Carrier and Radio

Articles Selected from

The *Lenkurt* Demodulator

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The Lenkurt Demodulator

Lenkurt Electric Co., Inc.

San Carlos, Calif. • Mexico, D.F. • Vancouver, B. C.

FOREWORD

The purpose of this volume is to present in bound form a selection of articles from past issues of The Lenkurt Demodulator covering subjects directly related to carrier and radio transmission. The selection was based on requests from readers for back issues of The Demodulator, which are no longer available, and on many suggestions which we received for a compilation of information regarding carrier and radio transmission techniques.

The articles have been grouped in chapters arranged more or less in the order of development of carrier communication. However, equipment testing and measuring techniques are discussed first. The next chapter deals with open-wire and cable facilities, including transmission line characteristics and application, line treatment and line equipment. Telegraph and digital data transmission are covered in the following chapter, while the last chapter is devoted to radio transmission including propagation characteristics of microwaves, engineering considerations for radio routes, and operation of certain basic high-frequency components.

A detailed index has been added for easy reference to any particular subject covered in the book.

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VOL. 3 NO. 1 JANUARY, 1954

An Explanation of DBA AND OTHER LOGARITHMIC UNITS

Logarithmic units are widely used in the communications industry as the most practical and convenient means of expressing the extremely wide range of power ratios encountered. Ratios between the strongest signals and noise may be as great as 100 million billion to one. The decibel is the most common logarithmic unit. Some units such as dba, dbRN, vu and dbx are related to the decibel but require more complex definitions.

In this article dba, dbRN and vu are explained in detail. Shorter explanations of other units are also included.

The decibel is commonly used in the communications industry to express a ratio between two quantities of power. Neither quantity needs to be defined to express the ratio in db, and for many purposes a knowledge of the ratio alone is sufficient. For example, the gain of a linear amplifier or the attenuation of a pad can be expressed in db without knowing either the input or output power of the device. Frequently, however, there is a need to know the ratio between the power at some point in a circuit and some fixed, known quantity of power. When this information is needed, it is customary to express the ratio as so many db above or below a reference power.

The most common reference power used in the telephone industry is the milliwatt. Because power in telephone systems is almost always undergoing attenuation (a division process) or amplification (a multiplication process) the expression of power directly in watts or milliwatts would often require lengthy and cumbersome calculations. A more convenient method of indicating an amount of power is to express it as being so many db above or below a reference power of one milliwatt because adding and subtracting decibels provides the same result as multiplying and dividing power. Because of its common usage, "decibels above or below one milliwatt" is usually abbreviated ± dbm.

In addition to dbm, there are several other logarithmic units in use in the telephone industry which are expressed as db above or below a reference power. The most important of these are dba, dbRN, dbx and vu.

Dba and dbRN are often used to show the relationship between the interfering effect of some noise frequency or band of noise frequencies and a fixed amount of noise power commonly called reference noise. Dbx is used as a measure of crosstalk coupling on transmission lines and expresses the relationship between some value of coupling and a reference coupling. Vu is used to express the ratio between the volume of spoken or musical sounds and a reference volume.

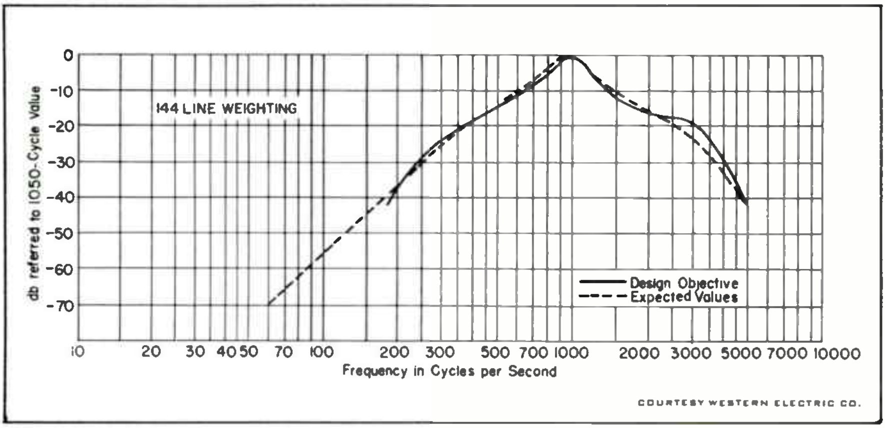
DBA and DBRN

Dba and dbRN are closely related units. In fact the abbreviation dba means, effectively, dbRN adjusted. Both terms originated as a result of research conducted by the Bell Telephone Laboratories and the Edison Electrical Institute to determine the transmission impairment caused by noise interfering with speech. Since noise may consist of random frequencies with widely varying amplitudes, it was necessary to evaluate the interfering effects of single frequencies or relatively narrow bands of frequencies to obtain usable data. A large number of listening tests were made with different tones introduced as interference. The degree of interference was determined by comparing the power of each tone with the power of a 1,000 cycle tone that created the same degree of interference.

For example, if the interfering effect of a particular tone was to be determined, that tone was superimposed on a specially selected conversation at a reference power level. When the interfering tone was turned off, a 1,000 cycle tone was superimposed on the same conversation and its power level adjusted until the listener determined that it had the same interfering effect. The difference noted between the power levels of the selected tone and the 1,000 cycle tone was then considered to be the difference in interfering effect.

When this same test was performed for a number of different tones in the voice frequency spectrum and for a number of different listeners using the same apparatus, it was possible to plot graphs such as shown in Figures 1 and 2 to show the relative interfering effects of different frequencies in the voice frequency spectrum compared to 1,000 cycles. These curves are called weighting curves. With this information available it was pos-

FIGURE 1. 144 Line Weighting



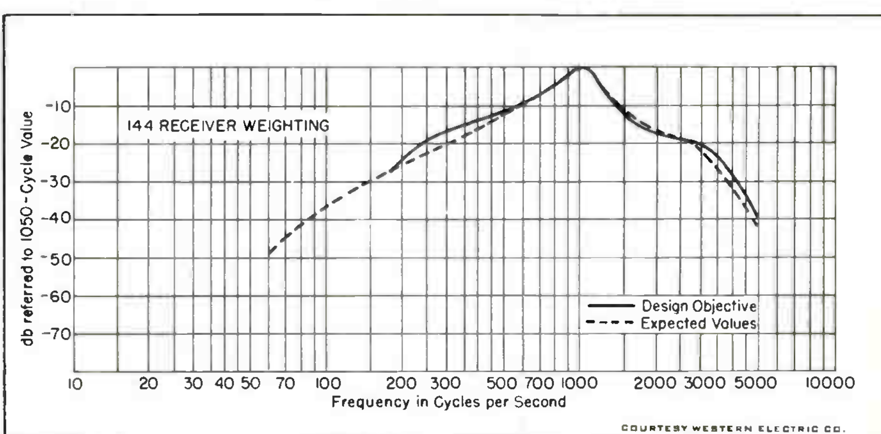
sible to construct equalizing networks such that each component frequency of the voice frequency spectrum was attenuated in the same manner as it appeared to be attenuated by the average ear with the listening test apparatus. By using these equalizers in conjunction with a suitable amplifier, rectifier and d-c meter it was further possible to measure electrically the interfering effect of any frequency or combination of frequencies. A simple block diagram of a noise meter is shown in Figure 3.

Since any noise or tone superimposed on a conversation has an interfering effect, it was desirable to express all quantities of interfering effect in positive numbers. To accomplish this, a power of 10^{-12} watts at 1,000 cycles was selected as the reference power because a 1,000 cycle tone having a power level of 10^{-12} watts or -90 dbm appeared to have negligible interfering effect. Therefore, all other noise powers likely to be encountered would have a positive interfering effect that could be expressed in db above reference noise of 10^{-12} watts at 1,000 cycles or, in abbreviated form, as dbRN.

These early experiments and tests to evaluate the interfering effect of noise were made with Western Electric Type 144 handsets and resulted in the weighting curves of Figures 1 or 2 depending on the point of measurement. The line weighting curve of Figure 1 includes the attenuating effect of a typical exchange circuit, subscriber loop, and telephone set. The receiver weighting curve of Figure 2 includes only the attenuation effects of a telephone set. Equalizers using these weighting curves were incorporated into a noise measuring set manufactured by the Western Electric Company. This was originally designated the 2A Noise Measuring Set, and measured transmission impairment in dbRN. A modification of the 2A set was later adopted and designated the 2B set.

Later an improved type of handset (Western Electric Type F1A) came into general use with a type F1 transmitter and HA1 receiver. When tests similar to those used for 144 telephone sets were conducted with this handset, a different set of weighting curves shown in Figures 4 and 5 were obtained and designated F1A and HA1 weighting. Equalizers designed from these curves were incorporated in a modification of the 2B Noise Measuring Set. The tests also indicated that the new handset gave approximately a 5-db improvement over

FIGURE 2. 144 Receiver Weighting



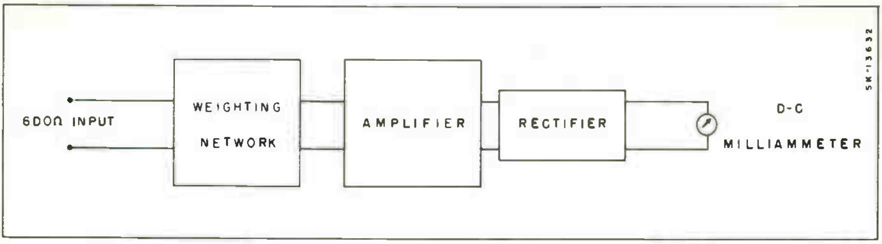
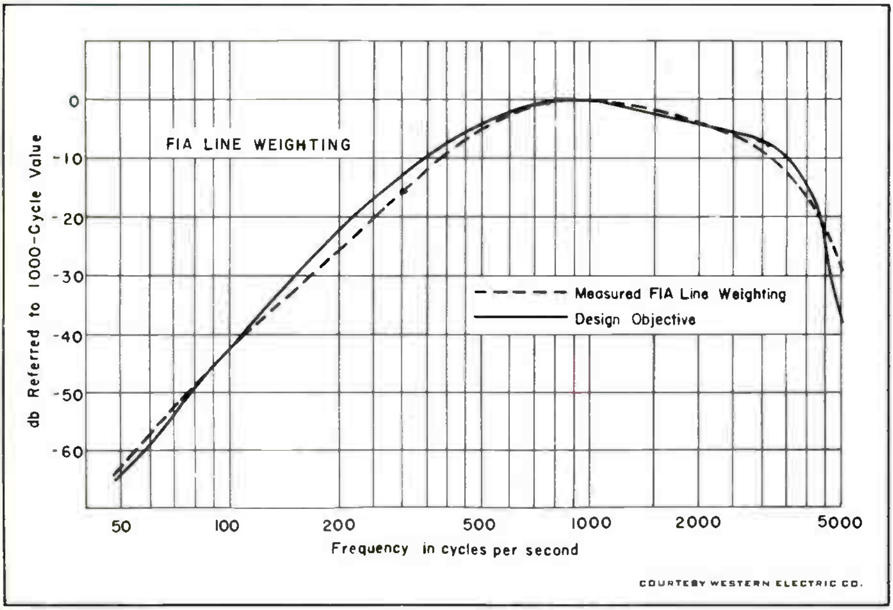


FIGURE 3. Block diagram of a noise meter. The combined characteristics of the weighting network, amplifier, rectifier, and meter mechanism produce a pointer deflection that is proportional to the interferring effect of the input noise.

the electrical and acoustical performance of the 144 handset when using line weighting. Therefore, a different reference noise power of $10^{-11.5}$ watts or -85 dbm was required to give equal noise measuring set readings for equal transmission impairments. The change in reference noise power necessitated a change in the units used to express interfering effect and resulted in the adoption of a new unit called dba. To avoid confusion, all noise measurements made with either 144 weighting or F1A weighting are expressed in dba. When using 144 weighting, dba and dbRN are numerically equal and represent the same amount of noise power and the same interfering effect. However, dba, F1A weighted and dba, 144 weighted represent the same interfering effect but they do not represent equal amounts of noise power. For example, 30 dba measured with F1A line weighting at 1,000 cycles represents a power level of -55 dbm but 30 dba or dbRN measured with 144 line weighting at 1,000 cycles represents a power level of -60 dbm. DbRN is not used with F1A weighting.

Except for certain special cases such as the one given above, there

FIGURE 4. F1A Line Weighting

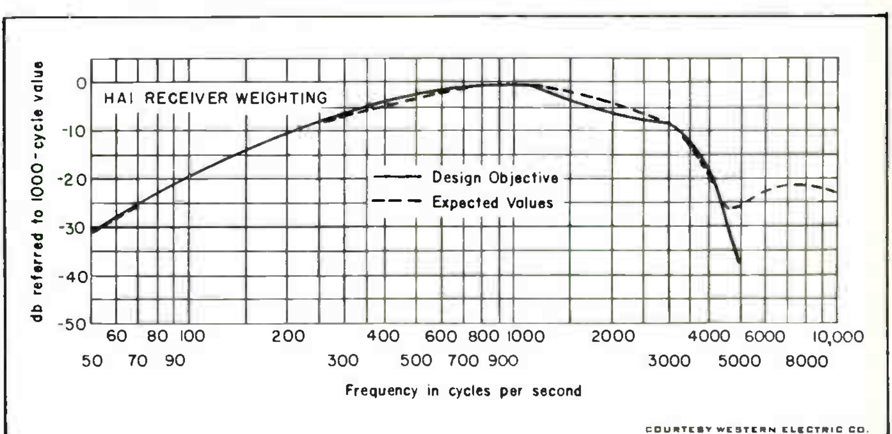


is no simple way of expressing an exact relationship between dba and dbm. If the noise consists of a single tone or an evenly distributed 3-kc band of frequencies, conversion from dbm to dba or vice versa is relatively simple. Otherwise conversion of dbm to dba can be quite complicated. Noise as found in transmission lines and electronic apparatus varies widely as to its component frequencies and their relative amplitudes. It seldom consists of a single frequency. The chief source of noise in open-wire voice frequency circuits is induction from power lines and apparatus. In open-wire carrier and radio systems, noise comes from a variety of sources including power lines, atmospheric disturbance and interference from radio transmitters. While cable circuits are not subject to any great degree of noise from atmospherics or power lines, the low power levels used make them subject to noise caused by the thermal agitation of the atoms and electrons in circuit components. Noise at carrier frequencies is often distributed evenly across a channel bandwidth. Open wire voice frequency circuits, however, more often have noise that is concentrated at certain frequencies. Usually these are odd harmonics of the local power frequency.

Where the noise is known to be evenly distributed across a 3,000 cycle voice frequency band, measurement can be made with a suitable transmission measuring set or voltmeter. Because the weighting network attenuates various frequencies differently one milliwatt of evenly distributed noise (flat noise) produces only 82 dba of interfering effect. This is approximately true for both 144 and F1A weighting. Therefore, measurements of flat noise in dbm can readily be converted to dba by adding 82 to the reading (-50 dbm +82 = +32 dba).

Single interfering frequencies other than 1,000 cycles can also be measured in dbm and converted to dba by use of the weighting curves. For example, if a 300-cycle interfering tone is measured to be -40 dbm, the F1A line weighting curve shows that it is 15 db less interfering than a 1,000 cycle tone of -40 dbm or in other words its interfering effect is the same as that of a 1,000-cycle tone of -55 dbm. A -55 dbm, 1,000-cycle tone is equivalent to 30 dba. Therefore,

FIGURE 5. HA1 Receiver Weighting



the -40 dbm, 300-cycle tone has an interfering effect of 30 dba.

Volume Units

Where programs and certain other types of speech or music are being transmitted, monitoring of the program volume level is necessary to maintain a constant average volume. Failure to maintain the program volume level constant may cause overmodulation of line amplifiers or radio transmitters and cause blasting and distortion from listeners' loudspeakers or handsets. If a simple db meter or voltmeter is bridged across the circuit to monitor the program volume level, the indicating needle will try to follow every fluctuation of program power and will be difficult to read and will have no real meaning. Also, different meters will probably read differently because of differences in their damping and ballistic characteristics.

To provide a standardized system of indicating volume, a special instrument and set of units were created. These instruments are called VU meters or volume indicators and the units are volume units or vu. Ordinarily, volume can be measured in vu only on these special instruments because the volume unit is based on the readings of these instruments under a specified set of conditions. One exception to this rule is the 2B noise measuring set which can be calibrated to read in vu with an error between zero and plus two vu.

The indicating instrument used in vu meters is a d-c milliammeter having a slow response time and damping slightly less than critical. If a steady sine wave is suddenly impressed on the meter, the pointer will move to within 99 percent of its steady state value in 0.3 seconds and overswing the steady state value by 1.0 to 1.5 percent.

A standard volume indicator (meter and associated attenuator) is calibrated to read 0 vu when connected across a 600-ohm circuit carrying 1 milliwatt of sine wave power having a frequency between 35 and 10,000 cycles per second. For complex waves such as speech, a vu meter will read some value between the average and the peak values of the complex wave. There is no simple relation between the volume measured in vu and the power of a complex wave. The actual reading will depend on the particular wave shape. For steady sine waves in the frequency range of the instrument, the reading in vu will be equal to the reading of a db meter in dbm connected across the same circuit.

Other Units

Various units other than db, dbm, dba, dbRN, and vu are in use in the telephone industry or are used in other sections of the communications field. They include dbx, dbw, dbRAP, dbv, and others. Of these, dbx is the most common in the telephone industry.

Dbx is used to indicate crosstalk coupling in telephone circuits. Like dba and dbRN, it is determined or measured with a type 2B noise measuring set. Dbx means decibels above reference coupling. Reference coupling is defined as the coupling necessary to cause a reading of 0 dba on the disturbed circuit when a test tone of 90 dba is impressed on the disturbing circuit. Both values of dba are for the same weighting.

The other units mentioned above are quite simple. Dbw are decibels referred to one watt. Dbk are decibels referred to one kilowatt. Both dbw and dbk are often used to indicate radiated power from radio transmitters. DbRAP means decibels above reference acoustical power which is defined as 10^{-16} watts. Dbv refers to decibels referred to one volt. Other logarithmic units in use are similar to these mentioned.

THE Lenkurt Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. I NO. 10 DECEMBER, 1952

A Discussion of

"LEVELS" AND "POWERS" IN A CARRIER SYSTEM

In this article the concept of "level" is discussed to show the difference between the use of this concept and the use of the defined quantity "power." Some of the reasons for common erroneous usage of the two terms are explained, and examples are given to show where and why both terms are properly used.

A brief discussion of the reasons why certain levels have been established in wire-line carrier equipment is also included.

Two of the most troublesome terms in communications language are "level" and "power." Although they are often used interchangeably in daily conversation, the two terms are not synonymous. A knowledge of the actual meaning and proper usage of each term will help show why both terms are necessary in communications work.

Basically, level is an expression of relative signal strength at various points in a communication circuit. Power, on the other hand, is an expression of absolute signal strength at a specific point in a circuit.

What is meant by "Level"

Generally speaking, the word "level" is used to indicate the relative value of a quantity with relation to an established value of the same quantity. The established value is known as the "zero reference level." This general conception of level has many applications. For example, in the aircraft industry the speed of supersonic aircraft is measured with respect to the speed of sound rather than in terms of distance per unit time. The speed of sound is arbitrarily called Mach 1, and the speed of any aircraft can then be stated as a Mach Number to express that speed with respect to the speed of sound.

In telephone work the term "level" is used in a similar manner to express the relative amount of power at various points in a circuit. Just as the speed of an aircraft is expressed as a multiple of the speed of sound, the amount of power at the output of a telephone repeater can be expressed as one-half, two, or three times the power at the zero reference level.

In practice, relative levels in a telephone circuit are expressed in db (decibels) rather than in arithmetic ratios. This is done because of the convenience of using this logarithmic expression for the relatively large ratios involved. They are sometimes as great as 100 million to one (80 db).

Unless some other reference is stated, the zero reference level for a signal in a telephone circuit is that amount of power which the signal has when measured at the two-wire input to the toll circuit. The concept of level is illustrated in Figure 1.

In Figure 1 the attenuation of each subscriber loop has been arbitrarily set at 6 db for illustrative purposes.

Since the level at the input to the toll switchboard has been defined as zero reference level, the level at the talking subscriber's subset is +6 db.

From the toll switchboard the transmitted speech passes to a carrier terminal which provides a gain of 17 db. The line attenuation between carrier terminals and the repeater is 42 db. Therefore, the level of the received signals at the repeater is –25 db. Since the gain of the repeater is 42 db, the transmitted level from the repeater is +17 db. With 42-db line attenuation between the repeater and the receiving carrier terminal, signals will be received at the terminal at a level of –25 db. The receiving branch of the carrier terminal provides a 19 db gain so the signals will be delivered to the toll switchboard at a level of –6 db. Then, since the loop attenuation is 6 db, the received level at the listener's subset will be –12 db. Thus, the range of speech signals from the talker will be heard by the listening subscriber at a level 18 db below the signal strength leaving the transmitter.

At all level points the strength of the transmitted speech has been clearly stated as having a definite ratio to the strength of the speech at the zero reference level. The statement of level at each point indicates only how much gain or loss the transmitted signals have received between the various points along the transmission path.

Level, therefore, is purely a relative term. Whenever level is expressed, the zero reference level is understood to be at the point where the circuit being considered becomes a toll circuit.

What is meant by "Power"

While level is always a ratio, "power" always designates a definite quantity. This quantity is defined in electrical terms as the rate at which electric energy is taken from or fed to a device. The most common unit for expressing power is the "watt."

In addition to the watt, a number of other defined units are commonly used for expressing the amount of power in telephone equipment. Among these are the *dbm*, the *dba* and the *vu*. Each of these units is based upon using the decibel to express the amount of power above or below a convenient amount of reference power.

Because of the use of the decibel and of a reference power in defining these units, powers expressed in them are sometimes erroneously called levels. They are not—because in every case a value stated in dbm, dba or vu can be readily converted to a value in watts.

The difference between power and level can be shown more clearly by considering the use of "dbm." This unit is perhaps the most common of the three mentioned. Stating that the power at a certain point is $\pm$X dbm simply means that the power is X db greater or less than one milliwatt.

A 1000-cps test tone with a power of one milliwatt is ordinarily available at toll switchboards. When this test tone is transmitted over a telephone circuit the test-tone power in dbm at any point in the circuit is numerically equal to the level in db at that point. It is this similarity which can cause confusion between proper usage of level and power.

The distinction between level and power can also be illustrated by considering the two terms with respect to a fixed gain amplifier as shown in Figure 2. Two conditions are shown. In the first, the input to the amplifier is .001 watts. In the second the input to the amplifier is .0005 watts. In both conditions the amplifier has a fixed gain of 30 db.

In this example the input in both cases is arbitrarily considered to be zero level. Therefore, the output level in both cases is +30 db and it cannot change unless the gain of the amplifier changes.

The power input and the power output change in both cases, however. In the first, the input signal of .001 watts is amplified 1000 times to produce an output of 1 watt. In the second, the input signal of .0005 watts is amplified

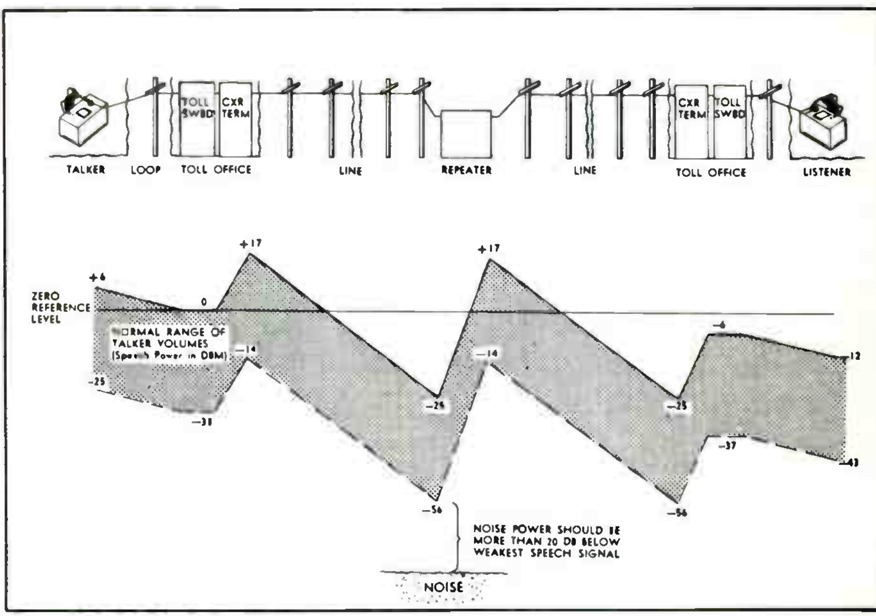


Fig. 1. A level diagram showing the relative strength of transmitted signals at various points in a typical toll circuit. Normal range of talker volumes is indicated, and relation of noise to signal strength is shown.

by the same amount since the gain of the amplifier is fixed at 30 db (or 1000). The output power is therefore .5 watts.

It is obvious that the power output of a fixed-gain amplifier will change when the input power changes. *But the level remains the same* so long as the gain or loss (in decibels) between zero or reference level and the output of the amplifier remains the same.

Both level and power are useful terms. Each is best suited for specific purposes in telephone circuit engineering and operation, and they serve to supplement each other.

Where and Why "Level" is Used

Transmitted speech consists of a large range of frequencies and powers which vary widely for different speakers. For this reason it is impossible to determine exactly what power will exist at any point in a circuit when the circuit is in use. However, regardless of the specific power at any point, the level, or in other words, loss or gain between the point in question and other points in the circuit, can be determined either by calculation or by measurements with a transmitted test tone.

When laying out telephone circuits it is necessary to know the net loss which the circuit will give to speech currents passing through it. It is neither necessary nor practical in this type of planning to know exactly what the actual power will be at any point, particularly since the power will vary over wide limits depending up the talker and the words spoken. The normal range of speech volumes transmitted over a telephone circuit is shown in Figure 1.

Since the gain or loss of a circuit is independent of power (within the power-handling capacity of the equipment) it is convenient to have the concept of level to express the relative strength of a signal at any point and to determine net loss of the circuit between any two points.

Where and Why "Power" is Used

Although level has definite value in circuit planning, it is necessary to consider actual power involved when designing and operating the electronic equipment used at voice and carrier frequency terminals and repeaters.

Operation of electronic equipment depends upon the minimum and maximum powers which can be supplied to the input of the equipment and delivered from the output. Equipment sensitivity and coincident power strength of noise and other disturbances usually determine the lowest practical input power. Maximum output power (and consequently the maximum input) depends upon the power-handling capacity of the equipment.

Specifications for carrier equipment normally give the test-tone power at the inputs and outputs of each channel. In some cases it is desirable or necessary to know the total peak power that may be delivered to common equipment or to the line by several channels.

Per-channel power is normally stated in dbm. Because dbm is a logarithmic value, two powers expressed in dbm cannot be added to obtain the total power. Instead, each channel power

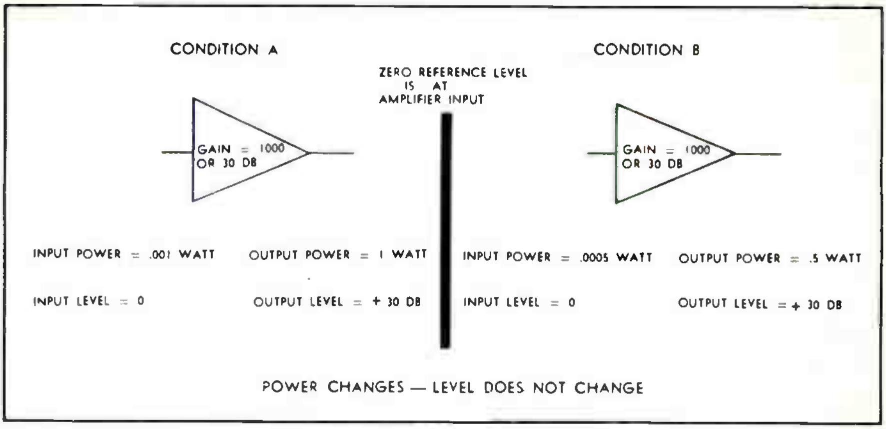


Fig. 2. The difference between level and power is shown by their use with a fixed gain amplifier.

must be converted to watts, added, and the total then reconverted to dbm.

If the per-channel power of all channels is the same, doubling the number of channels increases the total power by 3 db. Thus, if a system has 8 channels, each with a signal output power of +10 dbm, the total power delivered to the line is +19 dbm.

In speech-communication circuits it is unlikely that all channels of a carrier system will be transmitting signals of the maximum value simultaneously. Therefore, the common equipment is usually designed to handle the total expected power rather than the total possible power.

Levels and Powers in a Carrier System

Although levels are more important than actual power to the engineer laying out a telephone circuit, the transmission engineer interested in the installation or operation of a carrier system must usually know the actual power at the various points. Otherwise, there is a possibility of operating a circuit with either less input power or more output power than the equipment is designed to handle.

When test tone power at the point of zero reference level is one milliwatt, the power at any point in a circuit as measured in dbm is numerically equal to the level as expressed in db. When the test tone power is any other amount or when power is measured in any unit other than dbm, the numerical value of level is not the same as that of the power at that point.

Any consideration of power in a carrier system can be divided into two sections—the amount of power at the connections to the carrier equipment, and the amount of power at various points inside the carrier equipment. Power values inside the carrier equipment are of interest primarily to the design engineer.

The amounts of power at the line and drop terminations of the transmitting and receiving branches are of more importance to the operator of carrier equipment. These power values determine where carrier systems may be operated, how repeaters must be spaced, and how coordination may be achieved.

A typical two-wire carrier circuit with a four-wire termination at one drop and a two-wire hybrid termination at the other is shown in Figure 3.

Power at the V-F Input and Output

The v-f power at the input to the transmitting branch of a carrier system is primarily based on the normal amounts of power delivered to the line from the toll switchboard. Because many telephone offices are arranged for patching circuits on a four-wire basis at a –16 level, and the usual test tone power at the transmitting toll switchboard is 0 dbm, the input stages of carrier systems are often adjusted to receive a test tone power of –16 dbm on a four-wire basis. This really means that the input to the carrier system is at a –16 level or at a circuit point 16 db removed from the two-wire v-f level at the transmitting toll testboard.

The amount of v-f power obtained from the receiving branch of a carrier system is also determined primarily by switching requirements. If all of the other values indicated in Figure 3 are within proper limits, the v-f output power for each channel with 0 dbm test tone at zero level would normally be about +7 dbm on a four-wire basis.

Power Delivered To and Received From The Line

A number of factors influence the amount of power which can be transmitted or must be received from the line. Since the difference between these two values is the maximum span attenuation, these factors also influence repeater spacing for a carrier system. Among these factors are the noise level of the line, the line attenuation, the frequencies at which the system operates, the characteristics of the directional filters, and such things as line configuration which influence the amount of crosstalk encountered.

Basically, the amount of power transmitted must be high enough so that sufficient power will reach the receiving terminal to permit recovery of the transmitted intelligence unimpaired by excessive noise. The power received must be sufficient so that the proper v-f output power can be delivered with the receiving branch gain available, and so that the received power will be sufficiently higher than the line noise to maintain proper signal-to-noise ratio.

Transmitted and Received Power

The amounts of power now commonly transmitted have been established as a result of attenuation studies conducted during many years of experience with telephone lines used for carrier circuits. These studies have provided engineers with information concerning line characteristics and their effect on carrier systems under a variety of conditions.

Among the factors which determine the minimum amount of power which should be received from the line at a carrier terminal are the receiving branch gain and the noise level of the line.

The ultimate objective of a carrier circuit is to deliver a certain amount of power to the drop. Therefore, the

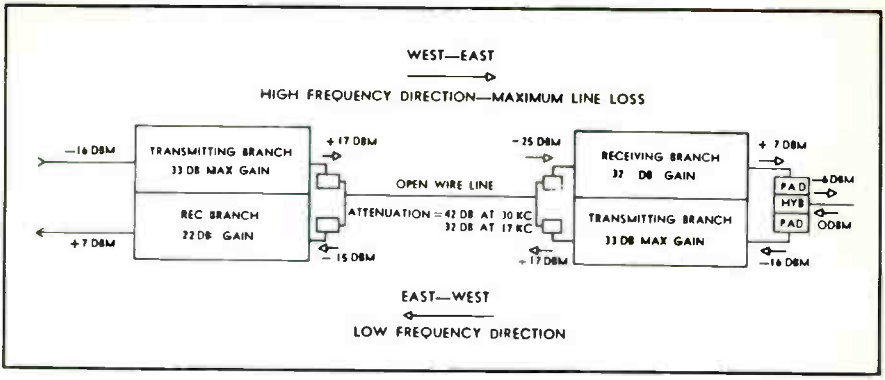


Fig. 3. Simplified block diagram of a two-wire carrier system illustrating typical test tone power values at the connections to the carrier equipment.

minimum received power must be such that, after being amplified an adequate amount, the signal will have the proper amount of power at the drop.

Higher receiving branch gain will not necessarily permit lower minimum amounts of power to be received since the received signal must be sufficiently greater than the noise level of the line to maintain the desired signal to noise ratio. Since noise is amplified by the same amount as the desired intelligence, the signal-to-noise ratio at the output of the receiving branch cannot be any better than it is at the input. This is illustrated in Figure 1.

Loop Gain and Level Coordination

Loop gain is defined as the sum of the gains which are given to a signal of a particular frequency in passing around a closed loop. The loop can be a carrier terminal, a repeater, or a complete carrier circuit.

Excessively high gains in the transmitting or receiving branches of a carrier system terminal or repeater can cause "singing." This occurs if the gain around the system, terminal, or repeater loop for any frequency is greater than the losses around the same loop for that frequency.

Loop gain is affected by a number of complex factors. Among them are the suppression by directional filters, the losses due to hybrid balance, and the effect of the other frequency-selective elements in a carrier system. All of these factors are considered by design engineers when they determine operating levels and the amounts of power which will be transmitted and received by a carrier system when operating under various conditions.

A further limiting factor which must be considered when determining the amounts of power which will be transmitted or received by a carrier system is coordination of the levels and powers between two or more systems operating at the same frequencies on the same lead. If all systems transmit the same amount of power they are not subjected to power differentials along the line. Any crosstalk between adjacent line conductors is then not further increased by a difference in power.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 7 NO. 1 JANUARY 1958

MEASURING VOLTAGE AND POWER

At Carrier Frequencies

Good measuring instruments and proper technique are important ingredients of any carrier maintenance work. They are of particular importance when measuring voltage or power level (db) at carrier frequencies.

This article discusses the theory of carrier frequency measurements, describes instrument requirements, and outlines some of the proper techniques to use. It also points out some of the most common errors committed.

Quality carrier equipment for long distance telephone and telegraph transmission must be as precise as a fine time-piece. Much thought and care goes into its design and manufacture. Carefully calibrated instruments are used in the factory to check over-all terminal performance and the frequency response, gain, or attenuation of its subassemblies.

Like an accurate clock, quality carrier equipment is made with high-quality components and is largely self-adjusting. Temperature-compensating circuits, negative-feedback amplifiers, signal-level regulators and other mechanical and electrical features simplify and reduce the amount of maintenance required in service. However, they do not

Fig. 1. Good instruments are essential to accurate, reliable measurements.

eliminate it entirely. To ensure proper performance in use, periodic checks must be made of such characteristics as system and channel power levels, channel frequency response, carrier leak, etc.

Of the measurements required to check system and channel performance, none are more important than those of signal voltage and power level. In contrast to direct-current or fixed power-frequency measurements, carrier frequency measurements involve a wide range of frequencies, voltages and impedances. For example, in Lenkurt 45-class carrier systems, power level measurements in a message channel may be at frequencies as low as 200 cycles per second or as high as 1196 kilocycles per second. Sometimes a single frequency is involved, but often the quantity being measured is the voltage or power level of a band of frequencies.

Moreover, such measurements may cover a voltage range from less than one-thousandth of a volt to over 100 volts, and a power range from about 25 millionths of a watt to nearly one watt. The impedance across which voltage is being measured or in which power is dissipated may range from a few ohms to several thousand ohms.

Despite the wide frequency, voltage, power, and impedance ranges encountered in carrier systems, measurements can be made simply and accurately if the proper instruments and techniques are used.

Voltage and Power

The correct choice and use of instruments for carrier frequency measurements often require a knowledge of the relationship between voltage and power and the definitions of alternating voltage.

The relationship between voltage and power is one of the basic concepts of

Fig. 2. Proper measurement technique helps prevent unexplainable errors.

electricity. It is expressed mathematically as:

$$\text{Power} = \frac{V^2}{R} \text{ or } \frac{V \times V}{R}$$

where V is voltage across a resistance R. In direct-current circuits, this relationship is always correct. But in alternating current circuits, it holds true only if the proper value is assigned to the alternating voltage.

Any alternating voltage wave has three principal values:

(1) The maximum instantaneous voltage or peak voltage

(2) The average voltage

(3) The effective voltage

Definitions of the peak and average values of an alternating wave are self-evident. The meaning of effective voltage, however, requires some explaining. Effective voltage is the numerical value that can be assigned to an alternating wave making it equal in power-producing ability to the same value of direct voltage. It is sometimes called *rms* voltage because mathematically it is the square root of the mean-squared instantaneous voltage (root-mean-squared).

The importance of effective voltage can be demonstrated by the following example. If a direct voltage of 10 volts is applied to a resistance of 10 ohms, 10 watts of power will be dissipated as heat. To create the same amount of power (average power per cycle) in the same resistance with an alternating voltage there must be an *effective* value of 10 volts, regardless of the peak or average values. Thus, when voltages are measured as a method of indicating actual or relative amounts of power, the instrument used should be calibrated to indicate effective voltage rather than peak or average voltage.

Although not directly related to power, peak and average values of alternating voltage are important in many carrier circuits. For example, the direct bias voltage on the grid of a tube is usually determined by peak signal voltage. On the other hand, the direct voltage component of a full-wave rectifier output voltage is the average value of the rectified wave.

For any specified wave shape there are definite, fixed ratios between peak, average, and effective values. A sine wave, for example, has a peak value that is 1.414 times the effective value. Its average value per cycle is zero. However, when rectified by a full-wave rectifier the average value is 0.637 times the peak value.

If the wave form is rectangular the average, peak and effective values per half-cycle are all equal. For other irregular wave forms, the ratios between the different values will vary widely.

Usually when writing or speaking of alternating voltages, expressions such as 1 volt, 110 volts, etc. mean effective volts.

Power Level

Although voltage and power are useful expressions in themselves, they are frequently not as useful in communications as *power level*. Expressed in decibels, power level is a comparison between two amounts of power. Some measurements in carrier maintenance involve comparisons between the signal power at some point in the circuit and the power at some other reference point. Others compare signal power and some fixed amount of power.

For many purposes, the exact amounts of power are not important. Only the ratio is important. For example, the gain of a linear amplifier or the attenuation of a pad can be measured without knowing either the input or output power of the device. This information is customarily expressed as a gain or loss of so many db.

When measurements involve comparison of measured power with a fixed reference power, the reference power customarily stated is one milliwatt. The measured power level is expressed as decibels above or below one milliwatt; or in abbreviated form as ± dbm.

Decibels are convenient units because they eliminate the drudgery of working with large numbers and change power multiplication (amplification) and division (attenuation) into simple addition and subtraction. Mathematically, decibels are ten times the logarithm of a power ratio. For example, a power ratio of unity is zero decibels; a power ratio of 2 is 3 db, a power ratio of 10 is 10 db, and a power ratio of 100 is 20 db.

If one of the two powers is a fixed reference of one milliwatt, then 1 milliwatt of measured power is 0 dbm, 2 milliwatts is 3 dbm, 10 milliwatts is 10 dbm, 100 milliwatts is 20 dbm and and 1,000 milliwatts (1 watt) is 30 dbm, and so on. Negative values of dbm indicate power quantities of less than 1 milliwatt. For example, –30 dbm is 1 microwatt.

The measurement of db or dbm is basically the same as the measurement of effective voltage. The voltage scale of a voltmeter can readily be calibrated in dbm by noting that 1 volt across 1,000 ohms is one milliwatt or 0 dbm. Thus the one-volt mark on the scale could be 0 db. Two volts would be 6 db; 4 volts, 12 db, 8 volts, 18 db; etc.

In telephone work, 1,000 ohms is not a common circuit resistance, but 600 ohms is. So, in most practical meters used for telephone measurements, 0 dbm corresponds to 0.775 volts.

Instruments

Carrier frequency voltmeters and power level meters (db meters) are of various types. Some are general purpose meters designed for a variety of uses. Others are designed for special types of measurement. Among the different types, there are considerable variations in operation, scales and accuracy.

Most meters operate such that the distance the needle is deflected is proportional to the average voltage of the measured wave. In others, needle deflection is proportional to the effective or peak voltage.

Peak and average voltage measuring instruments usually have scales graduated in effective volts and are accurate when the voltage measured has a pure

Fig. 3. Hewlett-Packard type 400D vacuum tube voltmeter. This meter has a flat frequency response from 10 cps to 4 mc. Input impedance is in excess of 10 megohms so that when connected across 500,000 ohms or less, there is negligible effect on the circuit being tested.

Fig. 4. Sierra type 121A wave analyzer (frequency selective VTVM). This meter measures only at the frequency to which it is tuned.

sine waveshape. If the wave form is complex—such as the waves of speech, or even simple combinations of two or more tones—a peak reading or average reading voltmeter will indicate differently than an *rms* instrument.

Although true *rms* indicating meters are the most likely to be accurate, the errors that may occur from using an average indicating meter are never great. For voltage, they can range from 11% too high to about 25% too low (+1 to -2 db). High readings result when the waveshape has a flat top such as a square wave. Low readings result when the waveshape is sharply peaked; for example, narrow pulses. Flat or gaussian noise measurements made with an average indicating voltmeter will be about 11.3% low (about 1db).

Average reading meters are generally more popular than true *rms* meters because they are rugged, economical and less subject to damage from overload. The small error that results when measuring complex waves is seldom objectionable and often unimportant.

Practical Meters

Maintenance instructions for Lenkurt carrier equipment generally recommend two types of voltmeter (or db meter) for voice and carrier frequency measurements. These are the Hewlett-Packard Type 400D wide-band meter shown in Fig. 3 and the Sierra Model 121A frequency selective meter shown in Fig. 4. Also suitable are the Fisher Research Laboratories Type VT-3 and the Alto Scientific Type 21A transistorized meters. Other manufacturers also make meters with similar capabilities.

The Hewlett-Packard 400D meter is an average reading instrument with a usable voltage range from 0.0001 to 300 volts or in dbm from -80 to +50 dbm. Frequency response is nearly constant from 10 cps to 4 megacycles per second. The scale is calibrated so that 0.775 volts in 600 ohms corresponds to 0 dbm.

Wide-band meters have one basic limitation: inability to measure directly the power levels of the different frequency components of a complex wave. Often in carrier line-up or maintenance, there is a need to select a single frequency from a band of frequencies and measure its power level. The best type of instrument for this kind of measurement is a frequency selective voltmeter (sometimes called wave analyzer) such as the Sierra model 121A.

The Sierra 121A meter is tunable like a radio receiver over a range of 15 kc to 500 kc and has a voltage range of about 0.00025 to 125 volts (–70 to +42 dbm). Input impedance is about 10,000 ohms in the passband.

Since the Sierra meter usually measures the power level of only one tone at a time, its indications are always approximately proportional to *rms* voltage.

A frequency selective meter has a variety of uses in carrier maintenance. Typical uses are setting channel transmitting level without disabling a complete channel bank and measuring carrier leak, signaling tones, regulating tones, etc. while a system is in service.

An example of frequency selective meter use might be as follows.

The channel unit of channel 9 at the west terminal of a 45A carrier system (NA allocation) has been replaced. A readjustment of channel transmitting level is required without disturbing the operation of the other channels. To make the adjustment, a 0 dbm 1,000 cycle test tone is applied to the voice-frequency two-wire input terminals of channel 9. A frequency selective voltmeter will then read the transmitting level of channel 9 if it is connected across the terminal output and tuned to 53 kc. Pads in the transmitting branch of the channel unit can then be adjusted to make the channel output power level meet requirements.

Measuring Techniques

The accuracy of voltage and decibel measurements is often as dependent on measuring technique as it is on the quality of instruments used. Factors which affect technique are: (1) the purpose of the measurement, (2) circuit impedance, (3) circuit balance, and (4) the nature of the measured signal.

Generally, measurements are made for one of three purposes: (1) test the performance of a circuit in service, (2) locate trouble, or (3) align the circuit to meet specifications. In-service and trouble-shooting measurements are usually made on a *bridging* basis. System alignment often involves power level measurements on a *terminated* basis.

In bridging measurements, the meter is connected directly across a functioning circuit. For example, in Fig. 5a the meter is connected across the terminals of a subscriber loop and telephone set to measure power level on a bridging basis. The accuracy of this measurement depends upon the impedance characteristic of the subscriber loop and telephone set.

In terminated measurements a fixed resistor replaces the load normally connected to the circuit. The resistor may be internal or external to the meter. In the Hewlett-Packard and Sierra meters, it is external. Instruments with the resistor built in are often called transmission measuring sets. In Fig. 5b, the telephone set has been replaced by a meter with a 600 ohm resistor across its input terminals. The meter is measuring the power (in dbm) dissipated in the 600-ohm resistor. The resistor, rather than the subscriber loop and telephone, terminates the circuit and accuracy is assured.

Both bridging and terminated measurements have their place in carrier maintenance. Bridging measurements are usually made where disabling the circuit is inconvenient, where the circuit impedance is known to be a definite fixed value, or where only voltage is to be measured.

For example, the carrier voltage applied to a modulator would normally be measured by bridging a voltmeter across the carrier frequency input to the modulator.

Usually, important internal circuit points are brought out to test receptacles to permit easy bridging measurements.

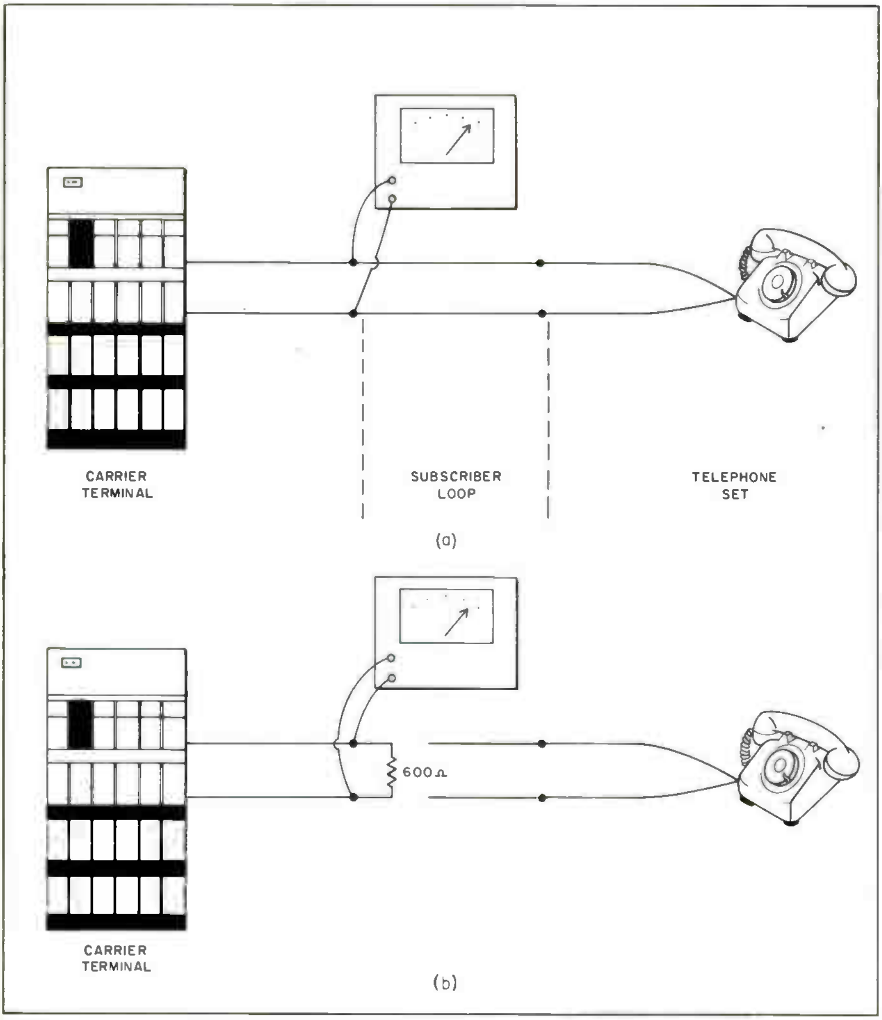


Fig. 5. Bridging and terminated techniques of measurement. (a) The bridging technique is accurate if the resistance of the terminating circuit is known. (b) Terminating the circuit with a fixed resistor assures accurate decibel measurements.

