±1 P/AR point and deviations in excess of 8 points from a calculated expected value, or ±4 from an initially measured value, provides sufficient reason to suspect that some characteristic has changed significantly.

2.7 Long Term Loss Variation

2.7.1 Due to normal aging and drift of amplifiers, temperature changes in cables, and changes in the physical makeup of a channel, changes in the loss of a circuit can be expected. Such changes are commonly referred to as seasonal changes, emphasizing the effects of temperature.

2.7.2 Routine tests are made on transmission facilities between central offices to maintain loss deviations within nominal values. Thus the greatest changes, from the user's point of view, normally come about due to temperature changes on the cable connecting the subscriber to his serving central office. No special test equipment, other than that used for frequency response or simple loss measurements, is required to detect these changes. The changes to be expected in 1000 Hz loss, for private line channels, for example, may be as large as ±4 dB (compared to the circuit installation measurement) over a period of a year.

3. NOISE (SIGNAL UNCORRELATED INTERFERENCE)

3.1 C-Message Noise and C-Message Notched Noise

3.1.1 Noise on telecommunications channels arises from numerous sources. Some of these are thermal noise from amplifiers, 60 Hz and its harmonics which may be picked up from power line induction, intelligible or unintelligible crosstalk, single frequency tones, and switching and signaling transients. All of these effects and others come under the general heading of noise.

3.1.2 Background noise has been referred to in various places as white noise, Gaussian noise, message circuit noise, etc. What is intended is a frequency weighted measure of the total power on a channel not arising from the desired signal.*

3.1.3 Primarily because of the existence of syllabic compandors[1] and digital transmission facilities which require quantizing, two types of background noise must be distinguished. The first is called C-message noise and is the total frequency weighted noise power measured on a channel in the absence of signal. Hence, it is signal uncorrelated interference.

3.1.4 The second is referred to as C-message notched noise[2] ("C-notched noise") and is a measure of unwanted power in the presence of a signal. When signal-to-noise ratios are discussed, C-notched noise is usually implied. Quantizing noise, inherent in PCM systems is measured as C-notched noise but is separately discussed in Section 3.4.

*The frequency weighting used is called "C-message" weighting. See Reference 2 for details.

3.1.5 C-message noise is measured by a quasi rms type of instrument with a time constant of about 200 milliseconds. The noise is measured through a filter called C-message weighting which shapes the noise in such a fashion as to make the measurement more meaningful in terms of annoyance to people listening to the noise with a telephone receiver.[2,3] The C-message weighting characteristic is shown in Figure 5. Noise measurements with this weighting have validity for data transmission since, as can be seen from Figure 5, the characteristic is relatively flat over most of the frequency range usually of concern for data transmission (600-3000 Hz). Thus this weighting is useful for data channels even though it was developed for voice applications. The accuracy of noise measuring sets of this type is typically ±1.0 dB as determined by its measurement of a known power source of 1000 Hz tone.

3.1.6 C-message noise and C-notched noise readings are expressed in units of dBrnC, or dB with respect to reference noise (-90 dBm, or -90 dB with respect to 1 milliwatt). The "C" refers to the C-weighting. For example, C-weighted noise having an rms power of -70 dBm would be expressed as 20 dBrnC (90 - 70 = 20).

3.1.7 C-notched noise is measured with the same type of instrument as message circuit noise except for the filter characteristic. In order to estimate signal-to-noise ratios, a 1004 Hz tone (called a "holding tone") is applied at the far end of the channel at a power approximating an actual signal. This tone activates compandors and quantizers. The tone is removed at the receiving noise measuring set by a notch filter which suppresses the holding tone by at least 50 dB. 1004 Hz is used instead of 1000 in order to avoid using a rational submultiple of the sampling frequency in PCM systems which can lead to problems.

3.1.8 When either message circuit or C-notched noise is measured, small swings of the meter (up to ± 6 dB) are mentally averaged. Occasional large jumps in the noise are ignored. These momentary large fluctuations are usually due to impulse noise which is measured in a different manner.

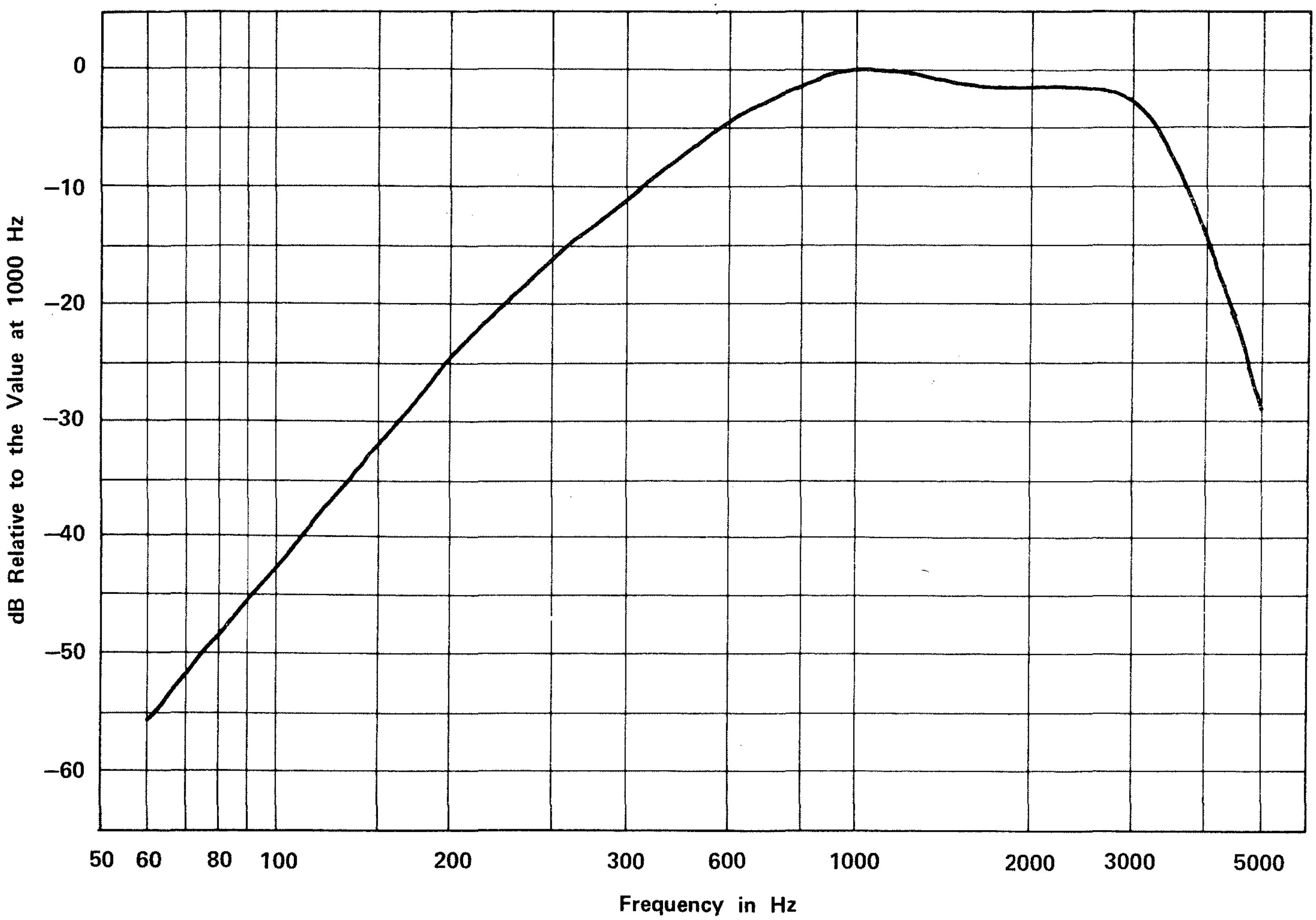
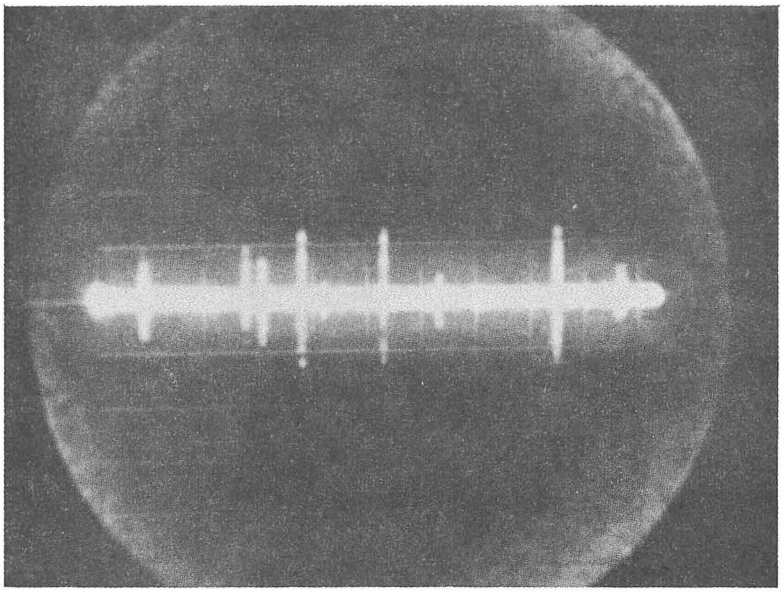


Fig. 5 — C-Message Weighting Characteristic

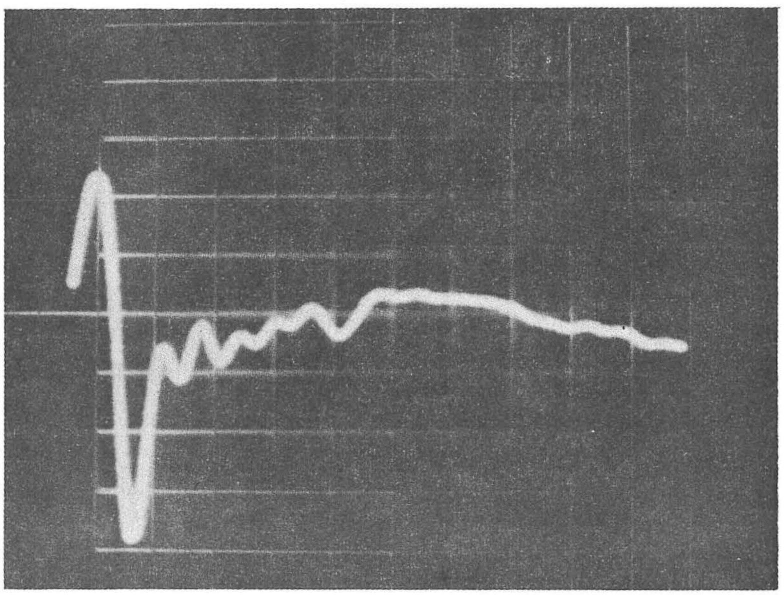
3.2 Impulse Noise

3.2.1 Impulse noise is characterized by large excursions of the total noise waveform which are much higher than the normal peaks of the message circuit noise.[4] It is measured with an instrument which responds to noise waveform excursions above a selectable power threshold. The instrument records these excursions on a counter which has a controlled dead time (defined below) of 140 milliseconds. The instrument uses C-message weighting with a notch to filter out a holding tone transmitted from the far end.

3.2.2 Impulse noise is a distinct phenomena from message circuit noise. Although some peaks of message circuit noise are registered on impulse noise counters, the vast majority of impulse noise arises from sources that are independent of message circuit noise sources. That this is the case is amply demonstrated by the photograph in Figure 6a which shows noise in a channel with a 48 kHz bandwidth photographed for 50 milliseconds. The difference between message circuit noise and impulse noise is easily seen. In this case, the impulse noise is so apparent because of the relatively large bandwidth involved. Figure 6b shows the result in a voice bandwidth channel of one of the impulses illustrated in Figure 6a. Note the smear in time (5 milliseconds full scale) and the ringing due to the baseband filters. It is noise of the type shown in Figure 6b with which we are primarily concerned.



(a) — 50 ms of Noise in a 48 KHz Channel



(b) - 5 ms of Impulse Noise in a Voiceband Channel

Fig. 6 — Examples of Impulse Noise on Two Different Bandwidth Channels

3.2.3 The most frustrating characterisitic of impulse noise is its time variability. When one is exposed to impulse noise measurements for the first time, the initial reaction may be uncertainty and lack of faith in any single measurement made in a reasonably short (5 to 15 minutes) time. This is primarily due to the fact that the number of impulses occurring during a fixed time interval at a fixed threshold is log-normally distributed.[14] A typical sample of six such numbers might be 5, 8, 52, 6, 6, 4. It is much more satisfying to look at impulse noise level, which is normally distributed[14] and thus it is worthwhile to discuss that first.

3.2.4 Impulse noise level is defined as the threshold (expressed in dBrnC) at which the median count from a number of observations (each having the same specified time interval) is equal to a specified number. Both the median number of counts and the specified time interval have changed over the years. In the mid-1950s, the number of counts (not exactly the median in this case) was 70 and the specified time interval was 1 hour. In recognition of the fact that 1 hour is inordinately long for measuring a parameter, the test interval has been steadily shortened as more knowledge has become available. The recommended interval is now 5 or 15 minutes, as discussed below.