Input and output circuits are usually connected to jacks to simplify terminated measurements and monitoring.

For terminated measurements in system alignment, a resistor is connected to replace the voice-frequency drop or carrier-frequency line. A standard test tone is applied to the input terminal (either carrier-frequency or voice frequency input, depending on the direction of transmission) and the meter reads the output power level in a standard resistance. In transmission measurements, a tone applied at the other end of the line is dissipated in the terminating resistor. The difference between input power level of the tone and the power level measured in the terminating resistor is the line loss. The advantage of the terminated measurement is that the impedance of the circuit is definitely fixed. This results in greater accuracy.

Circuit Impedance

When making voltage or power level measurements, particular attention must be paid to circuit impedances. If the impedance at the point of measurement is 600-ohms resistive, the Hewlett-Packard 400D and the Sierra 121 meters will read power levels directly in dbm. A special transformer adaptor is also available that makes these meters read correctly in dbm when they are connected across a 130-ohm circuit.

If the circuit impedance is a known value other than 600 or 130 ohms, power level in dbm can be computed from measured voltage and circuit im-

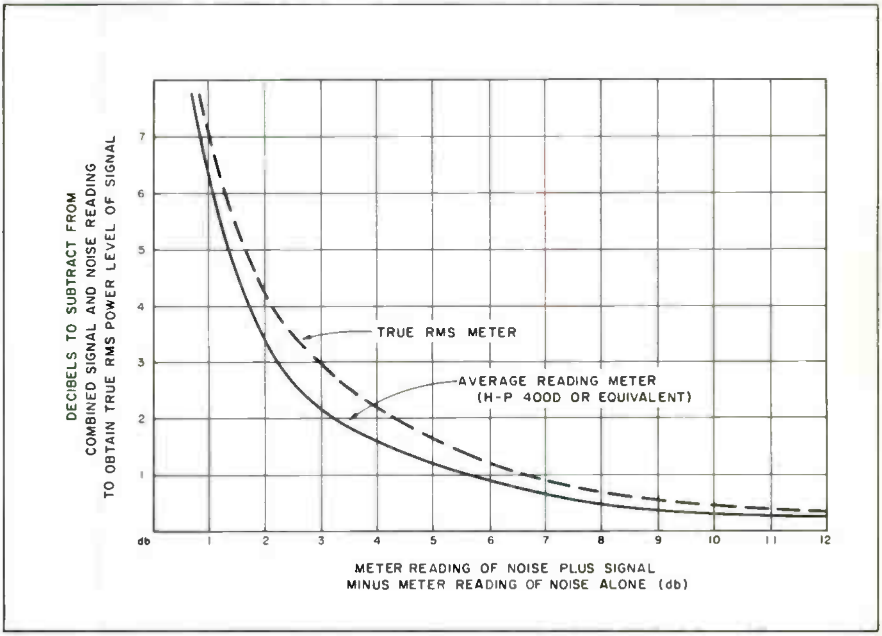


Fig. 6. Graph for separating signal from noise when using either a true rms meter or an average reading meter.

pedance. A nomograph that simplifies computations of power level from voltage measurements appeared in the November 1953, March 1954 issues of the Demodulator and in the Demodulator Carrier and Microwave Dictionary.

Circuit Balance

A potential source of error in carrier frequency measurements is the use of an unbalanced meter to measure voltage or power level in a balanced circuit. The receiving branch of a carrier channel is normally a balanced circuit. So is the carrier-frequency output circuit of wire line and cable carrier systems.

If an attempt is made to bridge or terminate such balanced circuits with a meter with unbalanced input terminals (one terminal connected directly to the meter chassis), stray longitudinal voltages on the wiring of the circuit may cause an erroneous reading. Also, cross-talk between the circuit under test and other circuits may be increased while the meter is connected. Unbalanced meter inputs can be readily converted to balanced input by connecting a suitable transformer between the meter and the circuit being tested. Such transformers are available as accessories to the Hewlett-Packard and Sierra voltmeters.

Signal Level and Noise

Another source of error in voltage or power level measurement is the possible presence of noise. When measuring small voltages or low power levels, noise voltages in the circuit may make the meter read too high. Therefore when measuring test tone voltages of less than 0.01 volts or power levels of less than −20 dbm, circuit voltage should be measured with and without the test tone to determine the amount of noise.

If the noise is 15 or more decibels below the test tone, it will not appreciably affect the accuracy of the measurement. If it is less than 15 decibels below the test tone, the meter reading can be corrected by the use of the graph shown in Fig. 6.

Some Useful Techniques

The best guarantee of accurate and reliable measurements is the use of good quality instruments according to the recommendations of their manufacturers. But since manufacturers' instruction books have a way of disappearing when needed most, here are some techniques that will be generally useful in making measurements at carrier frequencies:

(1) Never compare decibel readings unless both are made across the same value of resistance or a suitable correction factor is used to compensate for the difference in resistance.

(2) When measuring power level in dbm, be certain of the circuit resistance. Where possible, make terminated measurements.

(3) In bridging measurements, if meter input resistance is not at least 10 times circuit resistance, the reading will be in error.

(4) If the circuit to be measured is unbalanced (one side grounded) always connect the meter ground terminal to the ground side of the circuit.

(5) It is best to set the meter range switch so the needle deflects to the upper half of the scale. Most meters are more accurate when reading close to full scale.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 5 NO. 5 MAY, 1956

FREQUENCY MEASUREMENTS

in Communication Systems

An important factor in the development of modern carrier telephony has been the ability to generate electrical waves of precise frequencies. This ability has evolved from the further ability to measure electrical frequencies with great precision.

In this article, some of the methods currently used for the measurement of electrical frequencies are discussed and some of the more commonly used frequency measuring devices are described.

Frequency is the rate at which a cyclic event occurs. Its measurement, like the measurement of all quantities, ultimately involves a comparison with a fundamental unit. For frequency, this unit is an interval of time determined by the rotation of the earth on its axis. Hence the basic unit of measurement is one revolution of the earth per day.

Since a day—or even an hour or a minute—is too awkward an interval to use as a time base, one second is universally accepted as the fundamental unit of time for electrical measurements. Thus electrical frequency is normally expressed in cycles per second, kilocycles per second, or megacycles per second.

For purposes of discussion, the various methods of measuring frequency may be separated into five general categories: (1) comparison methods, (2) tuned circuit methods, (3) balanced bridge methods, (4) wave-length measuring methods, and (5) pulse counting methods. Each of these, in turn, encompasses a variety of different approaches to the same general method. Regardless of the approach or the method, however, every measurement of frequency involves either directly or indirectly a comparison with a standard interval of time.

Frequency Standards

Any stable oscillator whose frequency has been accurately determined may be used as a reference to measure other frequencies and is known as a frequency standard. Two important classes of these are primary and secondary standards. A primary standard is one whose frequency has been determined directly in terms of time. A secondary standard is one whose frequency has been determined by comparing directly with a primary standard.

A typical primary standard might consist of a high-quality, crystal-controlled oscillator which drives a very precise clock. A time interval as measured by the clock is compared with the same time interval as determined by astronomical observations. The number of seconds measured by the clock, multiplied by the frequency at which the clock is designed to operate, gives the total number of cycles which occurred in the astronomically determined interval of time. Thus the frequency of the primary standard is determined directly in terms of astronomical time.

A secondary standard is one whose frequency has been determined by comparing directly with a primary standard and thus indirectly with the frequency of the earth's rotation. Like primary standards, secondary standards are usually high-quality, crystal-controlled oscillators. However, some standards for the audio range use tuning fork control.

With suitable equipment, any person or organization can maintain a primary or secondary standard. The U.S. Naval Observatory transmits extremely precise time signals daily on several different radio frequencies for the purpose of calibrating primary frequency standards. Among the best known and most accurate primary standards are those maintained by the National Bureau of Standards of the United States Government.

The Bureau of Standards operates two radio stations, WWV in Washington and WWVH in the Hawaiian Islands, which transmit signals for measuring and calibrating the frequencies of secondary standards. The signals from WWV and WWVH are derived from the Bureau's very precise

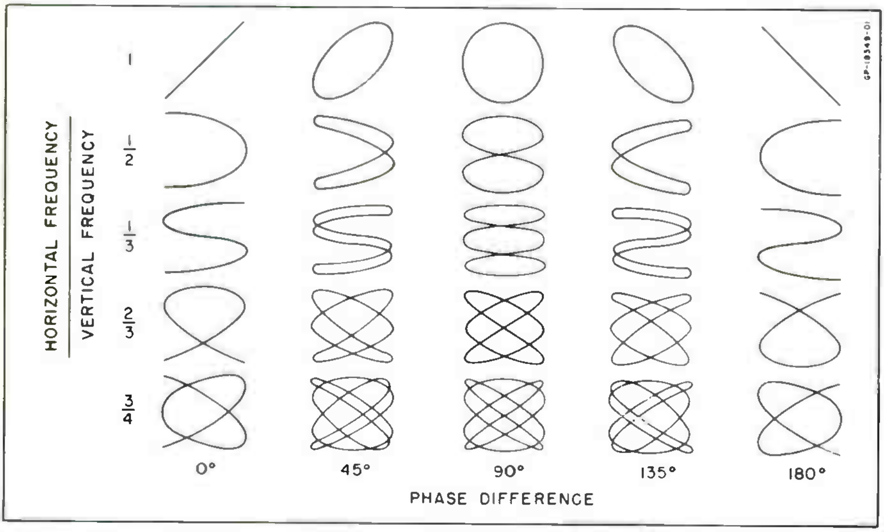


FIG. 1. Typical Lissajous figures for various frequency ratios and phase differences.

primary standards. They are transmitted daily in accordance with established schedules and provide a very convenient and highly accurate standard for measuring frequency. The average error of WWV is about one part in 100 million.

For many measurements, the high degree of accuracy provided by WWV and WWVH is not required and is, in fact, beyond the measuring capabilities of most instruments available. For such measurements, the carrier frequencies transmitted by commercial broadcast stations may often be used as standards. Broadcast stations in the United States are required by law to maintain their assigned frequencies within ± 20 cps and, as a matter of practice, usually maintain them much closer than that. Where such accuracy is tolerable, these frequencies furnish a standard of comparison that is easily accessible and almost continuously available.

To make possible the precise frequency measurements necessary in the manufacture of carrier equipment, the Lenkurt factory at San Carlos maintains both a primary and a secondary frequency standard. The primary standard is a crystal-controlled oscillator whose 100-kc output is divided down to 1 kc to drive a clock. The clock has facilities for comparing its time directly with time signals from the U.S. Naval Observatory.

The secondary standard is also a 100-kc crystal-controlled oscillator calibrated by comparison with the primary standard. The output frequency of the secondary is then divided down to provide frequencies of 10 kc, 1 kc, and 0.1 kc in addition to its fundamental frequency of 100 kc. These precise frequencies are "piped" throughout the factory and engineering laboratories to provide highly accurate standards of comparison for measuring the frequency characteristics of the various oscillators, filters, and amplifiers used in Lenkurt carrier equipment.

Measurement by Direct Comparison

Since frequency standards are so readily available, one of the most obvious methods of measuring frequency is the direct comparison of the unknown frequency with a known standard frequency. One method in common use is comparison by means of a cathode ray oscilloscope.

In this method, a voltage at the unknown frequency is applied to one set of deflection plates of the oscilloscope while a voltage at the known frequency is applied at the same time to the other set of deflection plates. When the relationship of the known to unknown frequency is a ratio of whole numbers, a stationary pattern called a Lissajous figure will be formed on the oscilloscope screen.

The configuration of the pattern will be determined by the ratio of the two frequencies and their phase relationship to each other. For the simpler ratios, the figures can be easily identified. When the ratio of the known and unknown frequencies is a fraction with large whole numbers in the numerator or denominator or both, the Lissajous figures are very complicated and difficult to identify. More involved methods are then necessary. Some typical Lissajous figures are shown in Fig. 1.

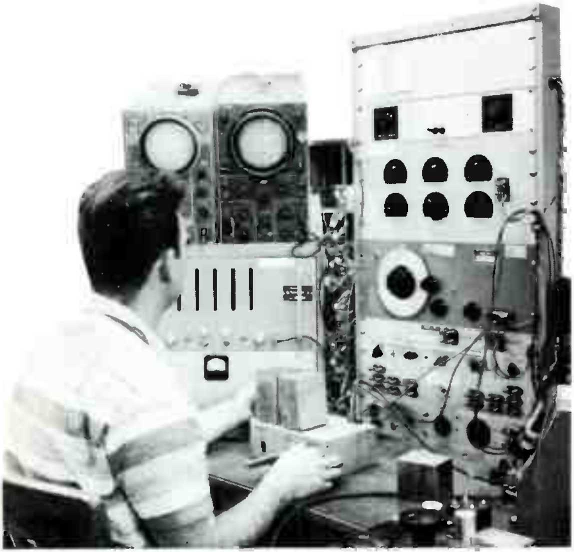


FIG 2. A test bench in the Lenkurt factory where frequency is measured by oscilloscopic comparison.

The measurement of frequency by means of oscilloscopic comparison has its most useful application when the unknown frequency and a known frequency are simply related to each other. Lenkurt production departments use large numbers of oscilloscopes for frequency measurement. Figure 2 shows frequency being measured by oscillographic comparison. The known frequency is derived from the precise Lenkurt frequency standard.

Another common form of comparison makes use of the *beat frequency* which is produced when two different frequencies are mixed. The beat frequency is the difference between the two original frequencies. An unknown frequency may be measured by mixing it with a known frequency and varying the known frequency until the beat frequency is zero. The unknown frequency is then equal to the known frequency.

One application of the beat frequency technique is used in heterodyne frequency meters. A heterodyne frequency meter is an instrument which produces a beat frequency by the nonlinear mixing (or heterodyning) of two frequencies, one of which is of a known value. The basic instrument consists of a stable local oscillator, a mixing device, and a monitoring device. The output of the oscillator is mixed with the unknown frequency and the beat frequency is monitored by headphones, a meter, or other device. The local oscillator is variable and usually has its tuning control calibrated in terms of frequency. In operation, the variable oscillator is tuned until a zero beat is obtained.

The accuracy of a heterodyne frequency meter is dependent on the ability of the user to detect the zero beat point, the precision of design of the oscillator and associated circuitry, the aging of circuit elements, and the accuracy to which the instrument can be calibrated. In general, it is possible to obtain an accuracy in the vicinity of a few parts in a million, stable over short periods of time, with frequency meters of this type.

Heterodyne frequency meters are often used in the measurement of the frequencies associated with microwave systems. Figure 3 shows a type that is used for the lineup and maintenance of Lenkurt Type 72 microwave radio equipment.

Tuned Frequency Methods

Among the simpler methods of frequency measurement are those using the principle of electrical resonance in tuned circuits which are series or parallel combinations of capacitance and inductance. By holding the inductance at a fixed value and varying the capacitor (or vice versa), the resonant frequency of a tuned circuit can be varied over a relatively wide frequency range.

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FIG. 3. Gertsch heterodyne frequency meter. This instrument has a conservative range of 20 to 640 megacycles and can be used for some purposes for frequencies as high as 1,000 megacycles.

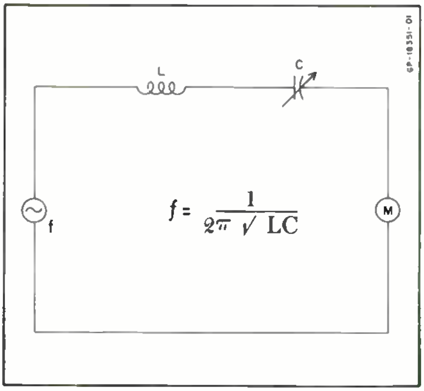

FIG. 4. Simple circuit to illustrate basic principle of a series wavemeter. Variable capacitor, C, is adjusted until meter, M, reads maximum current.

The variable capacitor (or inductor) can be calibrated to read directly in terms of frequency.

One of the simplest of tuned circuit frequency measuring devices is called a wavemeter. Wavemeters which read directly in terms of frequency may employ either a series or parallel circuit. A sensitive current-reading instrument is usually incorporated in the circuit as an indicating device. The source of the unknown frequency is connected to the wavemeter and the circuit is tuned through resonance as indicated by a maximum current reading for a series circuit or a minimum current reading for a parallel circuit. A rudimentary series wavemeter is shown in Fig. 4.

Many different types of measuring devices use the tuned frequency principle. Common among them are grid-dip meters, Q meters, and sharply tunable radio-like instruments, such as frequency selective voltmeters. In fact, a well designed radio receiver makes a frequency measuring device suitable for many purposes.

Balanced Bridge Method

Audio frequencies may be accurately measured by means of various bridge networks. A typical circuit for this application is the Wien bridge shown in Fig. 5. The circuit elements are so arranged that when a signal of unknown frequency is applied to the input, the bridge will be balanced for only one particular setting of two variable resistors. The point of balance is determined experimentally, usually by connecting a telephone headset or vacuum tube voltmeter across the output and

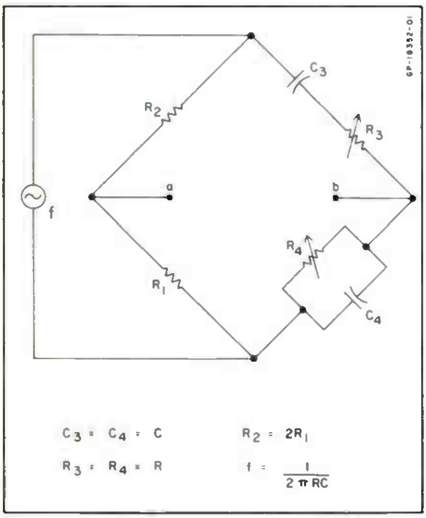

FIG. 5: Schematic diagram of a Wien bridge circuit for measuring frequency. Output of the bridge is across a-b.

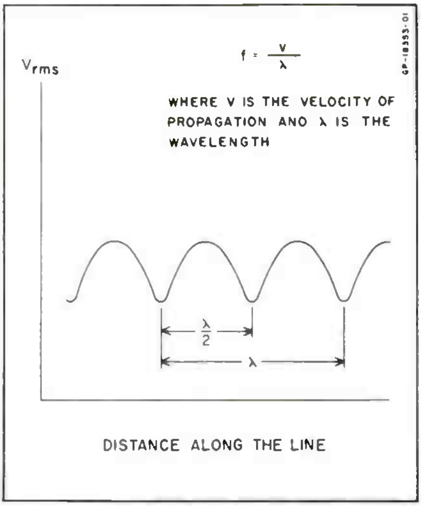

FIG. 6. Variation of amplitude of a voltage standing wave along a transmission line.

adjusting the variable resistances to obtain a null indication. The two variable resistances are normally constructed so that they can be adjusted to the same value simultaneously by means of a common dial. The dial can then be calibrated to read frequency directly.

An accuracy of less than 1 percent is not uncommon with carefully designed frequency - measuring bridges of this type. However, harmonics of the unknown frequency sometimes prove troublesome in the measuring process as they tend to be conspicuous in the output and may mask out the null point being sought.

Wavelength Measurements

An indirect method of frequency measurement involves the principle of resonance and the use of standing wave patterns on transmission lines. Since frequency and wavelength bear a fixed relationship to each other, frequency may be computed readily when the wavelength is known. Mathematically, frequency is equal to the velocity of wave propagation divided by the wavelength. The velocity of wave propagation is usually taken as the velocity of light, 300 million meters per second. At microwave frequencies, measurements may be made by measuring the wavelength along a transmission line.

The actual measurement consists of connecting a sensitive current-reading instrument across the line and adjusting its position along the line until a point of minimum voltage (maximum current) is obtained. The distance between any two such successive points when multiplied by two is equal to the wavelength. Figure 6 is a graphical representation of a voltage standing wave pattern on a transmission line.

The measurement of frequency by directly measuring wavelength has certain distinct advantages. One of these is the directness itself. The measurement obtained is a measurement of length and is independent of any comparison with a frequency standard. Also, since resonant lines normally have a high Q, the accuracy of such methods is quite good and may reach a precision that is accurate to within ± 0.1 percent. The basic principles of this method apply both to open-wire lines and coaxial cables. In the latter case, a slotted cable is used.

A variation of the above method uses a slotted waveguide. In this application, a probe is shifted along the length of a waveguide to determine the minimum voltage points of the standing wave pattern. The distance between these points is then read from a calibrated scale along the side of the waveguide. Figure 7 shows a slotted waveguide used to measure microwave frequencies. The scale on this type of instrument is usually equipped with a vernier and may be capable of measuring accurately to 0.1 millimeter.

FIG. 7. Slotted waveguide used to measure frequency in the 5.85 to 8.2 megacycle range.

Pulse Counting Method

Perhaps the most modern and one of the most convenient frequency measuring instruments is the electronic counter or scaler. The basis of such a device is a network which converts the signal to be measured into pulses and then counts these pulses against an accurate time base.

A circuit commonly used in such frequency measuring applications is the Eccles-Jordan circuit shown in Fig. 8. A circuit of this type has two stable conditions with a large area of unstable operation between. In one of these stable states, Tube 1 is conducting and Tube 2 is non-conducting and biased beyond cutoff. In the other stable condition, Tube 2 is conducting and Tube 1 is non-conducting and biased beyond cutoff. The shift from one stable state to another is almost instantaneous and is brought about when the grid of one of the tubes is excited by a voltage pulse of the proper polarity and sufficient amplitude. For this reason, such circuits are often referred to as trigger or flip-flop circuits.

When operating in one of its stable states, the circuit is triggered if a negative pulse of sufficient amplitude is applied to the grid of the conducting tube or a positive pulse to the grid of the non-conducting tube. The circuit will then jump suddenly to its other stable state. When another pulse of the proper polarity is applied to the proper grid, the circuit will jump suddenly back to its original stable state. Such a system therefore completes one cycle for every two pulses fed to it and may be said to count or scale by a factor of two.

Each cycle of the trigger circuit may then be converted to a pulse and fed to another trigger circuit. Thus, for two such trigger circuits, four original

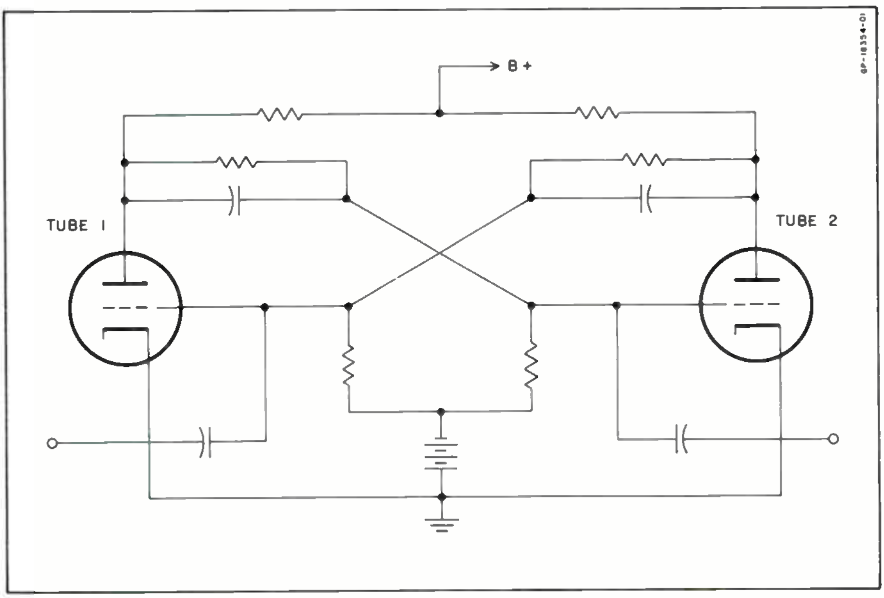

FIG. 8. Basic Eccles-Jordan trigger or flip-flop circuit.

FIG. 9. A frequency-counting system covering a very wide range. The counter is capable of measuring from 0 to 10 megacycles. With a frequency converter (shown mounted in front panel of counter) and transfer oscillator (top), the range can be extended to cover from 0 to 12,400 megacycles.

COURTESY OF HEWLETT-PACKARD CO.

pulses are required to complete one cycle of the second circuit. This process can be extended through several stages until the desired division factor is obtained. When used to generate lower frequencies than the original input frequency, this process is known as frequency division or subharmonic generation.

When such a circuit is used to measure frequency, the signal of unknown frequency is first converted to pulses which are then used to trigger the counting circuit. This count is then compared against an accurate time base furnished by a frequency standard which may be incorporated either in the counter instrument itself or external to it. Thus the counting circuit gives the number of pulses which occur in a given interval of time as determined by the standard. The result is the frequency of the signal being measured. Most frequency counters of this type are designed to display the reading by means of rows of neon lights which are energized in the correct order to provide a direct reading of the measured frequency at the end of each discrete sampling period. An electronic counting system covering a wide range of frequencies is shown in Fig. 9.

The accuracy of frequency counters cannot be any higher than the accuracy of the oscillator used as a standard to determine the time base. With high-quality oscillators for internal standards, instruments of this type can be designed to have an accuracy approaching one part in a million over short periods of time. Frequency-counting instruments may be used to cover a range of a fraction of a cycle to thousands of megacycles.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 7 NO. 3 MARCH 1958

MEASURING POWER AND FREQUENCY

At 6000 Mc

Because the wavelength is only about 2 inches, and because waveguide is used extensively in place of more conventional equipment components, the methods used for making measurements in the vicinity of 6000 Mc are considerably different from those used at lower frequencies. In carrier equipment, or in radio equipment operating at lower microwave frequencies, many measurements are made by connecting meter leads directly across points on the transmission path. This technique cannot be used with waveguide; instead, special apparatus must be employed to sample the energy passing through the waveguide.

Some of the special test instruments and measuring techniques used to measure power and frequency in waveguide circuits are discussed in this article.

Proper lineup and adjustment of microwave radio equipment requires that measurements of power and frequency be made at various points in the r-f circuits. When these measurements are made where waveguide is the transmission medium, some means must be provided to gain access to the energy passing through the circuit. Because the circumstances are not the same as at lower frequencies, microwave measurements require different types of test equipment. The individual items of test equipment utilize components which are considerably different in appearance from those used with carrier equipment. In this article, emphasis is placed on components of microwave test equipment. Among the individual components described are resonant cavities, crystal diodes, directional couplers, and bolometers.

Commercially available microwave test sets include the various components needed to make power and frequency measurements on waveguide circuits. The frequency meter portion of these test sets is normally based on the use of a built-in resonant cavity; the power meter portion is based on the use of a temperature sensitive device called a "bolometer." Test sets also normally supply a source of microwave test signal. A typical microwave test set is shown in Figure 1. Where test sets are not available, measurements can be

Fig. 1. A typical test set for measuring power and frequency on waveguide circuits.

made by setting up various arrangements of test equipment components.

Resonant Cavities

A typical resonant cavity of the type commonly used as a cavity wavemeter (to measure frequency) is shown in Figure 2. Is it essentially a cylindrical metal enclosure, one end of which can be moved by means of a micrometer screw adjustment. The micrometer scale normally reads in units of length. The reading is converted to frequency by reference to a calibration curve. Theoretically, a cavity wavemeter can be calibrated from its dimensions. In practice, however, the calibration is done by comparison with a standard meter. Although cavity dimensions can vary slightly with temperature, most wavemeters, once calibrated, will maintain sufficient accuracy under normal conditions to meet most field requirements. For critical applications, where extreme temperature variations are likely to be encountered, resonant cavities are made of materials (such as invar) with extremely low temperature coefficients. A crystal detector, an indicating meter, and a cavity wavemeter provide a means of measuring frequency equivalent to that provided in a standard microwave test set.

How a Resonant Cavity Works

A shorted quarter-wave transmission line is actually a resonant circuit. If two such resonant lines are connected in parallel, the resonant frequency is unchanged. In fact, connecting any number of shorted quarter-wave lines in parallel from the same two points

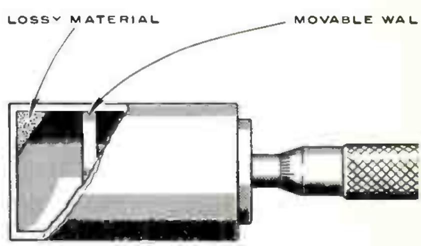

Fig. 2. A typical resonant cavity.

will not affect the resonant frequency. As more and more lines are connected in parallel, eventually a closed metal container will be formed—and this container is the simplest example of a resonant cavity. An example of this is shown in Figure 3.

A practical resonant cavity consists of a closed waveguide section with one dimension equal to an integral number of half-wavelengths of the resonant frequency. It can be excited in the same manner as any waveguide—i.e., by induction through a slot or from a probe inserted at the proper location.

Cavity Wavemeters

Resonant cavities used for frequency measurements are called *cavity wavemeters*. They are adjustable, and are calibrated to measure all frequencies within a certain band.

A special application of a cavity wavemeter, permanently adjusted to resonate at one frequency only, is called a *reference cavity*. In place of a micrometer dial, a reference cavity usually has only a simple screw adjustment which normally is set and locked once the cavity is tuned.

Three different types of cavity wavemeters are available—transmission, reaction, and absorption. They differ in the manner in which the resonant cavity is coupled to the waveguide. An example of each type is shown in Figure 4.

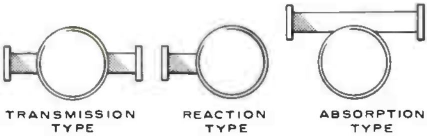

Fig. 4. Three types of cavity wavemeters.

Transmission- and absorption-type wavemeters are used with the signal source at one side of the cavity and a detector at the other. Resonance is determined by means of a microammeter connected to the detector. With the transmission type, the meter will peak at resonance, while with the absorption type, the meter will dip. Absorption-type wavemeters transmit maximum power at frequencies far from resonance. They can, therefore, be inserted directly into a transmission line and detuned when not in use. A transmission-type wavemeter, however, transmits maximum power only at resonance; it must be used with a directional coupler or some other arrangement that permits it to be removed from the transmission system when it is not in use. If left in the line, a transmission-type wavemeter would cause the r-f level to fluctuate with frequency changes.

Reaction-type wavemeters indicate resonance by a change in magnitude and phase of the reflection coefficient. The resonant frequency is best determined by instruments capable of detecting a

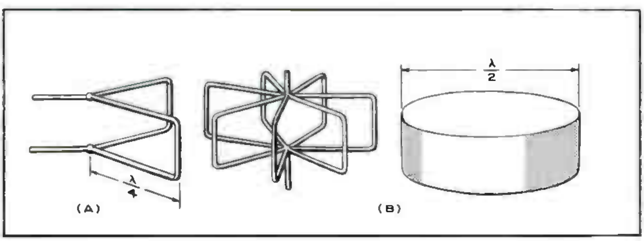

Fig. 3. Development of a Resonant Cavity.

change in phase. In addition to frequency measurements, the reaction-type cavity is often used as a reference cavity. The reference cavity in the transmitter afc circuit of the Type 74A Microtel system is essentially a reaction-type wavemeter. Because of its critical application, this cavity is made of invar.

Crystal Diodes

A crystal diode provides a convenient means of converting r-f energy in a waveguide into a measurable quantity. The crystal diode is connected to a probe (which may actually be a part of the crystal element) or loop which, in turn, is coupled to the electric or magnetic field in a waveguide. The sample of r-f energy intercepted by the probe is rectified by the crystal and can be measured with a microammeter.

Where the r-f energy in a waveguide is amplitude-modulated, a crystal diode can function as an AM detector. This characteristic is utilized in some types of measurements by deliberately inserting an AM signal into the transmission system. This signal can then be detected, amplified, and displayed on an oscilloscope to determine whether the desired characteristic has been obtained.

Bolometers

R-f power passing through a waveguide is normally measured by means of a *bolometer.* This device consists of a temperature-sensitive *bolometer element* in a *bolometer mount* which provides a means of connection to the waveguide. The bolometer element forms one leg of a bridge circuit. Since resistance of the bolometer will vary in accordance with the amount of power absorbed, power measurements can be made by determining the degree of balance of the bridge circuit.

There are two general classes of bolometer elements: (1) *barretters,* which have positive temperature coefficients and, (2) *thermistors,* which have negative coefficients. Any short piece of fine wire is a simple barretter. Instrument fuses are often used for the purpose, although the normal type of commercially available barretter consists of a piece of extremely fine platinum wire enclosed in a suitable capsule. Thermistors are made of metallic oxide materials selected because their resistance decreases as the temperature increases.

A bolometer can be mounted to absorb power directly from a waveguide, or from a probe which samples the r-f energy in the waveguide, coupling a definite portion of it to the bolometer.

Mounts

A crystal diode mount or bolometer mount provides means for coupling the element to the r-f energy, and for matching the impedance of the element to that of the transmission system. The mounts may be shorted sections of waveguide, connected to the transmission system by means of directional couplers, or they may be coaxial, and connected to the waveguide by means of a coaxial jack and probe.

A waveguide mount is essentially a short section of waveguide closed (shorted) at one end. Waveguide mounts can either be fixed to operate over a specific band of frequencies, or they may be tuned. Tunable mounts have tuning stubs which may be adjusted to exactly match the impedance of the element to that of the guide.

The bolometer element is held inside of the waveguide by the waveguide mount. Crystal diodes are mounted outside the guide, with the probe projecting into the waveguide.

Coaxial crystal diode and Bolometer mounts are also available, and are often used where a coaxial jack and probe are built into the main waveguide. Coaxial connectors are provided at each end of the mount so that, when connected to suitable instruments, the r-f power can be measured.

Directional Couplers

A directional coupler is a device used to sample the r-f energy traveling in one direction in a transmission system, with a minimum of interference from the r-f energy traveling in the other direction. There are two general classifications applied to waveguide directional couplers, the *multi-hole coupler* and the *cross-guide coupler*.

Multi-hole couplers are most often used for precision laboratory measurements and are sometimes called precision directional couplers. A typical multi-hole coupler is shown in Figure 5. It consists essentially of two parallel waveguide sections that have a common wall throughout most of their length. The main section is flanged at both ends and in use is a part of the r-f transmission system. The secondary section is used for measurement purposes. For example, a bolometer or crystal diode mount may be connected to the flanged end. The secondary section has a load termination at the blind end.

Wave energy traveling through the main section is induced into the secondary section through holes in the common waveguide wall. The holes are arranged in such a manner that wave energy in the main waveguide induces a wave traveling in the same direction in the secondary waveguide. An oppo-

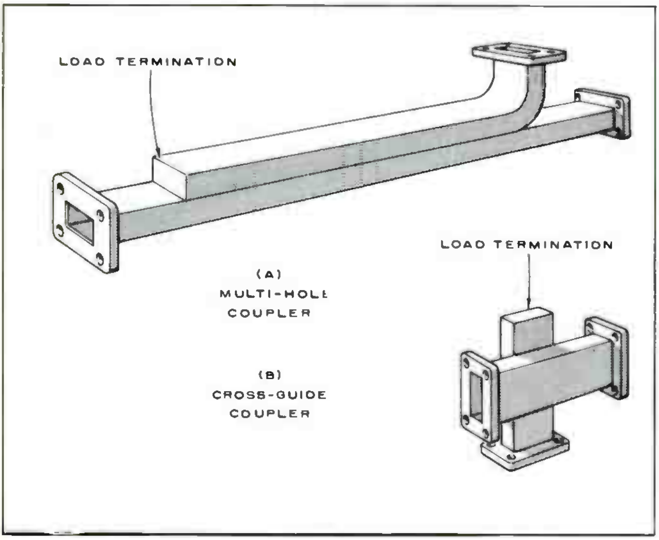

Fig. 5. Directional Couplers: At top, a typical example of a multi-hole coupler; at bottom, a typical example of a cross-guide coupler.

sitely directed wave is also induced in the secondary, but the energy of this wave is relatively small compared with the energy in the main waveguide.

There are two waves in the main guide. The wave carrying energy toward the load is called the preferred wave. Measurements in the secondary waveguide are normally related to the preferred wave. The preferred wave couples energy to the secondary flange, and the oppositely directed wave couples to the secondary load termination.

The power ratio between the preferred wave energy in the main guide and its component at the secondary flange is called the *coupling factor*, and is expressed in decibels.

A small amount of the power appearing at the secondary flange may be due to the energy of the oppositely-directed primary wave. The power ratio between the desired wave at the secondary flange and this undesired wave is called the *directivity*, assuming primary waves of equal magnitude. The directivity is expressed in decibels.

The definitions of coupling and directivity apply also to the cross-guide coupler, which is commonly used where there are space restrictions and where laboratory accuracy is not required. An example of a cross-guide directional coupler is shown in Figure 5. Although the axis of the waveguide sections which comprise the cross-guide coupler are at right angles, the operation is similar to that of the multi-hole coupler. Wave energy from the main guide is coupled into the secondary, and the direction of wave energy depends upon the direction of energy flow in the main guide.

Cross-guide couplers are available with flanges at both ends as well as with a flange at one end and a load termination at the other end. An advantage of a cross-guide coupler with a secondary load termination is that reflections from the termination to the secondary flange are minimized.

The secondary flanges should be covered by shorting plates or by waveguide terminations when they are not being used. Normally, shorting plates will be sufficient to prevent radiation which would cause deterioration in service. However, in some applications, matched terminations are used to provide optimum operation.

Typical values of coupling are 3 to 20 decibels for multi-hole, and 20 to 30 decibels for cross-guide couplers. Typical directivity for multi-hole couplers is 40 decibels or better, and for cross-guide couplers, 20 decibels or better.

Test Set

The components necessary for frequency and power measurements are incorporated in commercial microwave test sets. In addition to a cavity wavemeter, crystal detector, bolometer and power meter, a microwave signal generator is included in the test set, which greatly increases its versatility. The signal generator is similar to signal generators used at lower frequencies in that the oscillator may be tuned over a specific range, and the output power is adjustable. The signal generator is used where test and adjustments in a microwave system require a test signal of known value. Among the measurements which can be made are transmitter deviation, path loss and receiver sensitivity.

When power or frequency measurements are to be made, the waveguide energy is sampled by means of a probe or directional coupler, and is connected to the test set through a coaxial lead. The power meter includes a self balancing bridge circuit with the bolometer in one leg of the bridge. R-f power is read directly on the indicating meter. Frequency is measured in terms of the reading on the micrometer dial. To simplify field measurements, probes and coaxial jacks are often built into key spots in the waveguide system of microwave terminals and repeaters.

Frequency Measurement

The basic equipment for frequency measurements in the field include a cavity wavemeter, a crystal detector and mount, a potentiometer and a microammeter. These may be part of a microwave test set, or they may be used separately.

The sample of waveguide energy is applied to the cavity wavemeter. The crystal diode rectifies the energy so that an indication may be obtained on the microammeter. Rather than applying the rectified energy to the microammeter directly, a potentiometer is used to divide the voltage. This helps to prevent damage to this sensitive instrument. The cavity wavemeter is then adjusted to obtain the desired meter indication, either a peak or dip depending upon the type of cavity wavemeter used.

An oscilloscope can be used to facilitate the measurement. When connected to the output of the detector, the pattern on the oscilloscope (a straight line at frequencies off resonance) will jump as the resonant frequency is passed. This permits a rough frequency measurement to be done quite easily and rapidly. Once the rough micrometer setting has been made, the oscilloscope is replaced by the microammeter for the final measurement.

Power Measurement

For power measurement, a bolometer element and mount, and a microwave power meter are required. The power meter will include three legs of the bridge circuit, bridge power supply and the indicating meter. The bolometer element is the fourth leg of the bridge. R-f energy reaching the bolometer element changes its resistance and this tends to change the bridge balance. Self-balancing bridges are normally used with a power meter. The energy required to maintain the bridge balance is equal to the energy absorbed by the bolometer. The indicating instrument is direct reading and is calibrated in milliwatts or dbm.

Conclusion

While the components and measurement technique used in microwave measurements are somewhat different from those used at carrier frequencies, the quantities to be measured are not changed. In general, more precautions and a little more care must be exercised in making tests at microwave frequencies. However, they are not difficult to make, and only require an understanding of the measurement technique to be used and of the operation of the test equipment. Detailed descriptions of microwave test equipment and measurement techniques are included in the catalogs and instruction manuals of test equipment manufacturers.

THE
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NEWS FROM LENKURT ELECTRIC

VOL. 3 NO. 3 MARCH. 1954

Techniques for MEASURING OPEN-WIRE LINES *for Carrier Applications*

Part I

Measurement of transmission line characteristics should be an important step in planning the use of carrier over open wire lines. In making such measurements sufficient data should be taken so that the characteristics of the line in question are known throughout the frequency range of the carrier system.

Almost all of the tests required can be made very simply with inexpensive instruments which are also needed for normal carrier system maintenance. The purpose of this article is to describe and explain the tests necessary to determine transmission loss and to show how they can be accurately made with commonly available test equipment. Future articles will discuss measurement of crosstalk, noise, and other characteristics.

The basic worth of a transmission line for transmitting carrier frequencies depends on three factors. These are:

(1) The "transmission loss" between the sending equipment and the receiving equipment.

(2) The amount of crosstalk between the circuit in question and other circuits.

(3) The interfering effect of any external noise introduced into the circuit through the line facilities.

Plans for the installation of carrier equipment often include measurements of these three factors to determine the extent to which the line is usable for carrier.

The test equipment required to measure transmission loss need not be elaborate nor expensive. A variable frequency oscillator capable of generating the same range of frequencies to be used by the carrier equipment and with sufficient power output for convenient measurements should prove adequate as a signal source.

Two vacuum tube voltmeters having a wide frequency response and capable of measuring voltages in the range of 0.001 to 10 volts are also needed. It is desirable for the voltmeters to have a db scale. However, this feature is not absolutely necessary because loss in db can be determined from voltage measurements.

In addition to an oscillator and voltmeters, a calibrated adjustable resistor such as a decade resistance box or an assortment of re-

FIGURE 1. This voltmeter, oscillator, and decade resistance box are typical of the instruments that can be used for measuring transmission line characteristics.

sistors of known value are needed to accurately set the oscillator output power. With this small amount of test equipment, all the necessary information for determining the transmission loss of a line at carrier frequencies can be obtained. Figure 1 shows typical instruments that can be used for measurements at carrier frequencies.

Transmission Loss

When a signal source such as a carrier terminal or oscillator is connected to a transmission line, the power received at the other end of the line is less than the power that would be received if the load were connected directly to the signal source. The ratio of the power received from a transmission line by the receiving terminal to the power available from the sending equipment determines the "transmission loss" and is dependent on three factors. These factors are:

(1) The power dissipated by the line resistance as heat or radiated into space as radio waves.

(2) The power rejected by the line because of impedance mismatch at either of the line terminations or at intermediate points.

(3) The power accepted by the line but transferred to other circuits or conducting material by magnetic and electric induction.

These three factors which determine transmission loss are interdependent and difficult to measure separately. Their sum, however, is relatively easy to measure with simple and inexpensive instruments. Figure 2 shows a simple test set-up that can be used to measure transmission loss.

Determination of transmission loss requires the measurement of available power at the transmission line input and the power actually delivered to the receiving end equipment. The available input power is fixed by the internal resistance of the oscillator and the setting of the oscillator output control.

Impedance Matching

To obtain valid results when measuring carrier frequency transmission loss, the impedance matching conditions between the oscillator and the line must duplicate, as closely as possible, the impedance matching conditions that will exist when the carrier equipment is installed. In the usual case open wire lines are connected to a length of toll entrance cable at each end to bring the circuit into the telephone office. For carrier, a non-loaded pair in the toll entrance cable is usually used for this purpose. The junction of the open wire pair with the toll entrance cable results in an impedance mismatch, the severity of which depends on the length and type of cable. If the length of the cable is very short (less than 100 feet), the impedance mismatch may be tolerable. Longer lengths frequently cause reflection losses too large to be neglected.

Usually the carrier installation includes a line filter and impedance matching transformer at the terminal pole to separate the voice and low frequency carrier from the high frequency carrier and to match the open wire to the non-loaded pair of the entrance cable. However, since transmission loss measurements are usually made prior to the installation of carrier equipment, this line terminating equipment may not be available. Omission of the line filter will have little effect on the results. However, for those installations where the toll entrance cable is too long to be considered negligible, it will be necessary to either obtain suitable impedance matching transformers or make the transmission loss measurements on the open wire and toll entrance cable separately. Impedance matching transformers having the proper insertion loss and frequency response characteristics can be obtained from the carrier manufacturer or his distributor at minor cost.

To duplicate the conditions of impedance matching that will exist when the carrier equipment is installed, the oscillator internal resistance must duplicate the impedance of the circuit it simulates. Depending on the make and model of the test oscillator, its internal resistance may be anything from 20 to 1000 ohms.

Oscillators

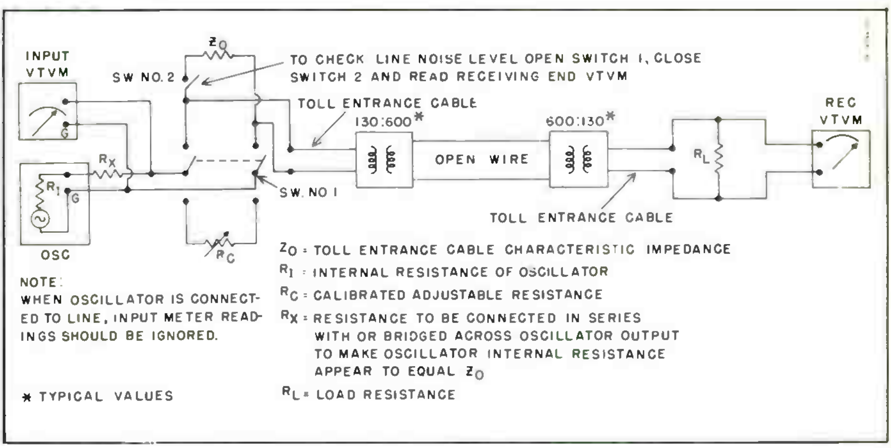
If the oscillator internal resistance is less than the characteristic impedance of the circuit it simulates, it can be increased to the correct value by connecting a resistance of suitable size in series with one of its output terminals.

If the oscillator impedance is greater than the impedance of the circuit it simulates, it can be reduced by connecting a resistance of suitable value directly across its output terminals. A simple diagram and a formula for determining the amount of resistance to connect across the output of the oscillator are shown in Figure 3.

Often the internal resistance of the test oscillator is unknown. However, this resistance can be measured quite simply as follows:

(1) Measure the open circuit out-

FIGURE 2. Suggested test setup for measuring transmission loss. With this one setup oscillator impedance, circuit noise and transmission loss can be readily determined.



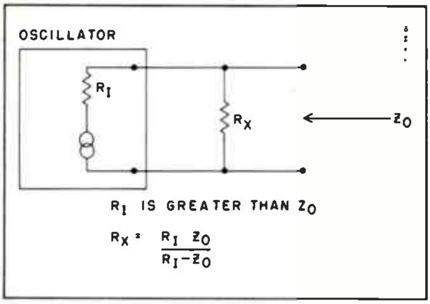


FIGURE 3. Circuit to match oscillator impedance to line characteristic impedance when oscillator impedance is greater than the line characteristic impedance.

put voltage of the oscillator.

(2) Load the oscillator with a calibrated, adjustable load resistance across its output terminals.

(3) Vary the load resistance until the voltage measured is one-half the open circuit voltage.

The load resistance is then equal to the internal resistance of the oscillator.

This method, though very simple, may not give accurate results on oscillators having a low internal resistance (below 100 ohms). Some of these oscillators have a distorted wave form when loaded too heavily. Their output voltage cannot be determined accurately because most vacuum tube voltmeters read accurately only when the measured quantity has a sinusoidal wave form.

If the oscillator internal impedance is found to be a low value, the accuracy of the measurement can be readily checked by the procedure outlined in Figure 4. This procedure can be used to determine the internal impedance of a great many electronic devices having a sinusoidal or d-c output voltage.

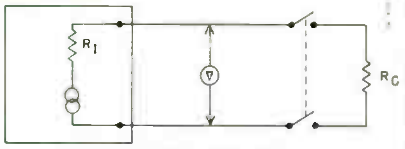
With the oscillator internal resistance properly matched to the desired impedance, the available power can be set to the desired value by the use of a calibrated resistor adjusted to equal the oscillator internal impedance. The desired available test power is set by substituting the calibrated resistor for the transmission line and adjusting the output of the oscillator to provide the desired power output.

In addition to calibrating and matching the oscillator to the input of the transmission line, the load resistance at the receiving end must also be equal to the impedance of the carrier equipment which will be connected to the line. This is necessary so that the impedance match between the line and load resistance will be the same as the impedance match between the line and the carrier terminal to be installed.

The transmission loss is measured by connecting the calibrated oscillator to the line input, varying its frequency over the desired range, and determining the power received at the other end of the line. Measurements should be made for both directions of transmission because the loss in one direction may be considerably greater than the loss in the other direction. Since open-wire carrier systems normally use different frequency bands for the two directions of transmission, it is sometimes possible, where coordination is not necessary, to arrange the terminal equipment to take advantage of differences that may occur between the transmission losses for the two directions of signal flow.

Units

It is customary and generally more useful to express transmission loss in decibels. However, it is not always convenient to make transmission loss measurements directly in db. Many vacuum tube voltmeters have a db scale that reads 0 db when the voltage is 0.775 volts. Others read 0 db when the voltage is 1.73 volts. Still others may have different db scale markings or none at all. Regardless of the scale markings of the



(1) WITH SWITCH OPEN, READ OPEN-CIRCUIT VOLTAGE V_{OC}
(2) WITH SWITCH CLOSED, READ LOAD VOLTAGE V_L
(3) THEN $R_I = R_C \left(\frac{V_{OC}}{V_L} - 1\right)$

R_C SHOULD BE LARGE ENOUGH SO V_L WILL BE 2/3 TO 3/4 OF V_{OC}

FIGURE 4. Circuit and procedure for determining internal resistance of an oscillator or other low impedance signal source.

instruments, transmission loss can be readily determined in db by use of the nomograph in Figure 5.

With this chart the power available from the oscillator, expressed in dbm, can be found by locating on the voltage and impedance scales the points corresponding to the internal resistance of the oscillator and to one-half the open circuit voltage. A straight line drawn between these two points intersects the dbm scale at a point corresponding to the power level, in dbm, available from the oscillator. The received power can be found by drawing a line between points on the voltage and impedance scales corresponding to the measured line voltage at the receiving end and the load impedance. The intersection of this line and the dbm scale gives the power received in dbm and the difference between the two dbm values is the transmission loss of the line (in db).

Since the available test power can be held constant for all frequencies by adjustment of the oscillator output control, its value in dbm need not be redetermined from the nomograph for each individual measurement. However, the received power in dbm must be determined at each frequency for which the transmission loss is measured.

Use of the nomograph for each individual measurement can be slow and inconvenient. If the receiving end meter has a db scale, the nomograph can be used to determine a correction factor to be applied to this scale. Then all the received power measurements can be made by reading the db scale of the voltmeter. The actual reading of the receiving end meter plus the correction factor gives the received power in dbm. For example, if the load resistance is 130 ohms and the db scale of the meter is marked so that 1 volt corresponds to 0 db, the nomograph shows that 1 volt across 130 ohms is +9 dbm. Then every reading of the db scale of the meter when connected across 130 ohms can be corrected to dbm by adding +9 to the reading. The correction factor for any db scale for any load resistance can be determined in the same manner.

In some cases, use of the nomograph or other charts or graphs is not necessary. If the voltmeter used at the sending end and the voltmeter used at the receiving end have the same scale markings and the internal resistance of the oscillator and the receiving end load resistance are equal, the transmission loss can be determined without converting to dbm. With the oscillator connected to a calibrated resistance, the value of which has been adjusted to equal the internal resistance of the oscillator, the db scale of the input meter connected across the oscillator terminals can be read directly. With the oscillator connected to the line, the db scale of the meter at the receiving end can also be read directly and the difference between the two readings is the transmission loss.

Sources of Error

Aside from purely human error, such as misreading instruments, improper connections, and errors

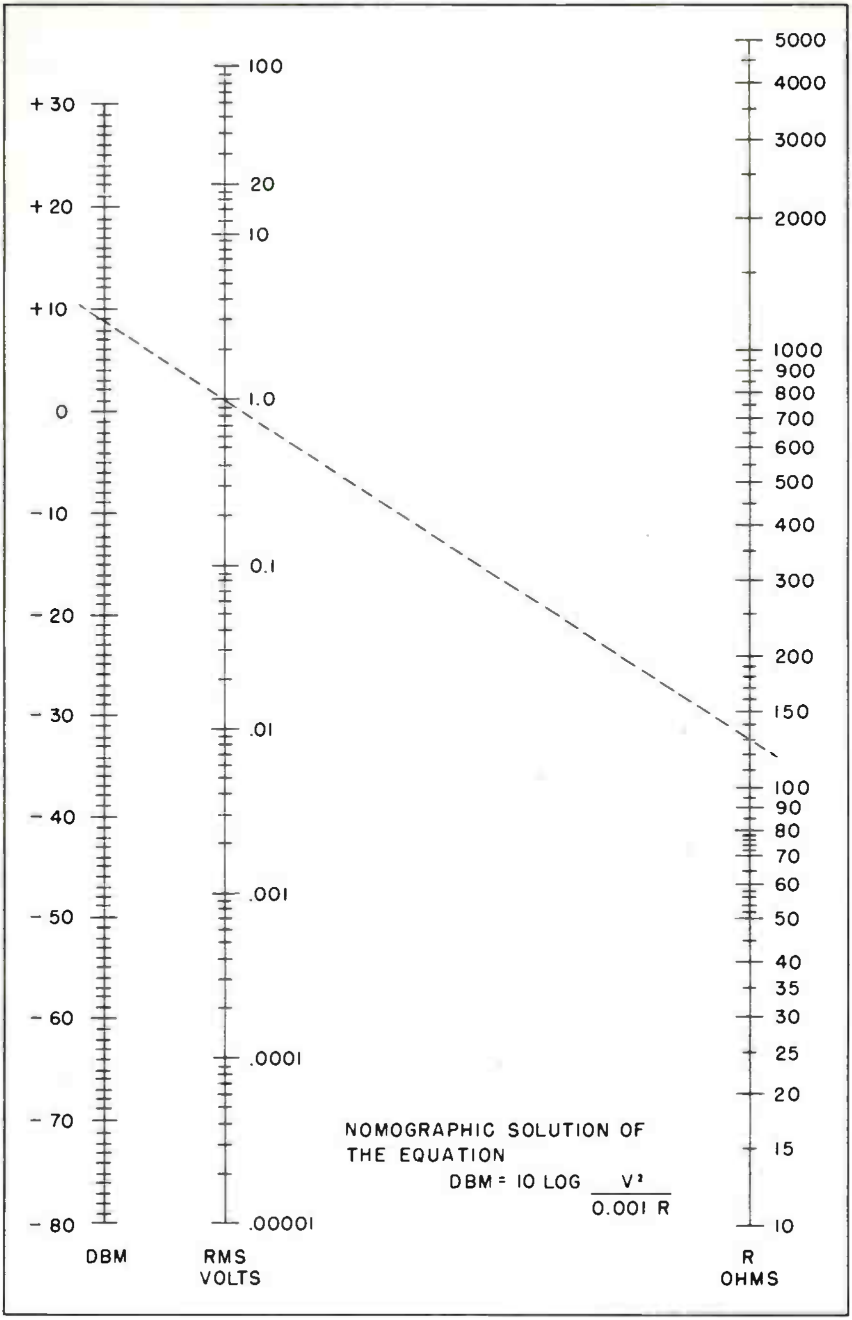


FIGURE 5. Determination of transmission loss from voltage readings is simplified by the use of this chart.

caused by test equipment imperfections, the chief source of error is the introduction of extraneous voltages (noise) into the circuit from external sources. On long circuits the error from this source can be considerable unless precautions are taken to reduce it as much as possible.

The simplest and most expedient precaution is the use of test tones of sufficient power so that the received signal will have a large amplitude compared to any noise that may be present. If available, a frequency selective voltmeter can be used that will eliminate most of the extraneous noise but allow the desired signal to be measured.

To determine the magnitude of any noise voltage that may be present, the oscillator at the sending end of the line should be replaced with a resistance equal to its internal resistance. Under these conditions, any voltage that appears across the receiving terminals is caused only by noise. The voltage that appears across the load when the oscillator is connected to the line is the combined signal and noise voltage. If the total voltage measured across the load resistance when a test tone is being transmitted is 15 db or more greater than the noise voltage, the error in received voltage measurement will be negligible. If the db difference between the combined signal and noise voltage and the noise voltage alone is less than 15, the oscillator output can usually be increased to provide a high enough received signal to restore the difference to 15 db or greater. If, however, the oscillator output cannot be sufficiently increased to override the noise by 15 db, Table I can be used to separate the test signal from the noise.

Test Results

When the transmission loss measurements have been completed, the results should be plotted in graph form for evaluation.

Evaluation of the results should take into consideration the weather conditions prevailing along the line at the time of test. If possible the test should be conducted under weather conditions similar to those the system is to be designed for. If this is not possible, allowance should be made for the worst weather conditions likely to be encountered.

Comparison of the test results with the regulation and gain specifications of the carrier equipment will help show the quality of circuit obtainable with carrier. The results may show (at least from a transmission loss standpoint) that the line is suitable for all carrier frequencies to be employed or that only part of the carrier frequency spectrum can be used. In the latter case it is usually possible to employ a partially equipped carrier system with those individual channels having too great a loss because of absorption peaks being deleted from the system. In a few cases, the tests may show that line repair and retransposition work is necessary before carrier can be installed. In any case, if there is a question, the test results should be made available to the distributor and manufacturer of the carrier equipment for their use in recommending the carrier equipment that would be satisfactory for the line conditions encountered.

TABLE I

DATA FOR SEPARATING SIGNAL FROM COMBINED SIGNAL AND NOISE WHEN MEASURING TRANSMISSION LOSS

DIFFERENCE BETWEEN DB OF COMBINED SIGNAL AND NOISE AND DB OF NOISE ALONE.	DECIBELS TO BE SUBTRACTED FROM DB OF COMBINED SIGNAL AND NOISE TO OBTAIN DB OF SIGNAL ALONE.
0.5	9.6
1.0	6.8
2.0	4.3
3.0	3.0
4.0	2.2
5.0	1.7
6.0	1.3
7.0	1.0
8.0	0.8
9.0	0.6
10.0	0.5
15.0	0.1

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NEWS FROM LENKURT ELECTRIC

VOL. 3 NO. 4 APRIL, 1954

Techniques for

MEASURING OPEN-WIRE LINES

for Carrier Applications

Part II

Increasing interest in carrier telephony has pointed up the need for more information concerning the application of high frequency carrier equipment to existing open wire lines. The only reliable method of obtaining specific information about how a particular line will perform at carrier frequencies is by measurement of the line characteristics at the carrier frequencies involved.

Part I of this discussion (March 1954. Vol. 3. No. 3) covered the measurement of transmission loss with inexpensive instruments that are also needed for normal carrier maintenance. In this part of the discussion, measurements of crosstalk and noise are described.

Measurement of crosstalk and noise involves many of the same techniques used for transmission loss measurements. With the exception of certain types of noise measurement, the instruments required are the same. Since most of the procedures required for transmission loss measurement are also required for crosstalk and noise measurements, the necessary steps for calibrating instruments and measuring transmission loss which were discussed in Part I have been summarized in Table I for easy reference.

Crosstalk Considerations

Crosstalk becomes important in planning for an installation of carrier over open wire lines under the following conditions:

(1) When more than one system of the same kind is to be installed on the same pole line.

(2) When the systems to be installed will parallel existing sys-

FIGURE 1. Near-end and far-end crosstalk paths.

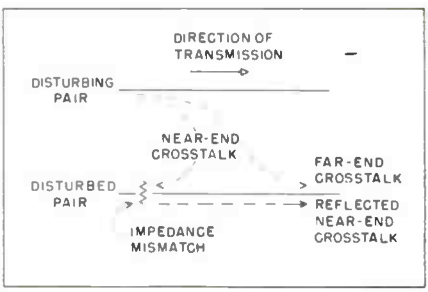
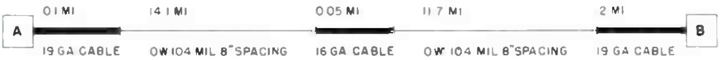


TABLE I

SUMMARIZED PROCEDURE FOR MAKING TRANSMISSION LOSS MEASUREMENTS AT CARRIER FREQUENCIES WITH COMMONLY AVAILABLE TEST EQUIPMENT

1. PREPARE DIAGRAM OF CIRCUIT LAYOUT:

TYPICAL EXAMPLE



2. TERMINATE LINE WITH PROPER VALUE OF RESISTANCE AND ARRANGE CIRCUIT TO SIMULATE IMPEDANCE MATCHING CONDITIONS THAT WILL EXIST WHEN THE CARRIER SYSTEM (S) IS INSTALLED.

3. DETERMINE OSCILLATOR INTERNAL IMPEDANCE FROM MANUFACTURER'S DATA OR BY MEASUREMENT.

4. IF NECESSARY ADD RESISTANCE IN SERIES OR PARALLEL WITH OSCILLATOR OUTPUT TERMINALS TO MAKE OSCILLATOR INTERNAL RESISTANCE APPEAR TO DUPLICATE THE RESISTANCE OF THE CIRCUIT OR EQUIPMENT IT SIMULATES.

5. MEASURE THE NOISE VOLTAGE APPEARING ACROSS THE LOAD RESISTANCE.

6. ADJUST OSCILLATOR OUTPUT SO THAT THE TOTAL RECEIVED VOLTAGE IS AT LEAST 15 DB ABOVE THE NOISE AT THE HIGHEST FREQUENCY TO BE USED.

7. FROM OSCILLATOR OPEN-CIRCUIT VOLTAGE AND NOMOGRAPH (MARCH ISSUE) DETERMINE POWER AVAILABLE FROM OSCILLATOR IN DBM.

8. FROM METER SCALE MARKINGS AND NOMOGRAPH DETERMINE CORRECTION FACTOR TO BE ADDED TO OR SUBTRACTED FROM RECEIVING END METER READINGS TO OBTAIN RECEIVED POWER IN DBM.

9. VARY OSCILLATOR FREQUENCY OVER DESIRED RANGE, TAKING MEASUREMENTS AT INTERVALS OF 4 KC. AT EACH FREQUENCY, CHECK AND READJUST OSCILLATOR OPEN-CIRCUIT VOLTAGE. ALSO CHECK CIRCUIT NOISE AT FREQUENT INTERVALS.

10. RECORD WEATHER CONDITIONS AT TIME OF TEST.

tems using the same frequency range.

(3) When expansion plans call for additional systems of the same type to be added to the lead at a future date.

Except in those cases where only one system will ever be installed, one or more of the conditions listed above is likely to be encountered.

Crosstalk between open wire pairs on a pole line can be divided into two main types: 'near-end' and 'far-end'. When signal energy introduced into one telephone circuit is detected on a paralleling circuit at the end opposite from the signal source, the detected signal is called far-end crosstalk; when detected at the same end as the signal source, the detected signal is called near-end crosstalk. Figure 1 illustrates the possible crosstalk paths between two parallel circuits.

Crosstalk is important only insofar as it interferes with desired speech signals or reduces the privacy of individual conversations. The ratio of desired signals to crosstalk signals at a v-f line termination determines the quality of a circuit from a crosstalk standpoint. This ratio, called crosstalk coupling loss, has the same significance as signal-to-noise ratio. If the crosstalk is intelligible, privacy may be the controlling factor, and the crosstalk coupling loss must necessarily be large to keep the crosstalk low enough so that it cannot be understood.

In carrier systems, crosstalk is largely unintelligible because of frequency staggering or inversion between individual carrier systems of the same type operating on the same pole line. When two individual carrier systems have staggered frequency allocations, corresponding channels of the two systems occupy a slightly different portion of the carrier spectrum. Crosstalk signals when demodulated appear to have been shifted in frequency by the amount of staggering and as a result are largely garbled and unintelligible. Inverted signals are generally classed as unintelligible noise.

Since two-wire carrier systems use completely different frequency bands for the two directions of transmission, near-end crosstalk is out-of-band compared to the desired signal and cannot be detected. Far-end crosstalk, however, will cause interference between systems because it is in the same frequency band as the desired signal.

The crosstalk performance of two parallel open-wire pairs can often be predicted from a knowledge of the pole line construction, transposition scheme, and impedance matching conditions. If a line is well constructed and transposed for the frequency range under consideration, and if impedances are matched at the line terminations and intermediate points, crosstalk performance should be satisfactory. If any of these conditions do not exist, or if transposition errors or irregular pole spacings are present, crosstalk performance may be poor and difficult to predict. To obtain the most satisfactory assignment of carrier to the various open-wire pairs, crosstalk measurements should be made for all possible pair combinations.

Crosstalk Measurements

Measurement of far-end crosstalk is much the same as measurement of transmission loss. In fact, both measurements can be made at the same time with considerable economy of time and effort.

When measuring crosstalk between two open-wire pairs, one pair can be designated A-B, with A representing the near-end and B the far-end of the pair (refer to Figure 2). In the same manner, the other pair can be designated C-D with C representing the near-end and D the far-end. If transmission loss measurements for a particular frequency are made from points A to D and from points C to D, then the decibel difference between these two measurements represents the far-end crosstalk coupling loss in db from line A-B to line C-D at the frequency used. In a similar manner, transmission loss measurements from C to B and A to B can be used to determine the crosstalk coupling loss from line C-D to A-B. In general, crosstalk measurements from A-B to C-D will differ from those made from C-D to A-B. Therefore, measurements in both directions are necessary. It is important in making crosstalk measurements that all pairs be terminated in their

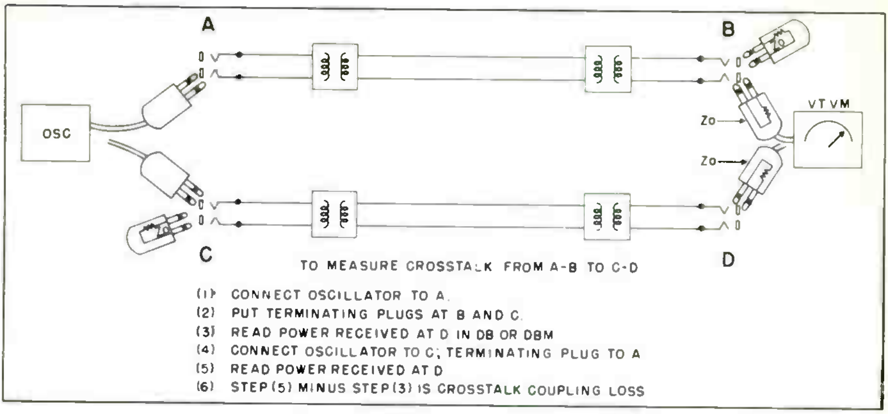


FIGURE 2. Suggested test setup for measuring crosstalk coupling losses. Measurements are necessary in only one direction of transmission.

characteristic impedance. It is equally important that impedance matching conditions in the lines simulate as closely as possible the conditions that will exist when the carrier equipment is installed. Figure 2 shows a simplified circuit layout and a suggested arrangement of test equipment to enable measurement of crosstalk and transmission loss at the same time.

Often two open-wire pairs on the same pole line have nearly identical transmission loss characteristics. In this case, crosstalk coupling loss from A-B to C-D can be determined more quickly by connecting the oscillator to point A and reading the received power at points B and D. (It can be assumed that the loss from C to D is the same as from A to B). The db difference between the power received at point B and the power received at point D is very nearly the crosstalk coupling loss from A-B to C-D. The crosstalk coupling loss from C-D to A-B can be determined in the same manner.

Corrections for Noise

The chief source of error in both transmission loss and crosstalk measurements is the presence of extraneous noise voltages in the circuits being measured. While in transmission loss measurements the test tone power can usually be increased sufficiently to override the noise, test equipment limitations and practical operating considerations often prevent the use of this expedient in measuring crosstalk signals. Even the use of a frequency selective voltmeter to measure received signals will sometimes prove insufficient to separate crosstalk signals from noise. If a difference can be detected between measurements of combined noise and crosstalk and noise alone, the graph of Figure 3 can be used to separate the crosstalk signals from the noise. (Procedures for measuring noise alone were outlined in Part I.) If no difference can be detected between these two measurements, the crosstalk signals are at least 20 db below the noise and normally insignificant.

Noise Measurements

An important asset of high frequency carrier telephony is its relative immunity to interference from many of the noise sources that affect voice and low frequency carrier. Electric power lines and other sources of audio frequency disturbances usually have little effect at carrier frequencies. At-

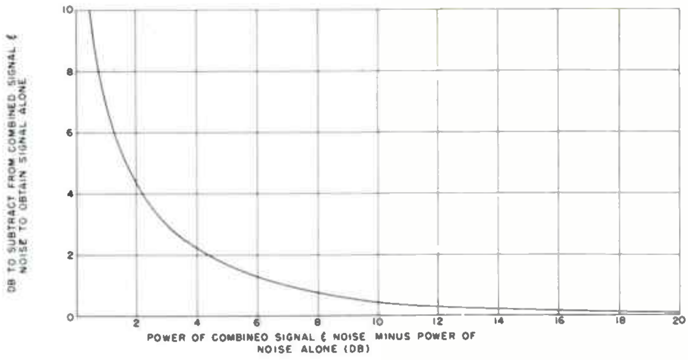


FIGURE 3. Graph for separating signal from noise when making transmission loss or cross-talk measurements.

mospheric disturbances cause noise of such a vagrant nature as to be impossible to predict accurately from a single set of measurements. If repeater sections are conservatively engineered in conformance with manufacturer's recommendations, noise from atmospheric disturbances will be well within tolerable limits.

Occasionally carrier transmission is subject to interference from various types of other apparatus generating frequencies that fall in the carrier range. Low-frequency radio transmitters and power line carrier systems are the principal sources. Such interference usually occurs as single frequency tones that affect only one or two channels per tone.

The presence and magnitude of single frequency tones can best be detected by a frequency selective vacuum tube voltmeter. This type of meter is in essence a calibrated radio receiver that can be tuned to any frequency in the carrier spectrum.

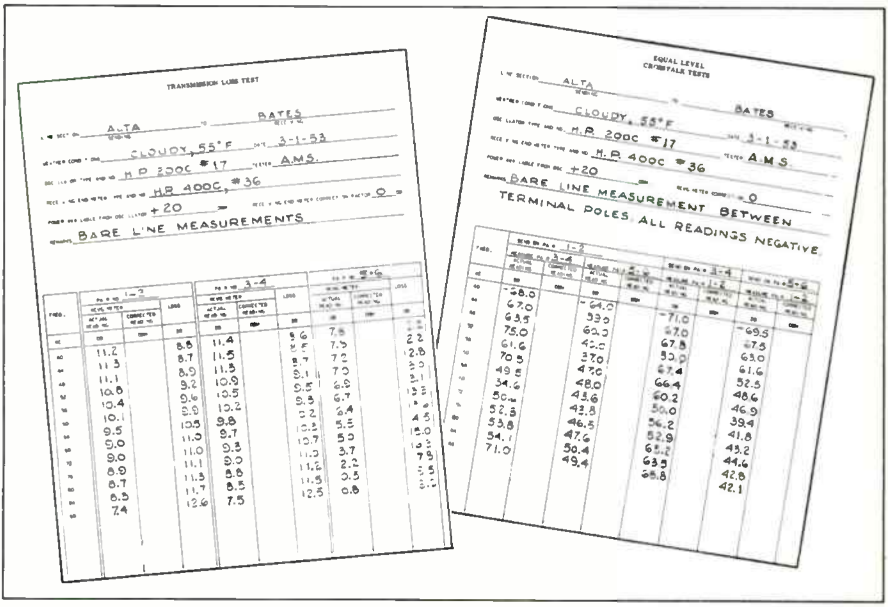
With the transmission line terminated at both ends with a resistance equal to the line characteristic impedance, a high impedance frequency selective voltmeter can be bridged across the receiving end and its tuning varied across the carrier spectrum in search of noise peaks and single frequency tones. The output of the meter can usually be monitored with headphones to determine the nature of any noise encountered.

If a frequency selective voltmeter is not available for determining noise peaks, an oscillator and a flat response vacuum tube voltmeter capable of measuring voltages as low as 0.0001 volts (-80 dbm in 600 ohms) can be used. The oscillator should be connected to the sending end of the line in the same manner as described for transmission loss measurements; the vacuum tube voltmeter is connected across the receiving end load resistance. The oscillator output control should be adjusted so that the receiving end meter can just detect the presence of test signal above the noise. Then by varying the oscillator very slowly across the carrier spectrum, the location of strong interfering single frequency tones can be detected. As the oscillator frequency approaches the frequency of a strong interfering tone, the oscillator tone and the interfering tone will beat against each other and cause the receiving end voltmeter to fluctuate. The oscillator frequency setting that causes the slowest fluctuation is approximately the frequency of the interfering tone. If the interfering noise is a steady single tone its magnitude can be determined approximately by adjusting the oscillator output until the amount of needle fluctuation of the receiving end voltmeter is a maximum (the needle should be fluctuating slowly so that it will not overswing.) The maximum fluctuation of the meter needle occurs when signal received from the oscillator is equal in magnitude to the interfering tone. When the meter needle swings to a minimum value, the oscillator and interfering tone are 180 degrees out of phase, the oscillator signal completely nullifies the single interfering tone, and the meter is reading all of the noise except the single tone being investigated. With the oscillator disconnected from the line and replaced by a resistance equal to the characteristic impedance of the line, the receiving end meter reads the total noise including the single interfering tone being analyzed. With these two values known, the graph of Figure 3 can be used to determine the magnitude of the interfering tone in the same manner as crosstalk magnitude in the presence of noise is determined.

Recording Data

The data acquired in making transmission measurements should be recorded in a systematic manner so that it becomes a complete and accurate record of the tests made. Every pertinent fact concerning the tests should be recorded. Important information includes identification of the line or lines under test, location of sending and receiving equipment, descriptions of test equipment, date of test, weather conditions, identification of test personnel, meter correction factors, oscillator frequency settings, and of course the actual readings of the sending end and receiving end meters. In addition, it is advisable to record a sketch (similar to Figure 2) of the test

FIGURE 4. Typical examples of data sheets for recording measurements of transmission loss and far-end crosstalk. Crosstalk coupling loss is calculated from the recorded data.



circuit used. The recorded data should be sufficiently complete so that even after a period of years it can be understood and interpreted.

Data should be recorded in a neat and legible manner. Erasures and mental computations should be avoided. When a mistake is made, the incorrect item should be lined out and a new line used for the correct version. Typical examples of recorded data are shown in Figure 4.

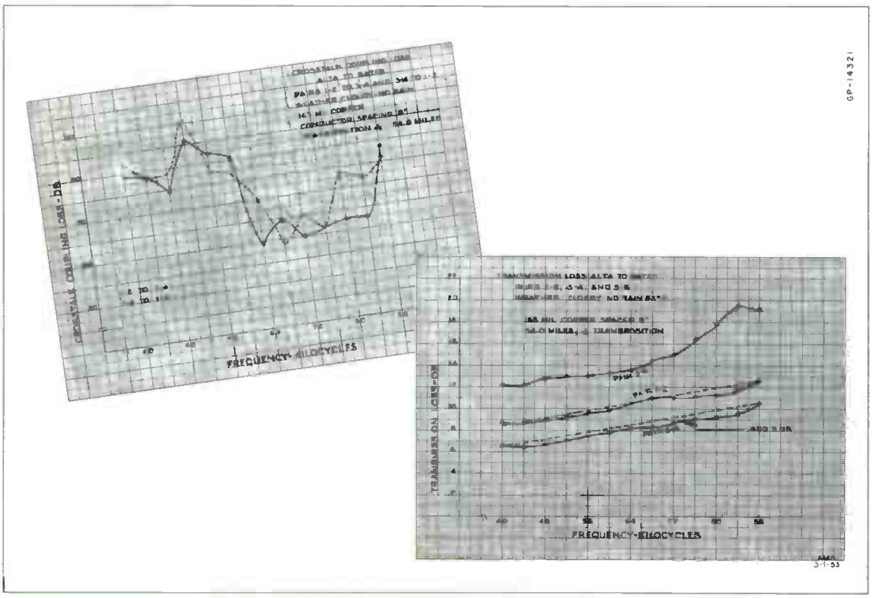
Test Results

When all the tests have been completed, crosstalk coupling loss, noise, and transmission loss should be plotted versus frequency in graph form for evaluation of the line performance (refer to Figure 5). Absorption peaks and poor crosstalk and noise performance can readily be detected and the performance of each channel predicted.

In the evaluation of plotted results of crosstalk and noise tests, the most important criterion is the quality of performance desired by the installing company or organization. Depending on a number of factors (frequency staggering, frequency inversion, type of service to be provided, and the length of the circuit), crosstalk coupling losses and signal-to-noise ratios greater than from 45 to 60 db should prove satisfactory. Telephone companies normally require very high standards of performance for long-haul domestic toll service. Performance requirements for other types of service will generally be less stringent.

In cases where the measured crosstalk and noise performance appears marginal, compandors can usually be used to provide up to 22 db improvement in crosstalk and noise performance. In any case, if there is a question concerning the evaluation of test data, the results of the tests and a description of the test procedure should be made available to the carrier distributor and manufacturer for their use in recommending the type of carrier installation that will best fit the user's requirements.

FIGURE 5. Graphs of transmission loss and crosstalk coupling loss plotted from the data shown in Figure 4.



THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 5 NO. 8 AUGUST 1956

Importance of Carrier in

NATIONWIDE TOLL-DIALING

The achievement of nationwide toll-dialing will require the careful planning and close co-operation of all segments of the telephone industry for the next several years. The new concepts in toll routing and design and the increased importance of carrier are significant aspects of the overall plan.

This article discusses the basic plan as it exists today and describes the important role that carrier will play in its achievement.

The ultimate objective of the nationwide toll-dialing plan is a system of communications throughout the United States and Canada which will enable any telephone customer to call any other customer merely by dialing his number. The accomplishment of the plan involves some new techniques and equipment. These in turn demand a high standard of transmission performance which will depend to a large extent on the use of carrier and microwave radio.

The basic structure of the plan has been set and the philosophy underlying it has been accepted by the major segments of the communications industry. However, telephony cannot afford the luxury of being a static art. As new developments occur, minor details of the plan may require revision; but the trend of any changes will remain directed toward an improved and more economical telephone service.

The Plan

The general plan is based on a principle called automatic alternate routing. By means of alternate routing, a call between any two points in the United States-Canada network may have many possible paths to its destination. These paths may vary in complexity from a direct connection to a connection containing as many as seven or eight intertoll links in tandem.

A call will be offered first to paths made up of the most direct routes between its two end points. These will be tried in logical order until either a

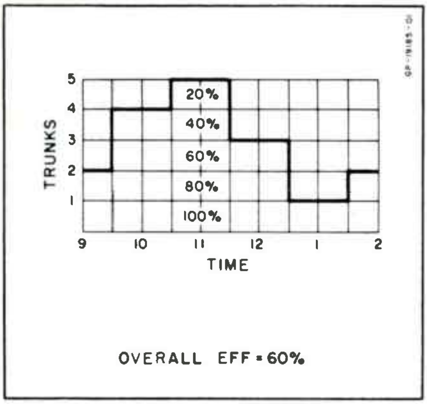


FIG. 1. Typical traffic distribution and trunk efficiencies over a period of several daytime hours.

through route is found or until it is determined that all trunks on these direct routes are busy. Since they are always the first choice, the trunks on such a route carry a heavy average traffic load and comprise what is termed a *high-usage group.*

If all high-usage groups are busy, the call will then be offered to a final group. A *final group* consists of trunks which have no alternate route. Such groups will have low-usage and be capable of handling the overflow of many high-usage groups with a small probability of lost calls.

The advantages of alternate routing stem from two fundamental facts associated with long-distance traffic: irregularity of flow, and diminished returns from additional trunks. Both of these are illustrated in Fig. 1 which shows a typical average variation of daytime traffic for a number of trunks between two points.

The curve consists of a peak of heavy traffic and valleys of light traffic. If this traffic is handled by directing the flow so that the trunks are used in ascending numerical order, the efficiency decreases with each succeeding trunk. At one extreme, trunk 1 is used to capacity throughout the entire time and operates at 100 per cent efficiency. At the other extreme, trunk 5 is idle most of the time and operates at only 20 per cent efficiency.

The over-all efficiency could be considerably improved by eliminating one or two trunks. However, this would mean that the calls represented by the top of the peak would be "lost" or at least delayed.

Long-distance traffic destined to one point in the country will differ in the positions of peaks and valleys due to time zone differences at the points of origin and for other reasons. If the peak traffic for several direct routes is allowed to overflow to a combined alternate route, the trunks on this alternate route may be operated at high efficiency as the peaks and valleys will tend to counteract each other and the curve will level out somewhat.

The trunks handling this peak traffic can then be eliminated from the direct routes. The net result is fewer trunks required and a higher operating efficiency for all trunks involved.

A hypothetical situation with and without alternate routing is shown in Figs. 2 and 3. Figure 2 shows the trunk requirements and traffic distribution between B and A and between C and A without alternate routing. Figure 3 shows how the same volume of traffic is handled by allowing the shaded portions of traffic to overflow to an alternate route between D and A.

Of course, the shaded portion of curve B-A must find available trunks between B and D and the shaded portion of curve C-A must find available trunks between C and D. These are provided either by direct high-usage groups or by final groups which are alternate routes for the combined traffic from several other points destined for D. Considerations of cost and existing traffic will determine the routing designation to preserve the efficiency and economy described for the initial situation above.

The example cited is a simple one involving one intertoll switching point (at D). Under the present general toll switching plan, as many as seven or eight intertoll links may be connected in tandem on some calls. The increased switching requirements are being handled by machines which automatically determine the destination of a call, search for an idle trunk, and build up a through circuit in the proper direction.

A large body of traffic theory and mathematical analysis underlies the techniques used in automatic switching. The final result, however, is simply that calls can be switched so quickly and so economically that switching is not a limiting factor in the achievement of the over-all plan.

Transmission Requirements

Alternate routing means that the path which any particular call may take will depend on the trunks available at the time. Thus successive calls between the same two points may vary widely in the number and types of links which make up the total path. For this reason it becomes important that the losses of each link be kept as low as practicable in order to prevent large differences in transmission levels among calls between the same two points.

Differences in transmission levels could be entirely eliminated if all links could be operated at zero loss. However, there is a practical limit to the minimum loss at which a circuit can be operated. This limit is a value of loss below which the factors of echo, singing, noise and crosstalk become objectionable. Of these factors, echo is usually the most critical and, in most cases, a circuit designed to hold echo within tolerance will meet the trans-

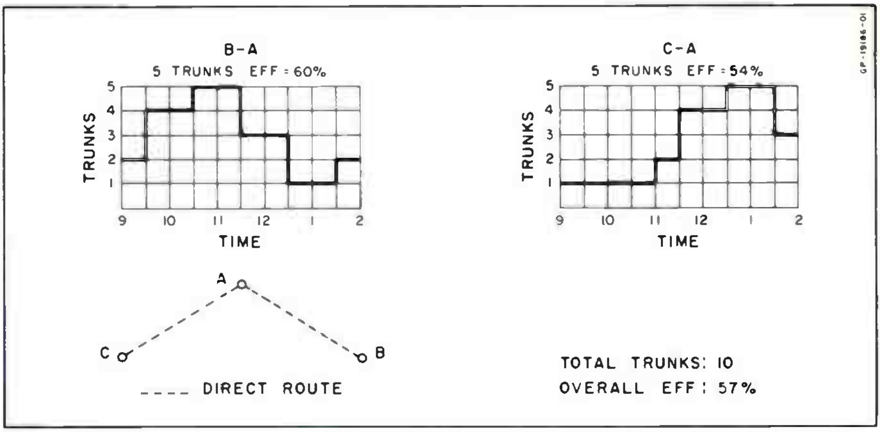


FIG. 2. Traffic distribution and trunk facilities required for a situation without alternate routing.

mission requirements for singing, noise and crosstalk.

The minimum loss of a circuit depends on the type of facility and the makeup of the total connection of which the circuit is a link. Since the makeup of the connection may vary from call to call, the minimum loss for a link will also vary. To maintain this minimum loss for each link for each possible makeup is far too cumbersome to be practicable. Therefore, a compromise is adopted which, with some approximation, gives a much simpler method of determining the lowest practicable loss for any link.

This method uses a figure called Via Net Loss (VNL) to define the loss which a particular circuit must maintain when it is used as an intermediate link, regardless of the total number of links in the connection. When the circuit is used as a terminal link rather than as an intermediate link, additional loss must be inserted. In either case, however, the VNL determines the lowest practicable loss at which a trunk can be operated without incurring objectionable effects of echo, singing, noise and crosstalk.

The VNL of a circuit is obtained by multiplying the circuit length by a value called Via Net Loss Factor (VNLF) and then adding 0.4 db to allow for individual variations. The VNLF is computed from the transmission characteristics of the circuit and a statistical analysis of customers' tolerance to echo. The relationship of tolerance to echo and VNLF is such that decreasing the objectionable nature of echo decreases the VNLF.

Tolerance to echo depends on the loudness of the echo and the time required for the talker's voice to make the round trip of the circuit and return to him. If this time delay is short, the effect of a given amount of echo is not too objectionable. As the time delay increases, however, the effect of the same amount of echo becomes more and more objectionable.

The primary transmission requirement of the toll-dialing plan—minimum loss on each intertoll link—becomes therefore a matter of achieving minimum VNLF for each link. This, in turn, means reducing the amount of echo in the circuit, the time delay of the echo, or both.

The computations of VNL and VNLF are based on the performance of four-wire circuits. For two-wire circuits, no simple solution exists and minimum loss must be determined from evaluations based in part on past experience.

Role of Carrier

Carrier has long since proved its worth over long-distance circuits by reason of the economies it affords. In addition, the inherent low noise level and high stability of carrier systems have continued to provide the quality of transmission required for toll use. As the nationwide toll-dialing plan progresses toward its goal, these contributions will be enhanced by other transmission characteristics necessary to the plan. Among these are low tendency to echo and high velocity of propagation.

The major share of echo stems from impedance mismatches at the hybrid junctions which must be used whenever a circuit is converted from two-wire to

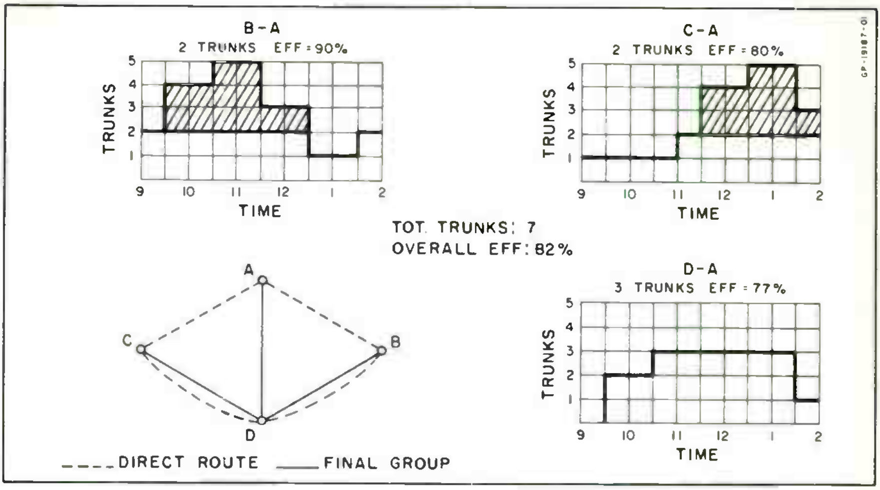


FIG. 3. Traffic distribution and trunk facilities required for same situation as in Fig. 2, but with alternate routing through D.

four-wire or from four-wire to two-wire. In two-wire circuits, these points of mismatch will occur at intermediate repeaters, terminal repeaters and switching connections.

Four-wire circuits with four-wire switching have much less tendency to echo since these points of impedance mismatch exist only at the terminal drops. In this respect, carrier circuits, which are electrically equivalent to four-wire circuits, serve the over-all plan by eliminating intermediate sources of echo and allowing circuits to be operated at low losses.

The high velocity of propagation of carrier circuits, on the other hand, acts to bring echo within tolerable limits by reducing the objectionable effect of whatever echo is present in the circuit. A higher velocity means a shorter time delay as the signals will require less time to make the round trip of the circuit.

Carrier and microwave circuits have a velocity above 100,000 miles per second. This assures a short time delay over even the longest intertoll circuits. Thus a higher echo level can be tolerated, or, in terms of transmission requirements, the circuit can be operated at a lower loss than it could if the delay were longer.

The transmission advantages of carrier and microwave are universally accepted throughout the industry. The vast majority of toll routes existing today make use of carrier and microwave circuits and as the plan progresses, more and more will be called into play. Also of importance, however, are the advantages carrier offers in providing some of the new short- to medium-haul facilities the intertoll-dialing plan will require.

One such application will come about when many small manual tributary offices are converted to automatic with the consequent transfer of operators to a toll center serving the area. Functions such as directory, information, complaints, and other services will require additional circuits. Since only a few extra channels are required and the distances are usually short, a carrier system such as the Lenkurt Type 33A or 45C, providing up to four channels and designed to prove-in over short distances, may result in greater savings than the alternate method of stringing extra wires or constructing extra facilities of other types.

Conclusion

Achievement of the nationwide toll-dialing plan will involve the careful and gradual integration of existing telephone plant into a complex long-distance network based on alternate routing and automatic toll switching. The new techniques necessary to accomplish this impose severe transmission requirements.

Probably the most important single role in the over-all transmission scheme is being played by carrier, not only in meeting these requirements but also in meeting the public's demands for more and better toll service resulting from a growing awareness of what constitutes good transmission and an increasing willingness to make use of long-distance facilities.

When the goal is reached, the plan will provide a continent-wide long-distance facility that is economical, convenient, fast, and reliable—a facility that will more than repay the effort and expense involved in its achievement.

THE
Lenkurt Demodulator
NEWS FROM LENKURT ELECTRIC

VOL. 2 NO. 10 OCTOBER 1953

TRANSMISSION LINE CHARACTERISTICS

And How They Affect Carrier Operation

A number of transmission methods are used for sending electrical signals over relatively long distances. Telephone signals are normally transmitted over open wire lines, wire pairs in a metallic sheath, coaxial cable, radio, or over a single wire with earth return. With the exception of radio transmission, all of the methods involve the use of metallic conductors.

Mechanically, a two-wire transmission line or coaxial cable is quite simple. However, the transmission line is one of the most complex circuits in the field of electricity. It is the purpose of this article to explain as simply as possible, the complex characteristics of transmission lines; and to show why carrier systems tend to take advantage of these characteristics.

Any transmission line has only four fundamental electrical properties. These are:

1. Series resistance
2. Series inductance
3. Shunt resistance
4. Shunt capacitance

The series resistance is the actual ohmic resistance of the transmission line conductors. The series inductance is the self-inductance of each conductor plus the mutual inductance between the individual conductors of the line. Shunt resistance is the total resistance of the leakage paths by which the current leaks across between conductors instead of going all the way to the end of the line through the terminating impedance and returning to the sending end. Shunt capacitance is the electrical capacitance between the two conductors of the line including the effect of capacitance to earth.

Each of the fundamental properties is normally considered to be constant for any particular set of conditions. Their values depend primarily on the physical configuration and the material used in the line construction. To a lesser degree, however, they also depend on frequency, temperature, and weath-

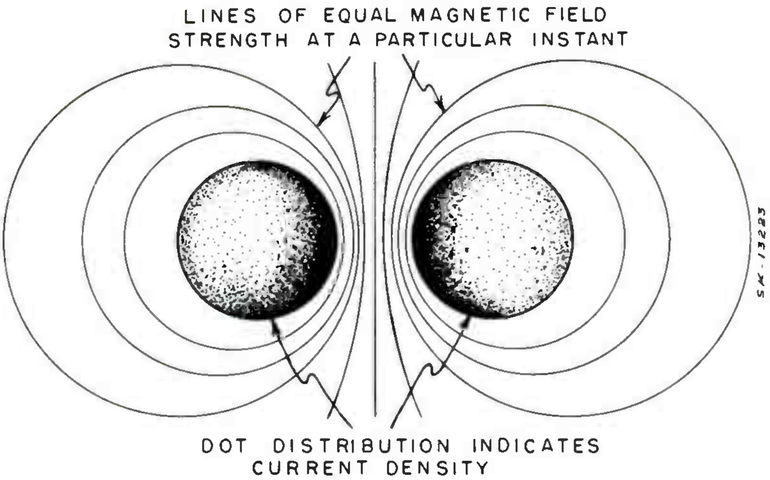


FIGURE 1. Increased line attenuation at high frequencies in cable pairs is partly caused by an increase in series resistance due to higher current density near the surface.

er conditions. Multi-pair sheathed cables and coaxial lines are not normally affected by weather conditions other than temperature.

Frequency Effects

The series resistance increases with increasing frequency because of a phenomenon called 'skin effect'. As the frequency is increased, an expanding and contracting magnetic field within the wire forces current to flow toward the outer surface of the wire rather than be distributed evenly within the wire. This action, in effect, reduces the total cross-sectional area of the wire and thereby increases the series resistance of the line.

In multi-conductor cable where the individual wires of each pair are very close together, series resistance is further increased by "proximity effect". The interlinking magnetic fields of the two wires force the current to flow in that portion of each wire that is closest to the other wire, causing a further reduction of the effective cross-sectional area of the conductors. Like skin effect, proximity effect in cable circuits increases with increasing frequency. The external magnetic fields and current distribution in a cable pair are shown in Figure 1.

A slight reduction of the self-inductance with increasing frequency is caused by the change in current distribution which reduces the net strength of the magnetic field within the wire.

Shunt capacitance changes so slightly with increasing frequency that the change is negligible at all frequencies below the micro-wave region.