3.2.5 By 1963, the measurement interval had been reduced to 30 minutes solely on the basis of a great deal of experience by numerous people who agreed that such an interval was adequate. On that basis, the 1963 impulse noise survey[15] was conducted by making 30-minute tape recordings of noise. The data collected continues to be the best source of information on this topic. The data were summarized in two ways that are pertinent for this document.

3.2.6 First, cumulative distribution functions (cdf's) of the noise counts on individual channels were made every 5 minutes from 5 to 30. Two examples are shown in Figures 7 and 8. The impulse noise level may be tracked over the 30-minute interval by observing the threshold at which five counts in 5 minutes occurred, ten in 10 minutes, etc.* Traces of the impulse noise level may then be made by connecting the points so identified on the cdf's. Figure 7 shows an example of very little movement of impulse noise level; Figure 8 shows an example of very great movement. In Figure 7 one is left with the impression that the impulse noise level is "settling down" by the end of the 30 minutes. In Figure 8, the noise level appears to be still on the move at the end of 30 minutes. Over 1000 such sets of cdf's were constructed and examined. In almost all cases, there was a qualitative feeling that the 30-minute measurement provided a reasonable estimate of impulse noise level. At least it is an average taken over 30 minutes and so is a better estimate than that provided by a shorter one. This is the only known supportive evidence that 30 minutes is an adequate measurement interval for impulse noise.

3.2.7 Examining this question from a different point of view, we have the following facts. Most impulse noise originates with actions of people. Customers initiating and terminating calls cause relays and switches to be activated and released giving rise to impulse noise from the associated electrical transients. Normal maintenance, installation, and repair activities within the telephone plant introduce impulse noise on systems. Since people use the telephone network more during the day than at night, impulse noise exhibits large diurnal variations. Actually then, it makes little sense to talk about the impulse noise level on a circuit. What is obtained in a measurement made during the normal business day is an estimate of the maximum impulse noise level achieved. Referring again to Figure 8, with this thought in mind, we have a 30-minute trace of impulse noise level during its peak period.

*Impulse noise level is here defined as that threshold at which the median count for a number of observations is equal to one per minute.

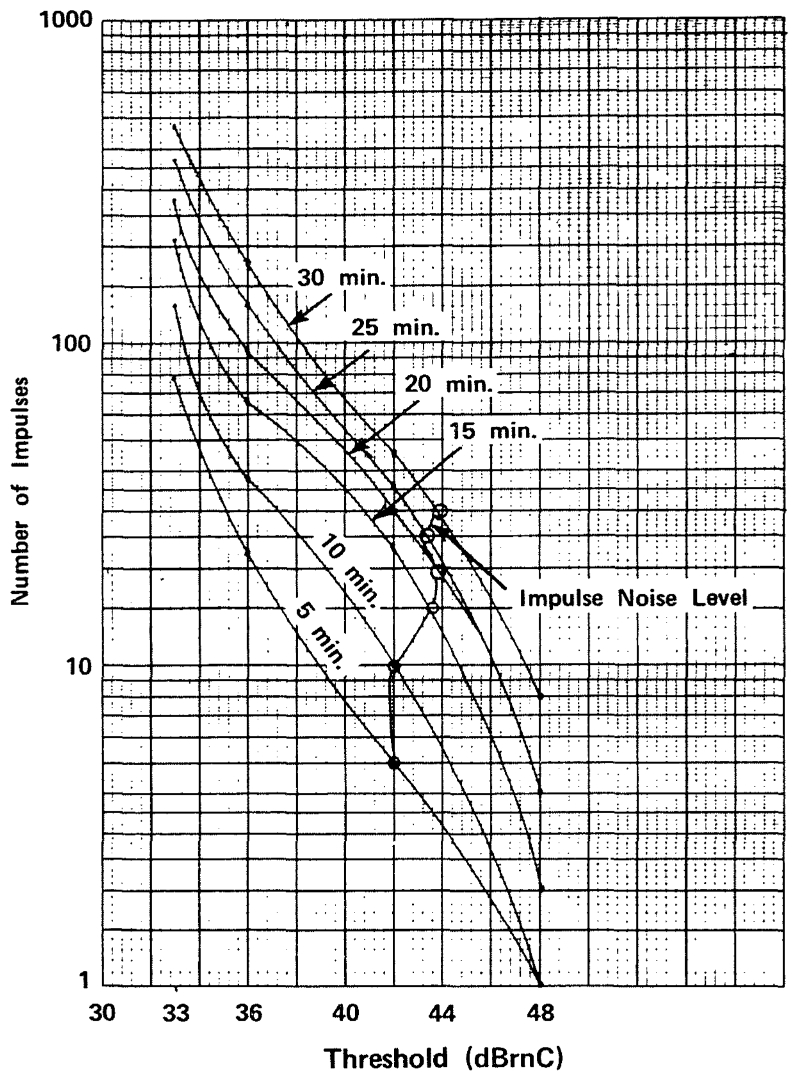


Fig. 7 — Cumulative Distributions of Impulse Noise Peaks In Successive Intervals From 5 to 30 Minutes.

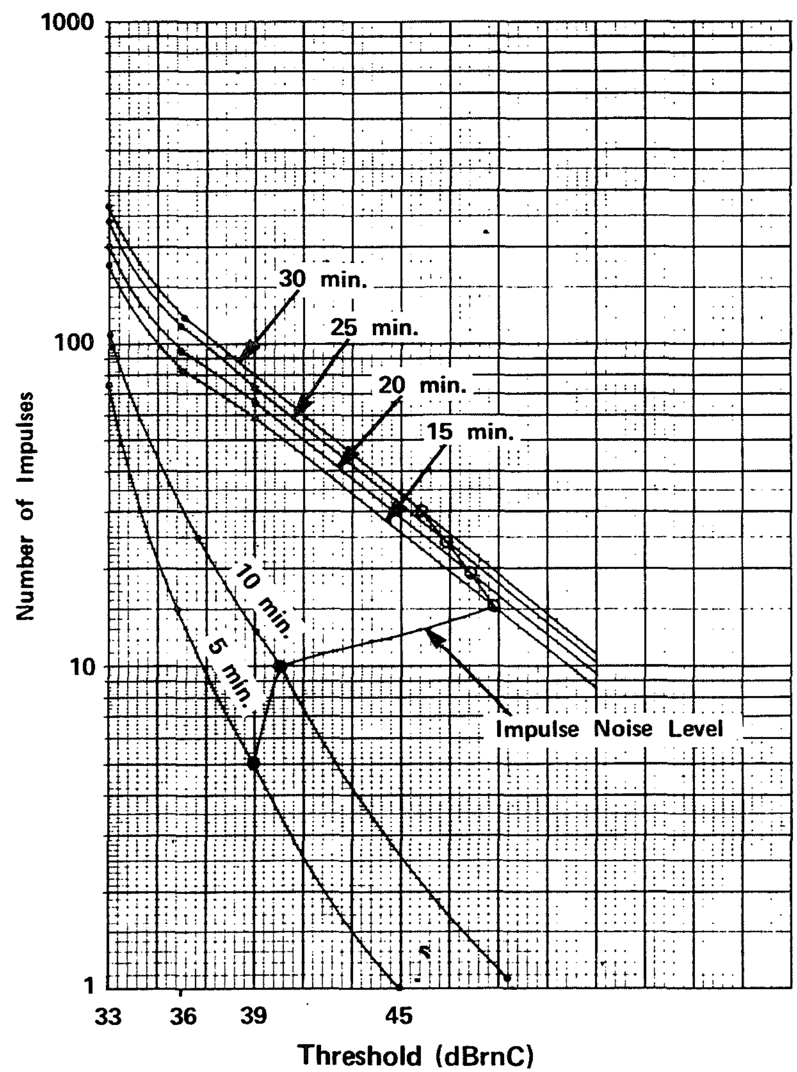


Fig. 8 — Cumulative Distributions of Impulse Noise Peaks In Successive Intervals from 5 to 30 Minutes.

3.2.8 Now, if we accept the premise that a 30-minute interval provides a usable estimate of peak impulse noise activity, the next logical question is simply, "Can an even shorter interval be used?" To answer this question, we examine the penalty in uncertainty incurred by using shorter intervals. To this end, we use the 30-minute estimate as the yardstick and look at the relative variance of estimated impulse noise level as a function of time. Two examples of the change in the variance of estimated level, compared to the 30-minute estimate, are reproduced here as Figures 9 and 10. Referring to Figure 10, we note that the variance falls from 4.8 dB^2 to 0.2 dB^2 as the interval increases from 5 to 25 minutes. Since the distribution of noise levels is normal, we can easily make statements of the expected error in a measurement of from 5 to 25 minutes long from the data in Figure 10. The standard deviation for 5-minute estimates is about 2.2 dB so we can say that 95 percent of all 5-minute measurements will be within ±3.6 dB of a 30-minute measurement. Similarly, 95 percent of the 15-minute measurements will be within ±2.7 dB of a 30-minute measurement. Most measurements will come closer to the 30-minute estimate than the numbers just given. For example, using the data in Figure 9 we find that 95 percent of 5-minute estimates will fall within ±2.3 dB of the 30-minute one. Impulse noise level is not simple to measure. It normally requires the use of a multithreshold counter and interpolation along a cdf of counts at some specified threshold or a maximum number of counts at some specified threshold. The requirements are always based, however, on a desire to control the maximum impulse noise level. As shown next, estimates of the 1-count/minute impulse noise level may be made solely from count distributions at an arbitrary level.

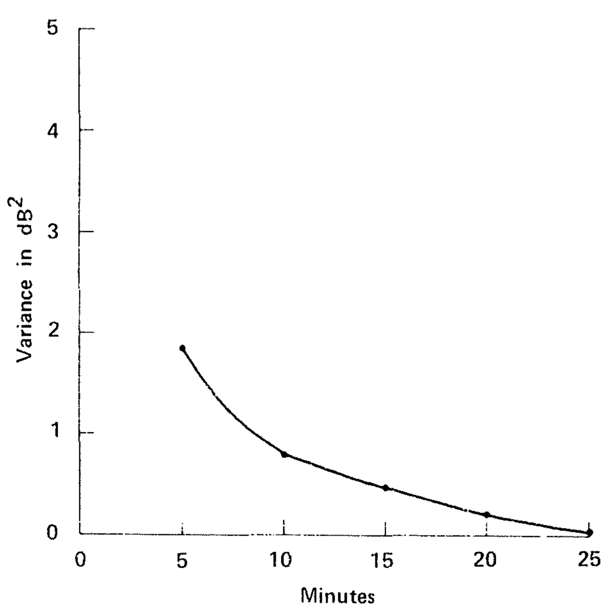


Fig.9 – Variance of Estimated Impulse Noise Levels in Increments From 5 to 25 Minutes When Compared With a 30 Minute Estimate–Cable Carrier Example.

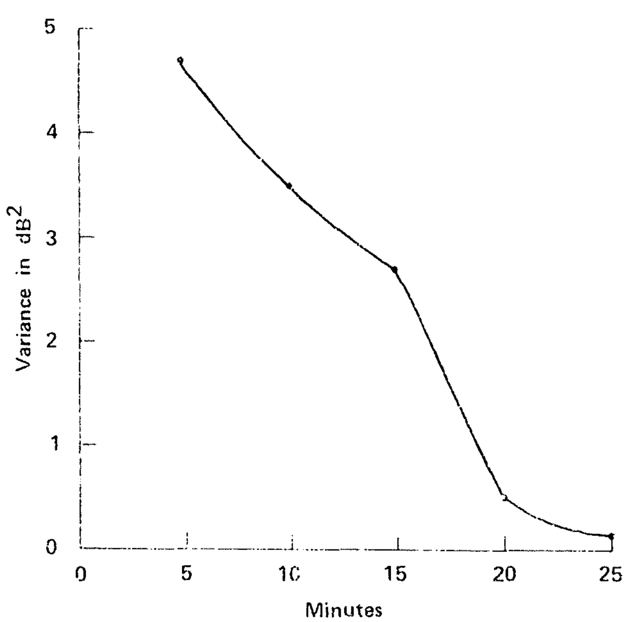


Fig.10 – Variance of Estimated Impulse Noise Levels in Increments From 5 to 25 Minutes When Compared With a 30 Minute Estimate–Radio Facility Example.

3.2.9 It has been stated that count distributions are log-normal and that impulse noise level distributions are normal. There are, fortunately, simple relationships to change from count distributions (readily measured) to impulse noise level distributions – the engineer's tool.[4] The mean of the normal impulse noise level distribution is simply the value, in dBrnC, at which the impulse noise test set recorded the count distribution. It has a count associated with it which is simply the median of the observed count distribution. σ_L, the standard deviation of the impulse noise level distribution may be estimated by the expression $\sigma_\ell = m\sigma_D$ where m is the inverse slope of the peak amplitude distribution in dB per decade of counts (averaging 7.0; see Figure 11) and σ_D is the standard deviation of the log-normal count distribution calculated by taking the square root of the $\log_{10}$ of the ratio of average number of counts to the median count. Thus:

$$\sigma_D = [\log_{10}(\text{avg. count/median count})]^{1/2}; \text{ median} \neq \text{zero.}$$

In the evaluation of impulse noise level on channels, the average and standard deviation of the impulse noise level may be estimated as above. Let the impulse noise level from a set of measurements be N counts per unit time. The threshold at which the median of a number of measurements would be one count per minute can be estimated by using Figure 11, which shows the peak amplitude distribution of impulse noise averaged over many facilities. For example, if N = 10 counts per minute at a given threshold, the one count per minute threshold would be 7 dB above the threshold used to measure 10 counts per minute.