Shunt resistance decreases considerably with increasing frequency. The total shunt resistance consists of two parts. The first and most familiar part is the actual ohmic resistance of the line insulation. The second part is an apparent resistance caused by internal heating of the line insulation. When an alternating voltage is impressed across the line, the insulation is stressed first in one direction and then the other. The heating that results is a power loss and appears to the line input voltage as a reduction of the shunt resistance. Since this loss increases with frequency the apparent shunt resistance decreases with frequency.

Characteristic Impedance

Various mathematical expressions based on the four fundamental properties are very helpful in determining the performance of transmission lines. One of the most useful of these expressions is called the characteristic impedance. A very common definition of characteristic impedance is that it is the impedance an infinitely long, smooth line would present to a source of alternating voltage connected to one end of the line. While this definition is quite adequate for many purposes, it is not very helpful for interpreting or understanding the performance of a line.

Characteristic impedance can be defined in terms of the fundamental properties as follows: If the effect of the series resistance and series inductance is the series impedance, and the effect of the shunt resistance and shunt capacitance is the shunt impedance, the characteristic impedance is the square root of the product of the series impedance and the shunt impedance.

Inasmuch as the series impedance and the shunt impedance vary with frequency, it would be natural to assume that the characteristic impedance would also vary with frequency. This assumption however, is not necessarily true. The series reactance ($2\pi fL$) varies directly with frequency while the shunt reactance ($1/2\pi fC$) decreases as the frequency increases. Likewise, the series and shunt resistances vary in a somewhat similar manner. The net effect is that the characteristic impedance tends to remain constant at all but the lower frequencies. Figure 2 shows that the characteristic impedance of an open wire transmission line varies only slightly in the carrier frequency range of 35 to 150 kc. The actual variation of the characteristic impedance depends in great part on the type of transmission line being considered. Figure 3 when compared with Figure 2 shows that there is a much greater variation in unloaded multi-conductor shielded cable than there is in open wire lines or coaxial cable. However, even for cable lines there is less variation at carrier frequencies than at voice frequencies.

For many uses of transmission lines it is sufficient to assume that the characteristic impedance remains constant regardless of frequency variation. However, for telephone work, particularly at the lower frequencies, the variation in characteristic impedance can affect the impedance match at the line terminations and lead to reflections, echos, and other unwanted effects. Because of the constancy of the characteristic impedance at carrier frequencies, elaborate impedance matching networks at line terminations are seldom required.

Attenuation and Velocity

Another important mathematical expression involving the series and shunt impedances is called the propagation factor. Defined as the square root of the series impedance divided by the shunt impedance, propagation factor is a measure of the attenuation and velocity (phase) characteristics of the transmission line.

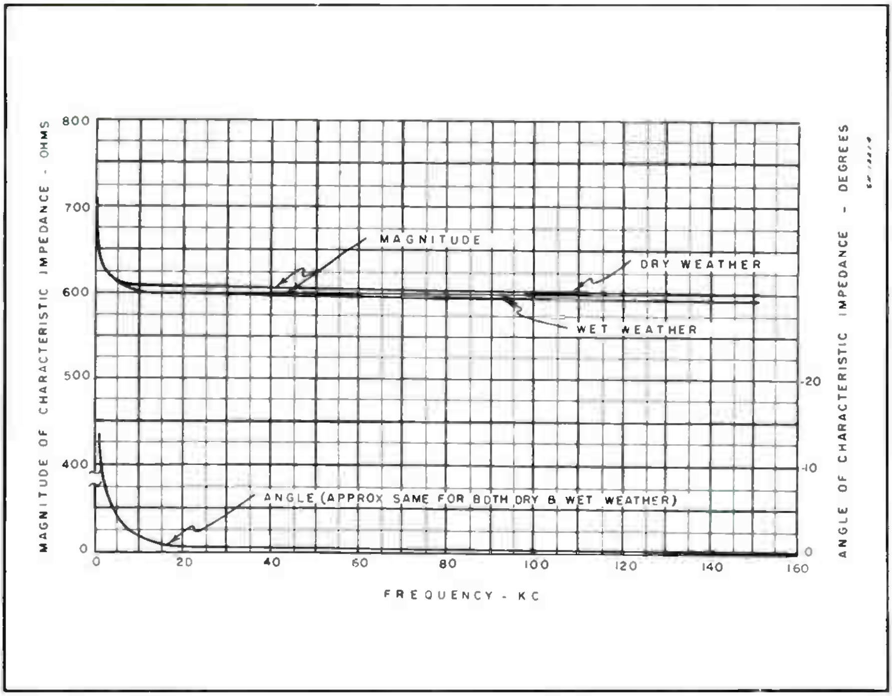
The propagation factor is composed of two parts which bear a relationship to each other similar to that of a resistance and reactance. The resistive part is proportional to the attenuation of the line in db. The reactive part is inversely proportional to the velocity of wave travel along the line and directly proportional to the frequency. It is sometimes called the phase constant or phase characteristic. It is desirable that the velocity of wave travel be as high as possible and that all frequencies have the same velocity. Relatively slow velocities such as occur in loaded cables cause echoes to be more troublesome. Because velocity usually increases sharply with frequency, in non-loaded cable long voice frequency lines may cause distortion by delaying the lower frequencies more than the high frequencies. The greater delay of some frequencies more than others is normally insignificant in open wire lines, loaded cable, and coaxial lines except for high fidelity program and picture transmission.

Influence of Weather

Open wire lines are often subject to a wide range of weather conditions. Cable lines are normally impervious to precipitation, although they are affected by temperature changes.

In the case of open wire lines,

FIGURE 2. Variation of Characteristic Impedance with frequency and weather of typical open-wire line with 8-inch spacing.

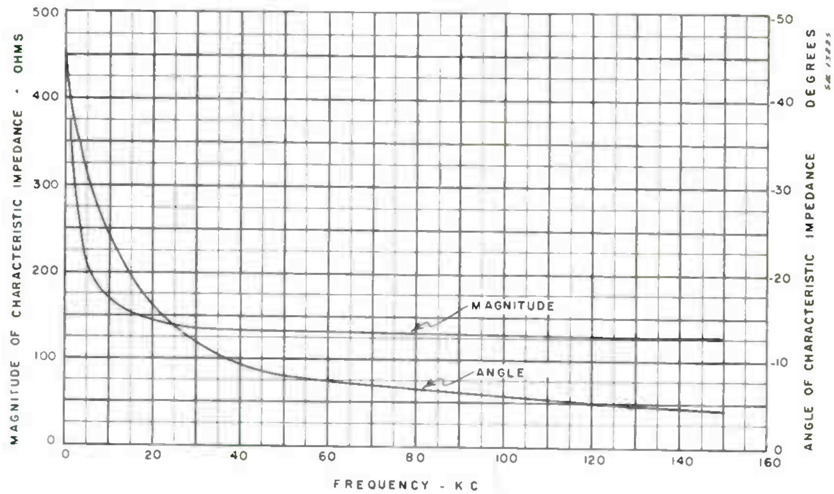


weather variations have significant effects on the series and shunt resistances and the shunt capacitance of the line. The series resistance of an open wire line is increased under high temperature conditions and also under wet weather conditions because of increased skin effect. During wet weather conditions the film of moisture on the wires is conductive to a certain extent. Since the magnetic field within the wires is forcing the current to flow more toward the surface, part of the current is crowded off the wire surface and flows in the film of water. The resistance of the water film is many times greater than the resistance of the copper wire. Therefore, the losses incurred by the current flowing in the water film are appreciable. During frosty or icy conditions, the coating of ice on the wires can become very thick with a considerable part of the current leaving the wire and flowing in the ice coating as shown in Figure 4.

Moisture and ice also affect the shunt resistance of the line. When the weather is dry, the shunt resistance is high and the loss is relatively low. During wet weather, dirt and dust collected on the insulators become much more conductive and the shunt resistance is lowered allowing more current to flow through the shunt leakage paths and thus increasing the attenuation. Under severe icing conditions the attenuation caused by skin effect and shunt leakage may be increased by as much as six or more times normal dry weather attenuation.

The age of open wire lines also tends to affect their operation. The continued effect of exposure to the weather causes the conductor sur-

FIGURE 3. Effect of frequency on magnitude and angle of characteristic impedance of typical 19-gauge toll cable transmission line.



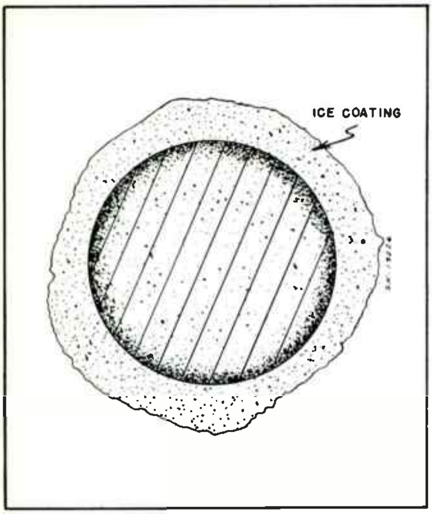


FIGURE 4. Current flow in a conductor covered with ice. Density of dot distribution indicates density of current flow.

faces to become rough and corroded. Splices, minor nicks and scratches that may have been incurred in the construction and subsequent repair of the line become filled with an accumulation of dirt, soot, and corrosion. Dirt and corrosion on the conductors increases skin effect and causes attenuation to increase with age. Since skin effect is more pronounced at carrier frequencies, aging will affect the attenuation of carrier frequencies. However, most carrier systems are automatically regulated or have adjustable networks that can be set to compensate for aging and other effects.

The effect of age on cable and coaxial circuits is not as great as it is on open wire circuits except for the possible deterioration of the insulation caused by moisture accumulation within the cable.

Line Balance

In addition to weather, temperature, and age, another factor that can influence the operation of transmission lines is the line balance. As previously mentioned, a magnetic field is associated with the current in the line. There is also an electric field associated with the voltage of the line. It is very important in telephone operation, especially at carrier frequencies, that the electric and magnetic fields about a transmission line be balanced with respect to earth and other nearby lines or conducting materials. When the electric and magnetic fields of both conductors of a two-wire line have the same coupling with the earth and other nearby lines and conducting materials, the line is said to have longitudinal balance. If the line is not balanced serious impairment of transmission can result. Noise, crosstalk, and absorption peaks can all be caused by unbalanced lines.

Noise can come from power lines adjacent to the telephone line, or from electrical apparatus and machinery operating near the line. Crosstalk is caused by coupling with other telephone circuits. Absorption peaks are caused by coupling with other circuits that are series resonant at some frequency in the range being transmitted. The resonant circuit absorbs power from the telephone line at the resonant frequency causing large attenuation at that frequency.

Effective balance is normally obtained by transposing the lines at regular intervals. By exchanging the positions of the conductors frequently along the line, unequal couplings between the two line wires and other circuits or earth tend to be canceled out and the average couplings of both conductors become very nearly the same. The number of transpositions required and the complexity of the transposition scheme depend on the frequency range to be transmitted and the number and nature of adjacent lines.

At carrier frequencies on open wire lines, more transpositions per mile are required than for the same lines at voice frequencies. Lines transposed for carrier may have a transposition point as often as every other pole. Where several wire pairs are strung on the same pole line, the different wire pairs must also be transposed to maintain balance. Reduction of absorption peaks and the general improvement of the attenuation characteristic effected by transposition is apparent in Figure 5.

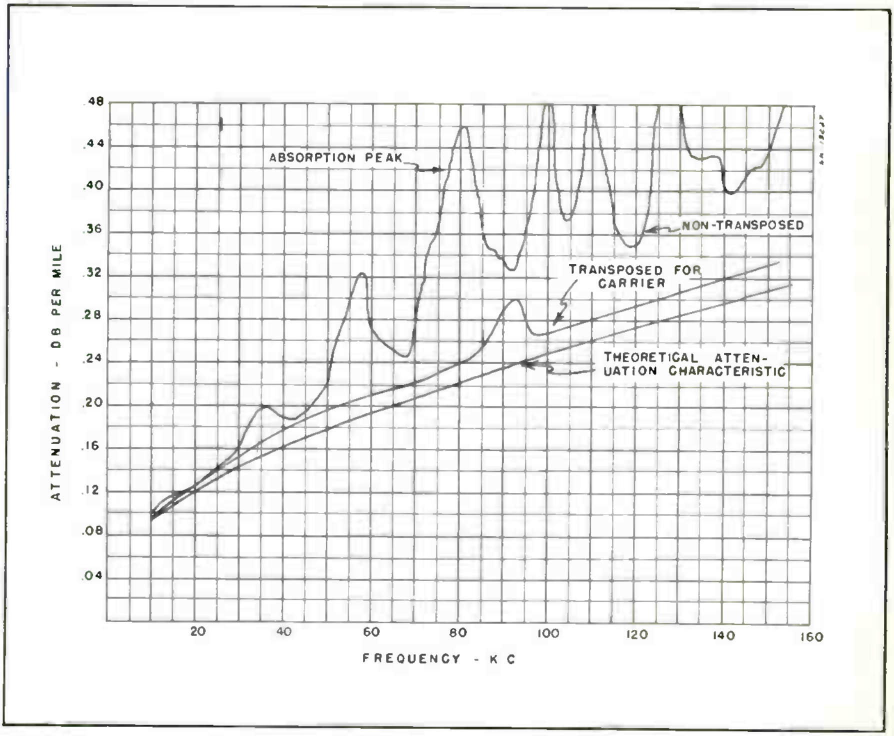
In cable circuits transposition is effected by using twisted wire pairs for the separate leads. The leads themselves are also transposed at frequent intervals.

A line transposed for carrier, whether it be cable or open wire, is also transposed for voice. However, the opposite is not necessarily true.

The Distortionless Line

Two principle types of distortion occur in telephone lines. They are frequency and phase or delay distortion. Frequency distortion refers to the unequal attenuation of signals depending on their frequency. It is shown graphically in Figure 5 as the increasing attenuation with frequency. A common name for frequency distortion is 'slope'. Changes in slope caused by vari-

FIGURE 5. Improvement in attenuation characteristic of open-wire line when transposed for carrier.



ations of weather and temperature are often referred to as 'twist'.

Delay distortion is caused by variation in velocity of wave travel. Higher frequencies usually travel faster than do low frequencies. In long non-loaded voice frequency cable circuits delay distortion can be so great as to render conversation virtually unintelligible.

If a line is so constructed that the ratio of series inductance to the shunt capacitance is equal to the ratio of the series resistance to conductance (for most telephone lines conductance is approximately the reciprocal of the shunt resistance) the line is distortionless. The characteristic impedance is purely resistive. Attenuation is a minimum and is the same for all frequencies. Velocity of wave travel is the same for all frequencies.

In practice it is impossible to build a distortionless line. The distortionless condition is most closely approximated in open wire lines of wide spacing and large conductor diameter. Such a line is of course very impractical for telephone use. In the usual telephone line the conductance is very small compared to the series resistance which requires that the inductance to capacitance ratio be very large to approach the distortionless condition. It is not practical to increase the conductance because that would increase the attenuation. To decrease the series resistance requires larger wires which is also impractical from the cost standpoint.

Series inductance can be increased in several ways. Spacing between wires can be increased which will increase the inductance and reduce shunt capacitance. This, however, is impractical for several reasons, the most important being cost and increased crosstalk. Wrapping the conductors with insulated permalloy tape has been used but is too expensive for most applications. Open wire line characteristics are sufficiently close to the distortionless condition so that further line treatment is unnecessary. Cable conductors, however, are much closer spaced and usually smaller guage than open-wire leads. Therefore, distortion effects are more pronounced in cable-pair circuits. The usual method for reducing distortion in cable lines is to 'load' the line by placing inductance coils in series with it at frequent intervals.

Loading of cable lines is not a perfect solution to the distortion problem. However, when properly loaded, line performance can be greatly improved. The biggest disadvantage of loading is that a loaded line acts like a low-pass filter. That is, the line remains relatively distortionless up to a certain frequency above which the attenuation and delay increase greatly with frequency. To raise the cutoff frequency, smaller inductance coils with more frequent spacing are required. Loading is almost essential in long cables at frequencies up to about 35 kc where large amounts of attenuation and delay distortion cannot be tolerated.

It should be pointed out that voice frequency loading and carrier loading requirements differ considerably. Carrier loading involves the use of smaller inductance coils placed more frequently in the lead. At frequencies above about 35 kc loading tends to become impractical because the loading coils required are too small and too closely spaced in the line for economy.

Conclusions

Many problems are encountered in maintaining adequate transmis-

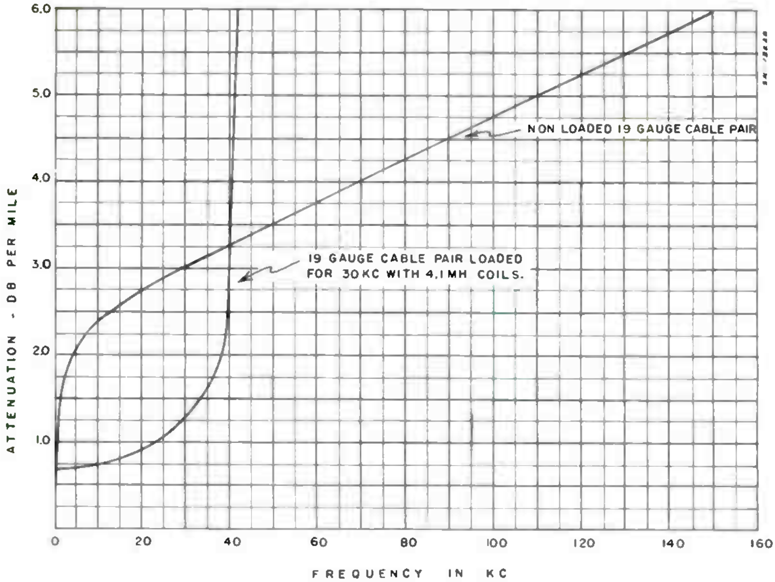


FIGURE 6. The loaded line acts like a low-pass filter. Attenuation is lower for frequencies below 40 kc but too great for use at carrier frequencies above 40 kc.

sion over open wire lines and cables. However, engineers are continually developing new techniques and devices to solve them. Techniques such as loading for voice and low-frequency carrier systems and more recently the development of high-frequency carrier make it possible to economically secure many telephone channels on the same pair of wires. Evidence of the successful solution of transmission line problems is found in the millions of miles of transmission lines criss-crossing nearly every country in the world.

Most transmission lines prove satisfactory for extensive use of multi-channel carrier systems, provided their characteristics are recognized and properly adapted for operation at the frequencies intended. Carrier equipment as furnished by the manufacturer usually includes provisions such as regulators and equalizers to completely compensate for many of the detrimental effects of distortion and weather. In some marginal cases where noise and crosstalk caused by poor line balance is objectionable, the application of compandors will bring the quality of transmission up to acceptable standards without additional transposing and repair.

THE *Lenkurt* ® **Demodulator**

NEWS FROM LENKURT ELECTRIC

VOL. 5 NO. 3 MARCH, 1956

Choice of

OPEN WIRE FACILITIES

for Carrier Operation

The choice of a suitable type and size of wire is one of the important factors involved in engineering open-wire communication facilities. With multi-channel carrier operation, considerations of strength as well as transmission have become more important than ever. This has led to increasing use of copper-steel instead of copper wire in major segments of the communication industry.

This article presents a brief review of the relative characteristics and costs of the two types of wire and some of the considerations involved in the use of copper-steel wire.

Before the advent of carrier communication and for a number of years thereafter, copper wire was employed almost exclusively for open-wire toll circuits of any substantial length and importance. For economic reasons it was general practice to employ the smaller sizes of wire (mostly 104-mil diameter), particularly after the vacuum tube repeater came into general use.

With the advent and increasing use of carrier superposed on the basic open-wire facility, it became more apparent than ever that small-size copper wire lacked the mechanical strength necessary to better safeguard service, particularly in areas subject to sleet storms, abnormally high winds, or other heavy loading conditions. This led to the use of the larger sizes of copper wire (usually 128-mil diameter) on lines which were to be developed for maximum carrier usage, particularly where the frequencies above 30 kilocycles were involved.

Shortly after the start of World War II, copper-steel instead of larger sized copper wire was employed for two important projects, mainly because of the need for conserving copper as a war measure. One of these involved the construction of a new line, with

104-mil 40 percent conductivity copper-steel wire arranged for maximum carrier usage, from Danby, California to Yakima, Washington—a distance of about 1,300 miles.

The other project was the Alaskan Highway communication system involving the construction of a similar type of line with 128-mil 30 percent conductivity copper-steel, between Edmonton, Alberta (Canada) and Fairbanks, Alaska—a distance of about 2,000 miles.

Both of these lines traverse very rugged terrain and are subject to severe weather conditions, particularly the Alaskan Highway line. In both instances, the information available indicates that copper-steel wire has given satisfactory performance.

Present Trends and Objectives

Since the end of World War II, outstanding progress has been made in the development of multi-channel open-wire carrier systems which are sufficiently flexible and economical for deriving both long-haul and short-haul communication channels. As a result, carrier exploitation of the basic open-wire facility to the maximum degree practicable has become a primary objective in many segments of the communication industry, not only for reasons of economy but also to provide better transmission performance. This has led to the concept of a substantially "all-carrier" open-wire plant.

As the industry works toward the all-carrier concept, service continuity of open-wire conductors becomes increasingly important. Consequently, the provision, among other things, of a maximum of wire strength consistent with economic considerations is desirable. Therefore, the advantages offered by the use of copper-steel wire are of considerable interest.

Relative Strength and Cost

Fig. 1 shows graphically the relative breaking strength of 104- and 128-mil copper and copper-steel wire in terms of 165-mil copper wire, and also their approximate recent material costs per pair-mile. For comparing strength, 165 copper is chosen because it is the largest size wire which has been used to any general extent in telephone plant and is in mechanical equilibrium with the standards used by a large segment of the industry in the mechanical design of the supporting structure; that is, poles, cross-arms, pins, guys, etc.

As indicated, 104 copper-steel has nearly the breaking strength of 165 copper wire; 128 copper-steel is substantially stronger. At the same time copper-steel wire is considerably cheaper. As a result fewer dollars buy more pounds of breaking strength. Under such circumstances there would appear to be little question as to the desirability of employing copper-steel instead of copper wire for new construction or major wire replacements. The practicability of doing this, however, is governed mainly by transmission considerations relating to voice frequency operation.

Relative Losses

Fig. 2 shows the approximate relative losses of copper and the two varieties of copper-steel (30 and 40 percent conductivity) wire. Data are given for 104-mil and 128-mil wire only, since these have become the wire sizes of principal interest in engineering toll open-wire plant. Smaller sizes are ordinarily undesirable for mechanical and transmission reasons; larger sizes are uneconomical for general use and, in the case of copper-steel, their strength is not in equilibrium with that of the ordinary types of line construction.

As indicated in Fig. 2, the losses of copper and 40 percent copper-steel wire do not differ substantially over the major portion of the carrier frequency range. Consequently, 40 percent copper-steel wire has proved to be suitable for carrier operation. Its somewhat greater loss below 30 kc has usually been offset by the availability of adequate carrier terminal or repeater gains, particularly when low slope basic equalizers are used.

The loss of 30 percent copper-steel is also sufficiently close to that of copper wire to permit its use at frequencies above 30 kc without incurring any substantial transmission or economic penalty. However, at 30 kc and lower frequencies, the greater loss of 30 percent copper-steel, in comparison with either copper or 40 percent copper-steel, is one of the factors which makes its use less desirable for toll applications notwithstanding its greater strength. For this reason, further consideration of 30 percent copper-steel wire is omitted in the discussion which follows.

Although copper-steel wire is generally suitable for carrier operation, its use for voice frequency facilities is subject to a number of transmission considerations.

Voice Frequency Message Circuits

Fig. 2 shows that the loss of copper-steel wire is much greater than that of copper in the voice frequency range. In addition, its impedance characteristics

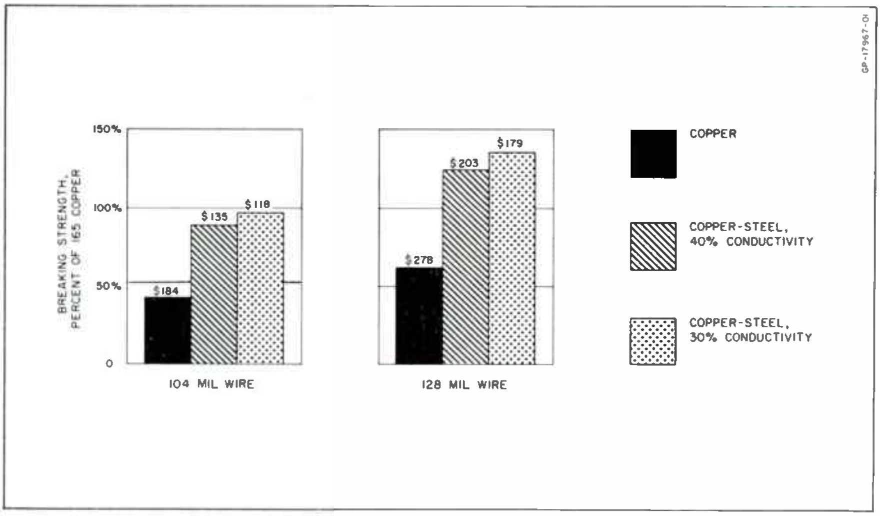


FIG. 1. A comparison of the relative breaking strength of copper and copper-steel wire. The approximate material cost per pair-mile is shown for each type of wire.

are known to be substantially different, particularly below 1 kc. These characteristics tend to make the use of copper-steel wire difficult where repeater layouts and entrance or intermediate cable loading systems based on copper wire are involved.

In situations of this kind and in the case of new open-wire routes, additional repeaters can be provided to offset the greater loss of copper-steel wire. This entails additional expense and the impedance matching problem remains a factor to be reckoned with. Considerations such as these, together with the poorer transmission performance of two-wire voice-repeatered circuits as compared to carrier facilities, make it desirable to dispense with voice repeatered message operation on open-wire pairs to be fully developed for multichannel carrier use whenever it is practicable to do so. This has become a recognized practice with a number of telephone companies.

When voice frequency message circuits are not repeatered, they tend to be so short (frequently less than the minimum economic length for short-haul carrier) that satisfactory transmission can ordinarily be obtained with either type of wire.

Phantoming

Copper-steel wire is somewhat more susceptible to series resistance unbalances than copper wire. These unbalances tend to increase noise, but past experience does not indicate this to be a controlling factor in the use of copper-steel wire for carrier operation.

The tendency of unbalances to cause crosstalk between a phantom and either or both of its component sides limits the operation of phantoms derived from copper-steel wire to the shorter lengths. However, present practices tend to avoid phantoming new open-wire facilities—either copper or copper-steel—whenever it is practicable to do so. This results from their susceptibility to the effects of noise and crosstalk, their restrictive influence on carrier and program operation, and their tendency to add to the cost and complexity of composite set arrangements for dial signaling or d-c telegraph operation.

Program Transmission

The low frequency loss and the impedance characteristics of copper-steel wire also offer some difficulty in program transmission and equalization. These difficulties can usually be overcome by spacing program repeaters at shorter intervals—particularly when new routes are being established—and the use of special equalization methods. Detailed consideration of the particular situation involved may also suggest other more desirable alternatives.

Telephoto

The characteristics of copper-steel wire make it difficult, if not impracticable, to meet the impedance matching requirements involved in telephoto transmission in the voice frequency range. This difficulty could be overcome through the use of impedance matching equipment which could be designed and made available for that specific purpose. However, the economy of this procedure appears questionable since it is feasible to employ carrier channels for telephoto transmission.

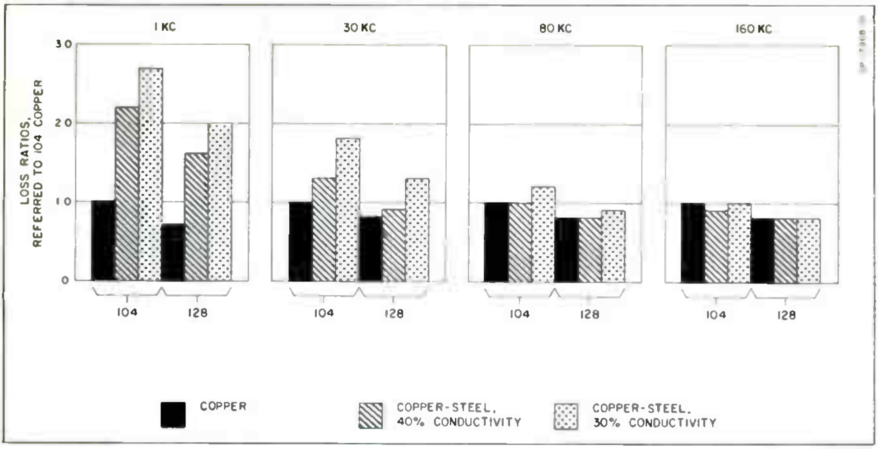


FIG. 2. Relative losses of copper and copper-steel wire for several frequencies.

As a matter of fact, telephoto is presently being operated on carrier channels to a substantial extent and the trend in that direction is increasing. Here again, present thinking emphasizes the use of carrier with a higher degree of mechanical dependability in the basic open-wire facility as compared to a sacrifice in such dependability in favor of voice frequency operation.

Long-Span Construction

The greater strength of copper-steel, as compared to copper wire, offers substantial economy through the use of long-span construction. Lines employing this type of construction can be developed for maximum carrier usage when the average pole spacing is not greater than about 300 feet. In such instances it is, of course, necessary to transpose on adjacent instead of alternate poles and in irregular terrain to employ floating transposition brackets somewhat more frequently than might be necessary with average pole spacings up to about 150 feet. These measures tend to increase transposition costs, but they do not preclude the use of long-span construction for carrier lines when that type of construction is otherwise practicable and results in worthwhile economy.

Conclusion

Because of the extent to which service can be interrupted by the failure of wire employed for multi-channel carrier operation, it is desirable to provide a maximum of strength consistent with economic considerations. Copper-steel wire can be employed successfully for such operation to obtain increased service security at lower material costs.

The use of copper-steel wire for several types of facilities operating in the voice frequency range is subject to some difficulties. These can be avoided through the alternative use of carrier with resulting worthwhile improvements in transmission and little, if any, substantial increase in overall costs. Consequently, voice frequency considerations should not, in general, prevent the use of copper-steel wire, particularly for new construction.

HE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6 NO. 1 JANUARY 1957

CARRIER SYSTEM CO-ORDINATION

Much of the thought that goes into the design of carrier systems is devoted to co-ordination requirements. By properly applying the results of this thought, the carrier system user can plan his installations for the most efficient operation.

This article discusses the basic factors which influence the design and application of co-ordinating carrier systems.

A simple definition of the term co-ordination is: that property of a carrier system which enables it to operate together with other systems without interference. Various FCC rulings and international agreements impose a certain amount of co-ordination on radio systems. The radio-frequency spectrum is public property and these rulings and agreements aim to provide the best use for all users. On the other hand, a wire-line facility is private property. Its most efficient use requires a self-imposed co-ordination.

Carrier system co-ordination is a joint effort of both manufacturer and user. The manufacturer works toward co-ordination by designing his equipment so that it will work together with other systems. The user completes the job by applying the systems to properly engineered facilities.

There are two broad types of co-ordination. One of these may be called *intra-pair co-ordination.* This exists when two or more carrier systems operate on the *same* pair of wires without interfering with each other. The other type may be called *inter-pair co-ordination.* This exists when two or more carrier systems operate on *different* wire pairs without interfering with each other.

Intra-Pair Co-ordination

The basic requirements for intra-pair co-ordination are relatively simple. To operate on the same pair without interference, different systems must operate

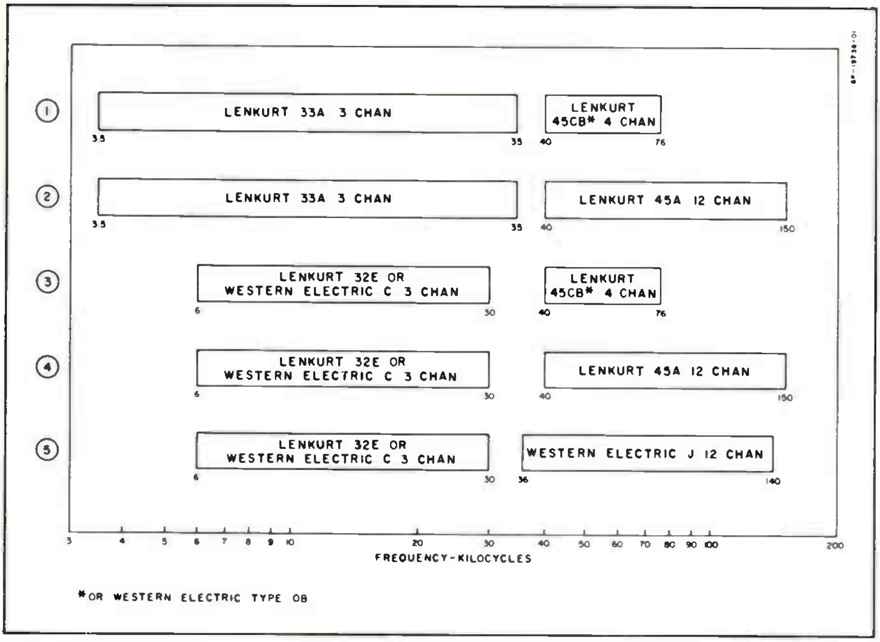


FIG. 1. Five combinations of open-wire carrier systems which co-ordinate on the same pair. All combinations shown may be superposed on physical voice-frequency circuits.

in different frequency ranges and at the proper levels.

The reason for frequency separation involves the first principle of carrier communications. The filters for one system pass only the frequencies assigned to that system and attenuate all others. Therefore, the frequency bands of different systems on the same pair cannot coincide or overlap.

Proper level means simply that one system cannot be so high in level that it leaks into another system through the line or directional filters. However, level difference is seldom a problem in intra-pair co-ordination. Modern, well-designed filters do a very effective job of suppressing out-of-band frequencies.

Carrier systems which co-ordinate on the same pair are available in many combinations. Figure 1 shows five co-ordinating open-wire combinations.

Inter-Pair Co-ordination

When two systems operate on separate wire pairs, their frequencies are physically separated. But a new source of interference arises from the inductive coupling that links paralleling wire pairs on an open-wire line or in a cable. As a result of this coupling, some of the signal energy in one pair transfers to an adjacent pair and may appear as *crosstalk*. When the crosstalk between two systems on separate wire pairs is kept to a satisfactory level, the systems may be said to co-ordinate on an inter-pair basis. The carrier manufacturer takes the first steps toward co-ordination with design that minimizes the effects of crosstalk coupling.

Crosstalk has its greatest effect at points where one system is at a high level and the other system at a low level. Such points exist at terminals and repeaters. The situation at a repeater station is shown in Fig. 2.

Here the west-east output of Repeater A consists of high-level signals. They have just been amplified by the repeater. The signals entering Repeater B in the east-west direction have been attenuated by the line and are at a low level. If the west-east frequency band of System A is the same as the east-west band of System B, the directional filters for the two directions will be the same. The near-end crosstalk from System A would then pass freely through the east-west branch of Repeater B and would be amplified and sent on to the west terminal of System B.

But if the west-east frequency band of System A differs from the east-west

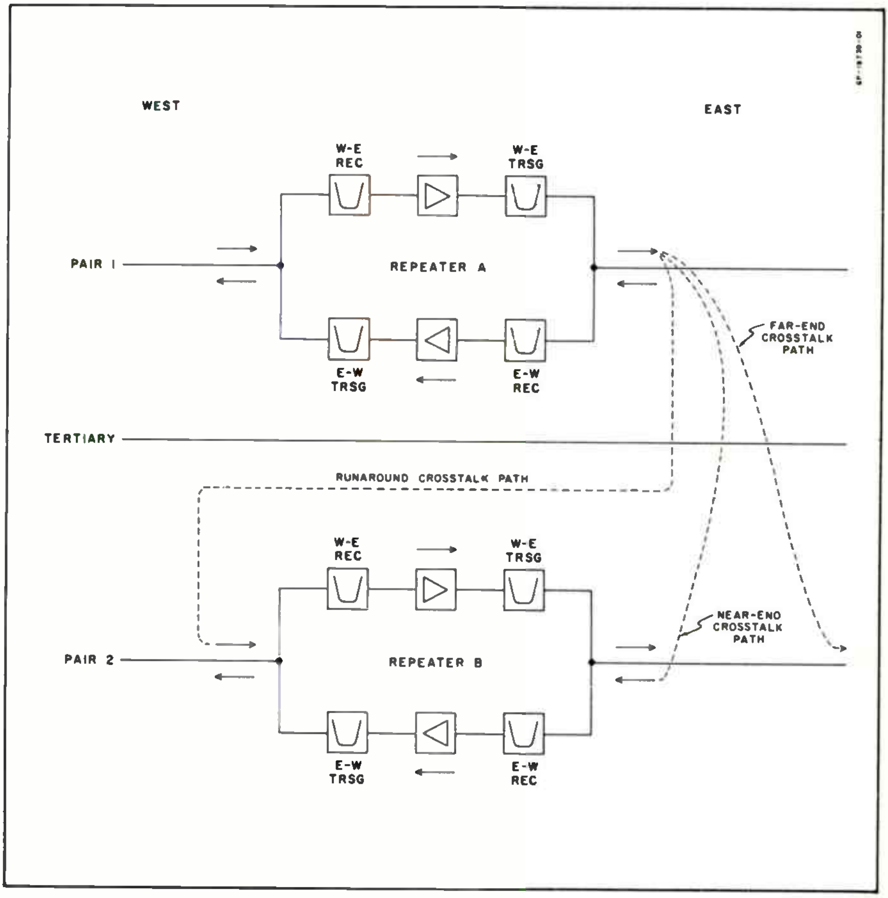


FIG. 2. Simplified block diagram of repeater station showing near-end, far-end and run-around crosstalk paths. Similar paths also exist for east-to-west transmission.

band of System B, the directional filters for these directions will be different. The directional filters of System B will then attenuate the crosstalk to a negligible level. Thus, to reduce near-end crosstalk, inter-pair co-ordination requires a basic frequency relationship between two systems: the east-west band of one system must differ from the west-east band of the other system.

Two-wire carrier systems use "grouped-frequency" operation in which channels in opposite directions use different frequency bands. This allows two carrier systems to co-ordinate on an inter-pair basis even though they both use the same over-all frequency allocation. In such cases, far-end crosstalk is also a factor. This is the crosstalk which is coupled from Pair 1 to Pair 2 (directly or through the tertiary) and is transmitted along Pair 2 toward the east. This crosstalk will be in-band to the west-east branch of the next terminal or repeater. However, after suffering coupling loss and line attenuation, it will arrive at a relatively low level.

Another form of interference that may occur at repeater stations is run-around crosstalk. Run-around crosstalk takes the path shown in Fig. 2. Here, the high-level output of Repeater A crosstalks to a tertiary, and the tertiary crosstalks to the west input of Repeater B. If the two systems use the same frequency allocations, the run-around crosstalk will be in-band to the west-east branch of Repeater B. In such cases, additional attenuation in the form of crosstalk suppression filters may have to be inserted in the run-around crosstalk path.

Another basic requirement of inter-pair co-ordination is co-ordination of energy levels. One system cannot be so high in level that its crosstalk on another pair is strong enough to cause objectionable interference. Any original difference in levels between sys-

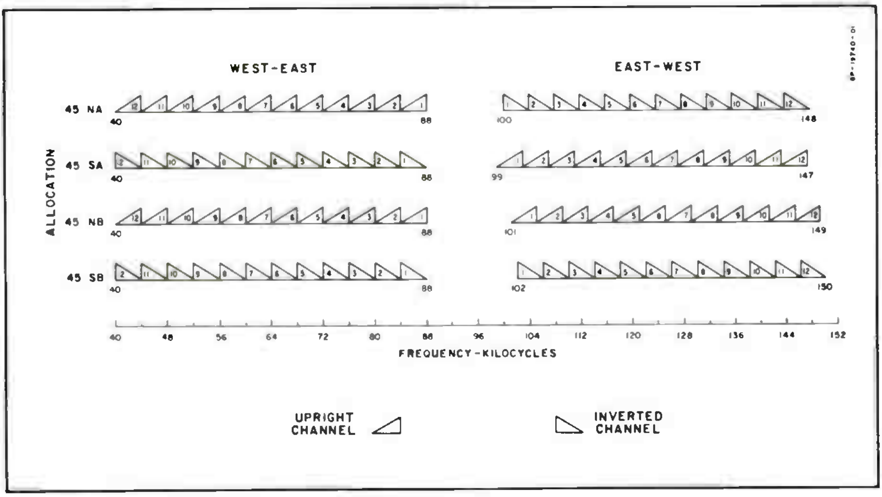


FIG. 3. Four frequency allocations for Lenkurt 45A open-wire system, showing channel inversion and staggering.

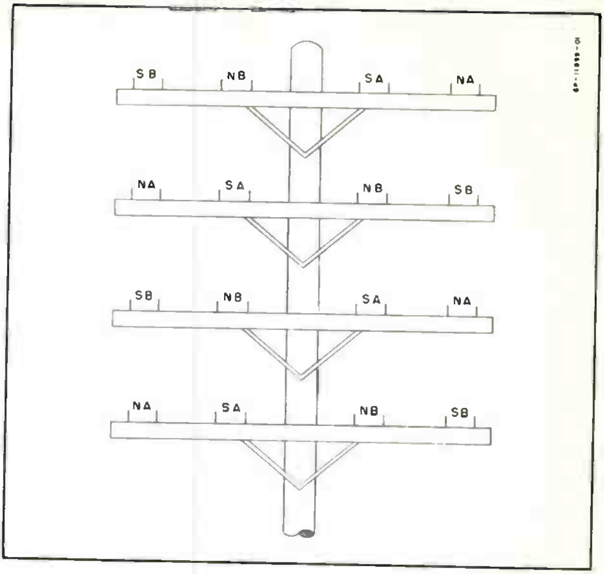


FIG. 4. Recommended pair assignments for Type 45A frequency allocations.

tems will only widen the difference that exists where one system is at a high level after amplification and the other is at a low level after line attenuation. Therefore, an important design step in co-ordination is keeping the transmitting level of the system as close as possible to the transmitting levels of the systems with which it is to co-ordinate.

But message power is not the only consideration in level co-ordination. The levels of signaling tones and carrier frequencies are also important. The carrier designer must plan his system so that signaling and pilot tones are at levels which do not interfere with other systems. When the carrier is transmitted, it must not be so high as to intrude on other systems. In systems where the carrier is not transmitted, the carrier must be sufficiently suppressed in the modulators and filters to prevent interference from carrier leak.

Level co-ordination is closely related to filter performance. Since filters are usually the ultimate stages in determining how much interference gets into the disturbed system, their design is important. Filters which offer high attenuation to out-of-band frequencies can ease the level co-ordination requirements for near-end crosstalk when two systems use the same frequency allocations, or for both near-end and far-end crosstalk when two systems use different allocations.

The carrier manufacturer can supplement his basic co-ordination efforts by designing his system to allow for different channel allocations. Figure 3 shows the various allocations of the Lenkurt 45A open-wire system. By proper choice of sidebands, the individual channels can be upright (upper sideband) or inverted (lower sideband). By a proper choice of carrier frequencies in the last stage of modulation, the allocations can be staggered. The facility can then be laid out so that adjacent pairs use systems with slightly different frequency allocations.

Channel inversion safeguards the

privacy of individual conversations. The signals which crosstalk from one pair to an adjacent pair are unintelligible. Staggering, on the other hand, reduces the effect of this unintelligible crosstalk. Crosstalk from one channel falls in a different portion of an adjacent disturbed channel and much of it will fall in the unused guard bands of the disturbed system.

Carrier designers can also further co-ordination by reducing the reflection effects of the system. Reflection occurs when signal energy meets an impedance mismatch in the transmission path. The reflected energy may then combine with crosstalk from other sources to increase the total crosstalk between systems. This reflected component of crosstalk can be controlled by closely matching the impedances of terminals and repeaters with the impedance of the transmission line.

Application of Co-ordinating Systems

All the effort that goes into the design of co-ordinating carrier systems is lost unless the systems are properly applied. This is where the user contributes his share to co-ordination. He can get the most efficient use out of the manufacturer's equipment only by applying it to properly engineered facilities.

One of the first considerations is line treatment. This means transposing, in the case of open-wire facilities, or capacitive balancing, in the case of cable facilities. Crosstalk increases with frequency. If new systems are to be added to existing facilities, the line treatment must be adequate to handle the crosstalk of the new frequencies.

In general, lines treated for a lower frequency will have excessive crosstalk coupling at a higher frequency. But compandors often allow a wide degree of leeway in this rule and their use should be considered. The 22-db crosstalk advantage between compandored systems has, in specific cases, allowed the use of frequencies as high as 150 kc on paralleling open-wire pairs transposed for only 30 kc.

Suppression measures are another consideration. Excessive crosstalk between two systems via a tertiary may be reduced to tolerable levels by the insertion of additional suppression in the tertiary path. As in the case of run-around crosstalk, crosstalk suppression filters may bring the interference under control.

The user can also contribute to co-ordination by keeping line reflections to a minimum. He does this by effectively reducing impedance mismatches in the line. This involves the proper matching of impedances at junctions where terminals and repeaters connect to the line or where open-wire connects to other open-wire or to cable.

Conclusion

The operating efficiency of a facility is largely the product of the combined co-ordination efforts of the carrier system manufacturer and the carrier system user. The manufacturer contributes by building certain co-ordinating features into his equipment. The user contributes by a proper choice of systems and proper engineering of the transmission path. But the goal of both manufacturer and user is the same—quality transmission at the lowest cost per channel-mile.

THE *Lenkurt* Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6, NO. 5 MAY, 1957

Crosstalk Advantages of

FREQUENCY INVERSION AND STAGGERING

The carrier systems on different pairs on an open-wire line usually operate in the same general frequency range. In such cases, transposing and co-ordinating signal levels can effectively reduce crosstalk between paralleling wire pairs. When systems of the same type are used on different pairs, the effect of crosstalk can be further reduced by the relatively simple techniques of frequency inversion and staggering.

Figure 1 shows the four frequency allocations for the Lenkurt Type 45A open-wire carrier system. By a proper choice of carrier-frequencies in the channel modulators and demodulators, the channels, as fed to the transmitting line, may be either upright (upper sideband) or inverted (lower sideband). Another choice of carrier-frequency for the final modulation stage permits the east-west band to be placed in any one of four slightly different locations in the frequency spectrum. This causes the east-west channels of one allocation to be staggered in frequency with respect to the east-west channels of another allocation.

In practice, open-wire lines using Type 45A systems are laid out so that adjacent pairs are equipped with systems having different frequency allocations. The recommended pair assignments are shown in Fig. 2.

The crosstalk coupling between two pairs is a function of the distance between them and their geometrical configuration. With the arrangement of Fig. 2, the pairs having the greatest coupling between them carry channels which are inverted with respect to each other. This means that any direct crosstalk between two such pairs will also be inverted in frequency.

Inverted crosstalk is unintelligible in the disturbed channel and therefore safeguards the privacy of individual telephone conversations. But it also has another important effect. Unintelligible crosstalk appears in the disturbed circuit as noise and, at levels tolerable in good practice, has less of an inter-

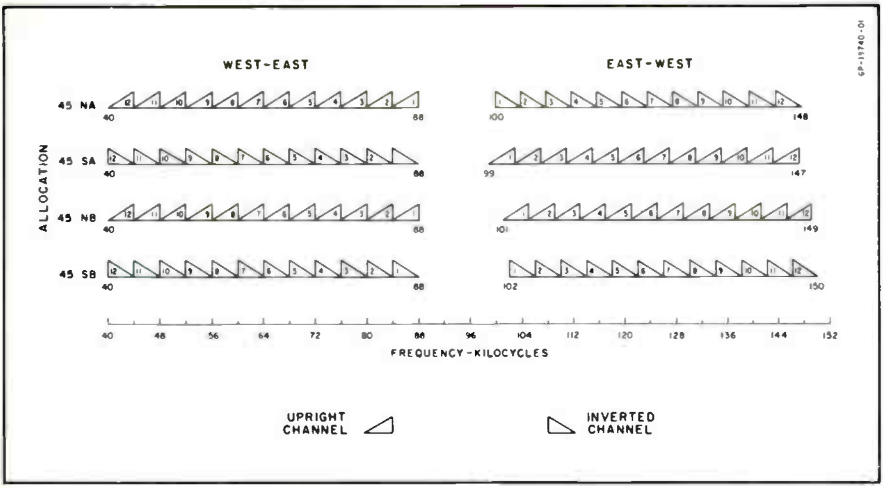


Fig. 1. Four frequency allocations for Type 45A system showing channel inversion and frequency staggering.

fering effect than intelligible crosstalk. For this reason, channel inversion alone is often considered to have the same effect as increasing the crosstalk coupling loss between two circuits by about 3 db.

Frequency staggering has a similar effect. Figure 3 shows the relative line frequencies for several channels of the NA and SA allocations of the Type 45A system. The bandwidths shown are the regions between the 3-db points of typical channel filters. The result of shifting the frequency of the over-all band is that no channel in one system coincides entirely in frequency with any other channel in the other system.

The shaded areas of Fig. 3 show the ranges of frequencies in which crosstalk could occur. Except for the end channels, crosstalk from one channel in the disturbing system will fall in two channels of the disturbed system. But much of it will also fall in the unused guard bands between channels and so will not appear at the output of the disturbed system. The effective staggering advan-

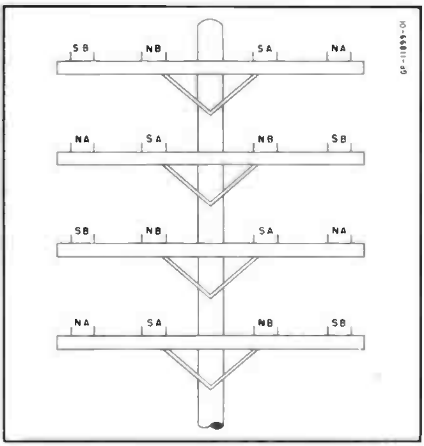


Fig. 2. Recommended pair assignments for 45A systems.

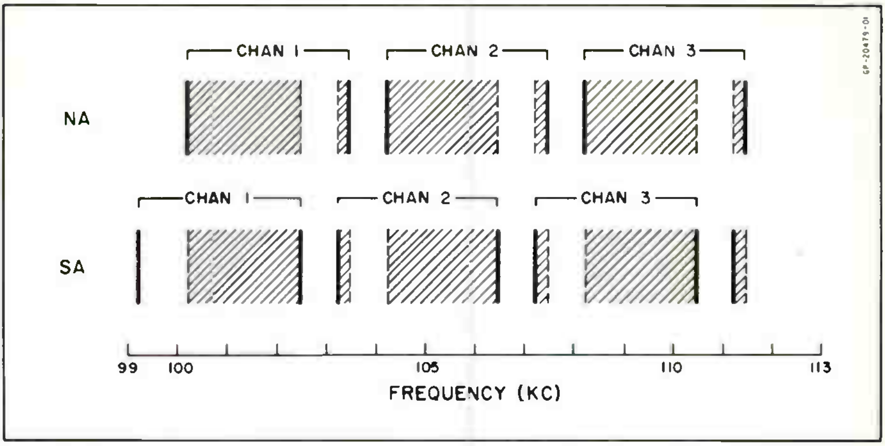


Fig. 3. Effect of frequency staggering. Only crosstalk at frequencies within shaded areas can appear in the disturbed channels.

tage of Type 45A systems can amount to as much as a 9-db increase in crosstalk coupling loss between pairs.

On open-wire lines equipped with Type 45A systems, frequency inversion and staggering are two of the most easily applied countermeasures against crosstalk. For example, in the 45A system, whether a channel is upright or inverted is determined by the carrier-frequency used in the channel modulator and demodulator. Since all the frequencies required are available at the carrier terminal, a system may be arranged for either upright or inverted sidebands simply by the proper distribution of carrier frequencies at the terminal. No changes in filters are necessary. Frequency staggering is determined by a plug-in carrier distribution connector and the proper plug-in crystal for the high-frequency oscillator.

THE Lenkurt Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 5 NO. 9 SEPTEMBER, 1956

TRANSPOSITIONS

For Open-Wire Lines

Crosstalk caused by inductive coupling is one of the problems involved in engineering open-wire carrier facilities. An effective solution in general use is the practice of transposition.

This article describes the basic causes of inductive crosstalk and discusses some of the general considerations pertaining to the theory and application of transposition arrangements.

Crosstalk in open-wire carrier facilities results from the inductive couplings which exist between paralleling wire pairs. This crosstalk can be reduced to tolerable dimensions by a method of line treatment known as *transposing* in which the pin positions of the wires of each pair are interchanged (transposed) at systematic intervals along the length of the facility.

Because of the many possible interactions among various wire pairs, the technique of transposition design tends to be involved, particularly where carrier operation is concerned. However, the extra engineering effort and construction costs of good transposition practice represent a sound investment in performance.

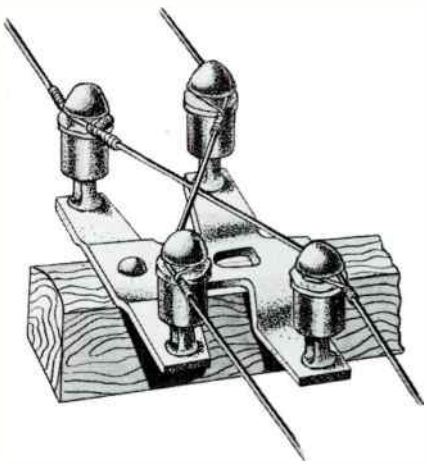


FIG. 1. Point-type transposition bracket. This is one of several methods used for interchanging pin positions of a pair of wires.

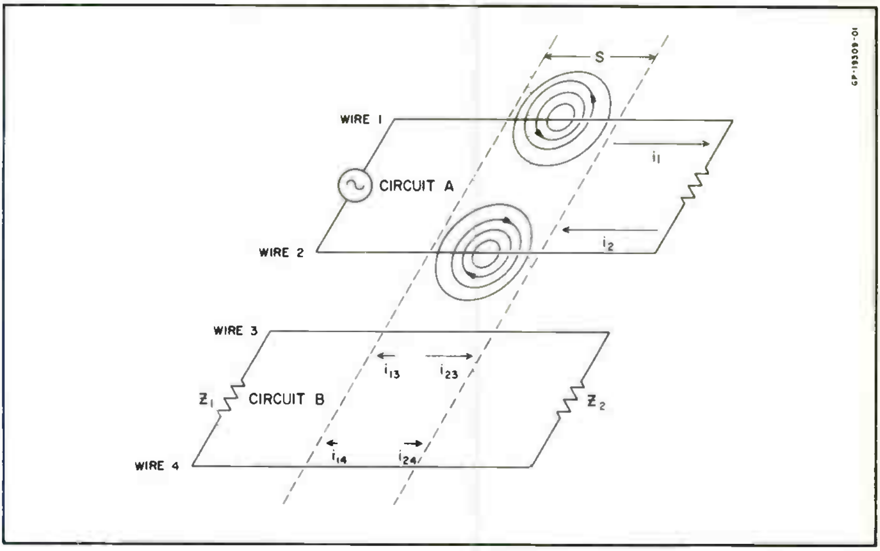


FIG. 2. Crosstalk by induction. Circles about wires 1 and 2 represent lines of magnetic induction and equipotential lines for a particular instant.

Source of Crosstalk

Inductive crosstalk between two wire pairs arises from the magnetic and electric fields that are set up when one or both of the pairs carry an alternating current. Although these fields differ somewhat in their actions in producing crosstalk, the over-all results of both are quite similar.

Figure 2 shows two pairs of telephone wires represented as two adjacent circuits. In a short segment, S, the alternating current in circuit A will set up an expanding and contracting magnetic field. The lines of this field will cut wires 3 and 4 of circuit B and induce voltages in them. These voltages, in turn, cause currents to flow in circuit B as shown in the figure. The induced currents are numbered to indicate that I_{13} is the current induced by wire 1 in wire 3, I_{14} is the current induced by wire 1 in wire 4, and so on.

Because of the differences in spacing between the wires and the directions of the inducing currents, the induced currents will differ in magnitude and phase. As a result, a small net current, the algebraic sum of all the induced currents, will circulate in circuit B. This resultant induced current, alternating at the frequency of the inducing current in circuit A, will set up near-end crosstalk voltages across Z_1 and far-end crosstalk voltages across Z_2.

The lines of magnetic induction of Fig. 2 may also be regarded as the equipotential lines of the electric field. This field will set up potentials on wires 3 and 4 which will not be equal. The resulting difference in potential will cause crosstalk currents to flow toward both ends of circuit B.

The short section S may be thought of as one of an infinite number of such sections contained along the length of

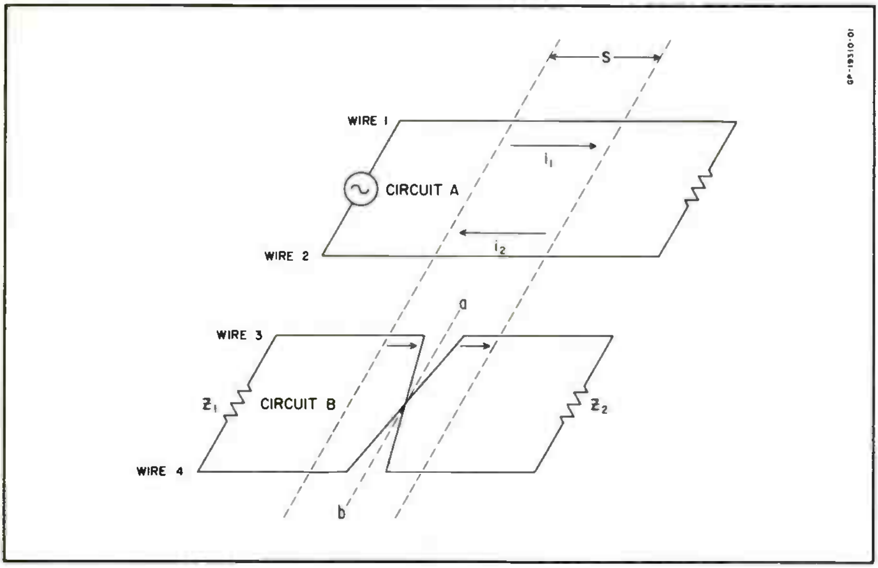


FIG. 3. Effect of transposing circuit B. Net induced current is made up of two components which oppose each other in circuit B.

the line, each contributing a share of the total crosstalk. In a similar manner, one circuit may crosstalk indirectly into another when the currents of the disturbing circuit induce currents in a third circuit (tertiary) which, in turn, induce currents in the disturbed circuit.

Principle of Transpositions

The basic principle of transpositions is shown in Fig. 3. Here the situation of Fig. 2 is modified by transposing the wires of circuit B at the center of segment S. The net induced current is shown as two components, one on each side of the line a-b. The transposition has now made these two components flow in opposite directions in circuit B so that they counteract each other.

The same effect would result from transposing circuit A instead of circuit B. If, however, both circuits were transposed at the same point, no relative transposition would exist between them and the original crosstalk situation of Fig. 2 would prevail.

Because of the propagation effects of line attenuation and phase change, the two components of induced currents in the transposed circuit B will not entirely cancel each other. Attenuation causes the inducing current to decrease in magnitude as it travels from the source toward the load. As a result, the induced currents to the left of line a-b will be greater than the induced currents to the right of a-b.

Phase change causes the instantaneous amplitude and direction of the inducing current to vary throughout each wavelength along the line, resulting in other inequalities. As a result of these combined propagation effects, a small residual induced current remains uncorrected by the transposition and the crosstalk which this current produces is known as *type unbalance* crosstalk.

Another component of the total crosstalk in an open-wire facility is introduced by structural irregularities. Lack of uniformity in wire sag, pole spacing or wire spacing throughout the length of the circuits will cause random unbalances. The crosstalk resulting from such factors is random in nature and is known as *irregularity* crosstalk.

By transposing the line a number of times within each wavelength, type unbalance crosstalk can be considerably reduced. Careful construction will hold irregularity crosstalk to a minimum. The over-all effect of a properly transposed and well-constructed open-wire facility is the reduction of crosstalk to a point where it is not a controlling transmission limitation.

Transposition Arrangements

The basic building block of a transposition plan is the *transposition section*. A transposition section is a segment of line of arbitrary length in which individual pairs have been systematically transposed, each in accordance with a definite pattern. The patterns chosen are those which will limit the crosstalk between any two pairs to a value which is within the design objectives of the system. Transposition sections are then connected in tandem to make up the total length of the facility.

The actual length of a transposition section is chosen by taking into account the frequency of operation and the crosstalk tendency of the patterns used. Ideally, it is a length which meets the crosstalk objectives of the system for all pairs using the minimum number of transpositions. In practice, transposition sections are usually designed first for the longest lengths practicable. Shorter sections are then designed to provide for situations where these longer sections cannot be used.

Transposition patterns have been standardized and classified by the Bell Telephone System into 32 basic types. A few of the simpler patterns for typical transposition sections are shown in Fig. 4. The most complex of the 32 basic patterns (type a, not shown) has 31 transpositions within one section length.

When more than 31 transpositions per section are required, *extra* types of the 32 basic types may be formed by dividing the transposition section into 32 equal segments called *intervals* and inserting one or more transpositions within each interval.

Extra types are designated by the letter of the basic type and a subscript numeral indicating the number of transpositions within each of the 32 intervals. For example, Type M_1 is a basic type M pattern with its three between-interval transpositions per section plus one transposition within each of the 32 intervals to give a total of 35 transpositions per section.

Any two pairs are transposed in relation to each other only at points where one pair is transposed and the other is not. The relationship existing between any two transposed pairs is expressed as a *relative* type. Thus, for example, the relative of a type O pair and a type N pair is type M and the two pairs are said to have an M *exposure*.

The design of an actual transposition section for an open-wire carrier facility begins with a tabulation of the requirements to be met. These are factors such as over-all length, maximum length of repeater sections, frequency of operation, wire configuration, wire gauge, and crosstalk objectives of the system.

The next step is the determination of the degree of crosstalk coupling existing between the various pairs on an untransposed basis. This is expressed as a *crosstalk coefficient* in units of crosstalk per mile per kilocycle and is a function of the configuration, gauge and material of the wires. Crosstalk coefficients may be determined from measurements, computations, or, for the more common configurations, from tables.

Equipped with this information, the transposition designer begins to lay out a transposition arrangement by choosing a pattern type for each pair of wires and a transposition section length. The pattern for each pair must be carefully selected so that the final arrangement will contain no two pairs which exceed the design objectives in crosstalk coupling, either directly or through a tertiary.

The procedure of choosing the pattern types is largely a trial and error method. For each tentative plan, the crosstalk (both type unbalance and irregularity) must be computed to ascertain that the relative types between all possible pair combinations are satisfactory for achieving design objectives. Whenever computations reveal that the crosstalk requirements are not met for any combination of pairs, the plan must be revised and the effects computed again. Fortunately, much of the data necessary for these computations is available in the form of tables and graphs.

Experience in transposition designing is a very valuable asset in this phase of the procedure. The seasoned designer often recognizes the pattern types which are not likely to meet his objectives. In general, type unbalance cross-

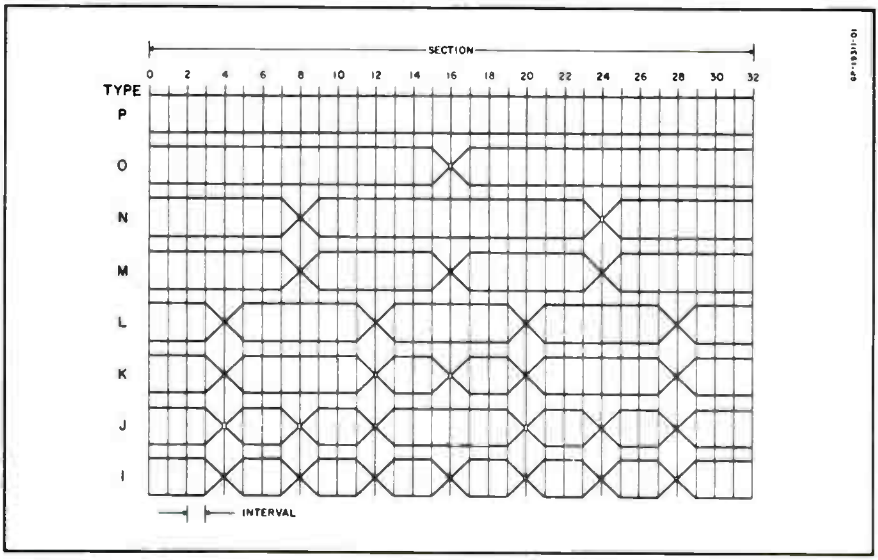


FIG. 4. Several of the 32 basic transposition pattern types.

talk will decrease with the number of transpositions in a section. Therefore, a good starting approach is to choose pattern types which give good crosstalk reduction (low type unbalance) for pairs with tighter coupling (higher crosstalk coefficient) and save the less effective types for pairs with looser couplings. In this manner, it is often possible to eliminate from consideration large blocks of types before the first tentative plan is chosen.

Throughout his analyses, the transposition designer must not lose sight of the relative importance of type unbalance versus irregularity crosstalk in the facility. Care must be taken not to overdesign from the standpoint of number of transpositions used. Transposition will not reduce the crosstalk due to random irregularities. Therefore, it is useless to transpose beyond a point where such crosstalk is controlling. In addition to increasing the cost, too many transpositions may even increase the crosstalk by introducing additional structural irregularities.

General

In applying transposition arrangements, it is important to keep in mind the possible future uses of the facility. The cost of transposing new wire is in the order of 30 per cent less than the cost of transposing existing wire on an out-of-service basis. Therefore, reasonable foresight in anticipating and transposing for future requirements may often result in substantial long-range savings.

The cost of transposing an open-wire carrier facility is frequently a large portion of the cost of the over-all project. However, this cost is outweighed by the advantages gained in the performance and economies of carrier. Therefore, transposing should be regarded not as an evil to be avoided but rather as a bargain in channel-miles of transmission performance.

THE *Lenkurt* Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 4 NO. 2 FEBRUARY 1955

CABLE TRANSMISSION CHARACTERISTICS

In the Carrier Frequency Range

In a previous issue of the Demodulator (October, 1953), some of the basic electrical properties common to all transmission lines were examined and discussed to show how they affect carrier transmission. In this issue, certain characteristics of multi-pair cable are described in greater detail to show how they influence the application of carrier transmission systems to such cable.

All two-wire transmission lines have four fundamental properties in common.

These are:
1. Series resistance.
2. Series inductance.
3. Shunt resistance (or conductance).
4. Shunt capacitance.

The values these properties assume for any particular transmission line depend primarily on the physical configuration (wire size, wire spacing, insulation, etc.) of the line. To some extent they also depend on other factors such as weather, temperature, and frequency. The performance of a transmission line, in turn, depends on how closely the relationships between these properties approach that ideal relationship which provides a low-loss distortionless line. This ideal relationship is realized when, for every frequency of interest, the ratio of the series inductance to the shunt capacitance is equal to the ratio of the series resistance to the conductance.

The theoretically ideal transmission line can be approached by a carefully constructed open-wire pair or coaxial cable. These two types of lines can be built with wire sizes and spacings designed for minimum attenuation and low distortion. The physical restrictions placed on multi-pair cable, however, are such that it is very difficult to manufacture such cable with transmission characteristics at carrier frequencies that even remotely approach those of open-wire or coaxial cable. Moreover, when many of the existing cables in the telephone plant were installed, their use at carrier frequencies was not contemplated. Because of these factors, the application of carrier to multi-pair cable is much different from simi-

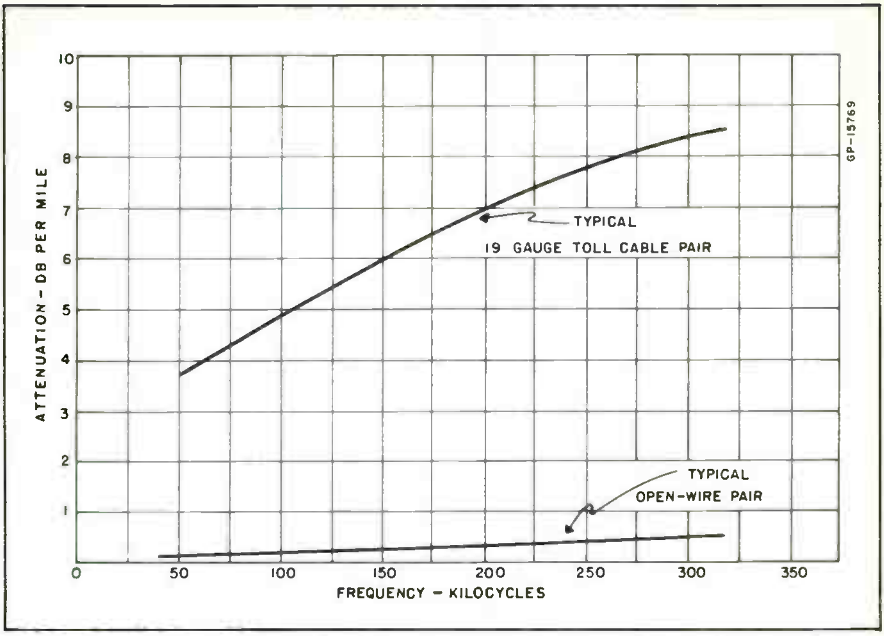


FIGURE 1. Attenuation at carrier frequencies is much greater in cable than in open wire.

lar applications to open-wire line, coaxial cable or radio.

Types of Cable

Although all types of multi-pair cable are subject to the basic restrictions of size and weight, there is a wide range of qualities and types in use. For the purposes of this article, all cable can be divided into two main classifications: exchange cable and toll cable. Exchange cable is often used for short inter-exchange office trunks where large numbers of circuits are required. Usually the lengths of exchange cables are so short that very small wire sizes can be used without too much degradation of transmission. Also, balance and uniform insulation of individual pairs are not critical.

Toll cable, on the other hand, is designed primarily for the long distance transmission of voice and carrier frequencies. Much greater care is taken in the manufacture of toll cable to insure uniform characteristics and minimum losses. Although all cable pairs are twisted to reduce noise and crosstalk, most toll cables are also quadded to further improve their electrical characteristics. A quad consists of two twisted pairs that are further twisted together to form a four-conductor group. The twisting and quadding of cable pairs has much the same effect on cable performance as transposition has on open-wire line. Attenuation, noise, and crosstalk are reduced because external fields are balanced out by the twisting process.

Cable Characteristics

The specific physical characteristics of both exchange and toll cable that affect the design and operation of cable carrier systems include:

1. Non-uniformity in spacing of wires and in distances between wires and sheath.

2. Small wire size.

3. Close spacing between individual wires and between pairs.

4. Relatively large changes in some electrical characteristics with normal changes in temperature.

These physical characteristics tend to cause high attenuation, wide and non-uniform variations of attenuation with frequency and temperature, and impedance variations. They also tend to increase the difficulty of noise and crosstalk control. A comparison between the attenuation characteristics of an open-wire pair and a 19-gauge pair of a quadded toll cable is shown in Figure 1.

The high attenuation of carrier frequencies in multi-pair cable is caused primarily by the small wire sizes and short leakage paths between conductors. In open wire lines and coaxial cable, conductors can usually be made large enough to keep the series resistance low. In toll or exchange cable, however, the necessity of keeping the total size and weight as low as possible requires a compromise between wire sizes that give low attenuation and wire sizes small enough for compact, lightweight cables.

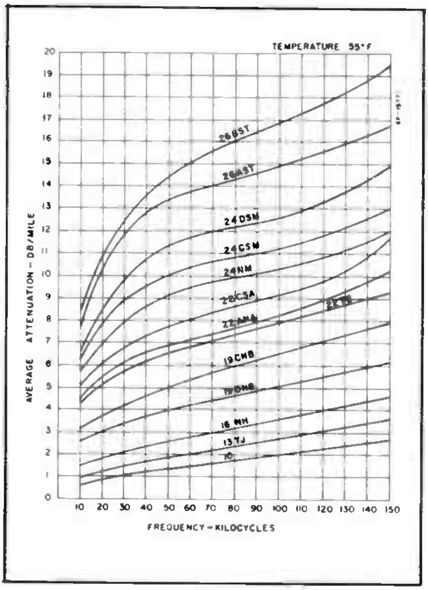
In general, wire sizes range from 10 gauge to 26 gauge depending on the type of cable. The most commonly used wire size in toll cable is 19 gauge; the smallest size usually considered practical for carrier transmission is 24-gauge. Figure 2 shows the effect of different conductor sizes on cable attenuation.

The type of line insulation has as much bearing on attenuation as does the size of the conductors. In open-wire lines and high quality coaxial cable, the dielectric (insulation) is usually air with solid insulators used only at regular intervals along the line for physical support of the conductors. Since air is one of the best insulating materials and has very low loss, these types of transmission lines have very low attenuation per unit of length. In multi-pair cable, however, the insulation is usually of paper or plastic which have relatively high losses compared to air, especially at carrier frequencies.

Several other important cable characteristics, besides high attenuation, that affect the design and operation of carrier systems are shown in the graphs of Figure 2. These include the marked increase in attenuation with frequency and the non-uniformity of attenuation increase with frequency. The general increase of attenuation with frequency is usually referred to as "slope" while the non-uniformity of the increase is commonly called "bulge".

Slope and bulge are characteristic of all transmission lines, open-wire as well as coaxial and multi-pair cable. They are more pronounced, however, in multi-

FIGURE 2. Cable carrier systems must be capable of operating over a wide range of cable characteristics



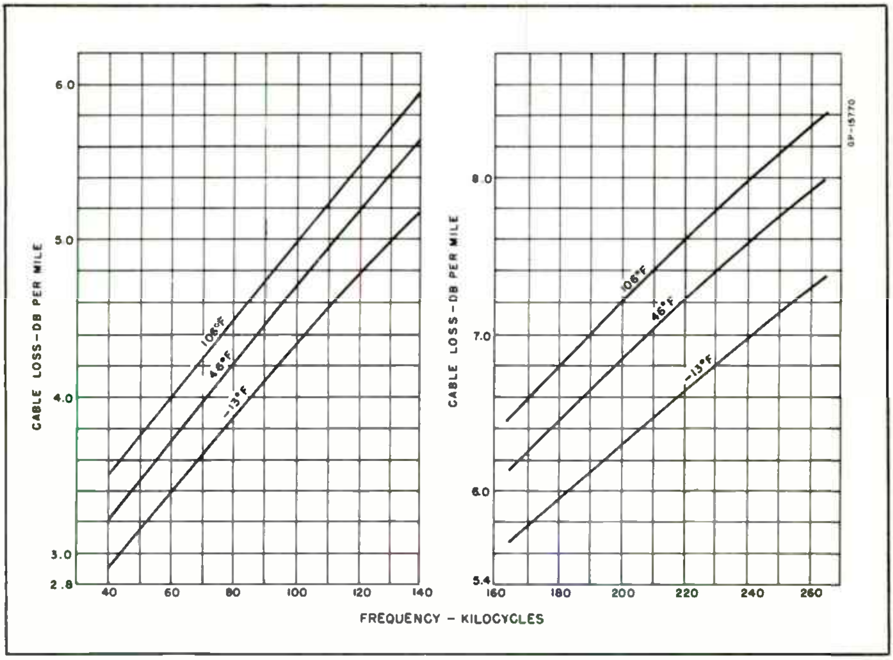


FIGURE 3. Typical variation of attenuation with temperature of a 19 gauge toll cable pair over the frequency ranges used in Lenkurt 45BN cable carrier.

pair cable. Slope compensation alone is possible by several different methods. Open-wire carrier systems usually make use of received regulating tones that adjust the gain-frequency characteristic of the terminal to compensate for slope caused by the transmission line. Another method involves the use of fixed equalizers that introduce a predetermined amount of negative slope across the transmission band. A third method, commonly used in modern cable carrier systems, is the inversion of frequency bands by modulation at each repeater point. In this method (usually referred to as frequency frogging), a channel occupying the lowest frequency space in one repeater section is inverted by the repeater to occupy the highest frequency space in the next repeater section. Thus, if two adjacent repeater sections are identical, the frequency frogging process will largely equalize the slope incurred in each section.

The frequency frogging process, though equalizing much of the slope in a cable characteristic, does not compensate for bulge. Except for very long systems, however, bulge has little effect on overall transmission. In carrier designed for long-haul use, special equalizers and regulating equipment are usually employed to eliminate bulge. In short haul systems, however, individual channel regulation corrects for most of the accumulated bulge.

Slope and bulge are variations of attenuation caused by variations of the fundamental properties of a transmission line with frequency. Another factor, temperature, also has a very pronounced effect on the attenuation of cable pairs. The amount of temperature variation that can be expected for a particular cable, depends on the climate and type of construction. If the cable is underground or under water, variations will tend to be seasonal and of smaller range than if the cable is strung on poles. Aerial cable, under certain climatic conditions, may have internal temperatures that range from below zero degrees Fahrenheit at night to over one hundred degrees Fahrenheit in the afternoon sun. Figure 3 shows typical variations of attenuation with temperature for the frequency bands normally used for carrier transmission.

A secondary effect of the variation of attenuation with temperature is the change in slope of the attenuation characteristic with change in line temperature. This variation of slope (usually called "twist") is most noticeable at frequencies below 40 kc. Above 40 kc, variation of slope with temperature is slight. In long haul, low frequency cable carrier systems, cumulative twist requires special compensating amplifiers to reduce distortion from this source. However, in modern short haul types of cable carrier operating above 40 kc, twist compensation is unnecessary.

Noise in Cables

Although cable carrier circuits are shielded from direct external induction by a metallic sheath, they are nevertheless subject to certain types of carrier frequency noise. The sources of carrier frequency noises that occur in cables are:

1. Unsoldered or poorly soldered cable splices.

2. Telegraph, dial signaling, and relay transients as well as other office generated noise voltages.

3. Atmospheric disturbance and radio transmitters.

4. Interchannel modulation and stray tones from carrier systems.

Splices made in cable pairs used for voice frequencies are usually unsoldered. While such splices normally form a good joint for voice frequency currents, they often generate carrier frequency noise voltages that may make the pair unsuitable for the application of carrier. A steady transmission of direct-current over the pair is often used to improve contact at such joints and thus reduce the noise.

Carrier frequency plant noise such as caused by telegraph, dial signaling, relays, and switches is almost always present in cable carrier circuits. This noise enters the carrier circuit by first being induced into office wiring of non-carrier pairs and then being transferred again by induction to carrier pairs within the cable sheath. Reduction of this type of noise is possible through the use of special suppression coils in the non-carrier pairs; however, a more routine approach is the use of short repeater sections adjacent to the noisy office so that received levels will be high enough to override the noise.

Noise from atmospheric disturbances and low-frequency radio transmitters enters cable carrier circuits in much the same manner as office noise. Noncarrier pairs in a cable often are connected to open-wire lines or non-shielded subscriber drops which act like radio antennas in receiving noise from external sources. Coupling between carrier and non-carrier pairs within the cable, transfers the noise to the carrier circuit. The same suppression measures that are effective in reducing office noise are also effective in reducing atmospheric noise or interference from external sources.

Carrier systems generate a certain amount of internal noise. Interchannel modulation, signaling tone leaks, and tube noise are all

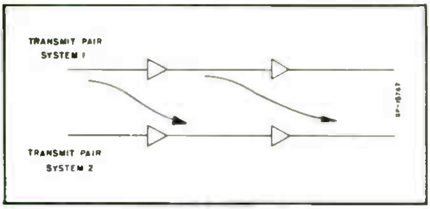


FIGURE 4. Far-end transverse crosstalk is caused by direct coupling between pairs of different systems transmitting in the same direction.

present in cable carrier systems.

The total noise that can be expected in a large number of exchange and toll cable pairs is large enough that special noise reduction techniques are a basic part of cable carrier engineering. Elaborate transposition and suppression schemes were used on early cable carrier systems to reduce noise to acceptable levels. In modern short-haul carrier systems, such expensive methods of reducing noise would make carrier uneconomical. Less expensive, yet perfectly adequate for short and medium haul carrier circuits, is a compandor used with the carrier channels which reduces the interfering effect of noise by as much as 22 db. This amount of reduction is sufficient to make the majority of exchange and toll cables suitable for carrier transmission over distances of 200 or more miles.

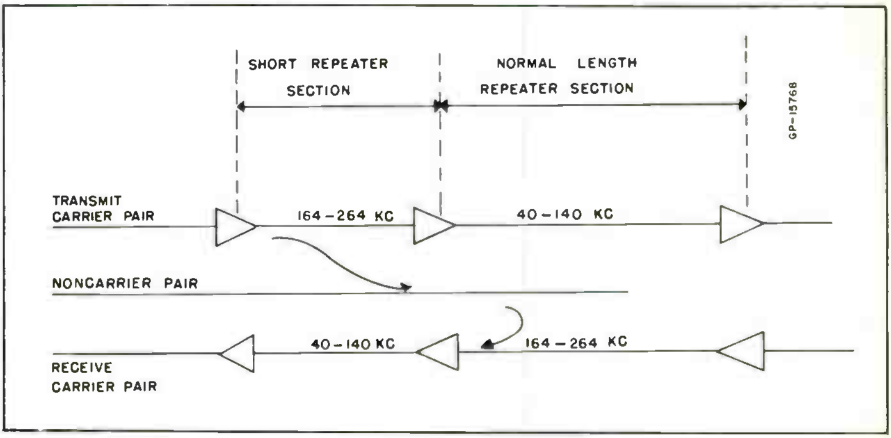
Crosstalk

Crosstalk problems in cable carrier systems are similar to crosstalk problems in open-wire carrier. For both types of systems, the close proximity of paralleling pairs creates the possibility of low-loss crosstalk paths. Slight unbalances in the electric and magnetic couplings between pairs and minor impedance variations are also contributing factors.

Two basic types of crosstalk occur between cable carrier systems operating within the same sheath: transverse crosstalk and near-end interaction crosstalk. Far-end transverse crosstalk is caused by direct coupling between paralleling systems. Near-end interaction crosstalk is caused by coupling between a carrier pair and a noncarrier pair which, in turn, is coupled to a second carrier pair in a different repeater section. This type of crosstalk is only of importance when adjacent repeater sections are of different lengths. Both types of crosstalk are illustrated in Figures 4 and 5.

Several methods are used to control the effect of crosstalk in

FIGURE 5. Near-end interaction crosstalk is caused by a noncarrier pair acting as a coupling link between a transmit pair in one repeater section and a receive pair in an adjacent section.



modern cable carrier systems. These methods include:

1. The use of compandors which reduce the effect of cross-talk by about 22 db.
2. Frequency frogging which eliminates far-end interaction crosstalk common in open-wire carrier repeater installations.
3. The use of separate pairs for opposite directions of transmission.
4. The use of different frequencies for opposite directions of transmission. This eliminates near-end transverse crosstalk as a problem.
5. Generation of masking noise to cover up low-level intelligible crosstalk.
6. Installation of improved cable types in new construction.

When all these measures are taken to reduce the effect of cross-talk, almost 100 percent of the pairs in a good quality toll cable can be used for cable carrier without excessive interference between systems.

Conclusions

Several recent advances in carrier techniques have made possible the economical application of short-haul carrier to multi-pair cable for distances as short as 15 miles. The use of compandors, frequency frogging repeaters, and low-cost channelizing equipment can multiply the message capacity of existing cables up to 12 times at less cost than the installation of additional cable.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 3 NO. 12 DECEMBER, 1954

Some Factors Affecting OPEN WIRE AND CABLE EXTENSIONS for Radio Carrier Systems

Radio equipment and carrier equipment for a multichannel radio system must often be installed in different locations with the multiplexed channels transmitted between them at carrier frequencies. Because radio carrier systems are designed to operate on a 4-wire basis special considerations must be taken when they are applied to open wire or cable. In some cases the carrier frequencies can be transmitted over open wire or cable directly from the radio receiver to the carrier equipment. In other cases, better performance may be obtained by interconnection of radio and land line carrier systems at the radio equipment site.

Some of the factors which affect land line extensions of radio carrier equipment and which influence the choice between direct extension or carrier frequency interconnection are discussed in this article.

Any of the frequency division carrier systems commonly used for channelizing radio circuits can be operated over land line (open wire or cable) extensions within certain limitations imposed by performance requirements. Because such operation is possible, the carrier terminals for a multichannel radio system need not be placed adjacent to the radio equipment.

Carrier terminals are normally installed in a centrally located toll office along with the other toll equipment. Radio transmitters and receivers, however, must be placed as close as possible to the antennas which are usually located where propagation conditions will be best. In some cases, radio equipment may be placed on the top floor of a building with a short length of cable extending to the carrier terminals several floors below. In other cases, because of propagation conditions, radio equipment must be located several miles away from the most desirable carrier terminal location; open wire or cable can then be used to transmit the carrier frequencies between the two equipment locations.

System Characteristics

Three Lenkurt carrier systems designed specifically for channelizing radio are presently available. Type 33B systems provide stackable channel equipment capable of providing up to 12 voice channels in the frequency spectrum below 48 kc. Type 33C systems provide up to 24 channels, stackable in groups of eight, in the frequency spectrum between 10 and 135 kc. Type 45BX systems provide up to 48 channels in the frequency spectrum between 40 and 264 kc. In some cases two different carrier systems (33B and 45BX for example) may be used on a single radio circuit to make maximum use of available bandwidth.

All carrier systems for radio have two basic characteristics that distinguish them from wire line and cable carrier systems. First, they are true four-wire systems with the same frequency allocation used in both directions of transmission. Second, their common equipment does not include the line amplifiers and system regulators commonly found on wire line systems. Also, carrier repeaters are not necessary on radio systems since radio repeaters provide the needed gain.

Because radio carrier systems use the same frequencies in both directions, operate at relatively low levels, and do not have system regulators, special consideration must be given to applying them on land lines. The attenuation and slope characteristics of land lines, and the amount of crosstalk which might be introduced in the land line section must be considered. These factors determine the ultimate transmission quality of channels transmitted for any distance over a land line extension. A number of other factors which affect the cost of land line extensions must also be considered in system engineering.

There are many conditions that affect engineering of wire line extensions. Because these conditions are seldom the same for different installations, no standard formulas can be given for determining the best means of transmitting carrier frequencies between carrier terminals and radio equipment. Only careful analysis can determine the most economical solution.

Engineering Considerations

Among the conditions to be considered are (1) the type of carrier equipment involved, (2) the type of radio equipment involved, (3) the types of land lines already available, (4) the cost of installing new land lines, (5) the possibility of future expansion of the radio system, (6) the requirements for coordination with other facilities, and (7) the length of the land line extension. Most of these conditions are interrelated; for example, a change in the type of carrier equipment may permit the use of a different and less expensive type of land line. On the other hand, if future expansion is contemplated, a more expensive immediate solution may be cheapest and most practical in the long run.

Total attenuation of the land line between the carrier terminal and radio equipment is limited by the maximum transmitting and minimum receiving levels of the equipment concerned. Table 1 shows the permissible attenuation (without auxiliary amplifiers) when using Lenkurt 72B radio equipment with the three different types of Lenkurt radio carrier systems. The permissible loss is not necessarily the same for both branches of a system. With 45BX systems, for example, more loss can be tolerated in the receiving branch than in the sending branch. Since both sending and receiving branches will normally be routed over the

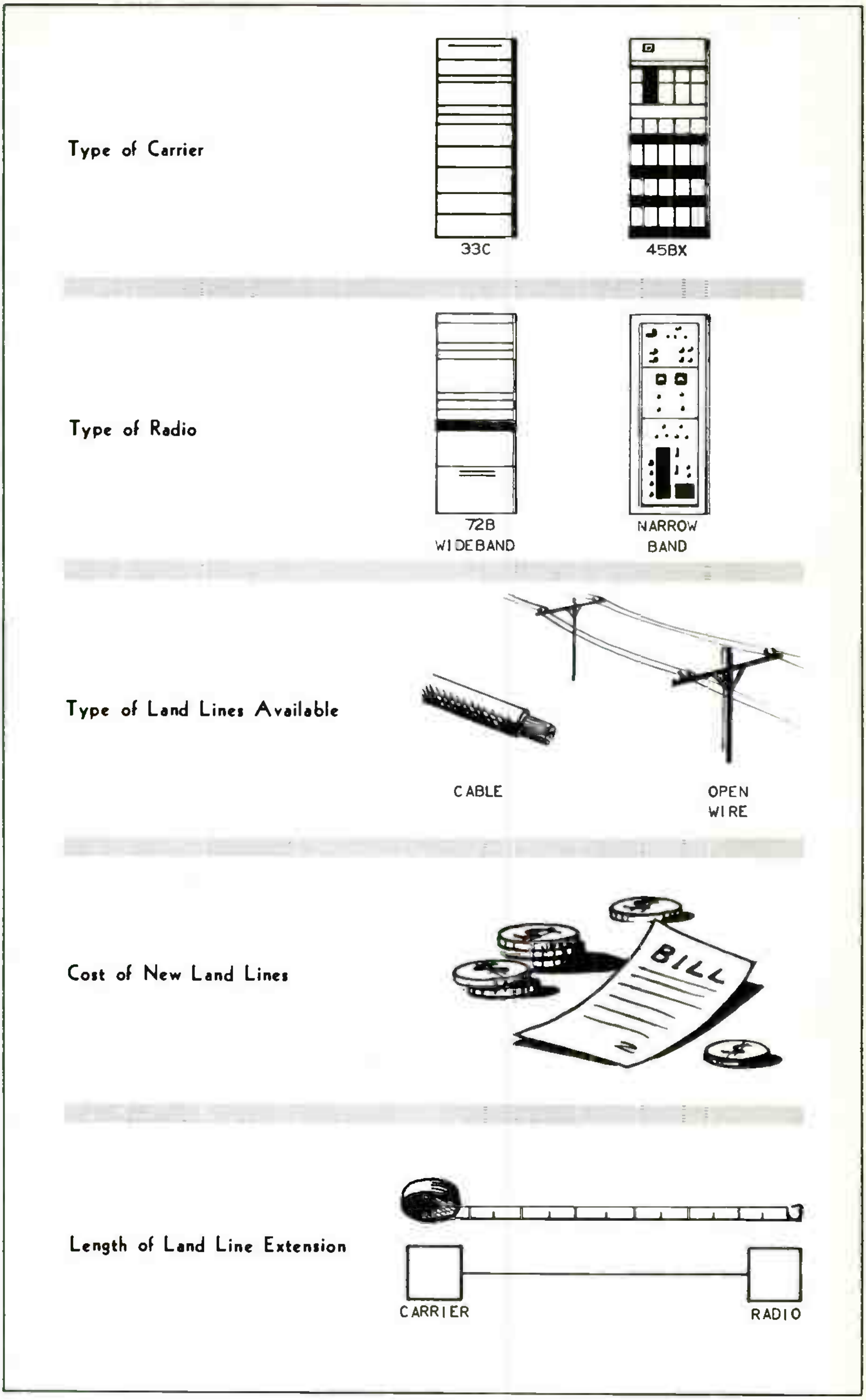


FIGURE 1. Many different factors must be considered when engineering land line extensions.

same facilities, the attenuation in the sending branch would be controlling for 45BX systems.

In some unusual cases, the difference in permissible loss may prove useful. For example, consider an installation of a 45BX system in a location where an aerial cable already available would introduce 30 db of loss. Although not common practice, it might prove most economical to route the receiving branch over one cable pair, and to install a new open wire pair on the cable poles. Since attenuation of the wire line would be less, the sending branch could then be routed over the open wire pair. Not only would this solution keep attenuation within acceptable limits, but it would also minimize crosstalk between sending and receiving branches of the system.

Frequency allocation of the carrier system also has a definite bearing on the type of facilities used for land line extensions. For example, either 45BX or 33C systems may be used to provide 24 channels, but the frequency allocations are different. The 33C system occupies frequencies from 10 to 135 kc while the 45BX system uses frequencies from 40 to 140 kc for 24 channels. Both branches of the 45BX system could be transmitted over open wire pairs already equipped with three-channel low-frequency wire line carrier systems. However, the 33C system could not be transmitted over the same pairs. Even if different pairs of the same pole line were used, interaction between the wire line and the radio system could result. If it is necessary to transmit a radio carrier system over a land line already equipped with other systems using the same frequencies, operating levels must also be coordinated to prevent intersystem crosstalk. Because four-wire open wire transmission is not conventional practice, applications of this type should be very carefully engineered.

The type of land line facilities already available, and the cost of installing new facilities, have an important bearing on the total cost of installing a complete new radio system. In general, three types of facilities are used as land line extensions between carrier terminals and radio equipment.

Exchange cables will very often prove suitable for fairly short distances. However, attenuation and noise of exchange cables is higher than that of toll cables or open wire lines which can, therefore, be used for somewhat longer extensions. Use of open wire lines is also limited to some extent by weather conditions. Increases in attenuation during rain or sleet storms could cause corresponding changes in level at the carrier terminal drops.

Attenuation-frequency charac-

TABLE 1. Maximum Permissible Attenuation for land line extensions between Lenkurt Type 72B radio equipment and Types 33B, 33C, and 45BX carrier equipment. All levels are on a per-channel basis.

	TYPE 72B RADIO Maximum Receiver Output Level -5 db Minimum Transmitter Input Level -30 db			
	Receiving Branch		Transmitting Branch	
	Min. level	Max. line loss	Max. level	Max. line loss
33B	-20 db	15 db	+10 db	40 db
33C (24 channel terminal)	-31 db	26 db	-4 db	26 db
45BX (24 channel terminal)	-40 db	35 db	-6 db	24 db

teristics for typical open wire lines, toll cables, and exchange cables are shown in Figure 2. Using the 33C carrier system as an example, the maximum length of wire line extension for typical exchange cable would be about 2 1/2 miles. For typical toll cable, extensions up to about 4 1/2 miles could be used, while for open wire line (assuming a nonsleet area) extensions up to about 65 miles might theoretically be possible.

In all cases the possibility of crosstalk between transmitting and receiving branches (and between the radio carrier system and other systems using the same facilities) would also be a factor limiting the length of the extension. If only one carrier system is involved, properly selected pairs of a single cable may provide adequate isolation between sending and receiving branches. In some cases, the most practical method of extension has been the use of separate cables for the two branches of the system.

When open wire lines are used, proper transposition and choice of pairs may permit transmission of both branches over the same pole line. Where more than one carrier system is involved, it is often practical to use two cables, one for sending and one for receiving. Then proper selection of cable pairs will help keep crosstalk between the systems to a minimum.

Attenuation and Slope Compensation

Auxiliary amplifiers and equalizers can compensate for slope and attenuation of land line extensions where the limitations imposed by system characteristics are exceeded. However, because all other considerations are related, amplifiers and equalizers cannot arbitrarily be inserted into the land line. In some cases crosstalk considerations might be the controlling factor; in other cases, the cost of special amplifiers or equalizers might be greater than the cost of some other solution.