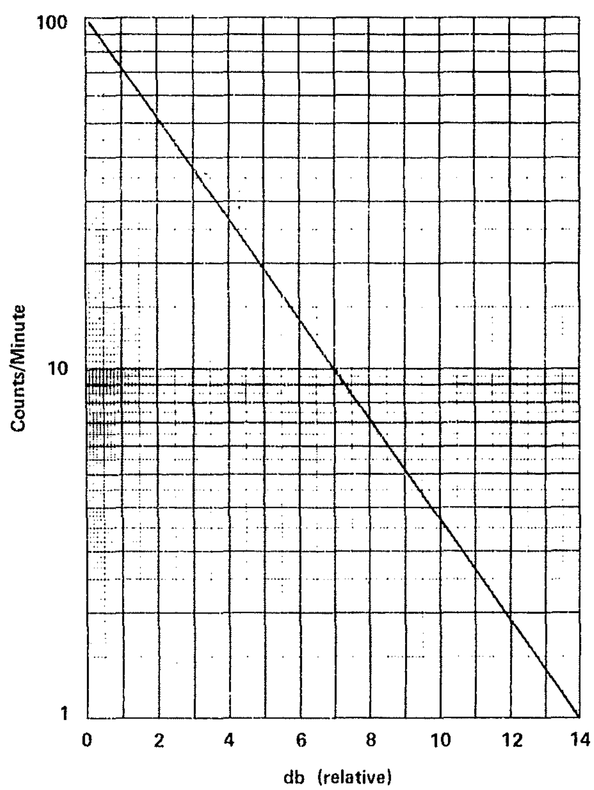


Fig. 11 — Average Impulse Noise Peak Amplitude Distribution

3.2.10 The impulse noise level on a transmission facility may be determined by making 5-minute measurements on a number of channels on the facility and computing N as discussed above. By the use of Figure 11, the measurements may be taken at any threshold and converted to the one count per minute threshold. Knowledge of the standard deviation of the impulse noise level is also useful in engineering facilities. For example, if the standard deviation is large, more margin against the effects of impulse noise may be provided by reducing the impulse noise level below normal objectives.

3.2.11 While the net noise on channels in a frequency division multiplexed carrier facility will be essentially the same, in general one cannot expect an external transient to cause a simultaneous pulse on all channels. This is because the noise is introduced to each channel by a process which essentially multiplies the external pulse with each channel carrier signal. In general, the carriers are not coherent, i.e., they do not all go through maxima and minima at the same time. Hence, one external influence will produce observable pulses on some channels but not on others. This same process also causes external transients of equal magnitude to be dispersed in amplitude on the carrier channel. Thus, fixed amplitude disturbers will be spread in amplitude over a range of about 20 dB on a double sideband carrier system and about 6 dB on a single sideband carrier system when measured at the output of a single carrier channel.

3.2.12 It is widely known that impulse noise frequently occurs in clusters.[16] Due to size and cost considerations, early instruments designed specifically to count noise pulses used electromechanical counters. It was recognized that some number of noise pulses would be missed because they occurred shortly after a noise pulse in the process of being recorded by the counter. The time it takes a counter to register one count is referred to as dead time. In order to collect comparative data with different sets, the dead time is electronically controlled to be very close to 140 ms. In order to evaluate the effects of not counting all pulses, comparative tests were run using an electronic counter and an electromechanical one. Distributions of the number of pulses missed in a 30-minute period due to dead time were constructed. One of these is redrawn as Figure 12 which shows that in some instances as many as 120 may be missed because of the dead time.* From data such as shown in Figure 12 it is possible to use an average number of missed impulses as a correction factor. It is, however, simpler to include such a factor in stated objectives for control of impulse noise rather than force the use of a correction

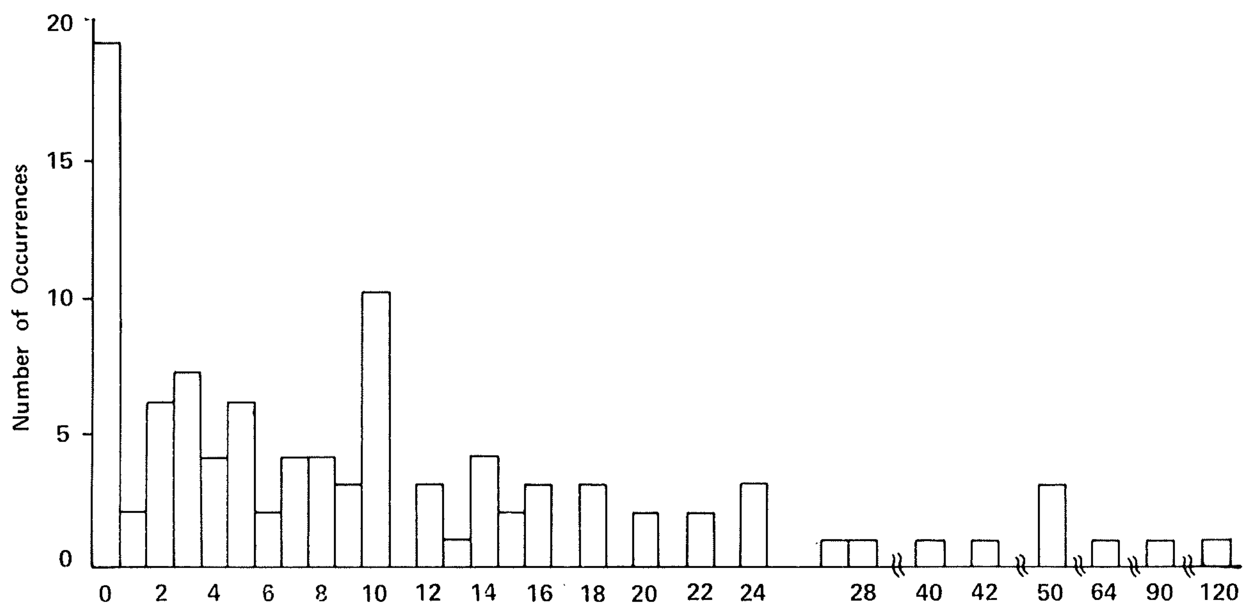


Fig.12 — Number of Impulses Missed in a 30 Minute Period by Electromechanical Counters. Threshold Always Adjusted to Produce Exactly 30 Counts in 30 Minutes on Counters.

factor after each measurement. This has been done for current Bell System objectives. Even though the distribution of missed noise pulses is very skew, years of experience in the use of these objectives has not uncovered any problems with this approach. Even though the number of missed impulses with an electromechanical counter may appear gross, the change in threshold (increased sensitivity) required from the electromechanical counter to achieve a count equal to the electronic counter is only about 0.9 dB on the average, with a standard deviation of 0.76 dB.

*When the electromechanical count is 30 in 30 minutes.

3.3 Single Frequency Interference

3.3.1 Single frequency (SF) interference refers to unwanted steady tones. Short bursts of tones which may occur from crosstalk of multifrequency signaling,[2] for example, do not fall in this category.

3.3.2 The requirements for SF interference is that, when measured through a C-message filter, it be at least 3 dB below C-message noise limits. A simple audio monitoring arrangement on the output of a C-message noise test set will usually detect this interference since tones exceeding the limit are easily heard if the C-message noise is within its normal range. The audio monitoring arrangement eliminates the need for sweeping the channel spectrum with a frequency selective voltmeter to detect tones.

3.3.3 Note that a foolproof test for SF interference is not possible with either technique. For example, if the SF tone source occurs before a compandored facility, chances are that it will not be detected with a C-message noise measuring set since it will be down approximately 18 dB from its power in the presence of a signal. Audio monitoring at the output of a

C-notched noise test set will detect SF interference for voiceband tones not in the rejection band of the notch filter.

3.3.4 Single frequency interference is potentially most disturbing to systems which frequncy multiplex several narrowband channels on one voice bandwidth channel. Voice frequency telegraph equipment is a common example. Assume that such a system subdivides a voice channel into 20 telegraph channels. The C-message noise in each narrowband channel will be down from that measured in the full voiceband channel by about 13 dB. SF tones, however, do not realize the reduction due to the narrower bandwidth and may be the controlling impairment in such a situation. This fact should be taken into account when such systems are contemplated.

3.4 Quantizing Distortion

3.4.1 To transmit analog signals over digital lines, the analog signals are first sampled in time. One of a finite number of amplitudes is then used to represent each sample by quantizing or rounding-off the sample to the nearest possible amplitude. The distortion caused by the round-off error, in conjunction with the sampling, is called quantizing distortion or quantizing noise. The reader is directed to Chapter 25 of Reference 1 for a more detailed discussion.

3.4.2 Quantizing distortion is measured in the same way as C-notched noise as described in Section 3.1. Although quantizing noise is signal correlated, it is considered to have the same effect on transmission quality as an equal (in power) amount of background noise. This is somewhat conservative since tests have indicated that the impairing effect is 0 to 3 dB less than Gaussian noise.

3.4.3 Quantizing distortion can appear as other impairments when viewed through certain test sets. For example, it can appear as harmonic or intermodulation distortion so, as discussed below, the test described above for nonlinear distortion is designed to minimize this effect. It may also appear as phase jitter. Figure 20 plots the reading produced on a phase jitter test set by quantizing noise as a function of the frequency of the test signal. These results are not consistent with the frequency independent results that are observed on analog systems and there is no reason to believe that these measurements have the same meaning that they have on analog systems. Indeed, it is perhaps best to interpret the results in Figure 20 as a "noise floor" for phase jitter measurements made over a PCM system.

3.4.4 The measurement of quantizing distortion reflects an additional component. System imperfections on PCM systems can cause the nonlinear distortion described below. Because of the sampling, the out-of-band energy from this nonlinear distortion is folded over and appears in-band (see Reference 1, page 568). Thus, when a 2800 Hz holding tone is used to measure quantizing noise, the second and third harmonic distortion caused by system imperfections appear at 5600 Hz and 8400 Hz. These tones are folded over and appear at 2400 and 400 Hz (assuming a sampling rate of 8000 samples per second) and add to the measurement since they are not removed by the 2800 Hz notch filter at the receiver.

4. MISCELLANEOUS IMPAIRMENTS

4.1 Nonlinear Distortion

4.1.1 Nonlinear distortion can be broadly defined as the generation of signal components from the transmitted signal that add to the transmitted signal, usually in an undesired manner. The main sources are electronic devices and other components comprising voiceband channels. The nonlinear distortion of concern here should not be confused with the intermodulation noise caused by nonlinearities in the terminal equipment and line amplifiers of a frequency division multiplex system. Although these nonlinearities can contribute to the nonlinear distortion at voice frequencies, their contribution is usually negligible.

4.1.2 Nonlinear distortion is commonly measured and identified by the effect it has on certain signals. For example, if the signal is a single tone having frequency A, the nonlinear distortion appears as harmonics of the input – that is, it appears as tones at 2A, 3A and so on. Since most of the distortion usually occurs at the second and third harmonic, it is often measured by the power of each of these harmonics, and is called second and third harmonic distortion. If the amount of nonlinear distortion is measured by the power sum of all the harmonics, the result is called total harmonic distortion. These distortion powers are not meaningful unless the power of the wanted signal (the fundamental) is known, so measurements are usually referred to the power of the fundamental and termed second, third, or total harmonic distortion.

4.1.3 For a multitone input signal, the nonlinear distortion is termed intermodulation distortion and appears as tones at frequencies which are linear combinations of the input frequencies. As an example, if the input consists of three tones at frequencies A, B and C, the nonlinear distortion appears at frequencies $k_1A \pm k_2B \pm k_3C$ where k_1, k_2, and k_3 are nonnegative integers. The distortion is called second order distortion for those nonnegative values of k_1, k_2, and k_3 such that $k_1 + k_2 + k_3 = 2$ and third order distortion for those values such that $k_1 + k_2 + k_3 = 3$. The distortion at a particular frequency is called a second or third order product. The second order product for $k_1 = k_2 = 1$ and $k_3 = 0$ is at A+B (or A-B) Hz and is called an A+B (or A-B) product. The third order product for $k_1 = 2$, $k_2 = 1$ and $k_3 = 0$ is 2A-B (or 2A+B). The amount of nonlinear distortion is measured by the magnitudes of various products with respect to the received fundamental.

4.1.4 A nonlinear channel might be modeled with a third degree polynomial

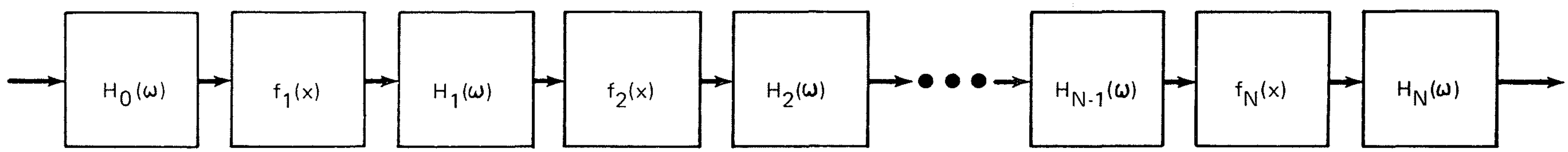
$$y = a_1x + a_2x^2 + a_3x^3. \qquad (4.1)$$

The quadratic distortion term, a_2x^2, generates second order distortion and the cubic distortion term, a_3x^3, generates third order distortion. If this is an accurate model, the harmonic and intermodulation distortion measurements are really equivalent for either one can be used to compute a_2^2 and a_3^2 and the measurements can be related to one another. (See, for example, Chapter 10 of Reference 1.) However, because Equation (4.1) is only an approximation, the different ways of measuring nonlinear distortion do not always yield the same a_2^2 and a_3^2. To select the best measurement, it is necessary to understand how telecommunication channel nonlinearities differ from Equation (4.1). The three main differences are (i) frequency dependency, (ii) time variability and (iii) the presence of nonlinear terms above third degree. The first two are most important and are discussed first. The third is discussed at the end of this section.

4.1.5 A telephone channel may consist of several sources of nonlinearity alternated with linear networks as shown in Figure 13a. A simplified model for estimating transmission performance is shown in Figure 13b. This model has the same linear characteristics as the actual channel but it has only one source of nonlinearity. This nonlinearity is chosen so that the system in Figure 13b produces the same ratios of signal-to-second and -third order nonlinear distortion for a complex signal as does the actual channel modeled in Figure 13a. The nonlinear distortion measurement is intended to determine these ratios.

4.1.6 The amount of second and third order distortion at the output of the model in Figure 13a depends on how the distortion from the several sources of nonlinearity "adds." This addition is illustrated below using two nonlinearities, each described by $x + a_2x^2 + a_3x^3$, with a linear network between them.

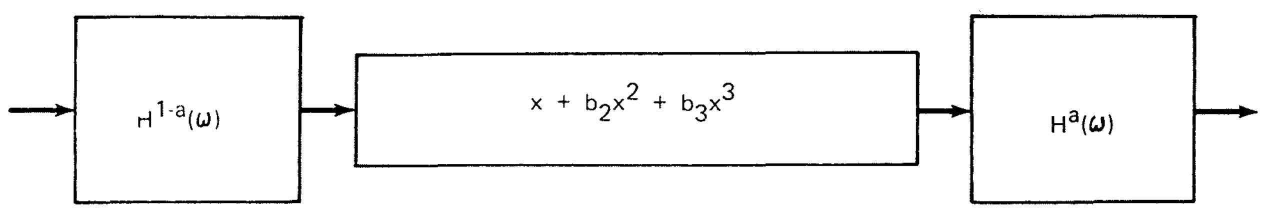
4.1.7 The phase angle of the linear network has the largest effect on addition of the nonlinear distortion. Envelope delay distortion is usually all that is known about the phase angle and this is known only above some radian frequency ω_1. Since the phase angle is obtained by integrating over the envelope delay distortion curve, the contribution below ω_1 is not known. This is left as a parameter and denoted K. On a single sideband (SSB) system K can be a function of time. If the system introduces a frequency shift of F Hz, K varies over the range 0 to 2π and does so F times a second. K can also vary in a random fashion. The effect of K on addition is discussed in Chapter 13 of Reference 1. Theoretical results in Reference 17 on the response of nonlinear systems to Gaussian inputs can be used to compute the combined effect of delay distortion and K. Although data signals are not Gaussian, the addition is influenced more by the fact that signal energy is distributed across the band than by the type of signal.



$$f_k(x) = x + a_{2k}x^2 + a_{3k}x^3$$

$H_k(\omega)$ = System Function of Linear System

(a) — Nonlinear System



$$H(\omega) = \prod_{k=0}^{N} H_k(\omega), \quad 0 < a < 1$$

(b) — Simplified Model

Fig.13— Two Models of a Transmission Line With Nonlinear Distortion

4.1.8 Results are first presented for third order distortion. The linear network is assumed to introduce delay distortion equal to that caused by one pair of filters commonly used on SSB systems. This amount of delay distortion is shown by the curve labeled "I" in Figure 14. The in-band third order distortion is plotted along the ordinate in Figure 15 in dB

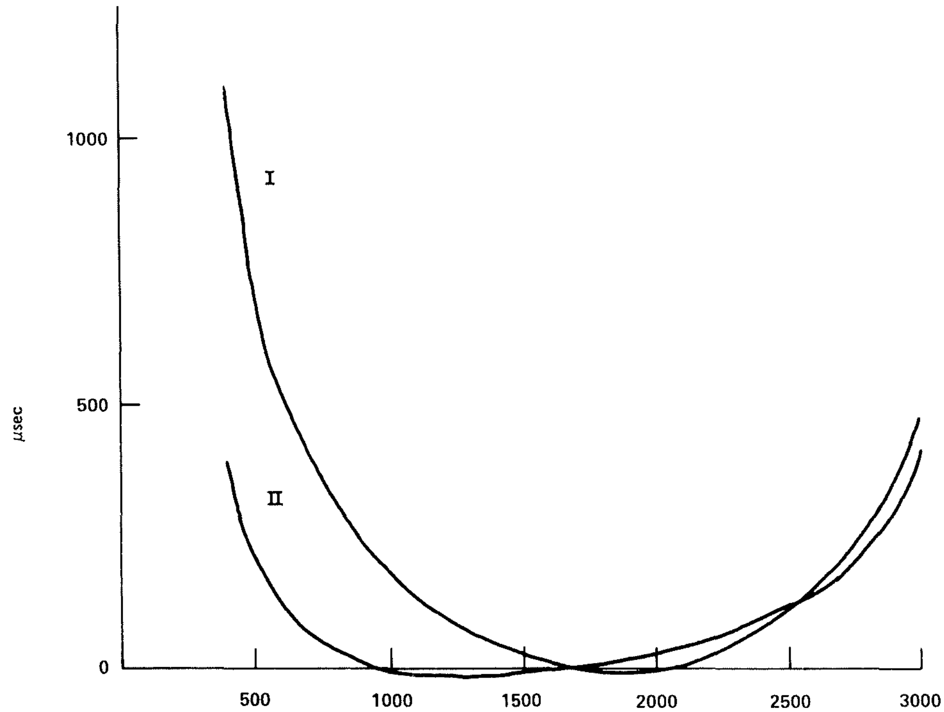


Fig. 14 – Envelope Delay Distortion as Measured on Two Different Transmission Lines

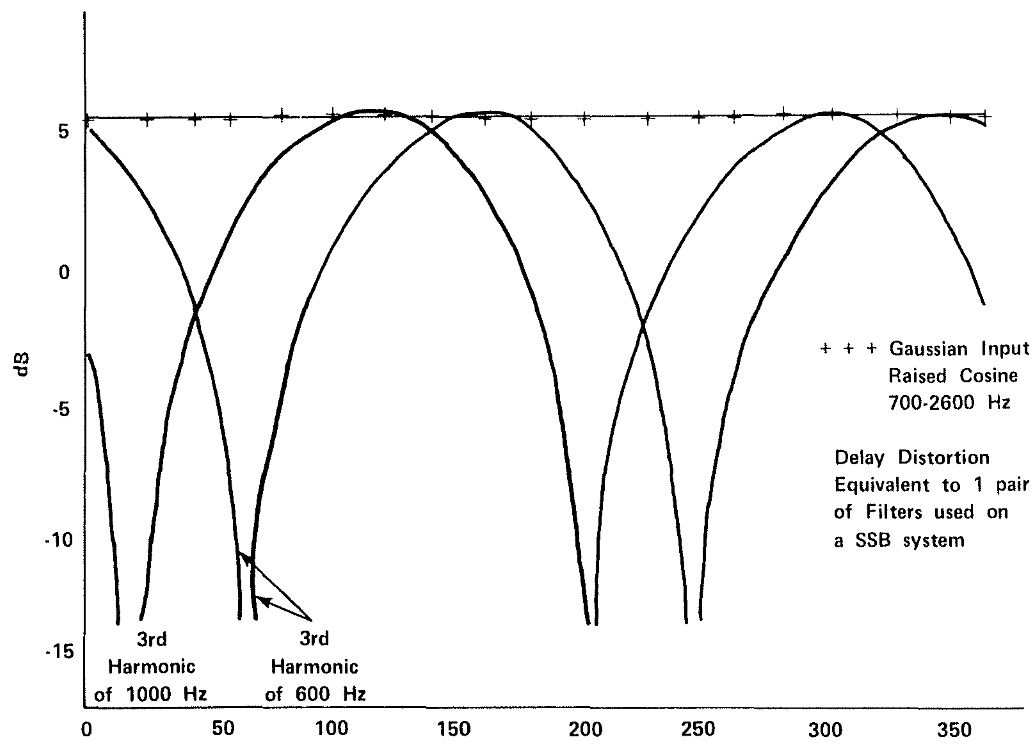


Fig. 15 – Some Third Order Products Compared to the Third Order Distortion for a Gaussian Input

with respect to the distortion generated by only one of the nonlinearities. Thus, 6 dB represents voltage addition and 3 dB power addition. K is plotted in degrees on the abscissa. A Gaussian input having a raised cosine spectrum (see Reference 5) from 700 Hz to 2600 Hz is assumed and in-band frequencies are taken to be 700 to 2600 Hz; however, results are very insensitive to spectrum shape and bandwidth. Figure 15 also shows the level of the third harmonic for two different input frequencies and the level of a 2B – A product. The third harmonics show that the nonlinear distortion is frequency dependent since different results are obtained for different input frequencies. Also, if K varies with time, the level of the third harmonic varies with time. The 2B – A product correlates best with third order distortion.

4.1.9 Several conditions are of interest for second order distortion but the following illustrate the important ones. One is for envelope delay distortion equal to that of the pair of filters described above. Since these occur on a SSB system K is a function of time due to the frequency shift mentioned above. The other is for envelope delay distortion equal to that caused by another commonly used short haul carrier system. This amount of delay distortion is shown by the curve labelled "II" in Figure 14. In this case the envelope delay distortion is less and K does not change with time. Figure 16 presents results for the latter case for two different input spectra. Notice that a null exists at some values of K and that differences between the two spectra are quite large in the null. A measurement which agrees with one spectrum will not agree with the other. However, it is important to recognize that, for physical systems, differences in this region are not serious because for these values of K the second order nonlinear distortion generated at the input of the linear network tends to cancel that generated at the output. Thus, the differences are large only when the amount of nonlinear distortion is insignificant.

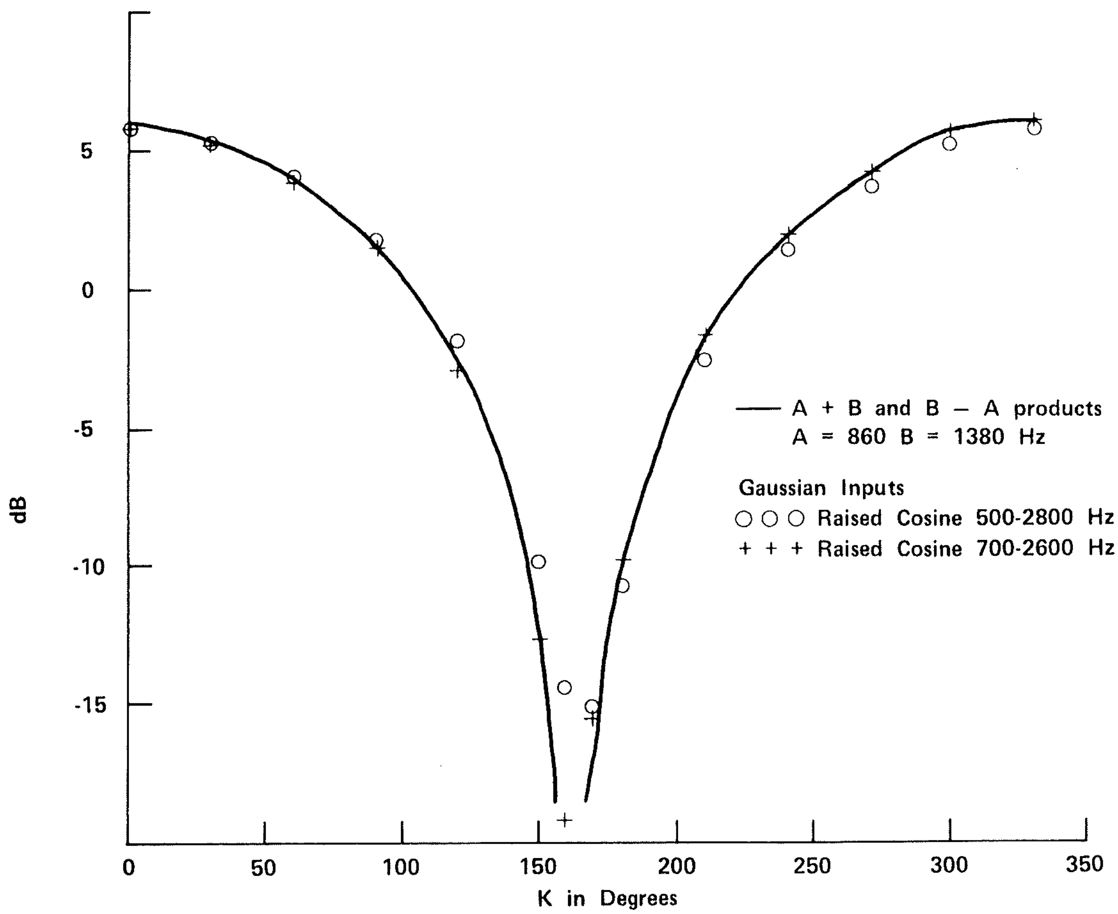


Fig. 16 – Power Sum of the A + B and B – A Products Compared to Second Order Distortion for Gaussian Inputs

4.1.10 Results in Figure 17 are for envelope delay distortion caused by the pair of filters used with SSB systems. Spectrum bandwidth has a larger effect than in Figure 16, but a null still exists at some values of K. Transmission performance in this case is a function of time since K is a function of time. However, if K is time variable there is no reason to suspect that it would spend more time at some values than at others. Thus, if the power were measured by averaging for a long enough time we would find that the second order distortion adds on a power basis. This value is labeled equivalent performance in Figure 17 and is used as a measure of quality for this circuit. This is the value the measurement should indicate when the second order distortion is time variable.