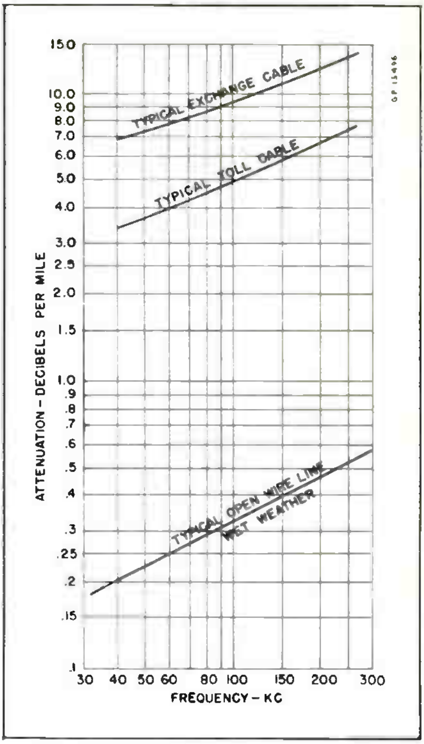


FIGURE 2. Attenuation-Frequency characteristics for typical open wire, toll cable, and exchange cable.

In most carrier terminals, a certain amount of slope correction can be obtained by special system alignment although this is not normally recommended. If the slope across each 3 kc of bandwidth does not reach a noticeable amount, the gain controls in the channels can be used to pre-emphasize the sending end and to equalize on the receiving end. Since slope compensation requirements will be different for each installation, special lineup procedures must be worked out for each case where channel gain adjustments are used for slope compensation.

Type 45BX systems have one feature which is not found in other carrier systems. Where land line extensions are used the range of individual regulators in 45BX chan-

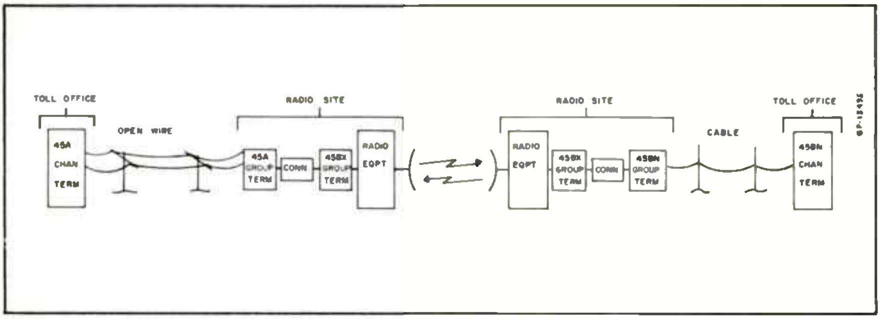


FIGURE 3. Channels transmitted over radio by Type 45BX carrier systems can be transferred at carrier frequencies for extension to distant offices by 45A or 45BN open wire and cable carrier systems.

nels (about plus or minus 7 db) can be used to compensate either for inherent slope in the connecting line or for minor variations in line characteristics with changes in weather.

Carrier Frequency Interconnection

With 45BX systems, operating companies also have another option which can eliminate all of the problems involved in extending non-coordinated carrier frequencies over land lines. Instead of directly transmitting the sending and receiving frequencies of the radio carrier terminals over land lines, the 45BX terminals which are required for radio transmission can be interconnected at the radio equipment site with 45BN (cable) or 45A (open wire) terminals. The carrier channels are then extended over cables or wire lines on systems specifically designed for these facilities.

Wherever any considerable distances are involved, interconnection may often be the only economically feasible course of action. For short distances, however, cost comparisons should be made to determine whether extension of the radio system or interconnection would provide the most satisfactory performance.

Interconnection is possible with 45-class systems because terminals for open wire, cable, and radio have identical channelizing equipment and identical twelve-channel carrier frequency base-groups. Twelve-channel carrier frequency groups can be transferred from 45BX terminals to terminals of other 45-class carrier systems designed specifically for use on the various types of land lines. A typical example of carrier frequency interconnection between 45BX and 45BN and between 45BX and 45A systems is shown in Figure 3.

Crosstalk Considerations

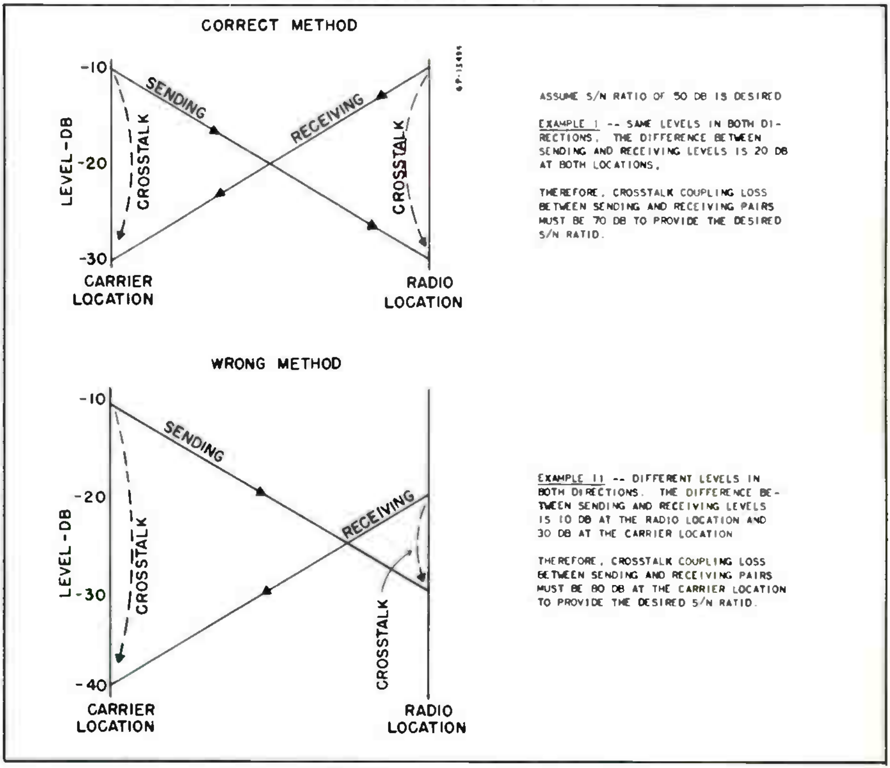
Because the same frequencies are used in both directions of transmission, near-end crosstalk between the sending and receiving branches of the same system is an important consideration in land line extensions of radio carrier systems. In contrast, near-end crosstalk is seldom of great importance with systems designed specifically for land lines because frequency allocations and operating levels are carefully coordinated for such systems. If only one radio carrier system is extended over a pole line or cable, the interference due to near end crosstalk will appear as sidetone to the talker. While requirements for sidetone suppression are not so severe as for intersystem or interchannel crosstalk, excessive sidetone can be quite objectionable. Also, objectionable interaction between sending and receiving signaling channels can be caused by insufficient isolation. Of course, where two radio carrier systems are extended over a single cable or open-wire lead, intersystem crosstalk will be a consideration.

Where near-end crosstalk is likely to occur, best operating results will be obtained by sending and receiving at the same levels at both ends. The total crosstalk suppression requirements are the sum of the desired signal-to-noise ratio and the level difference between signals being sent and received. Therefore, wherever possible, it is desirable to keep the total attenuation and total level differential as low as possible. The effect of near-end crosstalk for two different level differentials is illustrated in Figure 4.

Conclusions

Ability to extend frequency division multiplexed systems over land lines between carrier terminals and radio equipment offers a high degree of flexibility in system engineering. However, because transmission quality can be degraded in the land line extension, careful attention should be given to all of the factors which can influence both cost and performance of the system. Where the distances involved are too great to permit adequate performance with simple land line extensions, radio carrier systems can be interconnected at the radio site with land line carrier systems to assure proper overall system performance.

FIGURE 4. Near end crosstalk is important on wire line extensions because the same frequencies are used in both directions of transmission. Crosstalk is minimum when same levels are used in both directions.



THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6, NO. 11 NOVEMBER 1957

PROTECTING COMMUNICATION CIRCUITS

From High Voltages

Communication lines are subject to dangerous overvoltages from several sources, particularly high voltage power lines and lightning. Protection against these hazards involves line transpositions, shielding, voltage limiting devices and adequate grounding.

This article discusses some ways by which high voltages are induced in communication lines and the basic methods of controlling them to protect personnel and equipment.

Telephone communication lines normally live in a world of electrical energy that seldom exceeds 150 volts at a fraction of an ampere. Hence, the people who operate and maintain them, equipment connected to them, and the lines themselves must be protected against "intruder" potentials that may be up to a hundred or more times normal.

These overvoltages are of several origins. Some are induced in the circuit by lightning. But probably the most common overvoltages in communication circuits are those induced by unbalanced fault currents in power circuits running closely parallel to the communication lines. Actual accidental contact with the higher voltage power lines is also a frequent source of overvoltage.

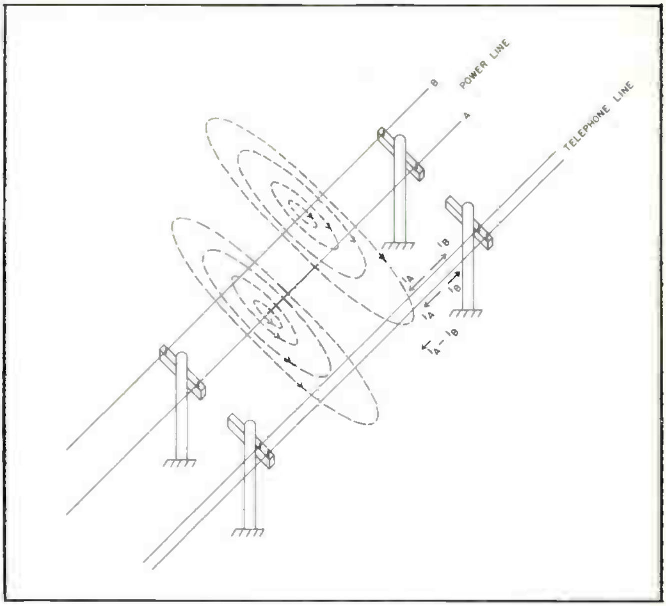
Induced Current and Voltage

Two kinds of current can be induced in a telephone line from external sources: *longitudinal* and *metallic*. Both types are usually produced by power line or lightning induction.

If the magnetic field of a power line couples with a telephone line in such a way that the currents induced flow in the same direction in each of the telephone line conductors, the currents are said to be longitudinal. Since they must return by some path other than the conductors themselves, they usually return through the earth. The voltage creating longitudinal currents thus appears between the telephone line and ground.

When the currents induced in the conductors of the telephone line flow in opposite directions, they are called metallic currents. They complete their path by flowing down one wire and returning by the other. Even though the currents induced in the separate wires of a telephone circuit flow in the same direction, if one is larger than the other,

Fig. 1a. Under normal conditions, the current induced in the telephone line by power conductor A nearly cancels the current induced by conductor B.



the difference between them will comprise a metallic current.

Normal voice or carrier currents in the telephone line are metallic currents. The voltage between wires can be called metallic voltage. Both metallic and longitudinal currents can be large under power line fault conditions, but generally the longitudinal voltages and currents are the largest.

Power Lines

Almost since the beginning of telephony, power lines have been a source of danger to telephone personnel and equipment. Recognizing this, many communication companies and power companies have been working together for more than 40 years to reduce and control the high voltages which may be transferred to telephone circuits from paralleling or crossing power lines.

High voltage energy is transferred between lines in one of two ways; by direct accidental contact between the power line and telephone line, or by magnetic and electric induction. Accidental contact is most likely to occur where the same pole line is used for both services or where power lines cross communication lines. Joint pole line use is generally restricted to power distribution lines carrying neither extremely high voltages nor large currents.

The chances of accidental contact are normally minimized by proper outside plant design and construction by both the power company and the communications company. However, despite preventative measures, accidental contacts do occur—especially during storms and as the result of unpredictable natural and man-made catastrophes.

The most common source of high voltage in a telephone line is magnetic induction from a paralleling power line. Under normal conditions, magnetically induced currents and voltages are held within safe limits by both telephone line and power line balance. If a power-

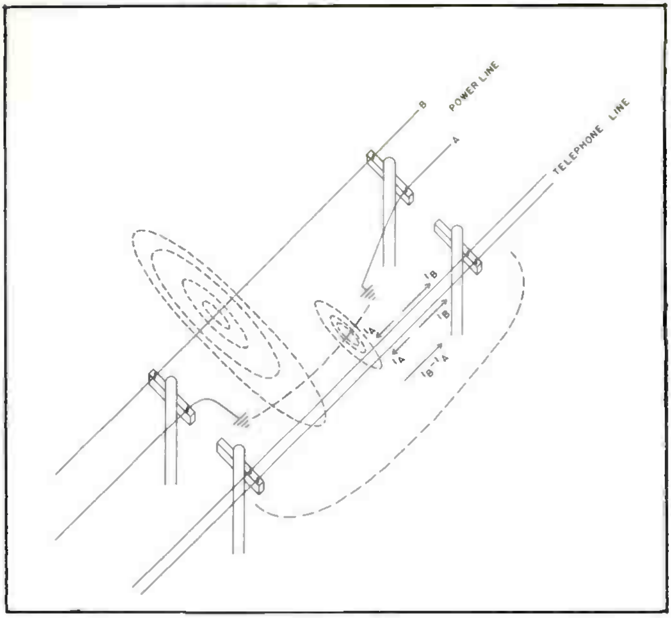


Fig. 1b. Under power line fault conditions, the magnetic couplings between power conductors A and B to the telephone line are no longer equal, resulting in a high induced voltage between the telephone line and ground.

line fault (short circuit) occurs between conductors or between one or more conductors and ground, very large voltages will be induced between the telephone line and ground. These voltages will tend to persist as long as fault current flows in the power line. Fig. 1 illustrates how the magnetic field of the power line couples with a parallel telephone line under normal and faulted conditions.

In some cases, even though a fault does not exist, power line induction may be dangerous. If the power is 3-phase serving an unbalanced load, fairly large steady voltages may be induced in telephone lines.

Lightning

Next to power line induction, lightning is the most frequent cause of overvoltage in communication circuits. In some areas, it is by far the most prevalent. When lightning strikes, extremely high voltage and current surges can result. Currents in a lightning discharge may rise at a rate of many tens of thousands of amperes per microsecond. Decline is generally at a much slower rate. Peak lightning currents as high as 345,000 amperes have been measured.

On most occasions a lightning stroke consists of more than one discharge. These occur so rapidly that the eye sees them as one stroke, although on rare occasions more than one discharge can be distinguished. Multiple strokes of six to eight separate discharges down the same channel are common; an eleven-stroke discharge has been recorded.

A direct stroke on a telephone line usually results in damage to the line and any equipment connected to it in the near vicinity. However, even a lightning stroke on a nearby fence, tree, building or power line can induce very high voltages and currents in a telephone line.

Protection Requirements

Whatever the origin of overvoltages, communication circuits must be protected against them to safeguard customers and operating personnel against injury and to prevent damage to the communication plant itself. Safeguarding persons is by all odds the most cogent reason for protection.

A protective system for telephone communication circuits has several basic requirements. It must be inexpensive because protection must be provided to a large number of circuits at many locations. Protective devices used must be physically small because space for them is usually limited. Likewise, they must be simple to install and require little or no servicing.

Needless to say, it is impractical to protect a communication line against all possible overvoltage hazards. A direct stroke of lightning near a central office or a direct cross between a telephone line and a high-voltage power line (66kv to 230kv) may create currents so large as to burn up telephone lines, carbon block protectors, line filters, or other equipment. The expense of completely protecting a communications plant against such catastrophes would cost more than the communications plant itself. Fortunately, such events are infrequent.

Protection Techniques

The basic elements of a protection system for communications lines may include transpositions, shielding, voltage limiters, grounding systems, and neutralizing transformers. In addition, power systems may include shielding, resistors or reactors in the power system ground return path, high-speed circuit breaker and relay systems, and line transpositions. These elements also reduce fault currents and thus help protect communications lines.

Transpositions are one of the most effective ways of reducing metallic currents induced in a telephone line by power line unbalances or faults. Fig. 2

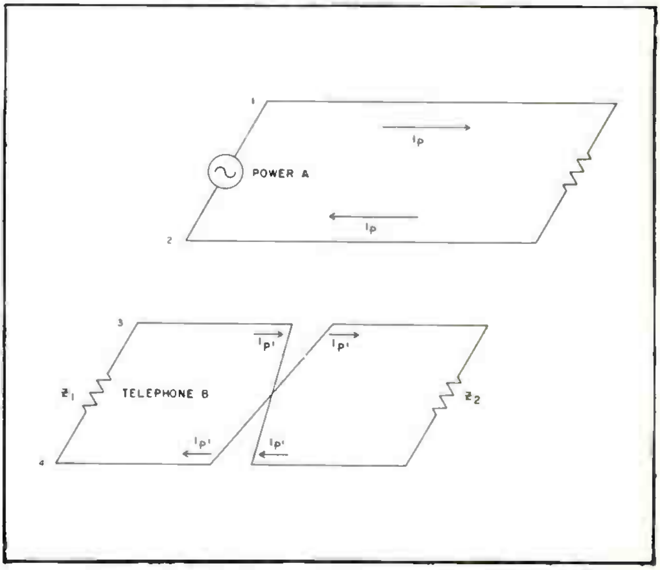


Fig. 2. Reduction of induced currents and voltage by transposing the telephone line. The net induced current is made up of two components which tend to cancel each other.

shows how metallic currents in one section of line are countered by currents in an adjacent section of line when the line is transposed at the junction of the two sections.

Another effective measure against overvoltages is shielding. Any nearby conducting material that provides a lower resistance path to ground than the communication line acts as a shield to draw off energy induced from power line or lightning currents. The lead covering on multipair cable is a partial shield. Even the messenger wire supporting an aerial cable line has shielding properties.

Protectors

One of the most common protective elements is the voltage-limiting carbon block protector. These are produced by several manufacturers in a variety of types and styles. However, in principle they are all similar.

The simplest carbon block protectors consist of two carbon electrodes separated from a metal ground strip by a narrow air gap. (Fig. 3) They may be completely or partially enclosed in an insulating housing. One carbon electrode is connected to each communication wire and the central metal strip is connected to ground. When a voltage appears between line and ground that is more than the air gap can withstand (generally about 450 volts to 3,000 volts depending on the type) the gap breaks down, allowing a current to flow between line and ground. Thus high voltages are prevented from reaching the central office or customer.

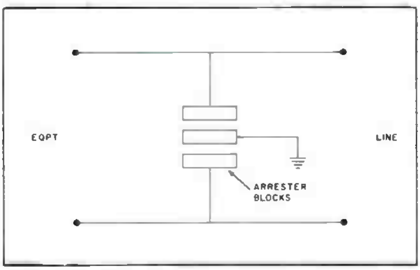


Fig. 3. Simple carbon block protector. Breakdown voltage is determined by the spacing between the carbon blocks and the center ground electrode.

A single carbon block protector is physically small, generally being less than two inches in its largest dimension. Fuses are sometimes used with protectors to interrupt currents so large that they might damage the protector or connected equipment.

Another type of voltage limiting device is the drainage coil or circuit. This device usually consists of coils and capacitors connected to provide a low-impedance path to ground for steady 60-cycle induction from power lines. The coils and capacitors form a simple tuned circuit that readily passes the 60-cycle induction current but has little effect on speech, carrier, or signaling frequencies. Drainage units usually have built-in carbon-block protectors to prevent high-voltage surges from damaging the coils and capacitors.

Good Grounding Important

The manner of grounding employed with protectors is of great importance. Good grounds are imperative. The impedance to ground should be as low as possible. Connections to ground should be made at many locations so as to provide many parallel paths for the excess current to take to ground. Also, the quality of the ground connections must be as good as can be made.

It is likewise important, if at all possible, for a communication circuit to be solidly connected to the same grounding system employed by the power circuit to which it is exposed. This is because surge or fault currents can rise so fast that even though two parts of a circuit or two neighboring circuits are both grounded, voltage differences of several thousand volts can appear between the two ground connections during the discharge.

Grounding of both power and communication circuits to the pipes of an extensive water system is preferred. Water piping comprises a superior ground not only because it provides a path of low impedance but also because it usually serves as a common ground for all electrical facilities in the vicinity.

On poles carrying both power wires and telephone cable, grounding can be effectively accomplished by connecting the cable sheath at regular intervals to the wires running to ground from the power system neutral. Because the metallic sheath of the cable provides a partial shield for the communication conductors, protectors can sometimes be omitted at the cable terminals.

Open-wire communication circuits are not directly grounded as that would nullify their purpose. It is here that protectors and drainage units must intervene between wires and ground to maintain the circuit to ground open until an unsafe abnormal voltage appears.

Where telephone cable terminates in open wire, it is general practice to use protectors to protect the wires in the cable from breakdown by overvoltages coming in over the exposed open wires.

Conclusions

The protection of communications personnel and equipment against high voltages is an important part of plant design. The amount of protection which should be used is determined partially by cost, but more importantly by the necessity of protecting operating and maintenance personnel.

So far not mentioned, but of great importance, is the protection afforded by educating personnel that any exposed communication line is a potential source of lethal high voltage and should be treated with caution.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 3 NO. 11 NOVEMBER, 1954

LINE FILTERS

For Carrier Telephone Installations

A single wire pair is often used to transmit two or more communications systems operating in different frequency bands. To prevent interference between them, some device is required at terminals, repeaters, and junction points to separate the different frequency bands so that the various signals in the line can be routed over the proper paths. The device used to separate the frequency bands is called a 'line filter'.

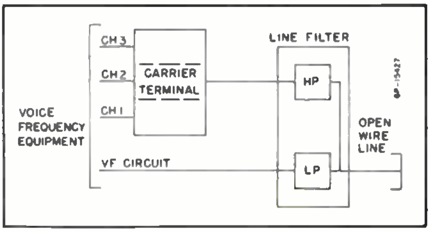
Some of the factors involved in the application of line filters to carrier lines are discussed in this article. Also included are a brief discussion of line filter theory, and descriptions of typical applications.

In the normal growth pattern of most telephone companies, the first toll facilities were open wire pairs strung on poles. Each pair of wires was used for one message at a time. When more message paths were needed, additional pairs were added to the existing pole line. Another method, often used for increasing the message handling capacity of a pole line, was the phantom circuit which gained one additional circuit for every two physical circuits.

When carrier telephone systems became available, it was usually more economical to increase the capacity of an open wire line by the addition of carrier channels. Many different types of carrier systems are now used to meet various circuit requirements.

The first step in the use of carrier is usually the addition of a low frequency system providing from one to four channels at frequencies below 35 kc. Demands for additional circuits can be met by high frequency carrier systems providing up to twelve more channels in the frequency range of 36 to 150 kc. Thus, through the use of carrier channels, up to 16 voice paths can be obtained from a single pair of wires. In addition, d-c telegraph or signaling pulses are often transmitted over the same pair without interfering with the voice

FIGURE 1. The primary purpose of the line filter is to electrically isolate separate frequency bands using the same wire pair.



or carrier circuits. In effect, the single wire pair becomes a common transmission medium for a number of different communications systems, each using a separate frequency band.

Why Line Filters are Needed

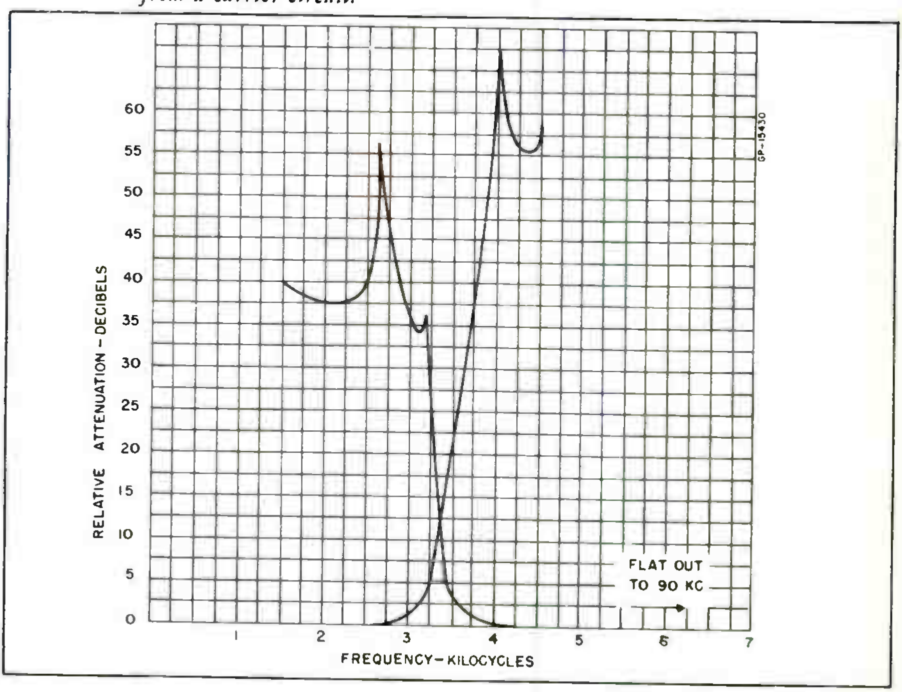
When a single pair of wires is used as the transmission medium for more than one communication system, three important requirements must be met. These are: (1) A person talking over one system should not hear any appreciable noise or crosstalk from another system using the same pair; (2) Different systems connected to the same wire pair should not absorb any power from each other (this is called mutual loading); and (3) the connection of two or more systems to the same pair should not create an impedance mismatch that would cause crosstalk, repeater unbalance, or attenuation. If these requirements are met, the different systems will be isolated from each other.

When two or more systems (two carrier, two carrier plus voice, or one carrier and voice) use the same pair, special filters are used to separate the frequency bands and thus isolate the individual systems. These filters are normally called 'line filters' or 'junction filters' depending on their characteristics. They are an important part of most carrier installations.

The need for line filters can be understood by considering the application of a low frequency carrier system to a wire pair that is also used to carry a voice frequency circuit. The range of frequencies required for the v-f circuit is from about 200 to 3000 cycles. The carrier may use frequencies from 3500 to 35,000 cycles.

If both the voice frequency and the carrier equipment were simply connected to the wire pair without any provision being made to isolate

FIGURE 2. Typical characteristic curves for a line filter designed to separate a voice circuit from a carrier circuit.



the two systems, interference and mutual loading would normally result. Interference with the voice circuit would be caused by spurious voice frequency tones generated in the carrier equipment. In a similar manner, high frequency components of speech may interfere with the lowest frequency carrier channel. Voice frequency speech energy might also cause distortion by overloading the receiving amplifiers of the carrier channels.

Fortunately, carrier and voice frequencies can easily be separated. If an appropriate line filter is used to isolate the two systems, each will operate independently of the other and no undesirable effects will occur.

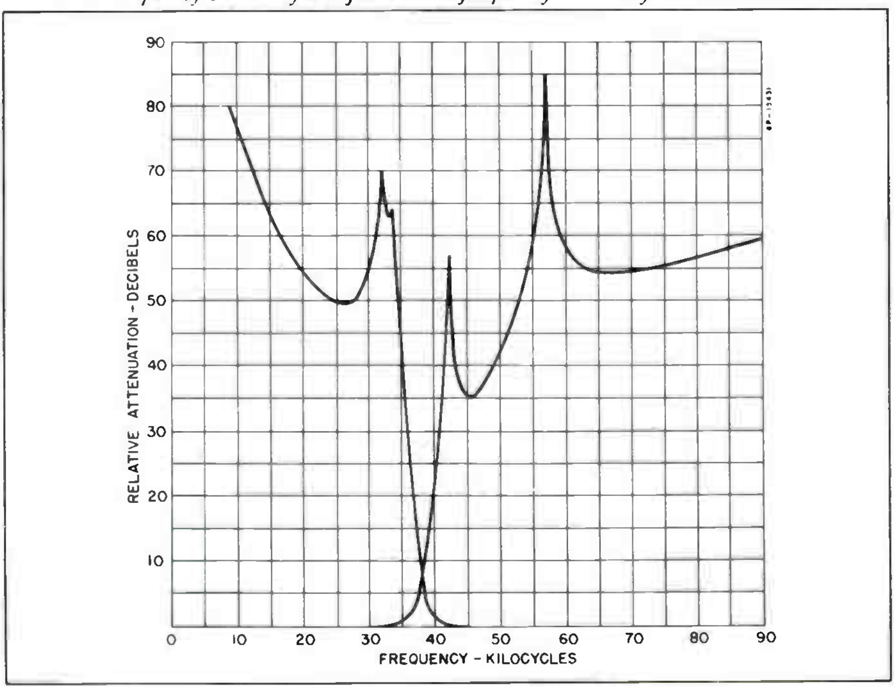
Line Filter Types

A line filter is in reality a combination of two filters; a low-pass filter and a high-pass filter. The two filters are normally mounted together in a common enclosure. The low-pass and high-pass filter sections are connected in parallel on the line side as shown in Figure 1.

In the case of a line filter used to separate a low frequency carrier system from a voice circuit, the carrier system is connected to the high-pass section and the voice circuit is connected to the low-pass section. Line filters are normally designed so that their low-pass and high-pass sections complement each other. That is, the cutoff frequency of the low-pass section is about the same as the cutoff frequency of the high-pass section. Hence, a line filter designed to separate a low-frequency carrier system from a voice circuit would have a cutoff or separation frequency in the vicinity of 3200 cycles. The exact frequency would be determined by the particular application.

Line filters are also designed

FIGURE 3. Typical characteristic curves for a line filter designed to separate a high frequency carrier system from a low frequency carrier system.



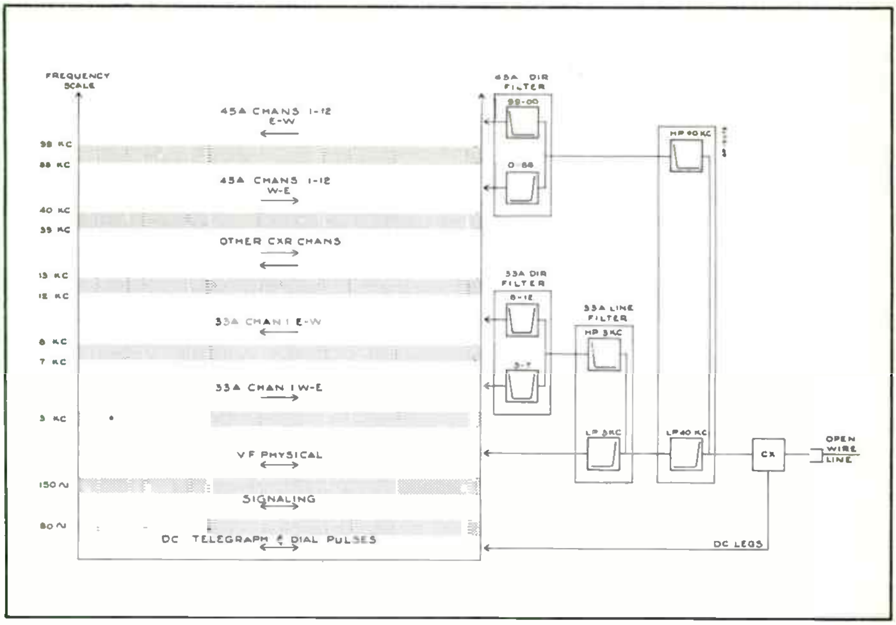


FIGURE 4. By use of appropiate line filters, a high frequency carrier system, low frequency carrier system, voice circuit, d-c telegraph, and signaling can all be impressed on the same pair of wires at the same time.

to separate low-frequency carrier systems from higher frequency systems.

Because line filters are used to separate voice frequencies from carrier frequencies or to separate two bands of carrier frequencies, they are often classified as either voice frequency separation filters or carrier frequency separation filters. When a line is used for high frequency carrier as well as for voice and low frequency carrier, both a voice frequency separation filter and a carrier frequency separation filter are usually required. Characteristic curves of both types are shown in Figures 2 and 3. Figure 4 illustrates the use of two line filters to permit the operation of both low and high frequency carrier systems as well as a voice circuit over a single pair of wires.

Filters for separating frequency bands are also classified according to their characteristics. For some types of applications, such filters must provide a high degree of isolation between the low and high frequency sections. For other applications, only enough isolation is required to prevent impedance mismatch. The filters used for such low isolation requirements are called 'junction filters'. They differ from the normal line filter in having simpler filter circuits. The name 'junction filter' is derived from their most common use; i. e., the junction between open wire and cable facilities.

When an open-wire telephone circuit passes through intermediate toll cable, it is often desirable to use a loaded pair for the voice frequencies and a non-loaded pair for the carrier frequencies. If a loaded pair and a non-loaded pair are simply connected in parallel, some impedance mismatching and excessive attenuation will result. If, however, a junction filter is used at both ends of the cable, the loaded and non-loaded pairs will be sufficiently isolated from each other so that the carrier frequencies will be transmitted over the non loaded pair and the voice frequencies over the loaded pair, each with minimum attenuation.

Line Filter Applications

The successful use of carrier equipment in the toll plant depends to a considerable extent on the proper application of line filters and associated equipment. The most important applications of line filters are at terminals, repeaters, and junctions.

The most common applications are at system terminals and at carrier repeaters. In many cases, an open wire line terminates in a toll entrance cable which carries the circuits into the central office. Since the attenuation and impedance characteristics of a wire pair in a cable are greatly different at voice frequencies than at carrier frequencies, a line filter is often mounted on the terminal pole so that the voice and carrier circuits can be brought to the central office over different pairs of the toll entrance cable.

If the distance from the terminal pole to the central office is very short, the line filter may be installed in the office and the small amount of attenuation caused by the short entrance cable neglected.

With the exception that toll entrance cable is not always involved, the use of line filters at carrier repeaters is essentially the same as that at line terminals.

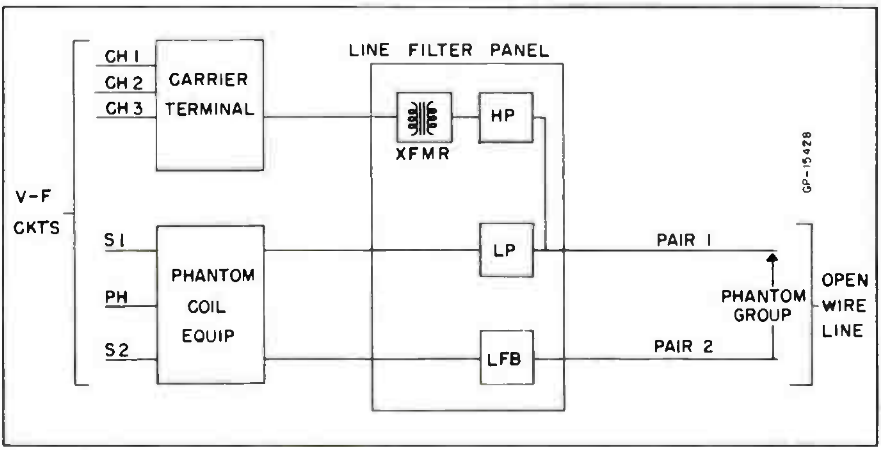
Line Filter Requirements

In the application of a line filter to a specific situation, two important factors must be considered. First, the isolation provided between frequency bands must be adequate. Second, the effect of the line filter on phantom circuits or other equipment connected to the pair must be considered. Isolation refers to the attenuation provided by the high and low-pass section of the line filter to frequencies outside their pass bands.

The amount of isolation necessary between any two systems using a common pair of wires as the transmission medium will depend on the particular application. In some cases, only 20 to 25 db will be required. In other cases, as many as 60 db of isolation may be necessary.

If isolation between two circuits is required only to prevent imped-

FIGURE 5. When a line filter is used on a wire pair that is also one side of a phantom circuit, a line filter balancing unit (LFB) should be placed in the other side of the phantom circuit.



ance mismatch and the resulting bridging loss, 20 db of isolation is usually adequate. Such a situation might occur when a way station is bridged across a toll line.

In some applications, however, the line filter must provide as much as 60 db of isolation between two circuits. This means that the stop bands of the low and high-pass filter sections must suppress undesired signals by at least 60 db. Usually such large amounts of isolation are required where intelligible crosstalk is likely to be a problem.

Effect on Other Circuits

Although line filters solve many of the problems involved in the application of a carrier system to a wire pair already carrying voice or other carrier channels, they sometimes, in solving one problem, create others. When a line filter is used in a wire pair that is part of a phantom circuit, a certain amount of unbalance is introduced into the phantom. To counteract the unbalance caused by the line filter it is necessary to add a filter balancing network to that half of the phantom circuit that does not contain a line filter. A typical example of the use of a line filter balancing network is illustrated in Figure 5.

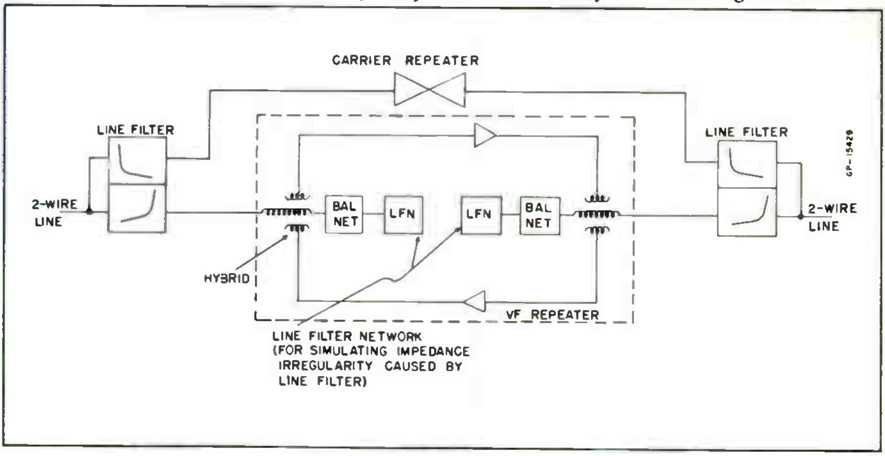
A second problem that is sometimes created by a line filter is the unbalance of voice frequency repeater hybrids. Unless the line filter matches the line characteristic impedance exactly, it represents an impedance irregularity to the repeater hybrid. If the unbalance is severe enough the repeater may sing.

To compensate for the unbalancing effect of the line filter, an electrically similar unit is added to the hybrid balancing network. Thus, the balancing network and the line impedances are made equal and the hybrid balance is maintained. Figure 6. illustrates the use of a line filter on a line containing a voice frequency repeater.

Conclusions

The line filter is a small but important part of any carrier installation in which more than one system operates over a single pair of wires. Through the proper application of line filters, a wide variety of combinations of carrier and voice frequency circuits are possible.

FIGURE 6. A line filter, when used on a pair that contains a repeatered voice circuit, tends to cause an unbalance of the repeater hybrid. To compensate for the unbalance, a special unit (LFN) is added to the hybrid balancing network.



THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 3 NO. 8 AUGUST, 1954

TRANSMISSION OF DIAL AND TELETYPEWRITER SIGNALS

Over Carrier Telephone Channels

Information is transmitted over carrier telephone systems in two basic forms: continuous waves and discrete pulses. Speech is the most important type of continuous wave transmission while telegraph and dial signals are the most common types of pulse transmission. In voice transmission, meaning or intelligence is conveyed by a constantly changing complex waveform. In pulse transmission, however, meaning is conveyed by the presence or absence of one of two different conditions of tone transmission. These two tone conditions have meanings of yes-no, mark-space, on-off, etc.

This article is a discussion of some of the factors affecting the transmission of dial and telegraph signals over carrier telephone channels. These factors include bandwidth, level variations, frequency stability, and noise.

In both telegraph and dial signal transmission, information is carried by trains of pulses such that each train conveys a single figure or letter. While basically similar, there are two distinct differences between dial signals and telegraph signals. The first difference is in the pulse repetition rate. Dial pulses have a normal repetition rate of about 10 pulses per second. (In some operator dialing a rate of 20 pps is used.) Telegraph pulses have a repetition rate of up to 60 pps or greater depending on the transmission rate in words per minute. (PPS $\approx$ 0.371 words per minute for standard teletypewriter coding.) Some circuits are designed for 60 words per minute, others for 40, 75, 120, 150, etc. The second difference is in the method of coding the pulse trains. In dial pulse trains the length of the pulse train (the number of pulses) determines the figure transmitted. Telegraph pulse trains, however, are all the same length. A coded arrangement of pulses and spaces during each interval determines the particular letter or character being sent. Examples of dial and telegraph pulse trains are shown in Figure 1.

The electrical similarity between dial and telegraph pulse trains suggests that both can be transmitted in the same manner with essentially the same types of equipment. In general, this is true. In both cases, d-c pulses originated by a dial or teletypewriter are transmitted to a carrier terminal

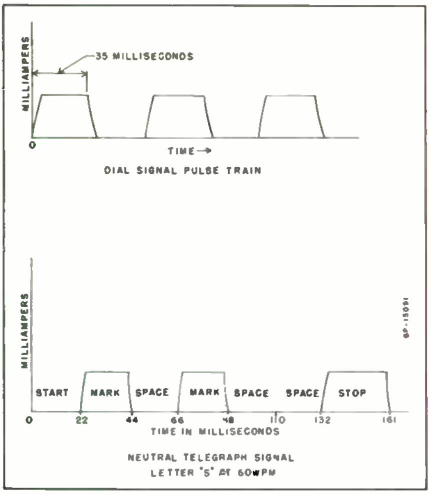


FIGURE 1. Electrically, dial and telegraph signal are very similar. Essentially the same methods are used for transmission of both types.

where they operate a relay or electronic keying circuit to cause transmission of the information over the carrier system.

The basic requirements of a carrier circuit to transmit pulse information include: (1) a translating circuit at the sending end to translate d-c pulses into tone information; (2) channelizing filters at the sending end to prevent interference between individual tone transmission channels; (3) receiving filters to select the proper tones from other transmitted signals; and (4) a second translating circuit to reconvert the tone information into d-c pulses for operation of the receiving teletypewriter or switching equipment.

For practical purposes, it can be considered that both dial and telegraph signals are transmitted over narrow-band carrier channels. In carrier telephone systems, a single narrow-band channel is normally associated with each voice channel for transmission of signaling information. For telegraph transmission, however, a number of closely spaced narrow-band channels are transmitted in the frequency spectrum normally occupied by a voice channel.

Transmission Methods

Two methods of translating d-c pulses into tone information are used in Lenkurt carrier systems. These are generally known as the on-off method and the frequency shift method. On-off signals are sometimes referred to as amplitude modulated (AM) signals, while frequency shift signals are often referred to as frequency modulated (FM) signals.

D-c pulses are translated into amplitude modulated signals by on-off switching of a single frequency tone. Frequency modulated signals are created by using the d-c pulses to shift the frequency of an oscillator so that one frequency indicates the presence of a marking pulse and the other frequency indicates its absence. In either case, the voice frequency tones produced are readily transmitted over a telephone voice channel.

Amplitude Modulation

In the transmission of amplitude modulated signals, the on-off keying of a single frequency produces short bursts of tone. If the original short tone bursts could be transmitted to the far end of the system without distortion, the received signals, when rectified and filtered, would be identical to the original on-off d-c pulses. This, although desirable, is neither necessary nor practical.

A tone keyed on and off is in essence a constant frequency carrier wave 100 percent amplitude modulated with a rectangular wave. It contains numerous frequency components. These include: (1) the carrier frequency; (2) modulation products of the carrier frequency with the fundamental frequency of the rectangular modulating wave; and (3) modulation products of the carrier frequency with all the harmonic components of the rectangular wave. Since these frequencies cover a relatively broad spectrum, to transmit all of them would be uneconomical. Fortunately, transmission of only a small part of these frequencies is necessary. The remaining side-frequencies are eliminated by the use of transmitting and receiving band-pass filters that allow only the desired frequencies to be transmitted without distortion. (Some signaling circuits omit the transmitting filter.)

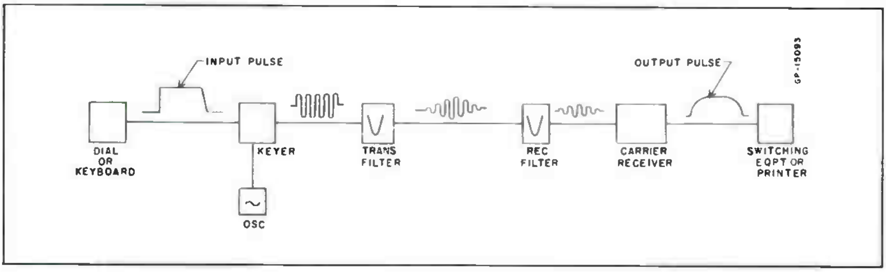
Restriction of the transmitting and receiving filter bandwidths plus the transient response of the filters themselves causes the envelopes of the received tone bursts to be 'cigar shaped'. The tone bursts, when reconverted into d-c pulses, are no longer rectangular; instead they are rounded. Figure 2 shows the wave shapes of a pulse as it travels from the keyer to the receiving end of the circuit.

Level Variations

If all the pulses received were of the same amplitude, rounding of the pulse wave shape would cause little impairment of signal reception. If, however, the amplitude of the received pulses varies because of changing transmission conditions; and if the receiving circuits do not properly compensate for the receiving level variations, distortion may occur. Shown in Figure 3 are typical received pulse wave forms with the current amplitudes necessary for closing and releasing of the receiving relay. If the pulse is reduced in size because of reduced signal level, the time interval during which the relay is closed is reduced. If the pulse amplitude is increased the time interval during which the relay armature is operated is increased. This type of distortion is called bias distortion. If the closed interval of the receiving relay is increased above normal, the distortion is called positive (or marking) bias distortion; if shortened, the distortion is called negative (or spacing) distortion. Figures 3b and 3c show how variations in received pulse amplitude cause bias distortion.

Bias distortion occurs in both telegraph and dial signaling circuits. Telegraph circuits, however, are also subject to a second type of distortion called characteristic distortion. This type of distortion can only occur when pulses of different lengths are transmitted. Under conditions of reduced signal level such as shown in Figure 3c, short pulses tend to be shortened considerably. A long pulse, such as would occur when two or more marks are transmitted consecutively, would not be shortened as much. If characteristic distortion is severe, the receiving relay will

FIGURE 2. Transmission of a pulse over a signaling or telegraph circuit. The d-c pulse is converted to a tone spurt, filtered, transmitted over the carrier circuit, detected, and reconverted into a d-c pulse.



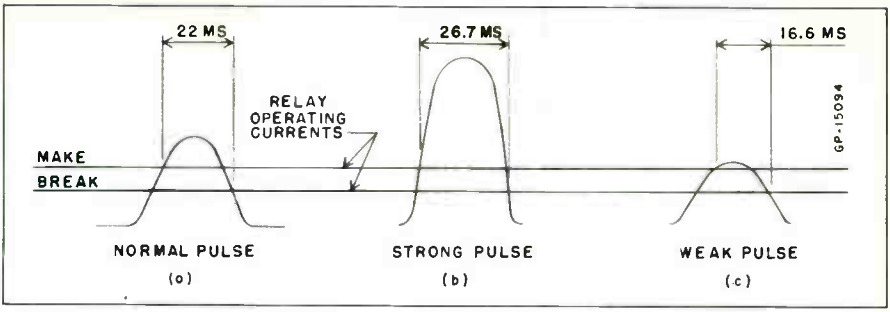


FIGURE 3. Typical received telegragh or dial pulses. In figure 3b, an increased pulse amplitude increases the effective pulse length and causes positive bias distortion. A weak pulse (Figure 3c) shortens the effective pulse length and causes negative bias distortion.

fail to respond to short pulses and garbled transmission will result.

A third type of distortion that can affect both telegraph and dial pulse transmission is known as fortuitous distortion. This type of distortion results from random causes that shorten, lengthen or delay individual pulses. Line hits, noise, or any other source of spurious impulse can cause fortuitous distortion.