4.1.11 The models presented have demonstrated time variable and frequency dependent nonlinear distortion. It is possible that the sources of nonlinearity themselves could be frequency dependent, however, it is felt that any such effects are small over the range of frequencies of interest.

4.1.12 Nonlinear distortion test sets recently developed use two pairs of tones as the fundamental signal. This is known as the 4-tone method. For this test, the four equal level tones are transmitted at a composite signal power of data level, - 13 dBmO.

4.1.13 The 4-tone method uses two pairs of tones. One pair consists of tones at 856 and 863 Hz (a 7 Hz spacing). The second pair uses frequencies of 1374 and 1385 Hz, an 11 Hz spacing. The frequency spacing within each pair of tones is not critical but should be different for each pair. Let these four tones be called A_1, A_2, B_1, and B_2. The second-order products (A+B) fall at A_1+B_1, A_1+B_2, A_2+B_1 and A_2+B_2. If the spacing between A_1 and A_2 is the same as that between B_1 and B_2 then $A_1+B_2 = A_2+B_1$ and these two components will add on a voltage basis and give an erroneous reading. The

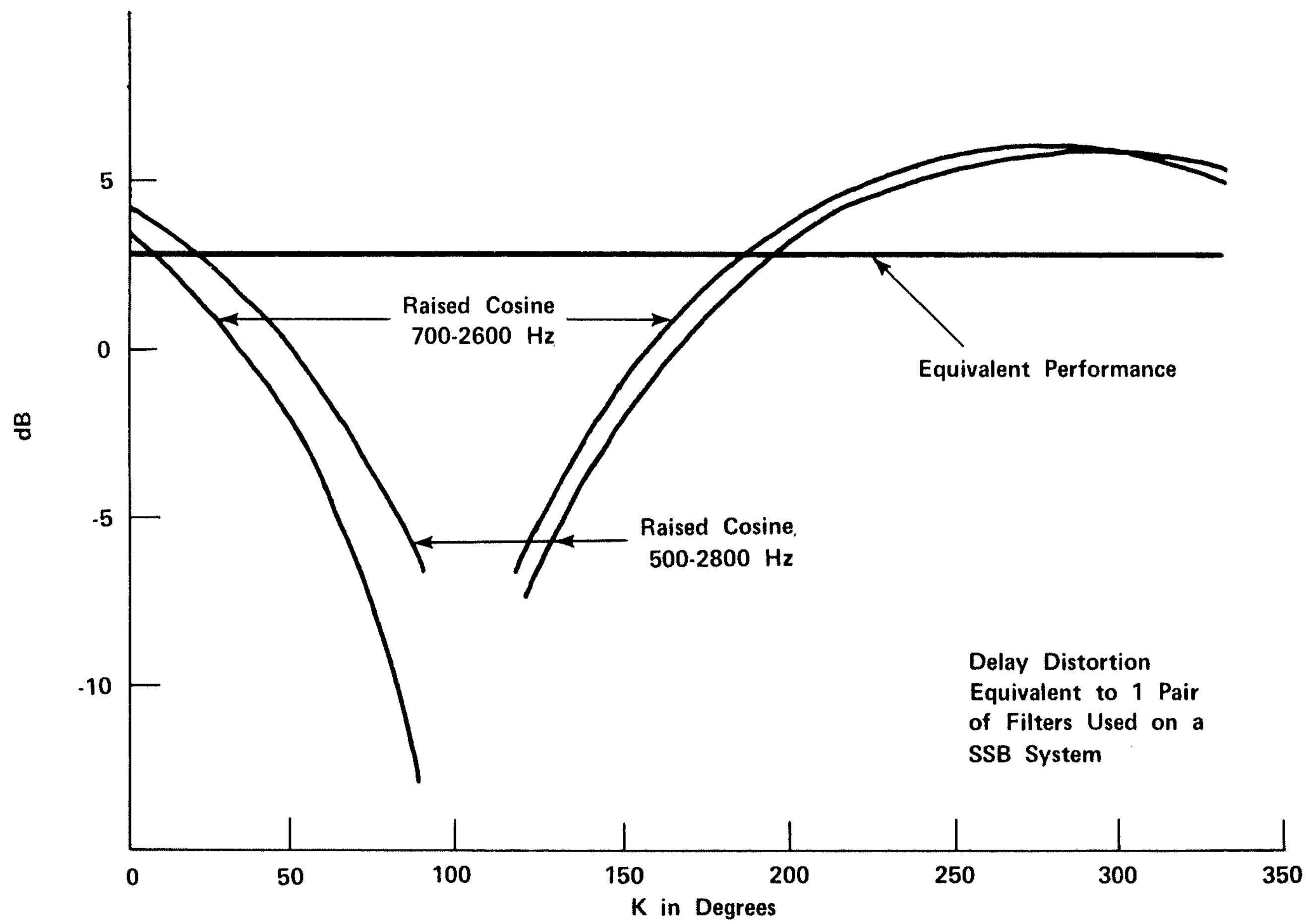


Fig. 17 — Second Order Distortion for Gaussian Inputs

third-order products (2B-A) fall at $2B_1-A_1$, $2B_1-A_2$, $2B_2-A_1$, $2B_2-A_2$, $B_1+B_2-A_1$, and $B_1+B_2-A_2$. The receiver uses 50 Hz wide filters to select the A+B, B-A, and 2B-A products. R2 is the ratio of received composite fundamentals to the power average of the A+B and B-A products. R3 is the ratio of received composite fundamentals to the 2B-A products. Two second order products are used to reduce time variability. The power sum of these two products are plotted in Figure 18 as a function of K for the amounts of delay distortion indicated on the figure. The measurement is not independent of K even though the depth of the null has been reduced. Thus, it is still necessary to average this reading for about 30 to 60 seconds. A slight risk exists that K will not change appreciably over 30 to 60 seconds and that the measurement will be in the null. The error could be up to 7 dB. This maximum error can be reduced by using a more complicated measurement scheme; however, more complicated schemes only reduce the maximum error by a dB or two. If the risk of being in the null is too great, after averaging 30 to 60 seconds, the measurement can be made twice – say 5 minutes apart – and the largest reading used.

4.1.14 For the condition used in Figure 16, in which case K is not variable but the amount of delay distortion is smaller, the power sum of the A+B and B-A products tracks the second order distortion for the Gaussian inputs.

4.1.15 The following considerations influence the choice of frequencies A and B. As seen from Figure 15, third order distortion adds on nearly a voltage basis even with envelope delay distortion present; thus, A and B must be chosen so that delay distortion has a negligible effect on the 2B – A product. For this to happen, A, B, and 2B – A should be close together and near 1700 Hz so that they appear in a relatively flat part of the envelope delay distortion curve. The A+B product is chosen less than 2300 Hz and the B-A product greater than 500 Hz to keep the effects of channel roll-off small. As B-A increases above 500 Hz, the depth of the null in Figure 18 (the maximum error) increases. Thus, B-A is kept close to 500 Hz. Finally, since phase jitter components might occur within 300 Hz of any signal component, the nonlinear distortion products must be at least that far removed from a signal component.

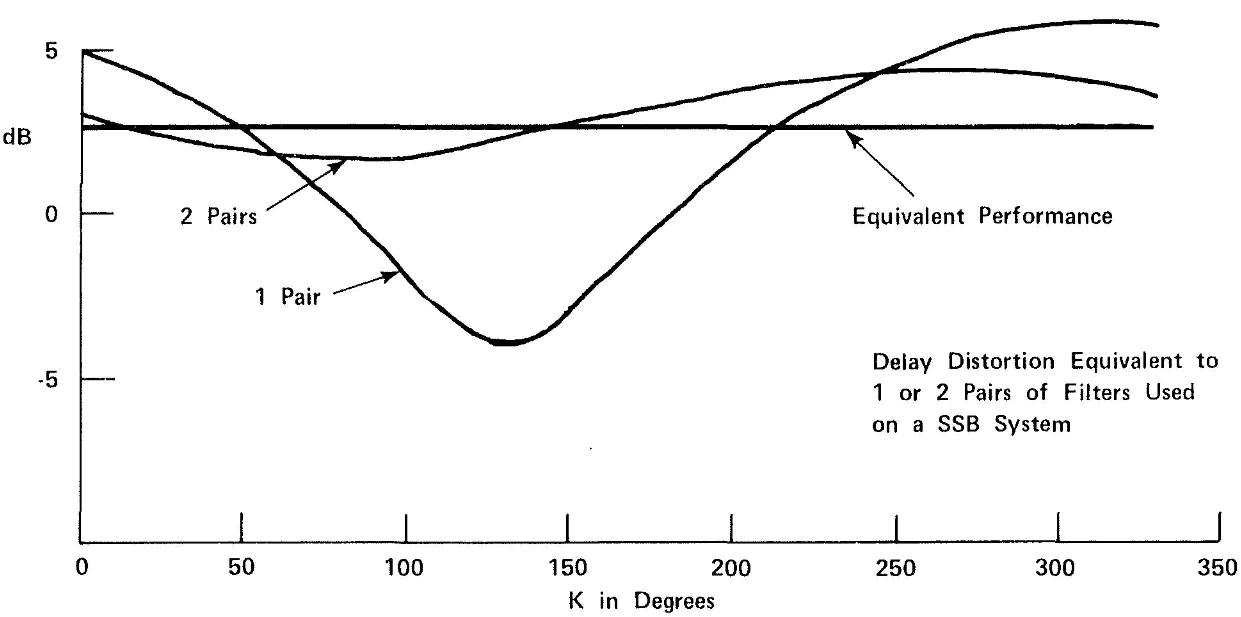


Fig. 18 – Power Sum of B – A and A + B Products
(A = 860 Hz, B = 1380 Hz)

4.1.16 The measurement must be able to identify the nonlinear distortion accurately in the presence of background noise and quantizing noise. The problem is especially severe for measuring third order distortion on part of a built-up connection because third order distortion adds approximately on a voltage basis and noise on a power basis. (Suppose a connection has third order distortion that is 6 dB above the noise and that the connection has four identical pieces. The third order distortion on each piece is 12 dB below that on the connection while the noise is only 6 dB below that on the connection. Thus, the third order distortion and noise are at the same level on each piece.)

4.1.17 Quantizing noise on a PCM system poses a special problem because when a PCM system is excited with tones, the noise spectrum is not flat and continuous but discrete. Some of these discrete components add or beat with the nonlinear distortion product being measured causing an inaccurate and time variable reading. Third order distortion is measured through a narrowband, 50 Hz, filter centered at 2B-A and second order distortion through narrowband, 50 Hz filters centered at B-A and A+B. These values are then referred to the received signal level.

4.1.18 To protect against false measurement because of high noise or an interfering tone in one of the measurement slots the following test is made. Disable the pair of tones at B and increase the others 3 dB. This loads the system properly for a noise measurement. The part of the measurement due to noise alone can now be determined. Figure 19 is then used to determine the correction factor which must be subtracted from the distortion measurement (or added to the signal-to-distortion measurement).

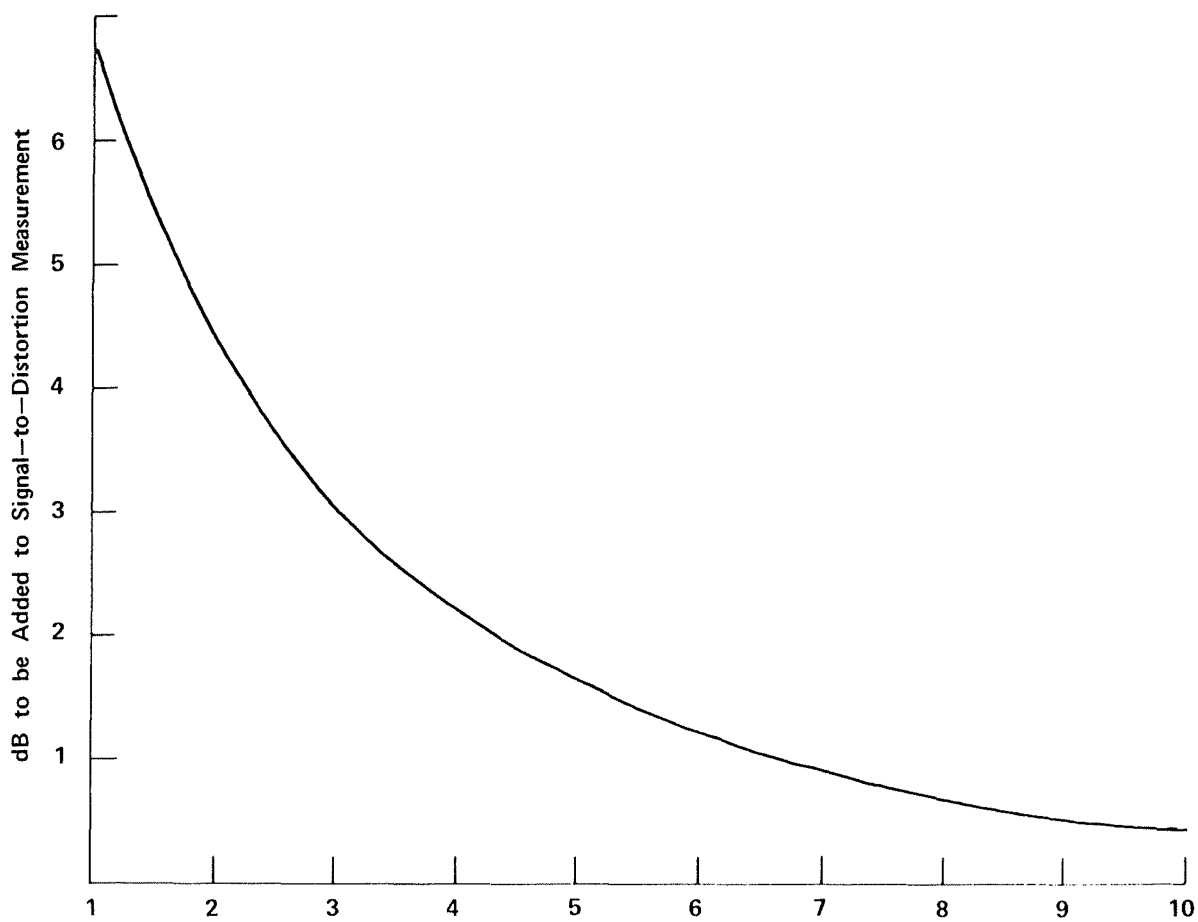


Fig. 19 – Correction For Noise

4.1.19 It is usually thought that harmonic distortion measurements follow what is called the "power series" law, i.e., the second and third harmonic increase 2 and 3 dB respectively per dB increase of the input power. This is based on the third degree polynomial model for nonlinearities. Measurements on telephone channels may not follow this law for two reasons: the presence of significant terms above third degree and the dependence of the magnitude of the nonlinearity on the input power.

4.1.20 An example of the first reason is a full wave rectifier nonlinearity that can occur on a PCM carrier system.[11] The second harmonic produced by a full wave rectifier changes only 1 dB per dB change in input level. However, this nonlinearity can be well approximated with a_2x^2 if a minimum mean-squared error approximation is used. The approximation then depends on the amplitude distribution of the signal. Practically, this means that to get a good approximation the amplitude distribution of the test signal should be similar to that of a typical data signal or that they should span roughly the same range of amplitudes. For example, a 2-tone test signal should be about 2 dB above the data signal. A 4-tone test signal should have the same rms power as a data signal.

4.1.21 The magnitude of the nonlinearity on syllabic compandored systems depends on the compandor operating point.[12] To produce the proper operating point, the test signal should have the same rms power as the data signal. Because of this, second and third order distortion might change only slightly more than 1 dB per dB change in input.

4.1.22 The signs of b_2 and b_3 in Figure 13b are not determined by the measurement. The sign of b_2 is just as likely to be positive as negative and in either case the effect on transmission performance will be the same. However, negative values of b_3 result in a compression of signal peaks and can be more interfering to data transmission than (enhancing) positive values. Although it is possible that a nonlinearity may be enhancing for certain input powers, it appears that most nonlinearities encountered are compressing.

4.2 Incidental Modulation

4.2.1 Incidental modulation is defined as any unwanted AM, PM, or FM imposed on the information carrying voiceband signal by a disturbing source other than itself. This is a signal correlated interference because it uses the signal as a carrier, but it is a parasitic rather than a reflexive interference like nonlinear distortion. Spurious AM imposed by electronic components with faulty power supplies and extraneous PM introduced via unstable carrier frequency sources are examples of incidental modulation. Jittering clock pulses in digital carrier systems may also contribute small amounts of phase modulation to a voiceband signal, but this is at most a second order effect.

4.2.2 To observe the important properties of this type of interference on frequency division multiplex systems, a single-frequency sinusoid* in the center of the voiceband is transmitted. Let this signal be represented by $V_t = A_O \cos 2\pi f_O t$, where $1000 \leqslant f_O \leqslant 2000$ Hz. The received waveform of this tone in the absence of nonlinear distortion products can be simply expressed as follows:

$$V_r = A_O G(f_O)[1+m(t)] \cos[2\pi f_O t + \phi(f_O) + \theta(t)] + n(t) \quad (4.2)$$

where

$G(f_O)$ and $\phi(f_O)$ are the channel amplitude and phase characteristics respectively at frequency f_O,

$n(t)$ is the total uncorrelated interference,

$m(t)$ is incidental AM, and

$\theta(t)$ is incidental PM $\left[\frac{d\theta(t)}{dt} \text{ is incidental FM}\right]$

The remainder of Section 4 is devoted to the characterization and measurement of $\theta(t)$ and $m(t)$. Transient excursions on $\theta(t)$ and $m(t)$, usually referred to as phase hits and gain hits will be treated separately from the steady-state components in the following sections.

*Single-frequency test tones are not recommended for systems using time division multiplexing for reasons given in Section 3.4.

4.3 Incidental Phase Modulation (PM and FM)

4.3.1 Incidental PM defined in Section 4.2 can be expressed in general as the following summation:

$$\theta(t)=\theta_O+[2\pi\Delta f(t)]+\sum_{j=1}^{N}\theta_j\cos(2\pi f_j t+\phi_j) \qquad (4.3)$$

θ_O when not equal to $n2\pi$; $n = 0,1,2,\ldots$ is phase intercept distortion. It contributes identical phase shift to all frequencies present in the voiceband signal; thus, it appears as the zero intercept on a graph of phase versus frequency for carrier derived channels. It may appear time variable due to contributions from the second and third terms.

Δf is steady-state frequency shift in Hz. It also contributes equally to all frequencies of the voiceband signal. It has the effect of shifting the signal spectrum.

θ_j, f_j, ϕ_j define N-independent sinusoidal components of the phase modulation. Note that the summation does not imply a Fourier expansion of $\theta(t)$, because the f_j are not necessarily harmonically related. Since the peak deviation in radians is usually small ($\Sigma\phi_j < 0.2$), this constitutes low index PM with a single set of sidebands ($f_O \pm f_j$) for each of the N components. The two sidebands are present for each frequency present in the voiceband signal, so that, in general, spectral symmetry is not guaranteed for the composite signal.

4.4 Phase Intercept Distortion

4.4.1 Phase intercept distortion is θ_O in Equation (4.3); it is the frequency and time invariant component of phase shift in the received signal waveform. For reasons already given in 2.4, under basic channel characteristics, measurement of the phase shift through a telephone channel is difficult to perform. Even on a looped back connection through the telephone network it is difficult to isolate phase intercept distortion from the phase nonlinearity discussed in Section 2.4. Phase intercept distortion is not easily controlled on carrier systems but it has no adverse effects on voice transmission. Designers of terminal modulating equipment for nonvoice signals have circumvented phase intercept distortion by frequency translation of baseband signals and derivation of the local demodulating carrier from received waveforms. No measurement techniques are discussed here, therefore, nor will any instrumentation be suggested.

4.5 Frequency Shift

4.5.1 The Δf in the second term of Equation (4.3) is frequency shift, which may be generated in the following manner. The Bell System makes extensive use of single sideband carrier transmission facilities. In these systems the carrier frequency is not transmitted with the signal so that the signal may be demodulated with a locally generated carrier that differs slightly from the modulating frequency. This introduces a fixed frequency shift for all single frequencies by an amount equal to the difference between the modulating and demodulating frequencies. In some older carrier systems still in use, frequency shifts greater than 5 Hz have been observed due to relatively poor control of the difference between modulating and demodulating carriers. These account for less than 1 percent of all carrier derived channels, however, and in newer systems frequency shift is held to less than 1 Hz per section. Measurement of carrier frequency shift requires a tone source near 1000 Hz stable to at least one cycle per million. Both the transmitted and the received waveform's zero crossings must then be observed with frequency counters accurate to within ±0.01 percent. The difference in zero crossings counts is frequency shift to the nearest 0.1 Hz.

4.5.2 Frequency counters are generally not balanced to ground. They may be sensitive to extraneous noise picked up on the test leads or from longitudinal currents on the line under test. To avoid disruption and insure repeatability of frequency measurements, it is desirable to place a 200-300 Hz bandpass filter centered at 1020 Hz just ahead of the counter. Alternatively, the de-modulated carrier from the phase locked loop, such as in a phase jitter set, may be employed for this measurement.

4.6 Phase Jitter

4.6.1 The third term in Equation (4.3) represents all of the AC components of incidental PM, which cause the zero crossings of a voiceband signal to "jitter." Phase jitter measurements also include the disturbing effects of uncorrelated interference and quantizing noise. In fact, phase jitter measurements should always be accompanied by a signal to noise measurement to ascertain what portion of the measurement is due to incidental PM and how much is noise. Figure 20 demonstrates jitter readings produced by quantizing noise on a time division multiplex system and Figure 21 illustrates the effect of uncorrelated white Gaussian

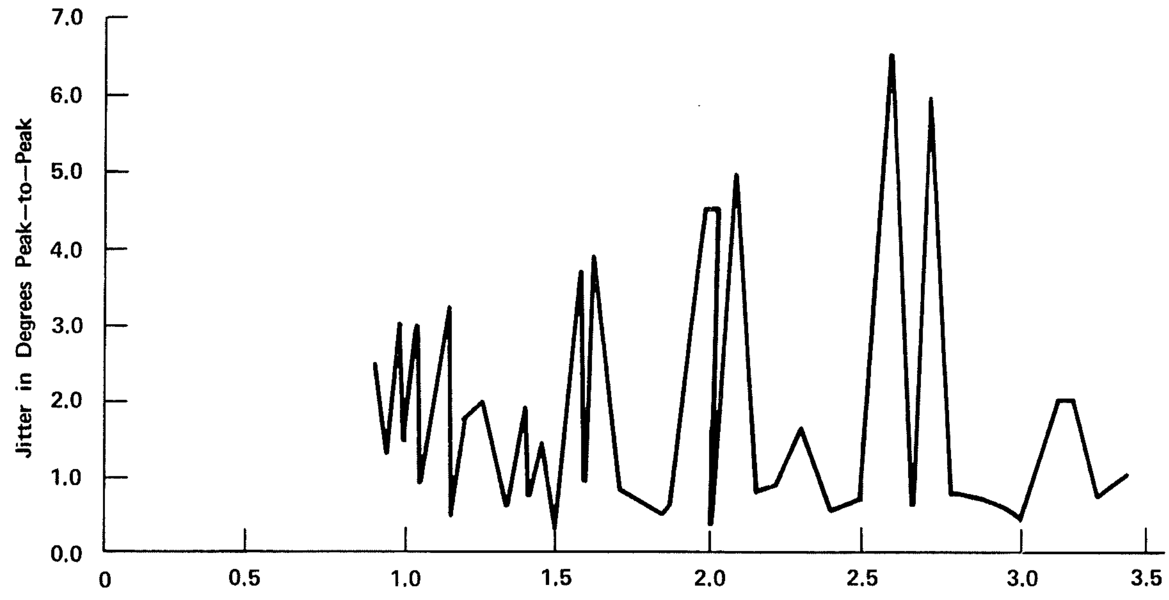


Fig. 20 – Typical Phase Jitter Measured on a PCM System as a Function of Carrier Frequency (110 Hz Bandwidth)

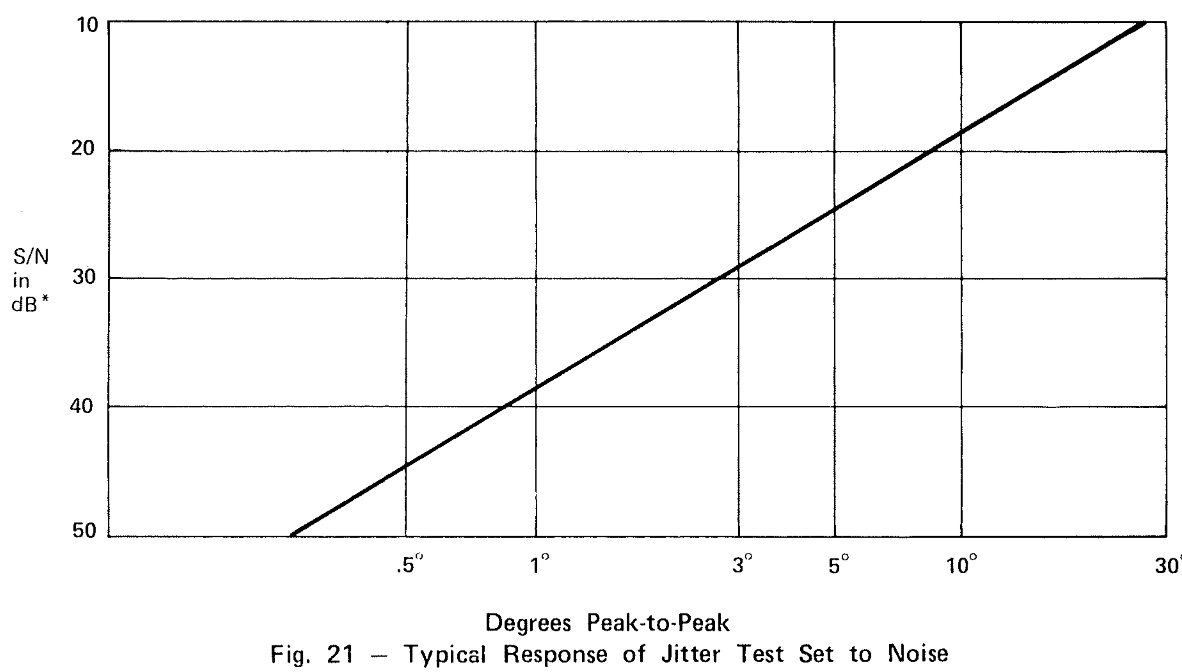


Fig. 21 – Typical Response of Jitter Test Set to Noise

*3.3kHz of White Gaussian Noise

noise on a typical phase jitter measuring set. A test set that would respond to pure phase modulation only and ignore noise is highly desirable but none is currently available.