In general, carrier signaling channels are transmitted under more carefully controlled conditions than are carrier telegraph channels; therefore less complex receiving circuits can be used for signaling channels. In AM telegraph circuits, such as a Lenkurt 24C system, however, additional level compensating circuits are usually provided to maintain the output level of the receiving amplifier more constant. The operation of the level compensating circuit is such that if the received signal level is increased, the average amplitude of the rectified pulses is increased. This rectified signal, when applied to the input of the amplifier tube, reduces the gain of the amplifier to keep the amplifier output nearly constant.

Frequency Modulation

While AM pulse transmission methods are subject to distortion from level variations and spurious impulses, FM systems have greater immunity to this type of interference. In FM transmission, keying of the sending circuit by d-c pulses from a dial or teletypewriter causes the frequency of an oscillator to be shifted. As in AM systems, sidebands are produced when a carrier wave is shifted from one frequency to another. Because, in effect, the carrier is frequency modulated by a rectangular wave, a very wide band of frequencies is produced. If it were necessary to transmit all the sidebands so that the original wave could be recreated undistorted by demodulating the received signal, very wide bandwidths would be required for each signal channel. However, as is the case for AM transmission, the transmission of all the side frequencies is undesirable for reasons of bandwidth economy, and because of the possibility of interference with voice frequencies or other pulse transmission channels.

The transmitting and receiving filter bandwidths required for FM transmission depend on the pulse repetition rate and the amount of frequency shift. In general, the requirements for transmitting filters are less stringent than for receiving filters because their primary function is to prevent mutual loading between individual channels or between the signaling channel and its associated voice channel. Receiving filters, however, must be much more selective because in addition to preventing mutual loading between receivers, they must also suppress unwanted interference from adjacent channels or other sources.

In the FM dial signaling used with Lenkurt 45-class carrier, the maximum pulse repetition rate that can be faithfully transmitted is at least 20 pps. The frequency shift is 150 cycles from 3400 to 3550 cycles. The signal receiving bandpass filters will pass frequencies from about 3360 cycles to 3600 cycles, a bandwidth of 240 cycles. This bandwidth is sufficient to pass a large number of the modulation products which results in the reception of the received pulses with very little rounding.

In telegraph transmission, the requirement of obtaining as many channels as possible in a given bandwidth dictates that the narrowest channel spacing practical be used. As a result the channel spacing in telegraph carrier systems is a compromise between bandwidth requirements for low distortion and maximum use of available spectrum.

In Lenkurt Type 22A FM telegraph equipment, pulse repetition rates of 23 and 30 pps occur for 60 and 75 wpm transmission respectively. The frequency shift for 60 wpm is 30 cycles and the frequency shift for 75 wpm is 40 cycles. For 60 wpm systems, filter bandwidths of about 70 cycles are used and for 75 wpm transmission, bandwidths of about 100 cycles are used. Because of the restricted bandwidths, the recovered d-c pulse at the descriminator output is considerably rounded. However, actual pulse shape is of less importance than the requirement that all pulses have the same shape and amplitude.

In the operation of a typical FM receiver, the desired signals are selected by a receiving filter, amplified, passed through a limiter, and then detected by the discriminator. The detected signals are again amplified and used to operate the receiving relay. Unlike AM transmission, the signal presented to the FM receiver is a continuous electrical wave instead of short bursts of tone with no signal between bursts. The on-off or Mark-Space conditions are determined by the frequency of the continuous wave. For the Mark condition, the received wave will have one frequency while for the Space condition the frequency will be shifted to a different value. But the wave remains continuous at all times. The desired information in the received wave is contained in its frequency shift aspects. Amplitude variations contain no useful information and are undesirable. Since, however, discriminators are sensitive to both frequency and amplitude variation, it is necessary to remove the amplitude variations from the received wave before it reaches the discriminator. Otherwise the waveform at the discriminator output would be distorted.

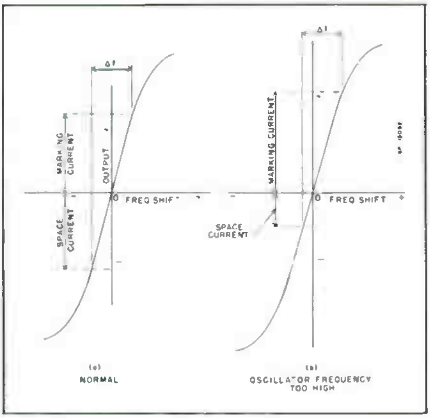
The function of the limiter is to remove any amplitude variation in the received wave before it reaches the discriminator. Its operation is such that each individual cycle of the received wave is 'limited' in amplitude to a fixed value. Thus the wave leaving the limiter and entering the discriminator is a constant amplitude wave of varying frequency. The discriminator translates the frequency shifts into d-c pulses for the operation of the receiving relay.

Frequency Drift

While FM telegraph and signaling systems are relatively insensitive to level variations, they tend to be more sensitive to variations in frequency than AM Systems. Because of the small frequency shift used in FM carrier telegraphy, the discriminator characteristic curve must have a steep slope for proper detection (refer to Figure 4). If the transmitting oscillator frequency drifts or the discriminator is misaligned, the operating point of the discriminator is shifted up or down on its characteristic curve. An increase in oscillator frequency shifts the operating point upward on the characteristic curve and the marking pulses are shortened while the spaces between pulses are lengthened. The result is, of course, negative bias distortion. A shift in the other direction causes the pulses to be lengthened causing positive bias distortion.

The small frequency shift and steep discriminator characteristic curve slope necessary in FM telegraphy require the use of very stable oscillators. In Lenkurt Type 22 FM telegraph carrier, the oscillator drifts less than 0.1 cycle for plate voltage variations of 20 percent.

FIGURE 4. Typical discriminator characteristics. In normal operation, marking and spacing currents are equal (5a) If the oscillator frequency increases above normal, bias distortion occurs (5b).



Not only must the telegraph oscillators be stable, but the carrier supply oscillators of the carrier telephone system over which the FM signals are transmitted must either be very stable or the sending terminal and receiving terminal carrier oscillators must be synchronized. Lenkurt 45-class carrier systems with oscillator frequency accuracies of better than one part in one million permit the operation of FM telegraph systems over the voice channels.

In some types of short wave radio telegraphy, where the radio receiver is manually tuned, very large frequency shifts are used (up to 850 cps) to permit good reception even though the receiver may be slightly mistuned. The wider frequency shift permits the use of a discriminator characteristic with a flatter slope that is less sensitive to frequency changes. In carrier transmission, however, bandwidth restrictions prevent the use of less critical discriminator characteristics.

AM vs. FM Transmission

Both AM and FM pulse transmission systems are capable of reliable high speed transmission. Except for the noise advantage of FM over AM, there is little difference between the two methods for dial signal transmission. For telegraph, however, the choice between the two depends upon the particular application. In general, a transmission medium having the following characteristics would tend to make an AM system the best choice. These are:

1. Levels not subject to extremely rapid level fluctuation.
2. Low noise levels expected.
3. Frequency stability relatively poor.

The following transmission conditions would tend to make an FM system the best choice. These are:

1. Considerable level variation expected.
2. Noisy circuits.
3. Frequency stability excellent.

The choice between AM and FM pulse transmission methods is determined largely by which of the above factors are likely to cause the most transmission impairment. In commercial transoceanic radio telegraphy, for example, wide level variations caused by fading are by far the most important factor. Even the best level compensation circuits will not completely eliminate the level variations experienced in radio transmission. As a result, FM radio telegraphy is in almost universal use.

On the other hand, telegraph dial signal transmission over carrier telephone channels is largely of the amplitude modulated type. Since many telephone transmission systems permit relatively large frequency errors between sending and receiving ends, and at the same time maintain very stable transmission levels, the use of AM telegraph is often preferred.

When telegraph tones are transmitted directly over wire lines, noise and level variations often make FM transmission the wisest choice.

Different operating companies tend to use one type more than the other because of differences in their types of operation. Many railroads and telegraph companies tend to use more FM telegraph because much of their traffic is solely telegraph and is transmitted directly over wire lines. Telephone companies, however, have tended to use more AM telegraph and signal transmission because much of their traffic has been over types of carrier telephone channels in which frequency drift was often a problem.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 5 NO. 4 APRIL, 1956

Telephone Channels for

DATA TRANSMISSION

Although telephone channels are used primarily for the transmission of voice conversations, they can also be used to transmit a variety of other kinds of signals. These other signals are sometimes given the broad name "data."

This article discusses the meaning of data, various methods of transmitting data, and the conditions limiting the operation of voice telephone channels for data transmission.

The purpose of a communication system is to transmit intelligence from one place to another. The intelligence may originate in many different ways, but before transmission it must be converted into either of two forms—continuous or pulse-type electrical signals. A generalization is sometimes made in which all communication by means of electrical impulses is called *data transmission.*

The most familiar forms of data transmission are telegraph, signaling, remote control, and telemetering. More recently, different forms have been developed which include photo facsimile, high-speed telegraph, air warning data from radar installations, and numerical information for the operation of electronic business machines. In each case, alphabetical, numerical, or symbolic data must be transmitted.

The high-speed operation of these new systems has created a need for rapid transmission of large amounts of data with minimum error. As a result of this need, scientists and engineers have been taking a closer look at what constitutes *information* and how it can be transmitted most efficiently with presently available types of facilities.

Information

One of the first steps in determining the exact nature of information was the selection of a unit, or yardstick, by which information could be measured. This unit had to be such that it could easily be determined and did not depend upon the importance of the message, since a message's importance is difficult to evaluate mathematically.

It turned out that the simplest and most basic unit was the amount of information necessary for a receiver (person or machine) to make the correct choice between two equally possible messages. This choice may be between the messages yes-or-no, on-or-off, A-or-B, 0-or-1, black-or-white, etc. Since the two possible messages correspond to the two symbols in the system of numbers called the binary digit system, a unit of information based on two symbols (messages) came to be called a binary digit and was abbreviated *bit*.

A simple electrical pulse has the informational value of one bit because the presence or absence of the pulse permits the receiver to choose the correct message from a set of two. As shown in Fig. 1, the transmission of two pulses permits the receiver to select the correct message from four equally possible messages. Three pulses, or bits, will enable the correct selection from a set of eight. This selection process gives the average amount of information which must be transmitted to specify a message from a set of equal possibilities. There is no method of transmitting information which uses less than this amount of information per message.

The letters of the alphabet are an example of a practical set of possible messages. The desired message might be any particular letter. If all the letters appeared equally as often, about five bits of information would be needed to select an individual letter. However, only about two bits per letter are needed because certain letters appear much more frequently than others.

Channel Capacity

An important factor in evaluating the utility of a communication channel for data transmission is its maximum information carrying capacity. Using the bit as a measure of information, the maximum capacity of a communication channel can be determined from

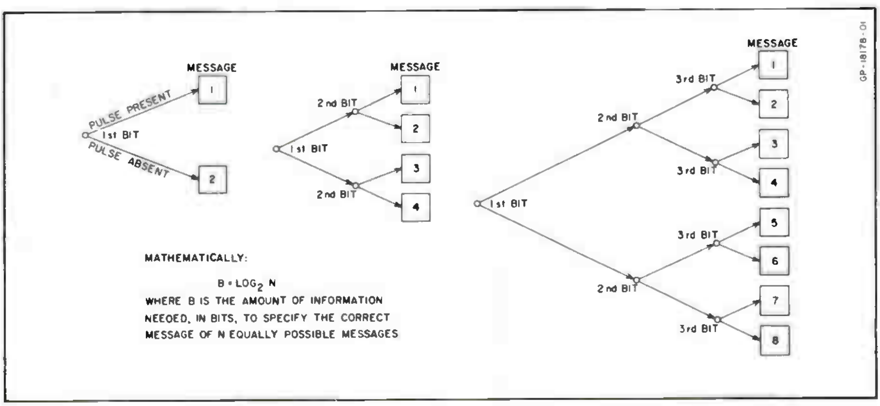


FIG. 1. Information requirements for specifying a particular message.

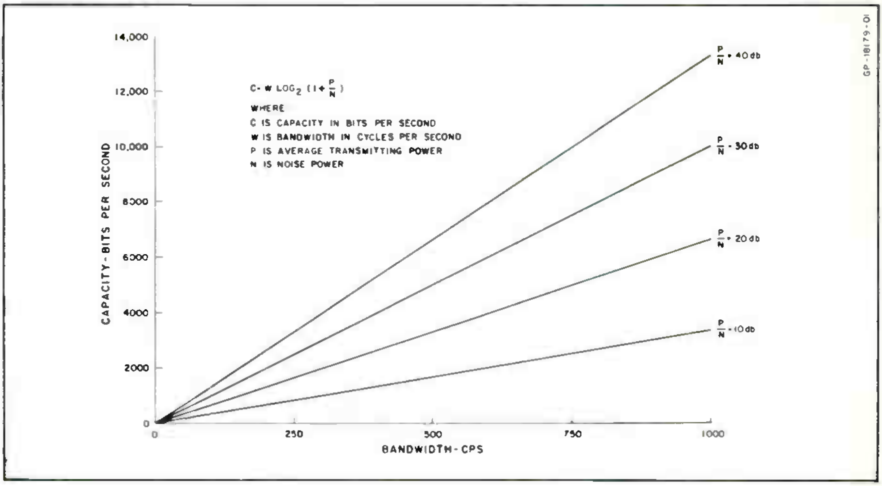


FIG. 2. The relationship between bandwidth, signal-to-noise ratio, and capacity.

the channel bandwidth, signal power, and noise.

The relationship between bandwidth, signal power, and noise is complex and depends upon many factors such as the kind of noise present in the channel, the nature of the power limitation, the type of modulation used, and the method of encoding the information. The relationship for the particular case where the noise in a channel of limited bandwidth is assumed to be random noise (equal noise power across the frequency band) and the channel is operating at a particular signal-to-noise ratio is shown in Fig. 2.

This figure shows that increasing the bandwidth or the signal-to-noise ratio will increase the capacity of the channel. A wider bandwidth permits shorter pulses to be transmitted well enough to be detected. Consequently, more pulses can be sent in the same period of time, thereby increasing the capacity of the system. Increasing the signal-to-noise ratio permits easier detection of the pulses. This permits faster transmission and results in an increase in capacity.

Because of the inter-relationship of factors affecting channel capacity, it is possible to exchange bandwidth for signal-to-noise ratio and still maintain the same channel capacity. For a given system, the occupied bandwidth can be decreased provided that the signal-to-noise ratio is increased sufficiently, or vice versa. Roughly, the bandwidth can be decreased one-half if the signal-to-noise ratio in db is doubled.

Transmission Methods

The most common method of transmitting alphabetical and numerical data is by binary pulse transmission; that is, pulses which are either present or absent. Familiar examples of this method are Morse and machine telegraphy. The transmission of information by Morse telegraphy is accomplished by means of coded pulse groups utilizing both long and short pulses (Fig. 3a). The more frequent characters are given the shortest code

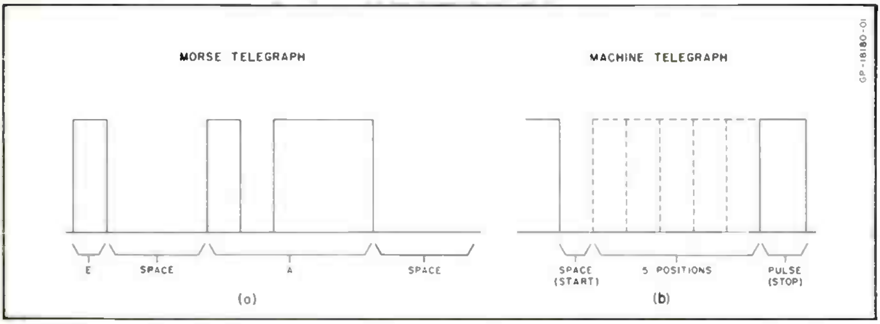


Fig. 3. Pulse code systems for two types of telegraph.

groups. For example, one short pulse represents E while a short and a long pulse represents A. The use of different length pulses permits the transmission of an extra bit of information which shortens the code groups.

Machine telegraphy is a somewhat different type of data transmission. It utilizes a standard length code group in which there are five possible information carrying pulse positions (Fig. 3b). A pulse may or may not appear in any of these positions. The five bits of information contained in the code group would permit selection of only 32 characters if each character appeared equally as often. However, since certain characters appear very infrequently, the information necessary to transmit over 50 different characters averages less than five bits per character.

Machine telegraph systems usually transmit information at a rate of 60 words per minute. This is achieved with a pulse rate of 30 pulses (bits) per second, plus the necessary synchronizing pulses, over a channel 120 cps wide.

The transmission rate of a binary system, such as machine telegraphy, can be increased by transmitting more pulses per unit of time. Information theory shows that this can be most easily accomplished by increasing the bandwidth of the channel, enabling the transmission of shorter pulses at a faster rate.

Another method of transmitting information more rapidly is by causing each pulse to contain more information. This permits the pulse repetition rate to remain constant, but more information is transmitted per pulse. The increase in the information carried by the pulse is accomplished by pulse modulation.

Pulse Modulation

The informational content of a pulse can be increased by modulating it with respect to amplitude or position. In pulse amplitude modulation, the distance between pulses remains constant, but the height of the pulse may be any one of a specified number of possible amplitudes. By modulating a pulse's position, its amplitude remains constant and its time of occurrence is varied over a range of discrete positions. These methods are shown in Fig. 4.

In either type of pulse modulation, each step of amplitude or position has

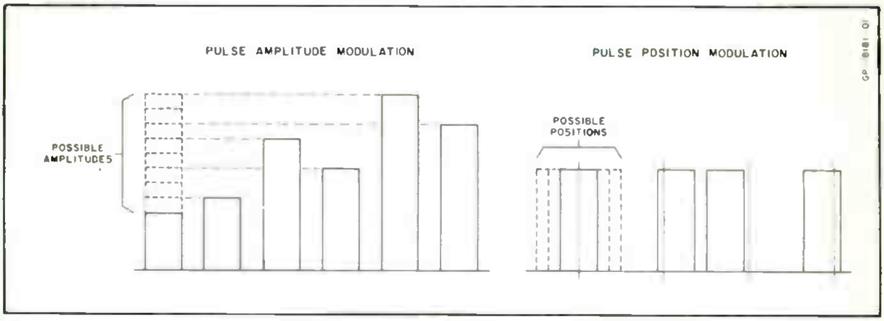


FIG. 4. Methods of increasing the informational content of a pulse by modulation.

a definite code value. As a pulse is permitted to occupy more steps, its informational capacity increases. For example, the correct message of a set of eight can be specified by a pulse which has eight possible steps. The pulse therefore contains three bits of information as compared to one bit when a single step pulse is transmitted.

Systems utilizing pulse modulation are relatively sensitive to noise because it obscures the transmitted amplitude of a pulse and also causes a slight variation in the precise location of the pulse. Therefore, as the number of modulation steps is increased, the signal-to-noise ratio must also be improved to enable the receiver to detect exactly the transmitted amplitude or position.

The limiting capacity of a channel can be approached most closely by using modulated pulses with a large number of steps. Of course, as this is done the transmitters and receivers must become more complex to interpret the large amount of information contained in each pulse. Consequently, a point is reached where the maximum theoretical capacity of a channel is not yet reached, but where further increase of the information rate is impracticable.

A Data Transmission System

A practical example of a high-speed data transmission system using telephone channels is the SAGE (Semi-Automatic Ground Environment) network. This network is being established to increase the speed of information processing and weapon direction in the national air defense system.

Data transmission for SAGE is basically similar to machine telegraphy; however, information must be transmitted at rates up to 1600 bits per second. This is equivalent to a telegraph system operating at a speed of over 2000 words per minute. While maintaining this high speed, the system must operate with a maximum error rate of 1 bit per 100,000.

SAGE transmission consists of pulse groups of standard length. The beginning of the pulse group is signified by a high-amplitude pulse, and the transmitted information is contained in pulses which may or may not be present in 11 positions following the start pulse. Also being transmitted continually at the pulse repetition rate is a low-amplitude pulse which maintains synchronization of the system.

Because the system is binary, the receiver must determine only if a pulse is present or absent at every position. Therefore, a well-shaped, square pulse is not necessary for proper detection. This results in conservation of bandwidth because the higher component frequencies of the pulse need not be transmitted. It is necessary, however, that the component frequencies which are transmitted remain at the same amplitude and that they all be delayed by the same amount of time.

Most telephone channels can transmit the component frequencies without amplitude distortion because they normally have a flat attenuation characteristic over their passband; however, they may require delay equalization.

For SAGE data transmission, telephone channels are required to be relatively flat between 500 and 2500 cps (no more than 3 db deviation from midband to band edge). Also, the difference in transmission time of any two frequencies in the band from 1000 to 2500 cps should not exceed 500 microseconds, and the number of impulse noise peaks which come within 18 db below the synchronizing signal level should not exceed 1 per minute.

SAGE-type data transmission usually can operate over any type of facility which provides these necessary requirements. However, certain types of facilities require special treatment such as delay equalization and impulse noise reduction to make them suitable. When the circuits are provided by carrier, the delay equalization requirement is not too great a problem since most of the data circuits are expected to be relatively short and simple in make-up. Thus, the primary consideration on carrier data circuits is the suppression of impulse-type noise.

The control of impulse noise on carrier channels is principally a matter of plant layout, since very little such noise is contributed by the carrier equipment itself. In general, a conservatively designed system which provides high-grade toll service without compandors is probably satisfactory for data transmission. In some cases, data transmission may be possible on normally compandored circuits if the impulse noise level can be controlled.

All Lenkurt 45-class equipment, if installed in a suitable plant layout, can meet SAGE requirements. In addition, it has certain particularly desirable features—significant when used for data transmission—such as channel regulation and carrier frequency interconnection.

THE Lenkurt Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6, NO. 9 SEPTEMBER 1957

Some Practical Aspects of

DIGITAL DATA TRANSMISSION

Over Telephone Facilities

The increasing demand for data circuits is becoming of great importance to telephone and telegraph operating companies. In the future, data transmission may become a major source of revenue. A previous issue of the Demodulator (April 1956, Vol. 5 No. 4) discussed the theoretical nature of data and the basic methods by which it is transmitted.

This article discusses some of the practical problems of transmitting high-speed data over ordinary telephone facilities and lists briefly some of the data transmission systems presently employed or proposed.

Digital data is information in the form of successive discrete signals. Present-day concern and controversy over data transmission often obscures the fact that its transmission is as old as man. Fig. 1 illustrates a few of the early methods of digital data transmission. Variations of some of these are still in use.

The simplest message that can be transmitted by digital means is the information necessary for a receiver to make a correct choice between two different equally possible events or conditions: yes or no, on or off, black or white, something or nothing, 0 or 1, etc. Since the two possible messages correspond to the two symbols in the system of numbers called the binary digit system, a unit of information based on two symbols came to be called a binary digit and was abbreviated *bit*.

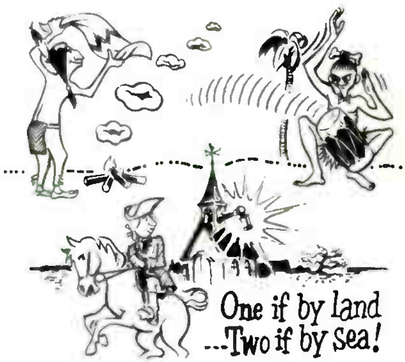


Fig. 1. Early forms of digital data transmission.

Modern digital data transmission began in 1832 when Samuel F. B. Morse invented the telegraph. Machine transmission and reception occurred just before the beginning of the Twentieth century. The earliest machine transmitters and receivers were teletypewriters and dial telephones. Until recently, the only other major uses of digital data transmission have been for telemetering, remote control, and indication.

During the past ten years, the tremendous increase in electronic business machine and computer usage has raised the question, "Why can't business machines and computors 'talk' directly to each other over telephone circuits without converting the messages into human language?"

The answer is, "They can to a limited extent." In fact, it seems probable that "conversation" between business machines or computors will someday become as commonplace as telephone calls between people.

Transmission Rates

The practical aspects of digital data transmission over telephone channels are related to two theoretical questions. First, what is the theoretical maximum rate at which data that can be transmitted without error over a channel of a certain bandwidth and signal to noise ratio? Second, what are the characteristics of telephone channels that affect data transmission?

C. E. Shannon provided the answer to the first question by a mathematical analysis of the nature of data or information. His results, illustrated graphically in Fig. 2, show that the maximum data transmission rate can be increased by increasing the bandwidth or signal to noise ratio. However, to achieve the transmission rates indicated, three conditions must be met.

First, the transmission medium must be distortionless. This applies to both phase and amplitude distortion. Second, the noise in the channel must be random (equal noise power across the frequency band). Third, the encoding system must be so complex that no possible combination of noise impulses will ever cause erroneous information to be transmitted. None of these conditions can be met exactly or even closely ap-

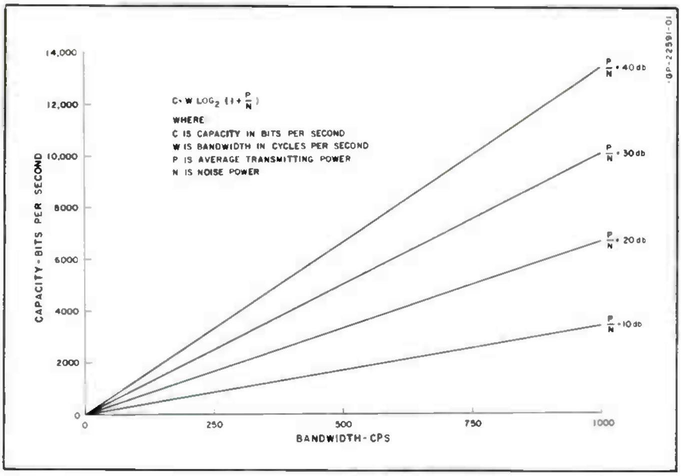


Fig. 2. The relationship between maximum transmission rate (capacity), bandwidth, and signal-to-noise ratio.

proached with present day techniques. Hence, the information rates attainable are much slower than the maximum rates indicated by Fig. 2.

Telephone Channel Characteristics

If telephone channels were designed specifically for data circuits, strong emphasis would be placed on achieving characteristics that would permit a maximum transmission rate with minimum error. Such characteristics would include small phase distortion, freedom from noise—especially impulse noise, level stability, and frequency stability.

These characteristics are also desirable for speech transmission, but the lack of them does not necessarily prohibit successful transmission. For example, we talk on circuits that are so noisy they cause us to repeat words, but we still get the information through. We talk on circuits with high impulse noise peaks but never know they are there. We talk on circuits with enormous amounts of phase distortion, but our ears fail to detect it. We talk on circuits that attenuate the message so greatly that we have to shout, but we adapt ourselves to the condition.

The reason we can talk on such imperfect circuits is that speech is very redundant. That is, it contains much repetition and extraneous information not needed to convey simple messages. This has resulted in the use of many different types of transmission facilities with widely varying characteristics. Figures 3 and 4 show some of the amplitude and phase distortion characteristics of different telephone facilities in common use.

Data Transmission

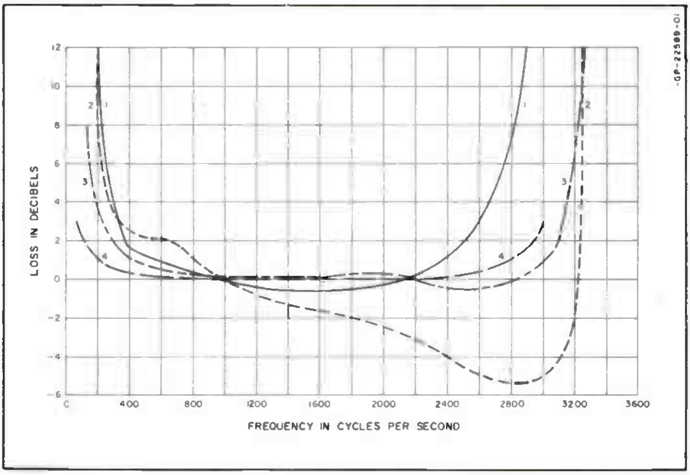
When telephone circuits are used for data transmission, the imperfections which do not greatly affect voice transmission have to be carefully considered.

For example, noise in its various forms has a much greater effect on data circuits than it does on voice circuits. A commonly stated noise objective for long-distance telephony is that a circuit should have an *average* noise of no more than 38 dba (a signal power to noise ratio of about 44 db). This means that the circuit will have more than 38 dba of noise half of the time and less half of the time. For short periods of time, the noise may greatly exceed 38 dba.

Such performance is entirely adequate for speech transmission but could cause serious errors in the transmission of digital data. To avoid errors from noise, data circuits must be designed on the basis of adequate signal-to-peak noise ratio rather than signal-to-average noise ratio. If the noise present on a circuit is impulse noise (a common type on wire and cable circuits) the noise peaks may be of such great amplitude that they make data transmission impractical.

One of the most serious difficulties encountered in transmitting data over telephone circuits is the unequal transmission velocities of different frequencies. This is called phase distortion or sometimes phase delay. The human ear is relatively insensitive to phase distortion and does not recognize it as a speech impairment. But to data circuits, this is one of the most severe types of distortion that can occur. The reason for this is that the short bursts of energy necessary for data pulse transmission

Fig. 3. *Attenuation - frequency characteristics of typical carrier channels.*



contain many frequencies having fixed relationships to each other in time. If some of these frequencies are transmitted at a slower rate than others, the received pulse does not have the same shape as the transmitted pulse and errors can occur—especially if the pulses are short.

Compandors and Echo Suppressors

The interfering effects of noise, crosstalk, and echo on speech are often reduced by the use of compandors and echo suppressors. These devices take advantage of certain peculiarities of human conversation. Their use, however, often renders a circuit unfit for many types of data transmission.

Compandors give an apparent reduction in noise on voice circuits by increasing the circuit loss between syllables and words of speech. If amplitude-modulated (on-off) type data signals are transmitted through a compandor, some of the pulses may be badly distorted by the varying loss characteristic. Frequency-shift type data signals are not appreciably distorted by a compandor since their power level remains essentially constant. But, neither are they provided with any signal-to-noise ratio improvement.

Echo suppressors are frequently used on long circuits to prevent the echo of reflected speech from annoying the talker. They are simple devices which short-circuit the return path when a person is talking. In effect echo suppressors permit transmission in only one direction at a time. They limit the use of a circuit for data transmission since they do not permit a data receiver to ask for the repeat of a message in which an error has been detected.

Practical Data Transmission Systems

Despite the shortcomings of present-day telephone facilities for digital data transmission, there are a number of practical systems in use or about to be introduced. Generally, these systems can be divided into two broad catagories: (1) slow-speed narrow-band systems originally designed for tele-

TABLE I

Typical Data Transmission Systems

Type	Number of Channels	Spacing	Type of Keying	Total Band Required	Bits per Second	Notes
BASIC TELEGRAPH SYSTEMS						
A. T. & T. Lenkurt Signal Corps	16	170 cps	F-M & A-M	340-3060 (16 ch)	74 per ch 1187	100 wpm/channel
Western Union Telegraph	20	150 cps	F-M	300-3300 (20 ch)	57 per ch 1136 74 per ch 148[illegible]	75 wpm/channel 100 wpm/channel
Collins Synchronous Telegraph	Up to 40	110 cps for each 2 ch.	Phase-shift	550-2750 delay equalized (40 ch)	74 per ch 2968	All channels must be synchronized. Makes use of special synchronizing signal to effectively regenerate all channels. All channels must originate and terminate together.
VOICE CHANNEL DATA SYSTEMS						
IBM Card-to-Card Transceiver	4	450 cps	A-M	650-2450 (4 ch)	180 bits/channel Transmits 11 cards per minute	Used on leased toll circuits with echo suppressors removed. Uses code which includes two error checking bits for each 6 information bits.
A. T. & T. (experimental)	1	--	F-M	700-1600	750	Designed to work over any telephone circuit which may be dialed. Error rate between 1 in 10, 000 and 1 in 100, 000.
A. T. & T. Teletypesetter	1	--	A-M	1000-2800	510	Phase delay equalization required.
SAGE	1	--	A-M (Vestigial Sideband)	500-3000	1600	Phase delay equalization required from 1000-2500 cps. Sensitive to impulse noise. Error rate of less than 1 in 100, 000.
Western Union Sub-band	2	1500	F-M	300-3300 (2 ch)	1800 delay equalized 1500 unequalized	Under development. For operation on same voice channels as Western Union Telegraph Systems. These voice channels are not the ordinary "dialed-up" channels but are special for W. U. Telegraph.
Signal Corps AN/TSQ-7 AN/TSQ-8	1	--	A-M Double Sideband	975-2500	750	Delay equalization required.

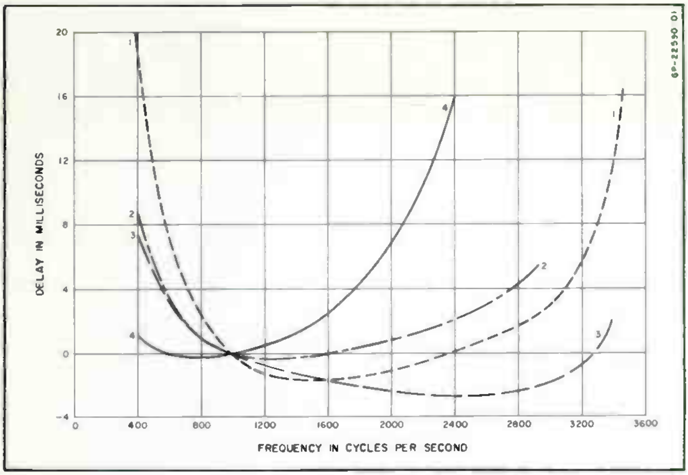


Fig. 4. Delay characteristics of typical telephone channels.

typewriter service, and (2) high-speed systems using a major part of the speech channel bandwidth.

Among the slow-speed systems there are three transmission rates in common use: 45.5, 57, and 74 bits per second. These correspond to teletypewriter speeds of 60, 75, and 100 words per minute.

Slow-speed data transmission systems can be used for many applications where the time required to transmit the information is not too important. Signals from the transmitting business machine or computor can be stored on paper or magnetic tape and then retransmitted at will. At the receiving end, the information can again be stored on tape and fed into the receiving computor or business machine at any convenient rate and time.

The most recent data transmission methods involve the use of all or a major portion of a voice channel. They usually require that the phase delay of frequencies within the channel be equalized—either by design or auxiliary devices—and that channel noise be maintained below certain levels. One system, still in the experimental stage, is being designed to operate over the majority of existing telephone channels without phase delay equalization or special noise treatment. Table I gives the basic characteristics of the typical data systems of the various types.

Conclusions

The immediate future should see greatly expanded use of existing communications channels—both telephone and telegraph—for dialed data circuits. The use of "private line" circuits requiring special engineering will also become more common. A large increase in private line usage is already evident in applications such as SAGE circuits for the Air Force and data circuits for brokerage houses, railroads, airlines, and banks.

In the distant future, when more complex business machines and computors are put to use, higher data transmission rates than any shown in Table I will become economically desirable.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 7 NO. 2 FEBRUARY 1958

AM VERSUS FM

For Digital Data Transmission

Information can be transmitted long distances most efficiently by selection of the proper vehicle to act as the information carrier. The fastest and most versatile carrier is an alternating sine wave of electrical current or radio energy.

This article includes a discussion of two techniques, amplitude modulation and frequency modulation, which are commonly used for imparting information to a sine wave. The advantages and disadvantages of each technique are described for different applications and different transmitting conditions.

Digital data such as alphabetical, numerical or symbolic information is normally transmitted over long distances by varying (modulating) one or more of the characteristics of a sine wave of electric current. The characteristics that can be varied are: (1) amplitude, (2) frequency, and (3) phase with respect to some reference wave.

The two most common techniques are *amplitude modulation* (AM) and *frequency modulation* (FM). Frequency modulation is sometimes called *frequency shift* (FS). Other techniques combining two or more types of modulation have also been used to a limited extent.

Equipment for both AM transmission and FM transmission is readily available from a number of manufacturers. Some supply only one type. Others offer both types. Questions that are often asked are: Which type is best? and why?

Before either question can be answered, a basis for comparison must be decided upon and the nature of amplitude modulation and frequency modulation must be examined.

Basis for Comparing AM and FM

There are many ways to compare AM with FM. But perhaps the most useful comparison would involve the performance and cost of AM and FM channels under equal operating conditions. Performance of a communication channel is measured primarily by (1) transmitting rate and (2) error rate. Both of these rates are affected by

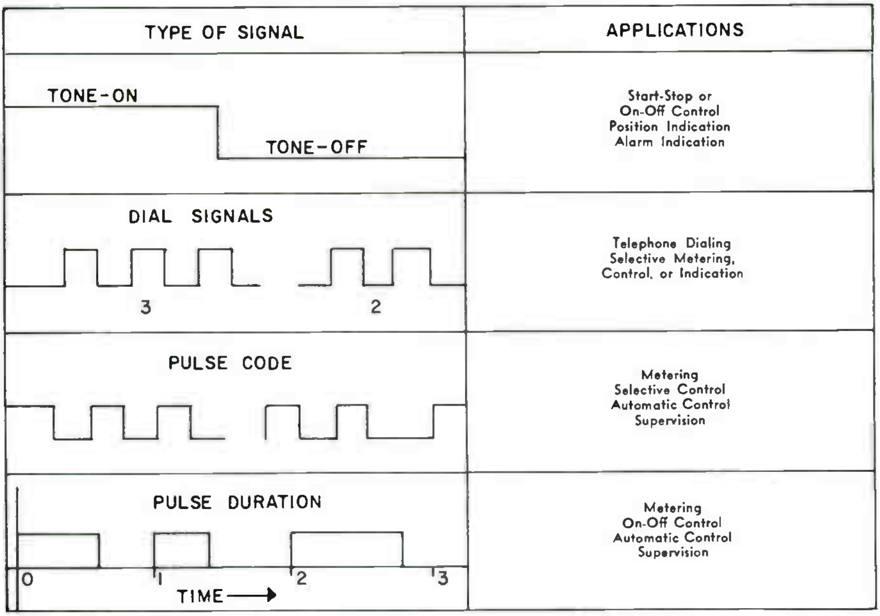


Fig. 1. Four methods of encoding on-off signals for the transmission of numerical, alphabetical, and symbolic information.

channel bandwidth, signal power, noise, and distortion. These, in turn, affect the cost of equipment used for transmitting information.

If AM and FM transmission are examined on the basis of equal bandwidth per channel, equal peak signal power per channel, and equal amounts of noise, then the resultant error rates, equipment costs and transmission rates should give an indication of which type of modulation is best for different situations.

Amplitude Modulation

A first look at amplitude modulation might indicate that the easiest way to impart numerical information to a wave would be by varying its amplitude in direct proportion to the number being transmitted. A more careful look would reveal a number of reasons why this is impractical. Perhaps the most obvious is that any variation in level caused by the transmission medium would result in the reception of different numbers than those transmitted.

To avoid this problem, numerical, alphabetical and symbolic information is converted to on-off or pulse type signals. Normally, the information is first converted to d-c pulses which are then translated into amplitude modulated signals by on-off switching of a single-frequency tone. Various types of d-c pulses used to modulate a sine wave are shown in Figure 1. Regardless of the code by which a tone is switched on and off, the factors affecting transmission remain the same.

In the unmodulated state, a sinusoidal tone has but one frequency component. But when modulated with a rectangular wave, it has many. They include: (1) the carrier frequency, (2) modulation products of the carrier frequency and the fundamental frequency of the rectangular wave, and (3) modulation products of the carrier frequency with all the harmonic components of the rectangular wave. Figure 2 shows the frequency components of an on-off modulated tone.

Since these frequencies cover a relatively broad spectrum, to transmit all of them would be uneconomical. If the carrier and first two side frequencies are transmitted, the remaining modulation products can be eliminated by sending and receiving bandpass filters with little degradation of received signal intelligibility but with considerable savings in frequency spectrum.

Restriction of the transmitted bandwidth causes the envelopes of the received tone pulses to be rounded.

Distortion

If all the pulses received were of the same amplitude, rounding of the wave shape would cause little impairment of reception. However, if their amplitude changes because of changing transmission conditions, or if the signal-to-noise ratio is low, the receiver may misread the received intelligence and give an erroneous output.

If the pulse is reduced in size because of reduced signal level, the time interval during which the receiver is able to recognize the pulse is also reduced. If the signal level is higher than normal, the receiver may detect the pulse for a longer period than it should. The over-all effect of pulse amplitude variations is the shortening or lengthening of individual pulses. This is called *bias distortion* and is illustrated in Figure 3. A similar type of distortion is called *characteristic distortion.* It can occur when pulses of different lengths are transmitted. Under conditions of reduced signal level, short pulses tend to be shortened a greater percentage of their length than long pulses. If severe, characteristic distortion can cause errors in pulse length telemetering systems or garble telegraph messages.

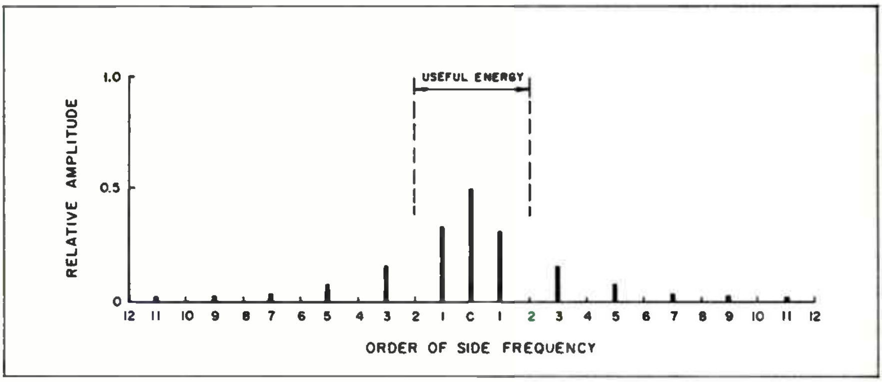


Fig. 2. Frequency spectrum of a square-wave amplitude modulated tone.

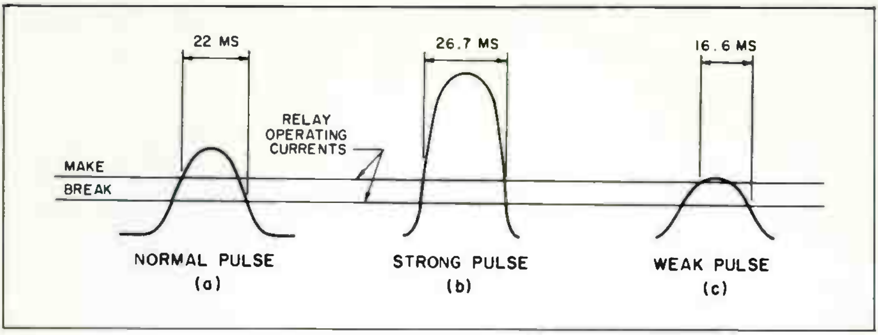


Fig. 3. Typical received telegraph or dial pulses. In figure 3b, an increased pulse amplitude increases the effective pulse length and causes positive bias distortion. A weak pulse (figure 3c) shortens the effective pulse length and causes negative bias distortion.

When random causes shorten, lengthen or delay individual pulses, the distortion is said to be *fortuitous*. Line hits, lightning, noise or any other source of spurious impulse can cause fortuitous distortion.

The usual method of reducing bias and characteristic distortion is to automatically regulate the output of the pulse receiver. All high quality AM telegraph systems have level compensating circuits. However, the circuits increase equipment cost and are not completely foolproof. They are relatively ineffective against fortuitous distortion.

Noise in AM Circuits

Next to level variations, noise is the greatest source of trouble in amplitude modulation systems. If the noise level is high, a receiver may not be able to distinguish between the presence and absence of a tone.

When ordinary telephone circuits are used as the transmission medium the noise characteristics may be such that the circuit is perfectly usable for voice transmission but too noisy for reliable data or telemetering circuits. Voice circuit quality is based on *average* noise levels over a period of time. Sharp noise impulses or high noise levels for short periods of time are not particularly objectionable to the human ear. However, if a voice circuit with impulse noise present is used to transmit amplitude modulated tone pulses, errors in the received information may be numerous.

Frequency Modulation

In a frequency modulation system, a constant amplitude, high-frequency carrier is caused to vary in frequency in accordance with the modulating signal. D-c pulses may be used to vary the frequency or, under carefully controlled conditions, the frequency may be made to vary directly with a numerical quantity being transmitted. For example, some telemetering systems produce a low-frequency tone that changes frequency in direct proportion to the quantity being metered (pressure, flow, etc.).

When AM and FM systems of equal bandwidth are compared, the transmission rates of the two systems will be about equal when the frequency shift of the FM system is equal to the pulse repetition rate. However, for the same unmodulated carrier powers, the transmitted sideband power (signal power) of an FM channel is about 3 db greater than that of an AM channel. This is evident by comparing the FM sidebands in Figure 4 with the AM sidebands in Figure 2.

Restriction of the transmitted bandwidth causes rounding of the received information pulses the same as in AM. Level variations, unless removed, cause bias and characteristic distortion. However, they are much more easily removed from FM than from AM. A simple limiter circuit in the receiver will completely remove all but the most extreme fluctuations in signal level.

Unlike AM transmission, the signal presented to an FM receiver is a continuous electrical wave instead of short bursts of tone. The on-off conditions are determined by the frequency of the continuous wave. For the *on* condition, the wave will have one frequency while for the *off* condition it will be shifted to a different value. But the wave remains continuous at all times.

This characteristic of FM is extremely valuable in data transmission, telemetering, remote control, and supervision. It adds another bit of information to the transmitted wave. It can be used to operate alarms or initiate an equipment transfer in the event of complete signal failure.

Noise in FM Circuits

One of the most outstanding characteristics of frequency modulation is its performance in the presence of noise. In general, noise is of random frequency, amplitude, and phase. When impressed on a carrier wave, it will cause the wave to vary somewhat in amplitude and phase. The FM receiver, however, responds only to the variation in phase. Because of the limiter, it is insensitive to the amplitude variations.

Numerous experiments have shown that FM is at least 3 db less susceptible than AM to noise near the failure point and up to 10 db less susceptible at noise

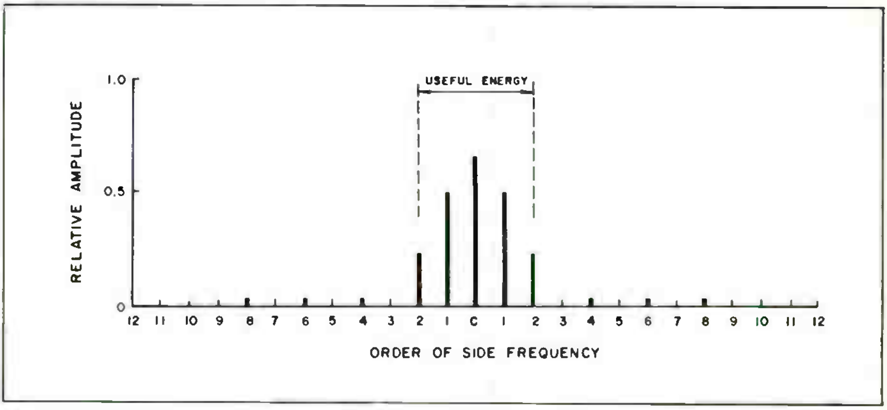


Fig. 4. Frequency spectrum of a square-wave frequency modulated tone. Pulse repetition rate is equal to maximum carrier frequency swing.