4.6.2 The most commonly found single-frequency components of phase jitter are 20 Hz (Ringing Current), 60 Hz (Commercial Power) and the second through fifth harmonics of these. Since the peak phase deviation caused by AC components of PM rarely exceeds 0.2 radians (low index phase modulation) only one pair of significant sidebands are produced for each sinusoidal component. Hence, a bandwidth of about 600 Hz centered about a carrier at or near 1020 Hz suffices to recover the major suspected sinusoidal PM without incurring large amounts of uncorrelated interference.

4.7 Incidental Amplitude Modulation (AM)

4.7.1 Incidental AM on telephone channels takes the form of low index double sideband modulation of voice-band signals. Thus, referring to Equation (4.2), m(t) is envelope modulation on the sinusoidal carrier of frequency of f_0 with received amplitude $A_0G(f_0)$. The incidental AM waveform may be expressed as follows:

$$m(t) = \sum_{a=1}^{M} m_a \cos(2\pi f_a t + \phi_a) \qquad (4.7)$$

where $m_a, f_a \phi_a$ define the M-independent sinusoidal components of the amplitude modulation with low peak index ($\Sigma m_a < 0.2$). (Note again that use of the summation does not mean to imply a Fourier series.) A single set of sidebands, $f_0 \pm f_a$, are produced for each AM component but, except for their phase relationship to the carrier signal, these are indistinguishable from PM sidebands. Also, since incidental AM is low index, only small peak-to-peak excursions of carrier are evident and simple observation by oscilloscope or envelope detection makes the AM extremely difficult to distinguish from additive uncorrelated interference (especially hum and other single frequency interference). Hence, as was the case for incidental PM, band limiting and removal of other interfering modulation by a circuit analogous to the limiter for PM are implied.

4.7.2 Little concern has been expressed up to the present time over incidental AM on telecommunications channels. Its effect on voice transmission is negligible but with the advent of high speed (9600 bps) modems it may become a parameter of interest. Some high speed modems show a sensitivity to incidental AM when it reaches a magnitude of about 10%. Almost no data exist to estimate the magnitude of this impairment in the network but there have been no reported cases of it causing problems. Therefore, at this time, no standard measuring technique will be described.

4.8 Rapid Gain and Phase Changes

4.8.1 Gain and phase changes that occur very rapidly may be encountered on telecommunications channels. These are transient phenomena that might be thought of as components of the m(t) and $\phi(t)$ of Equation (4.2), but have not been discussed thus far. Some of the more common causes of these phenomena are automatic switching to standby facilities or carrier supplies, patching out working facilities to perform routine maintenance, fades or path changes in microwave facilities and noise transients coupled into carrier frequency sources. The channel gain and phase (or frequency) shift may return to its original value in a short time or remain at the new values indefinitely.

4.8.2 Gain changes are typically detected by changes in an AGC circuit and phase changes by means of a phase locked loop. In order to provide protection against the detectors falsely operating on peaks of uncorrelated noise (impulse noise), a guard interval of 4 milliseconds is designed into the peak indicating instrument. Unfortunately, such a guard interval will also effectively mask out true phase impulses shorter than 4 milliseconds that are not also accompanied by a peak amplitude excursion large enough to be detected by the threshold devices described in Section 3.2. This risk is considered justified at this time when the known relative frequencies of phase jumps are compared with those for impulse noise.

4.8.3 Instruments used to measure gain and phase hits, as the rapid gain and phase changes are usually called, do so by monitoring the magnitude and phase of a sinusoidal tone. Hits are recorded and accumulated on counters with adjustable threshold levels. Gain hit counters typically accumulate events exceeding a threshold of 3 dB, although they do not distinguish an increase from a decrease of magnitude. Similarly, phase hit counters accumulate changes at thresholds from 10 to 45 degrees in 5-degree steps. They respond to any hits equal to or in excess of the selected threshold. A switch which removes the impulse noise blanking feature under the user's discretion may be desirable, when impulsive phase hit activity is suspected. As with the impulse noise counters discussed in Section 3.2, a controlled dead time of 140 ms should be built into the counters in order to obtain consistent readings with sets of different manufacture.

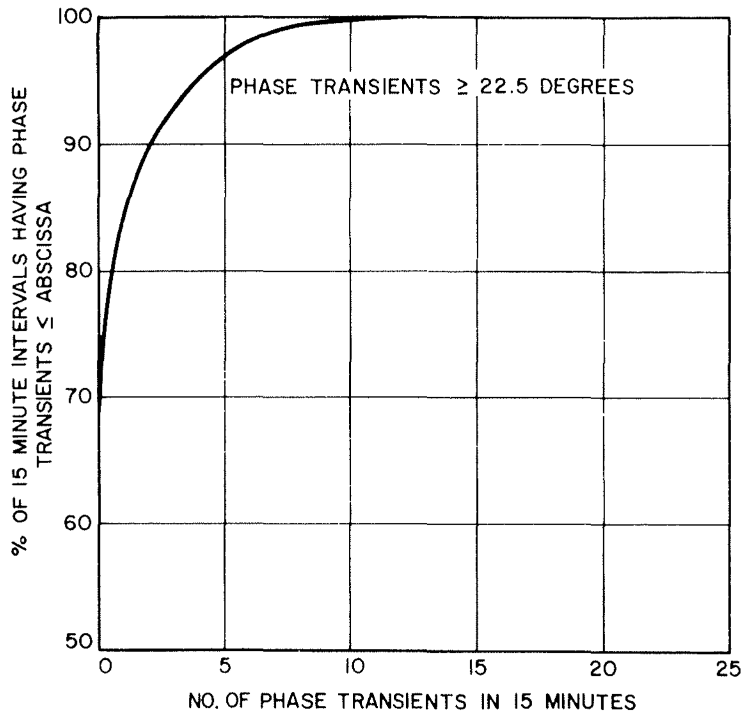
4.8.4 One serious problem with gain and phase hit counters occurs under the following condition: A signal drop out (see Section 4.9) of on the order of 20 to 30 dB may occur and the background noise may simultaneously rise to a value near the original signal level. The phase locked loop will still recognize the desired signal but erroneous gain and phase hits are recorded due to the influence of the noise. This problem is partially mitigated by blanking the phase and gain hit counters during a drop out.

4.8.5 Phase Hits and Gain Hits

The following statistics on phase hits (and gain hits) were collected by monitoring five long-haul channels for a total of about 70 hours. They are meant to be descriptive only and do not quantify the occurrence of these events in Bell System channels.

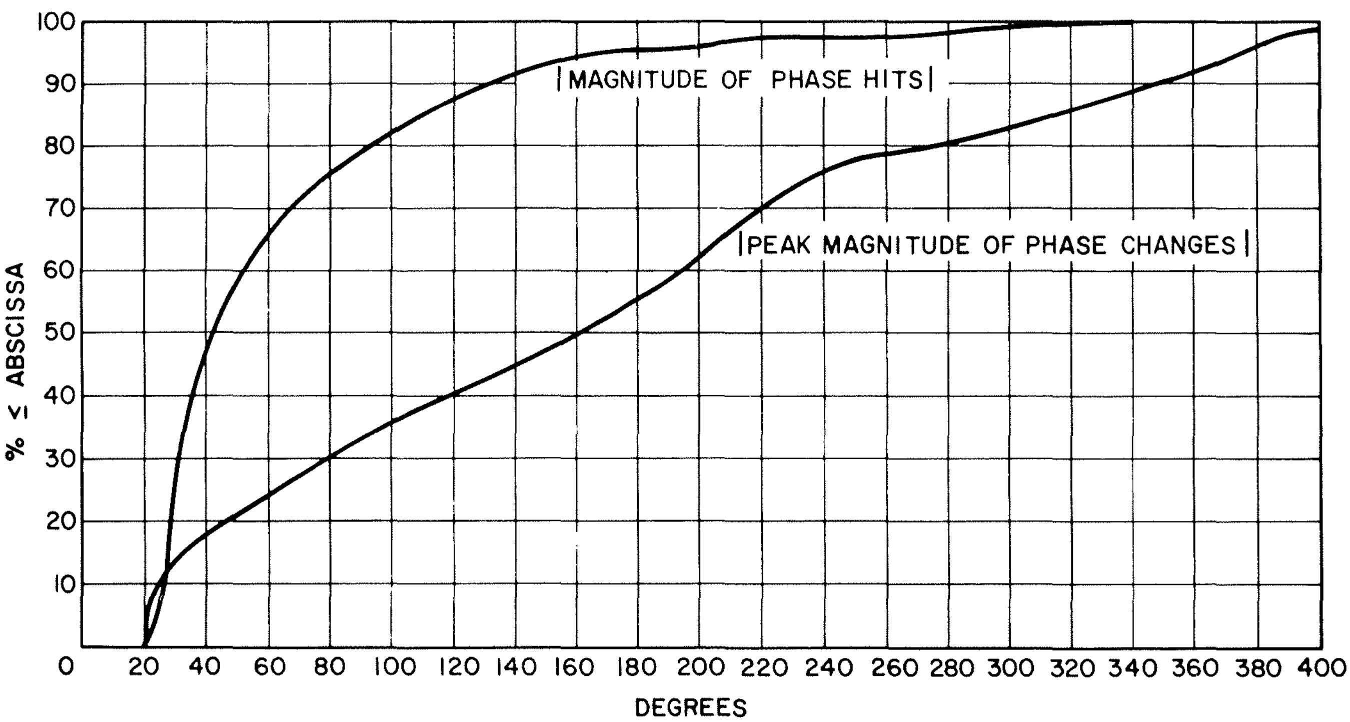
4.8.6 When the 70 hours was broken down into 15-minute intervals, it was observed that 70 percent of the intervals had no phase hits in excess of 22.5 degrees. The interval with the largest number of phase hits had 12 of them; see Figure 22. The distributions of the magnitudes of phase hits and phase changes are shown in Figure 23. A phase hit is defined as being a phase change which lasts for a short period of time after which the signal switches back to its original phase. A phase change is defined as one which occurs in the signal and it then remains at this new phase for an indeterminate amount of time. Both plots in Figure 23 start at 22.5 degrees. This was an instrumentation limitation. Phase changes less than 22.5 degrees were not recorded. The distribution of the durations of phase hits is shown in Figure 24. The distribution shown is a conditional one. The phase hit duration had to be between 1.6 and 220 ms to be entered into the distribution. Most data modems will have accommodated the new phase before 220 ms and be operating error free again. This also was an instrumentation limitation. Note that approximately 13 percent of the phase hits are shorter than 4 ms in duration and would be missed by a phase hit counter which has the 4-ms guard interval in it.

4.8.7 Gain hits appear to be more common in the network than phase hits. The distributions of occurrences in 15-minute intervals are shown in Figure 25. Note, that at a 2-dB detection level, only 58 percent of the intervals had no gain hits. The largest number of gain hits observed in 15 minutes was 27. Also, as might be expected, the larger gain hits occur less frequently than the small ones. Figure 26 shows three distributions of the amplitudes of gain hits. Figure 27 shows distributions of the durations of gain hits for various detection levels. Note that most (95 percent) gain hits in excess of 10 dB in fact, last longer than 10 ms.



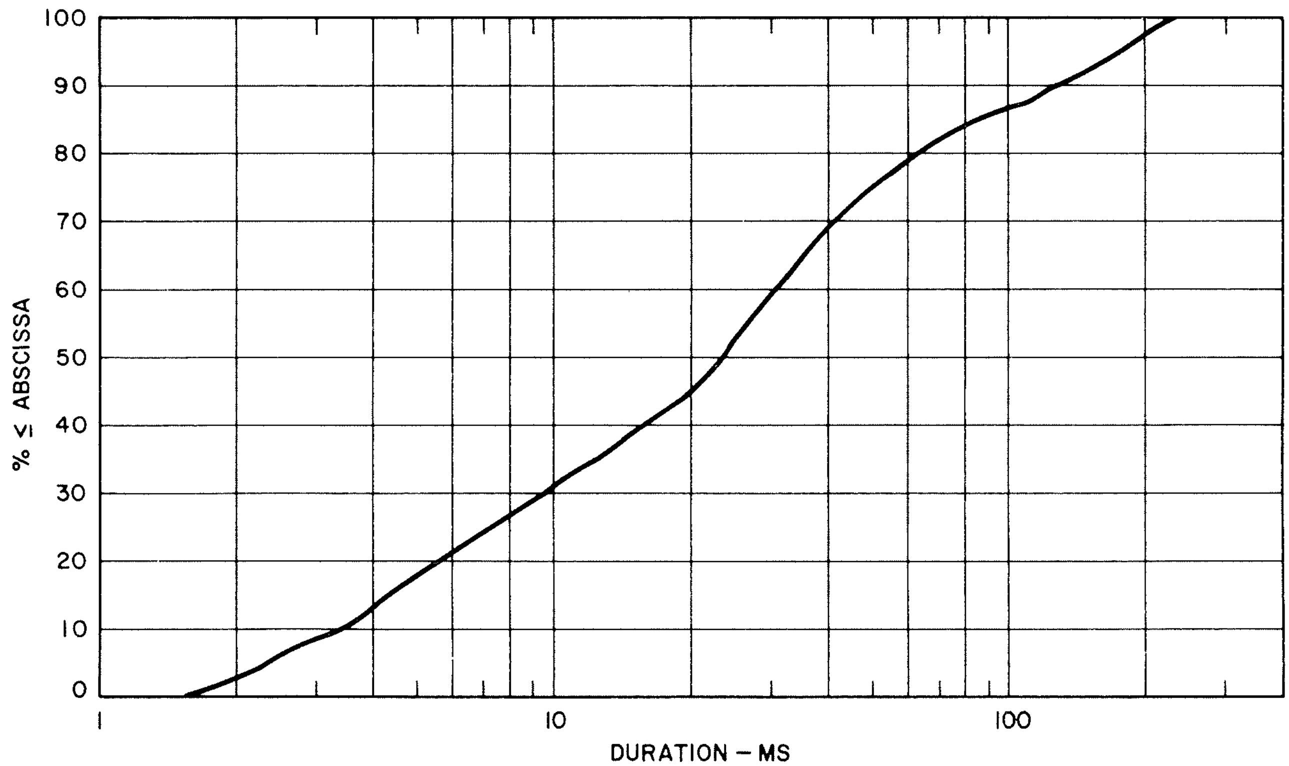
DISTRIBUTION OF PHASE TRANSIENT COUNTS IN 15 MINUTES

Fig. 22



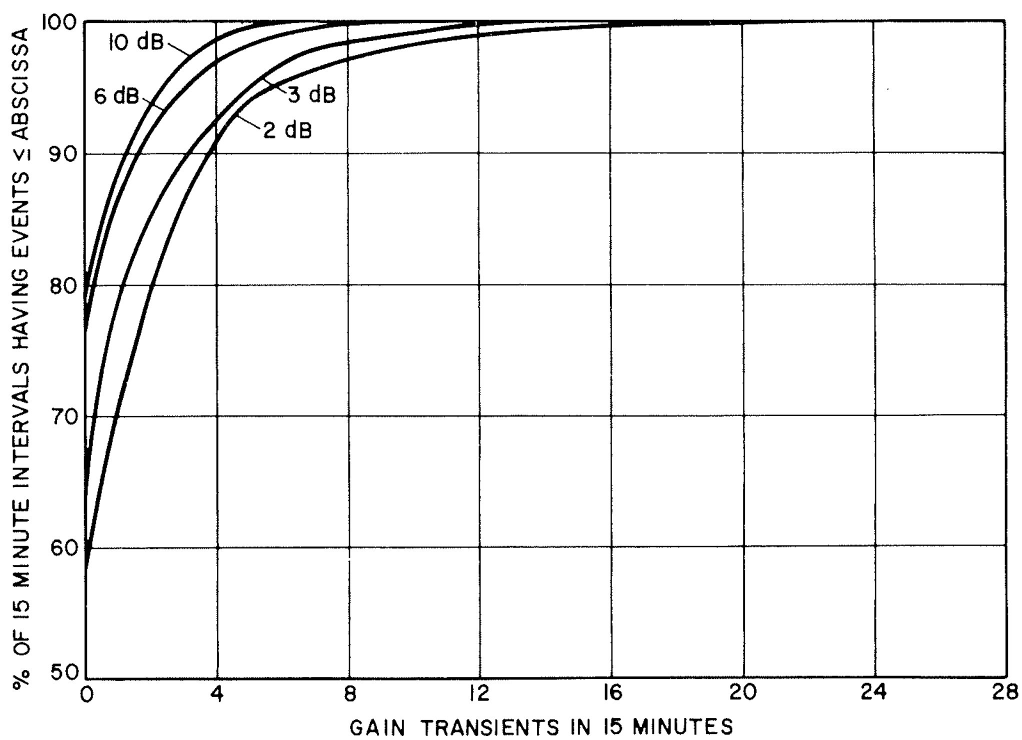
DISTRIBUTIONS OF MAGNITUDES OF PHASE HITS AND PHASE CHANGES

Fig. 23



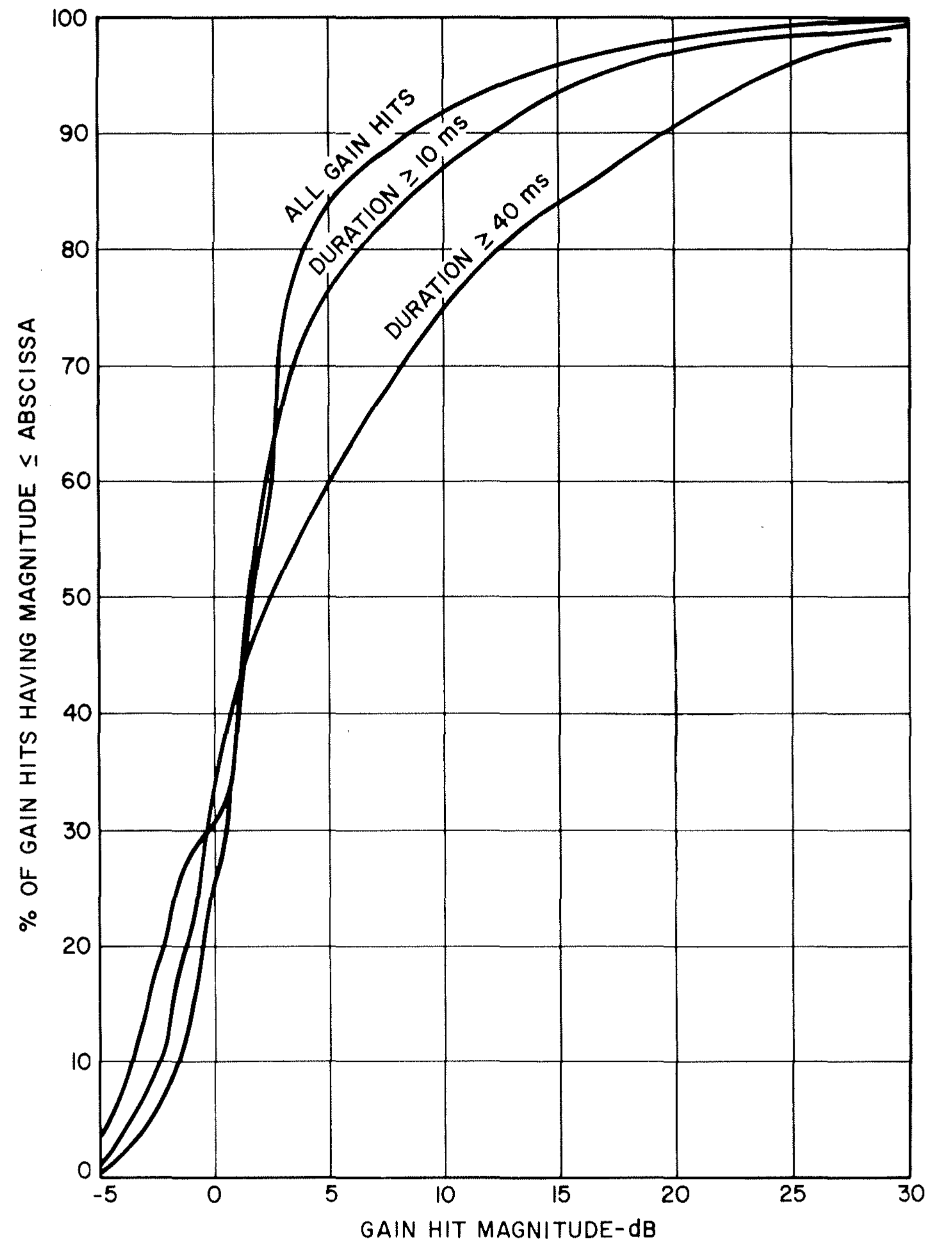
DISTRIBUTION OF DURATION OF PHASE HITS > 22.5°
GIVEN 1.6 ≤ DURATIONS ≤ 220ms

Fig. 24



DISTRIBUTION OF NUMBER OF GAIN HITS IN 15 MINUTES

Fig. 25



DISTRIBUTIONS OF MAGNITUDES OF GAIN HITS

Fig. 26

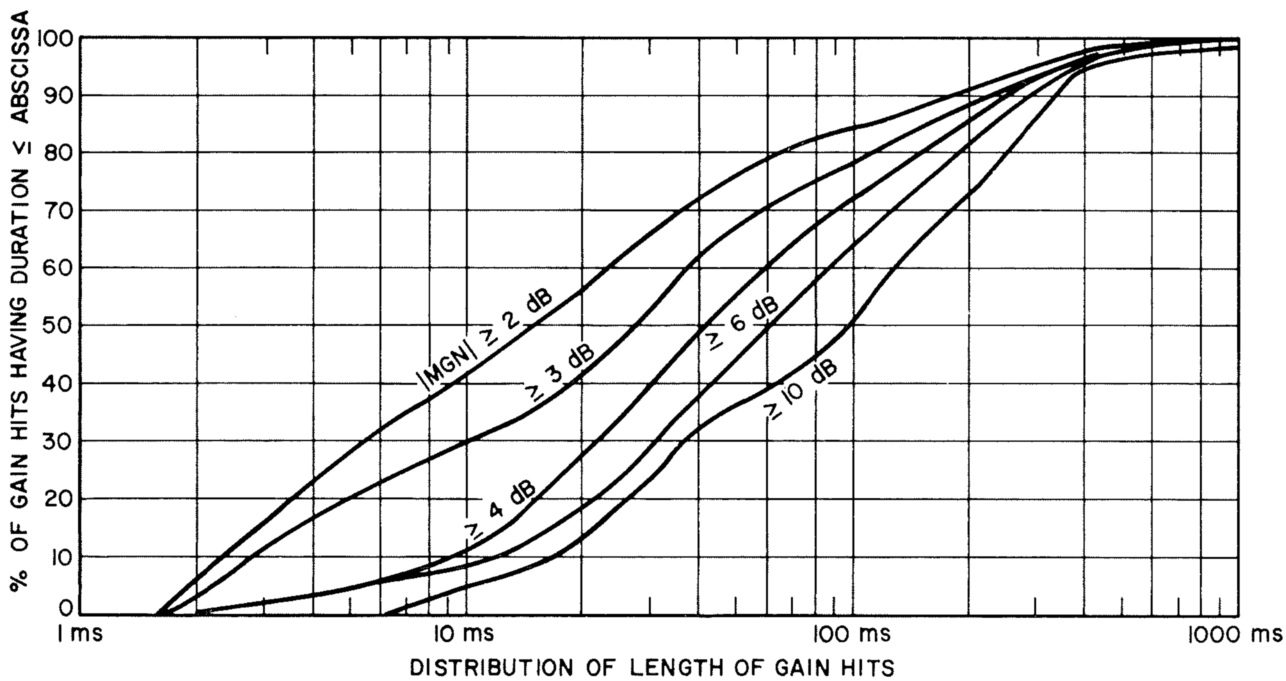


Fig. 27

4.9 Dropouts

4.9.1 Given the A.G.C. range of most modems and from observations made on the channels mentioned previously in Section 4.8.6, it would appear that a most reasonable threshold for defining a dropout would be 12 dB. This would allow moderate decreases in signal level and still permit the modem to operate. Therefore, 12 dB is defined to be the threshold for calling large negative gain hits dropouts.

4.9.2 The data in Figure 27 can be extrapolated to estimate the minimum length of a 12 dB dropout. This occurs at about 10 ms. This then completes the definition of a dropout. It is any negative gain hit which is equal to or greater than 12 dB and lasts for at least 10 ms.

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