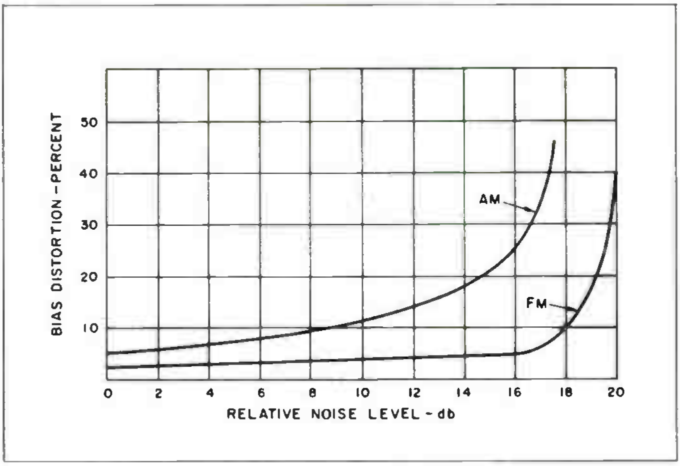


Fig. 5. A comparison of FM and AM in the presence of thermal noise.

levels below the failure point. Figure 5 shows typical curves of distortion versus relative noise level for approximately equivalent AM and FM data channels.

Frequency Stability

While FM signals can be transmitted through noise more readily than AM signals and can be made virtually immune to level variations, they tend to be sensitive to any difference between the unmodulated received carrier wave and the center frequency of the FM discriminator or detector. This difference can occur in several ways: (1) frequency drift of the carrier oscillator, (2) misalignment of the discriminator, (3) lack of frequency synchronization between the transmitter and receiver if transmitted by single sideband suppressed carrier or radio.

If the transmission medium is radio and one or both of the terminals is moving with respect to the other (for example, an airplane or missile), an apparent frequency shift will be observed depending on the relative motion between the transmitter and receiver. This is called *Doppler effect* and is an important consideration when using frequency modulation for air-to-air or air-to-ground communications.

The effect of frequency drift or discriminator misalignment in FM systems is much the same as level variations in AM systems. It results in bias distortion. A shift in one direction results in positive bias distortion and in the other, negative bias distortion.

Frequency drift of the carrier oscillator and discriminator misalignment are seldom a problem. Oscillator stability at the frequencies normally used for telemetering, telegraph, or signaling can easily be maintained within a fraction of a cycle. Discriminator alignment, once set, does not appreciably change with time or temperature. The chief difficulties are encountered when the data channel is transported on a single sideband suppressed carrier telephone channel or radio circuit.

Single sideband carrier systems may employ two or more free-running crystal oscillators. Depending on the system these may generate frequencies up to one megacycle or more. Any drift in a single sideband carrier oscillator can show up as a frequency error in the data channel.

FM Versus AM

Both amplitude modulation and frequency modulation have their advantages and disadvantages. To say that one is always better than the other is unrealistic. Both are capable of reliable transmission under the proper circumstances. The choice between the two depends on the particular application.

With the exception of sensitivity to frequency instability, FM is technically superior to AM. However, when frequency instability does exist and cannot be readily corrected, an AM system is the logical choice. On the basis of equal channel bandwidths, equal peak signal powers per channel, and equal amounts of channel noise, the following comparisons can be made:

1. Cost—FM equipment usually costs slightly more than AM equipment.

2. Transmission rate—Under ideal conditions, both systems are capable of the same transmission rates. However, since the carrier wave in FM transmission is continuous, its presence or absence provides an additional bit of information.

3. Effect of noise—FM will operate satisfactorily through from 3 to 10 db more noise than AM.

4. Effect of received level variations—FM is unaffected by all but extreme level variations. In all transmission, uncompensated level variations (some are bound to occur) cause bias and fortuitous distortion.

5. Effect of frequency instability—AM is superior to FM provided the frequency drift is not too great. Where frequency instabilities can be corrected, FM is not at a disadvantage.

A good indication of the general superiority of FM is its rapidly growing use. The Western Union Telegraph Company has been using FM transmission exclusively for many years. Since 1956, the Bell Telephone System has installed thousands of channels of FM telegraph.

FM signaling was chosen for Lenkurt 45-Class carrier systems because its immunity to noise permitted the signaling tones to be transmitted at much lower levels, without loss of reliability. It also permitted the signaling carrier to be used for channel level regulation in addition to its primary function.

In commercial radio telegraphy, FM is in almost universal use. The wide level variations caused by fading tend to make AM radio telegraphy unreliable. Even the best regulating circuits will not completely eliminate the rapid level changes characteristic of long range radio reception.

Despite the growing popularity of FM for telegraphy, data transmission, and telemetering, there are still applications where AM is best. During the past 20 years, many thousands of carrier channels have been installed that do not have sufficient end-to-end frequency stability for FM transmission. Yet they are perfectly suitable for voice communication and have relatively stable transmission levels. When pulse type data is to be transmitted over these channels, amplitude modulation is by far the best technique.

THE *Lenkurt* Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 3 NO. 2 FEBRUARY, 1954

TRANSMISSION OF FDM* CHANNELS

by FM Point to Point Radio

Point-to-point FM radio systems provide a convenient method for transmitting large numbers of telephone channels over relatively long distances. FM systems have both operational and economic advantages over comparable AM systems because they can transmit large amounts of information with less power and better signal-to-noise ratio.

In this article some of the factors involved in the use of FM radio to transmit frequency division multiplexed telephone channels are discussed. Of particular importance are bandwidth, modulation index, noise, and power considerations.

In any type of multichannel radio system, the information handling capacity is determined largely by the type of multiplex equipment used and by the bandwidth which the radio equipment can transmit. Important characteristics of the multiplex equipment are frequency spectrum per channel and the number and power level of constant signaling or carrier tones. For the radio equipment the chief limiting factor is the effect of noise on the carrier channels.

For comparable microwave systems with equal transmitted powers and bandwidths, a frequency modulated system with single sideband suppressed carrier multiplexing equipment can economically carry more information with less transmission impairment than other combinations of radio and multiplexing equipment available today.

Since it is desired to transmit intelligence without introducing objectionable noise, the factors which determine the effect of noise on the transmitted signal must be carefully analyzed and controlled.

Frequency Modulation

In a frequency modulation system, a constant amplitude high frequency carrier is caused to vary in frequency in accordance with a modulating signal. The maximum deviation of the carrier frequency is proportional to the maximum amplitude of the modulating frequency.

The essential elements of an FM

* FREQUENCY DIVISION MULTIPLEX

radio system are a means of varying the carrier frequency in accordance with the modulating signal and a means of recovering the modulating signal at the receiver. Characteristics of the 'modulator' and 'detector' largely determine the effect of noise and crosstalk on the transmitted intelligence.

When a radio frequency wave train is amplitude modulated by a single sine wave, the resultant wave contains only three frequencies. These are the carrier frequency and the sum and difference of the carrier frequency and the modulating frequency (side frequencies). When an r-f wave train is frequency modulated, however, the resultant wave will contain a large number of side frequencies. These are shown in Figure 1.

Not all of the side frequencies are of equal importance. Mathematical analysis of an FM wave shows that as the modulating frequency increases, the desired intelligence becomes concentrated in fewer sidebands. Therefore, for a given system with a fixed maximum frequency deviation, the number of sidebands required for effective transmission increases as the frequency of the modulating signal decreases.

Mathematical analysis also indicates that the bandwidth which an FM system must transmit and receive to keep distortion within tolerable limits is approximately twice the sum of the maximum frequency deviation and the highest modulating frequency. For example, consider a system with a 900 megacycle carrier frequency and a maximum frequency deviation of 300 kc transmitting 15 kc and 300 kc signals. The sidebands due to the 15 kc signal would be 900 mc ±15 kc, 900 mc ±30 kc etc. The important sidebands of the 15 kc signal would be within 900 mc ±315 kc. Only two sideband pairs would be required for the 300 kc signal -- at 900 mc ±300 kc and at 900 mc ±600 kc. This means that the total bandwidth required for the system would be 1.2 mc. In actual practice, transmitting and receiving equipment is designed for somewhat wider bandwidths to further reduce distortion and allow for possible misalignment.

An inherent characteristic of FM systems is that the effects of noise can be reduced by increasing the bandwidth. However, because of practical considerations, particularly the need to conserve frequency spectrum, equipment designers attempt to keep bandwidth requirements as low as possible.

Modulation Index and Deviation Ratio

When the modulating signal to an FM transmitter is the output of a multichannel frequency division carrier system, it contains a very broad band of frequencies of random phase and amplitude characteristics. For such a modulating signal, the ratio of the maximum frequency deviation of the r-f carrier to the highest frequency contained in the modulating band is usually called the deviation ratio.

When a carrier is frequency modulated by a pure sine wave, the ratio of the maximum frequency deviation of the r-f carrier to the frequency of the modulating sine wave is often called the modulation index. Since the modulating frequencies of an FDM system consist of a number of 4-kc channels located at different bands in the spectrum, the channels occupying the higher end of the spectrum will have a smaller modulation index than the lower frequency channels.

This can be seen by considering modulation in a typical wideband FM radio system transmitting a large number of frequency division multiplexed channels. In a Lenkurt Type 72B system, for example, a group of carrier channels occupying frequencies up to 300 kc are spaced at 4-kc intervals. In a typi-

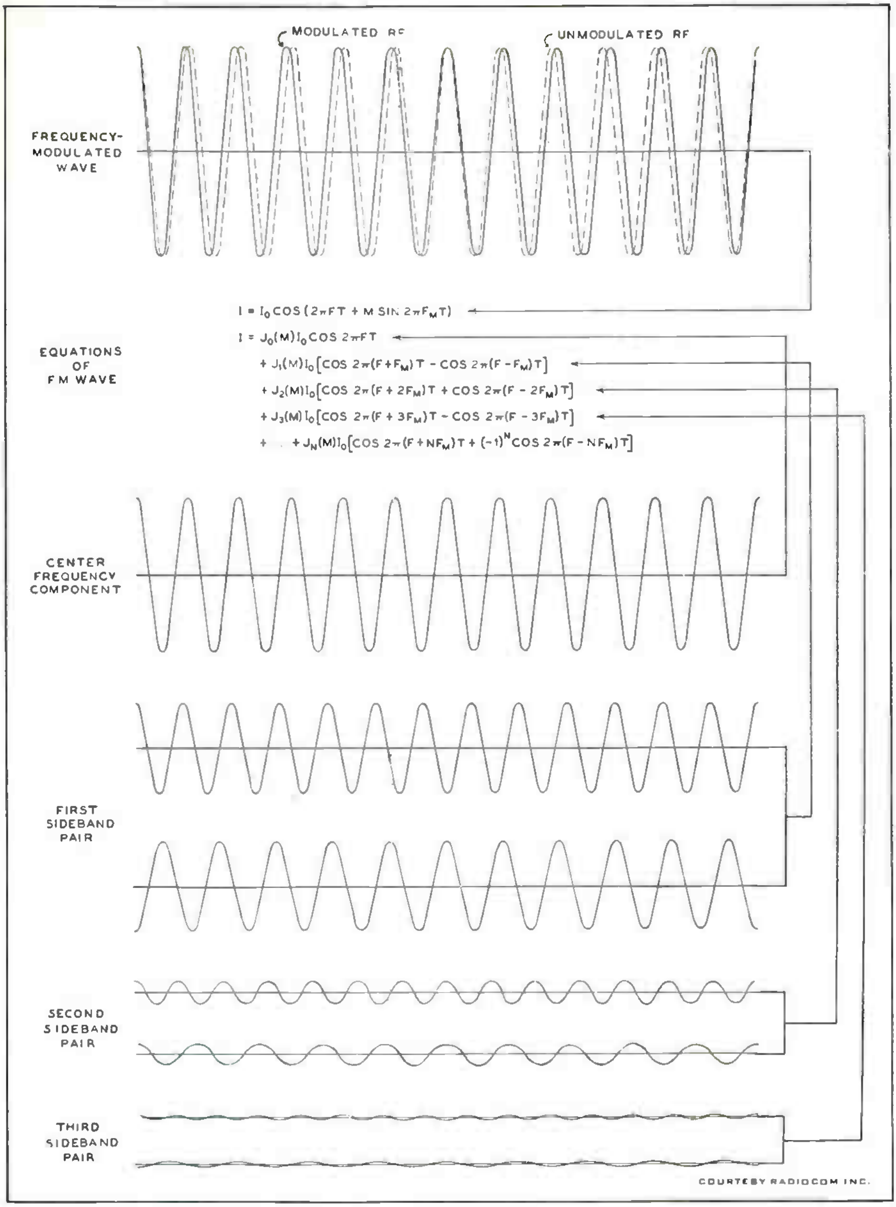


FIGURE 1. The Frequency-Modulated Wave and its Components When the Modulation index is 1.

cal system, channel 1 might occupy 10-14 kc and channel 72 might occupy 296-300 kc.

The maximum frequency deviation in a 72B system is 500 kc. Since the highest frequency of the modulating band is 300 kc, the modulation index of the highest frequency is 1.67. However, the modulation indices of the other individual channels are greater than 1.67.

The total frequency deviation of the system is related to the sum of the amplitudes of the signals in all of the individual channels. If the system were arranged so a signal of equal amplitude in any channel caused equal deviation, either a 15-kc signal or a 300-kc signal of equal amplitude might cause a deviation of 500 kc. However, the modulation index of the 15 kc signal would be 500/15 or 33.3, while the modulation index of the 300 kc signal would only be 1.67.

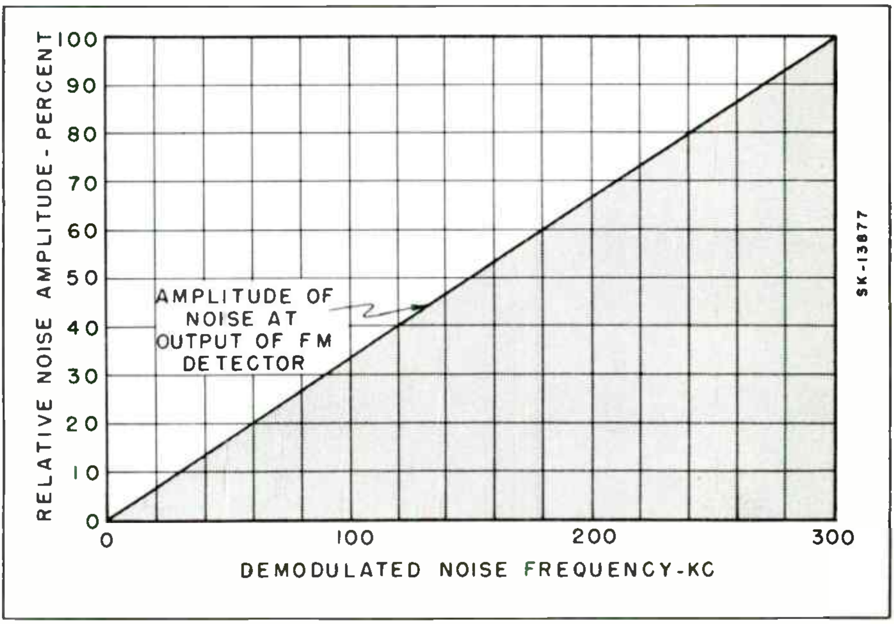
Just as greater bandwidth will decrease the effects of noise, so will higher modulation index. In fact, with greater r-f bandwidth for a system handling a given modulation band, higher modulation indices are possible. Modulation index is, of course, limited by the same factors which limit bandwidth.

Noise

Noise in an FM system has several possible sources. If the modulating signal has a noise component, that component will also appear in the output of the FM detector. However, if the noise is imposed on the modulated radio signal either from external sources or in the radio equipment, its effect on the received signal is much reduced by the frequency modulation process.

In general, noise is of random frequency and amplitude. When a noise signal is impressed on the carrier, it will cause the carrier frequency to vary somewhat in amplitude and phase. The FM receiver, however, responds only to the variation in phase, being insensitive to the amplitude variations. If the modulation index is sufficiently large, phase change or frequency swing caused by the modulating signal is thousands of cycles whereas the phase shift caused by the noise will generally be only a few cycles. The noise signal, therefore, causes only a small amount of frequency swing compared to the desired signal. If the modulation index is large enough, the noise can have an amplitude nearly one-half as large as the carrier without creating large amounts of interference.

FIGURE 2. The interfering noise frequencies, when detected, have amplitudes that vary directly with separation between the interfering noise frequency and the carrier.



If the frequency of the interfering noise signal is close to the unmodulated carrier frequency, the effect of the noise is much less than if the noise frequency occurs near the maximum frequency deviation of the system. That is, the effect of noise increases as its frequency is separated from the frequency of the carrier.

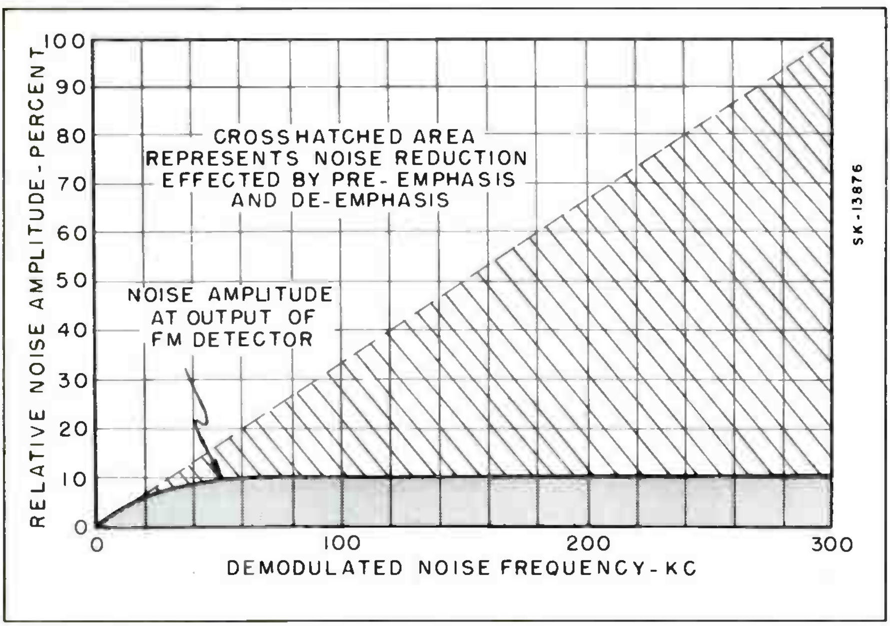
In the detecting circuits of an FM receiver, the amplitude of the detected signal is proportional to the deviation of the received radio frequency carrier. Since the detector circuits translate frequency deviation of the carrier into amplitude of the received signal, noise voltages with frequencies near the maximum deviation of the system appear in the receiver output with higher amplitudes than would noise voltages with frequencies near the carrier. When plotted graphically, noise amplitude at the output of an FM receiver has the triangular shape shown in Figure 2.

To compensate for the 'triangular noise spectrum' of an FM system, it is common to adjust the response of amplifiers in the radio equipment so that the higher modulating frequencies have a greater amplitude than lower modulating frequencies. This process, commonly called pre-emphasis, tends to create a uniform signal-to-noise ratio across the entire modulating frequency spectrum. After the signal has been detected by the receiver, the recovered signal is passed through a special network that de-emphasizes the higher frequencies and restores all frequency components of the received information to the original relative level. The results of pre-emphasis are shown in Figure 3.

Power Considerations

In any given FM radio system, the amount of power radiated from the transmitting antenna remains constant. Changes in input power affect only the deviation of the carrier frequency. If the total power

FIGURE 3. Pre-emphasis of the higher modulating frequencies tends to make the signal-to-noise ratio uniform for all carrier channels.



of all channels should exceed the design limitations of the system to cause frequency deviation greater than the maximum designed for, the radio equipment would be overloaded. This would cause distortion products (noise) to be introduced into the transmitted intelligence.

It is, therefore, desirable in a multichannel radio system to keep the necessary constant tones as low as possible. At the same time it is desirable to maintain the speech level high to obtain the maximum possible signal-to-noise ratio at the receiver output.

The use of low constant tones reduces the total power content of the modulating wave and allows much more effective use of the speech signals. Consequently, an FM radio system will provide better speech transmission quality as the power levels of the constant tones are reduced.

With single sideband multiplexing equipment designed so carrier leak is effectively suppressed, the constant tones necessary for transmission of dial signals are of prime importance. By reducing these tones to the lowest practical power level, the best possible overall performance of the entire system can be obtained. Where carrier frequencies or high level signaling tones are transmitted or where double sideband carrier systems are employed, the performance of the entire system will be adversely affected by the additional constant tones.

Conclusions

Signal-to-noise ratio of the carrier channels at the output of the radio receiver determines the usefulness of a multichannel radio system. Because so many factors can influence the effect of noise, any FM system represents a compromise between theoretical principles and practical considerations.

However, three important factors apply to most situations. First that the bandwidth and deviation ratio, which determine the basic noise characteristics of a system, are selected by the system designer to provide the best compromise between theoretical requirements and practical considerations. Second, that for any given combination of radio and multiplexing equipment pre-emphasis can be employed to obtain optimum performance. And, third, that once the characteristics of the radio equipment are fixed, signal-to-noise ratio for individual channels can be improved by using the lowest practical constant tone power levels.

THE *Lenkurt* Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6, NO. 3 MARCH, 1957

DETECTION OF FM SIGNALS

Day by day, radio as a transmission medium grows more important to the telephone industry. A major factor in this growth is the superior quality of transmission that can be achieved through the technique of frequency modulation.

This article describes some of the common methods used to demodulate FM signals.

The function of any detector (demodulator) is to recover the amplitude and frequency of the modulating signal that enters the input to a transmitter. An amplitude-modulated (AM) wave consists of a constant-frequency carrier which varies in amplitude as the amplitude of the modulating signal varies. The over-all shape or *envelope* of the AM wave therefore takes on the shape of the original signal.

Detection consists of rectifying this wave and filtering out the carrier frequency. The resulting wave is a varying d-c wave which has the same shape as the original signal. When the d-c component is eliminated by a capacitor, the final wave duplicates the original modulating signal.

However, the nature of a frequency-modulated (FM) wave is quite different and its detection requires an additional process.

Nature of an FM Wave

An FM wave varies in instantaneous frequency above and below a certain center (carrier) frequency. The amplitude of the modulating signal determines how much the FM wave varies from the center frequency. And the frequency of the modulating signal determines how fast these variations occur.

For example, suppose a sine wave signal of peak amplitude one volt and frequency of 200 kc modulates a center frequency of 100 mc. As the signal starts its cycle, its amplitude is zero. The

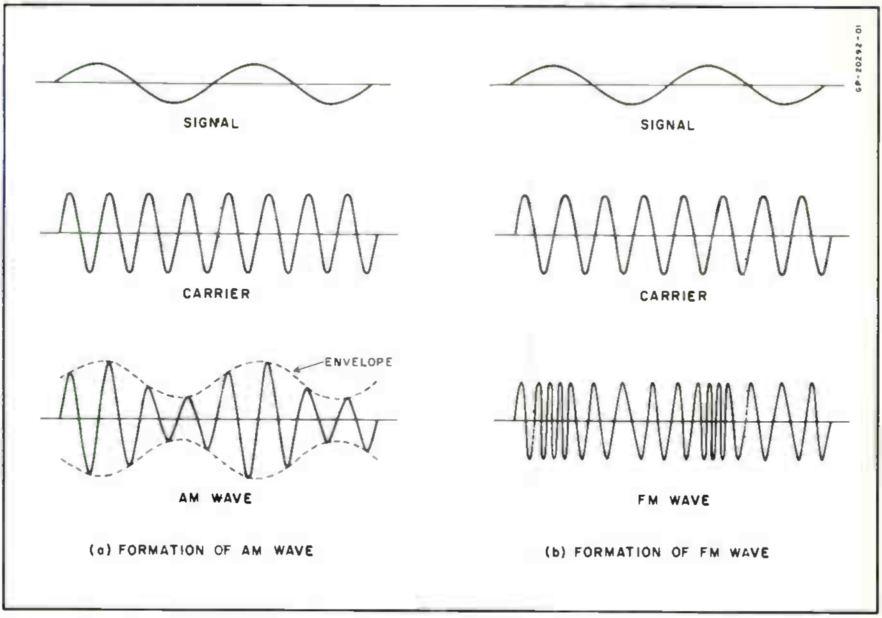


Fig. 1. Formation of AM and FM waves from a sine wave signal and a carrier.

carrier frequency is then an unmodulated 100 mc. As the amplitude of the modulating signal begins to rise, the frequency of the carrier wave begins to increase. It reaches its highest frequency when the amplitude of the signal is at its peak value of one volt. Then as the amplitude of the signal begins to decrease, the frequency of the carrier wave also begins to decrease. When the signal has completed one-half its cycle, its amplitude is again zero. The frequency of the carrier wave is again at the center frequency of 100 mc.

On the second half of its cycle, the amplitude of the signal decreases toward its negative peak. The frequency of the carrier wave then decreases below the center frequency. At the negative peak of the signal, the carrier-wave frequency reaches its lowest value. Then as the signal swings upward toward zero, the frequency of the carrier wave increases. When the signal has completed one cycle and its amplitude is again zero, the frequency of the carrier wave will be back to the center frequency of 100 mc.

If, in the above example, one volt caused a 100-kc change in frequency, the FM wave would go from its center frequency to a maximum frequency of 100.1 mc (100 kc above the center frequency), down past the center frequency to a minimum frequency of 99.9 kc (100 kc below the center frequency) and back up to the center frequency again. Since the frequency of the signal is 200 kc, this would happen at a rate of 200,000 times per second.

The amount by which the frequency differs from the center frequency is proportional to the amplitude of the signal. If the amplitude of the signal is doubled, the amount of frequency swing will double. The frequency would then change from 100 mc to 100.2 mc to 99.8 mc to 100 mc at a rate of 200,000 times per second.

If the peak amplitude remains at one volt and the signal frequency is doubled, the frequency would still vary between the limits of 100.1 and 99.9 mc. But the rate at which it varied would increase to 400,000 times per second.

The amount by which the modulated wave differs from the center frequency is called the *deviation* of the wave. The maximum amplitude of the modulating signal determines the maximum deviation of an FM wave. For the preceding example of the one volt signal, the maximum deviation is 100 kc.

Figure 1 compares the steps in the formation of AM and FM waves. In the resultant FM wave, the amplitude remains constant. Only the frequency changes. The job of an FM detector is to convert these variations in frequency into variations in amplitude and to extract from the resulting AM wave the original modulating signal.

Slope Filter Discriminator

One of the simplest forms of FM detectors is the slope filter discriminator shown in Fig. 2. The FM input is coupled to a parallel LC circuit which is tuned to a frequency either above or below the center frequency of the incoming wave.

The voltage across a parallel tuned circuit has the characteristic of Fig. 3 for frequencies near the resonant frequency. This figure shows that the voltage is a maximum at the resonant frequency. As the frequency deviates above or below resonance, the voltage decreases.

Figure 3 also shows what happens when the tuned circuit is resonant at a frequency above the center frequency

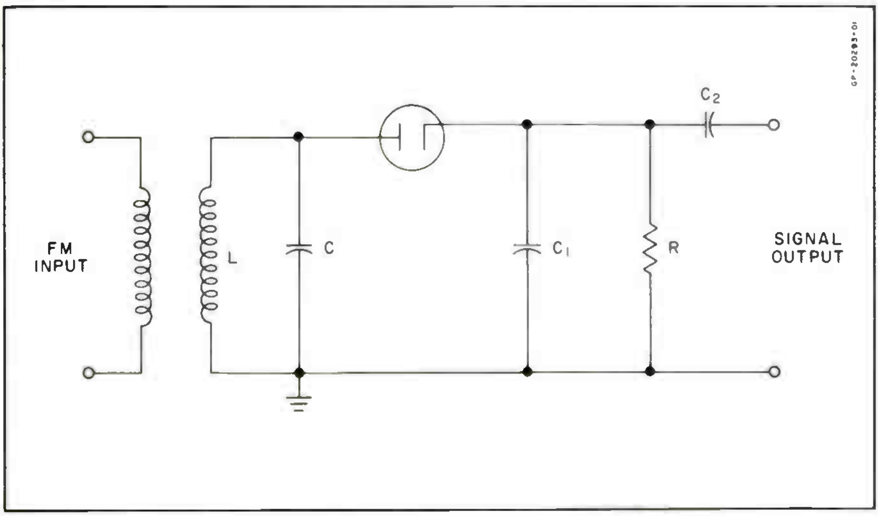


Fig. 2. Simple slope filter discriminator. Capacitor C_1 bypasses the high center frequency and its variations. Capacitor C_2 removes d-c component from output.

and its highest deviation frequency. All frequencies of the incoming FM wave will then fall on the sloping segment of the curve. The center frequency will produce a given voltage across the resonant circuit. Higher frequencies will produce higher voltages and lower frequencies will produce lower voltages.

As long as the segment of the curve is linear, the voltage variations across the resonant circuit are proportional to the input frequency. The circuit has converted the FM wave of constant amplitude and varying frequency into a wave which now varies in amplitude and in frequency.

On positive halves of the cycle, the diode of Fig. 2 conducts. The resulting current through the diode takes the form of d-c pulses which are proportional to the amplitude of the voltage across the resonant circuit. This current flowing across the load resistor sets up a voltage proportional to the current. Therefore, the voltage across the load resistor will be proportional to the voltage across the tuned circuit. Since the voltage across the tuned circuit is proportional to the input frequency, the voltage across the load resistor is proportional to the input frequency.

Bypass capacitor C_1 shunts off the

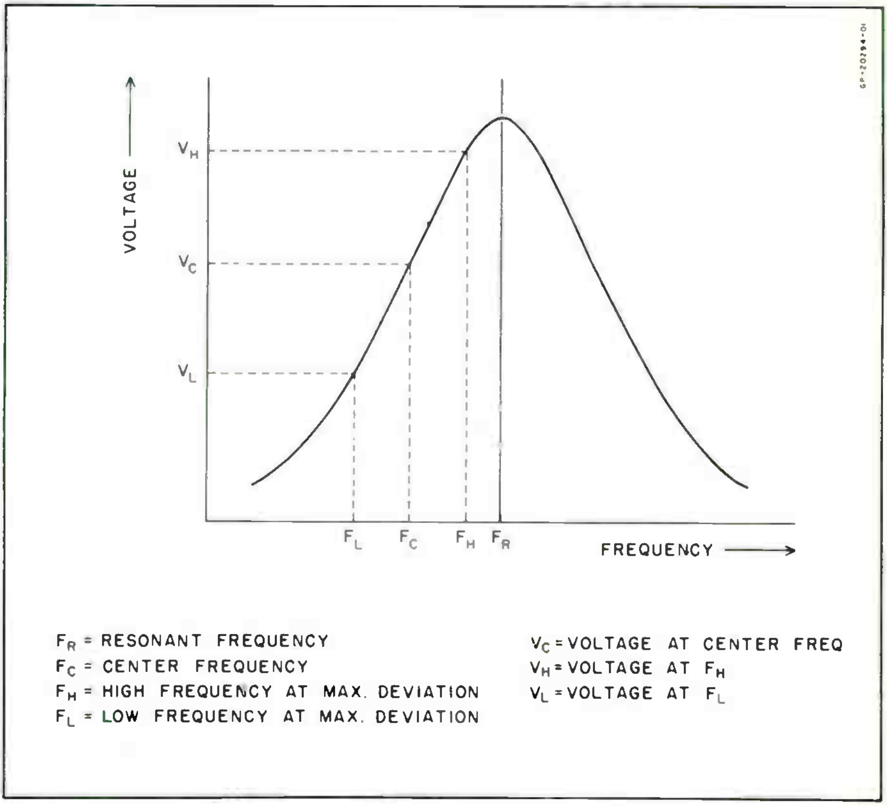


Fig. 3. Voltage-frequency characteristic of a parallel tuned circuit.

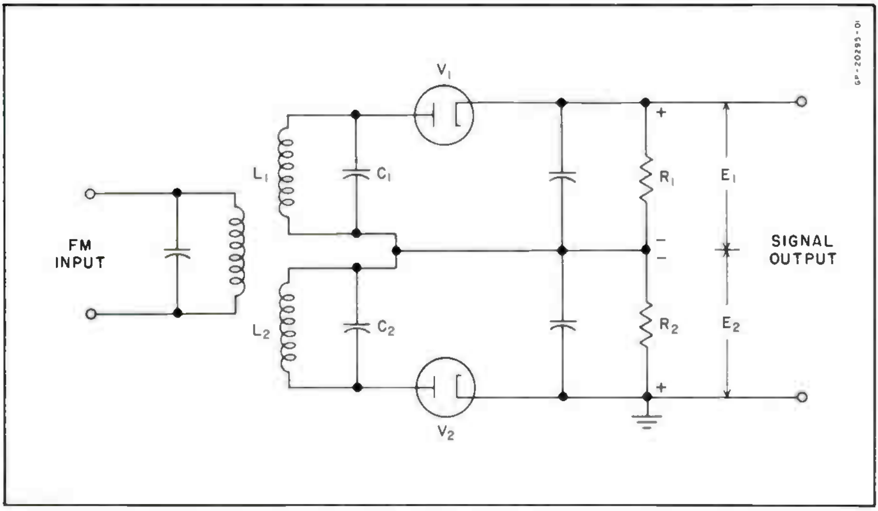


Fig. 4. Circuit of a double-tuned FM detector.

high frequencies in the vicinity of the center frequency. The voltage across the load then appears as a varying direct voltage having the shape of the envelope of the original modulating signal. Capacitor C_2 removes the d-c component and the output of the circuit is the original modulating signal. The process of frequency modulation has been reversed.

Double-Tuned Detector

A more commonly used detector is the double-tuned circuit shown in Fig. 4. The FM input is through a parallel LC circuit tuned to the center frequency. This is coupled to two secondary tuned circuits. One secondary resonates at a higher frequency than the center frequency and the other resonates at a lower frequency than the center frequency.

At the center frequency, the induced voltages across both secondaries are equal. The diodes then conduct equal currents. These currents flowing across the two load resistors set up equal and opposite voltages which cancel each other out. The net output voltage is zero.

At frequencies above the center frequency, the voltage across the high-tuned circuit is greater than the voltage across the low-tuned circuit. This upsets the balance of voltages across the two load resistors and makes the positive voltage (E_1) greater than the negative voltage (E_2). The net output is a *positive* voltage which reaches its peak amplitude when the input frequency reaches its maximum deviation above the center frequency.

At frequencies below the center frequency, the voltage across the low-tuned circuit is greater than the voltage across the high-tuned circuit. The net output is a *negative* voltage which reaches its peak amplitude when the input frequency reaches its maximum deviation below the center frequency.

For any specific frequency of the FM carrier wave, the output is therefore a

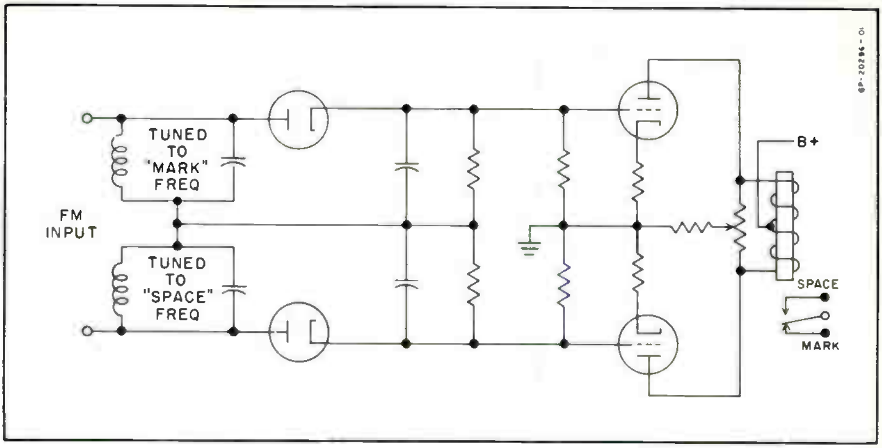


Fig. 5. Simplified form of detector circuit for FM telegraph receiver.

direct voltage. But since the frequency of the wave is varying at a frequency of the modulating signal, the output shows a voltage which varies in amplitude at the frequency of the modulating signal.

A variation of the double-tuned discriminator is used to detect FM telegraph signals. A simplified schematic of such a circuit is shown in Fig. 5. In this case, a specific frequency is used to denote the "mark" signal and another frequency to denote the "space" signal. One of the tuned circuits is then resonant at the "mark" frequency and the other is resonant at the "space" frequency.

When a "mark" frequency is received, the output is a direct voltage of one polarity. When a "space" frequency is received, the output is the other polarity. These output voltages control a polar relay which then indicates "mark" or "space" to the receiving apparatus.

In FM telegraphy, the input frequency does not vary continuously but shifts from one frequency to another. For this reason, this method is often called frequency-shift (FS) telegraphy.

Phase-Shift Discriminator

The most widely used form of FM detector is the phase-shift or Foster-Seeley discriminator. This circuit can be designed to give a very linear response over the total range of frequency deviations of the incoming wave. It also eliminates the very delicate adjustments of the double-tuned discriminator where three resonant circuits (a primary and two secondaries) must each be tuned to a different frequency. The detectors used in Lenkurt Type 72 radio equipment are based on the phase-shift discriminator circuit.

The phase-shift discriminator works on the principle that the voltage induced in a resonant circuit by another resonant circuit will vary in phase as the frequency of the inducing voltage varies. Figure 6 shows two circuits tuned to the same frequency and coupled together inductively. At the resonant frequency the output leads the input by 90°. As the frequency goes below resonance, the impedance of the circuit changes and the phase angle increases to more than 90°. As the frequency goes above resonance, the impedance of the circuit changes in the other direction and the phase angle decreases to less than 90°.

Figure 7 shows the same circuit with a center tap on the secondary coil. The center tap divides the total induced voltage across this coil into two separate voltages—one between the top of the coil and the center tap (E_3) and the other between the center tap and the bottom of the coil (E_4). These voltages are equal. If the center tap is taken as a reference, E_3 is a rise in voltage and E_4 is a drop in voltage. These are the voltages which could be measured between terminals 2 and 1 and between terminals 2 and 3. Thus the two voltages referred to the center tap are equal and opposite (180° out of phase).

The above relationship will hold true for all variations of primary voltage amplitude and frequency since it is based on the physical configuration of the circuit. Voltage E_3 and E_4 will always be equal and 180° out of phase. Because of the frequency-impedance relationships of coupled tuned circuits, these voltages will change their phase angles with respect to the primary voltage as the frequency changes. But with respect to each other, they will always stay equal and 180° out of phase. The voltage diagrams for frequencies at, above, and below resonance are also shown in Fig. 7.

Changes in input frequency vary the phase angles between the two induced voltages and the primary voltages. The phase-shift discriminator converts these variations of phase into variations of amplitude. The circuit to do this is arranged by coupling the primary voltage to the center tap through a capaci-

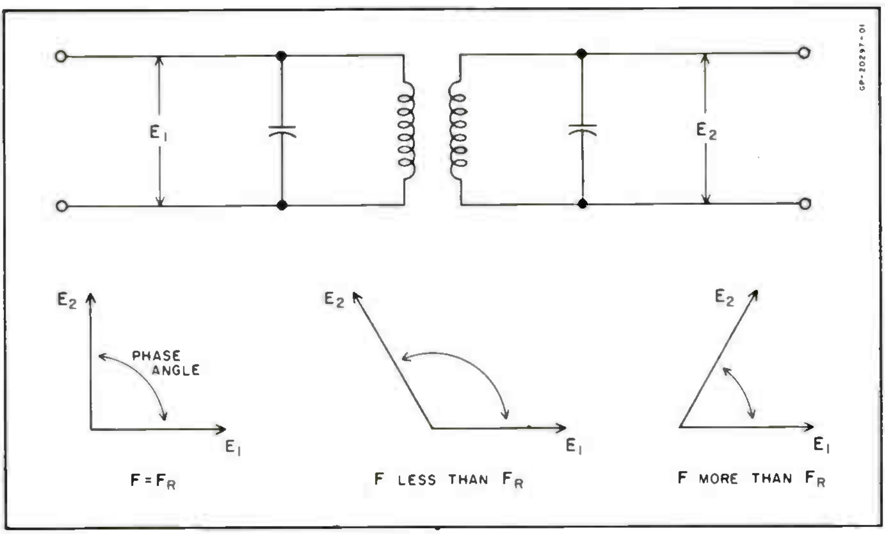


Fig. 6. Two coupled tuned circuits showing the phase shift that occurs as the frequency varies.

tor as shown in Fig. 8. The capacitor must be large enough to provide an effective bypass so that E_1 is always present at the center tap. It also serves to block any d-c from the preceding stage.

The voltage across the top half of the secondary coil now consists of two components—E_1 plus the *positive* induced voltage E_3. And the voltage across the bottom half of the secondary coil consists of two components—E_1 plus the *negative* induced voltage E_4.

The voltage diagrams for frequencies about resonance are also shown in Fig. 8. These diagrams show the resultant voltages E_{21} and E_{23} which appear at the anodes of the two diodes. Voltage E_{21} is the combination of E_1 and E_3 and voltage E_{23} is the combination of E_1 and E_4.

At the resonant frequency, E_{21} and E_{23} are equal in magnitude. (The arrows indicating their magnitude are equal in length.) But as the frequency changes, the phase changes of E_3 and E_4 will cause these voltages to add with E_1 so as to make their magnitudes unequal. At frequencies below resonance, E_{23} is greater than E_{21}. While at frequencies above resonance, E_{21} is greater than E_{23}. Thus the circuit of Fig. 8 has converted changes in frequency of the input signal into changes in the amplitudes of two voltages.

The relationships shown in Fig. 8 are not strictly accurate. Actually, E_3, E_4 and E_1 change in magnitude as well as in relative phase as the frequency varies. But the diagram shows the general way in which amplitudes of the voltages vary with frequency.

Voltages E_{21} and E_{23} are then applied to the anodes of two separate diodes. The cathodes of the two diodes are tied together through two equal resistances, R_1 and R_2.

The voltages at the anodes of the two

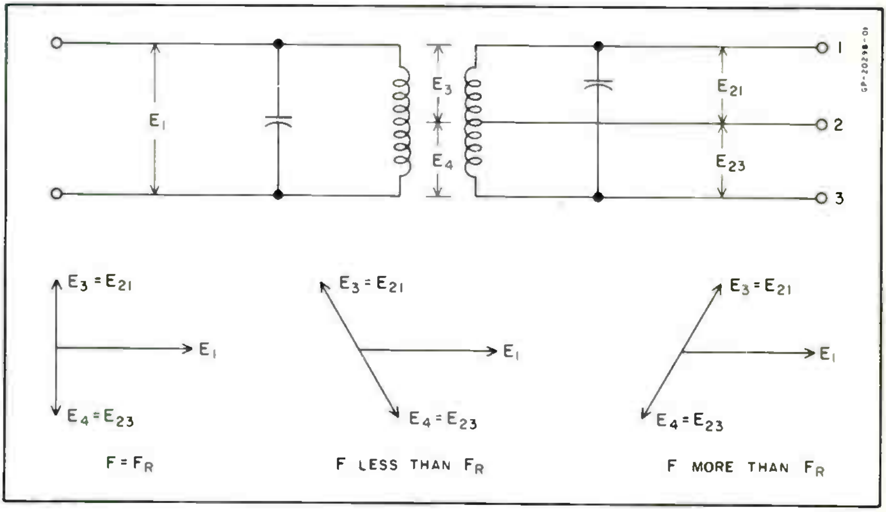


Fig. 7. Two coupled tuned circuits with a center tap on the secondary coil. Diagrams show the phase relationships between the two induced secondary voltages and the primary voltage.

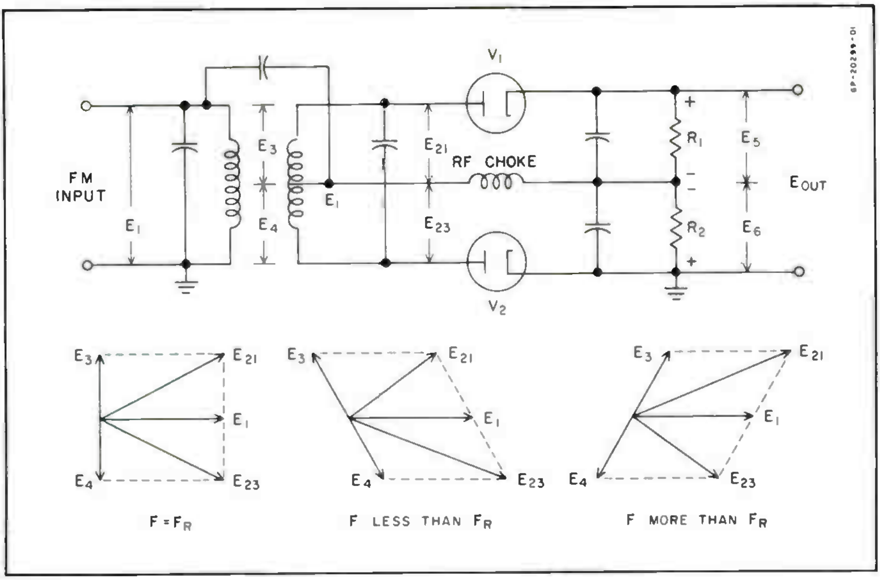


Fig. 8. Typical phase-shift discriminator circuit. This is also known as a Foster-Seeley discriminator.

diodes will cause them to conduct. The currents will set up voltages across R_1 and R_2. Since the currents through these resistors will be opposite in direction, the two voltages across them will be opposite in direction. This means that the total output voltage will be E_5 minus E_6.

At the resonant (center) frequency, the voltages at the anodes of the diodes will be equal. They will then conduct the same amount of current and E_5 will equal E_6. The total output voltage will then be zero since the two voltages will cancel each other completely.

As the incoming FM wave starts to swing above the center frequency, the voltage at the plate of V_1 starts to increase and the voltage at the plate of V_2 starts to decrease. Diode V_1 will draw more current than V_2 and the voltage across R_1 will be greater than the voltage across R_2. The total output of E_5 minus E_6 will be a positive voltage which reaches its maximum value when the incoming wave reaches its maximum frequency deviation above the center frequency.

As the incoming wave starts to swing back toward the center frequency, E_5 will start to decrease and E_6 will start to increase. The net voltage will be a decreasing positive voltage. When the input passes the center frequency, the total output will again be zero.

The reverse takes place as the input frequency swings toward its maximum deviation below the center frequency and then back toward the center frequency. On this half of the cycle, the output is a negative voltage that reaches its maximum value when the frequency is at its maximum deviation and returns to zero when the frequency returns to the center frequency.

The phase-shift discriminator thus

converts an FM input of constant amplitude and changing frequency to an output which varies in amplitude in accordance with the input's variations in frequency. The capacitors in parallel with the two load resistors bypass the high center frequency and its variations. The frequencies which appear at the output consist only of the relatively lower modulating frequencies. A typical output voltage-frequency curve is shown in Fig. 9.

To give an undistorted output, the input to the phase-shift discriminator must be free of any variations in amplitude. If this is not so, the amplitude variations will cause E_1, E_3, and E_4 to have different values for the same frequency. These variations will appear at the output of the discriminator. The output will not then be a true duplication of the original modulating wave. For this reason, FM receivers using phase-shift discriminators for detectors usually have one or more limiting stages to clip off any variations in amplitude.

This additional stage is eliminated in some types of receivers by using a variation of the phase-shift discriminator called the ratio detector. The general circuit differs in that the polarity of the two diodes are arranged so that their load voltages are additive. The output leads are then connected to pick off a portion of the total load voltage which is proportional to the ratio of the voltages at the diodes. Amplitude variations in the input signal will increase or decrease these voltages but their ratio will remain constant for a particular frequency. This makes the ratio detector relatively unresponsive to amplitude variations of the input.

The need for limiting stages does

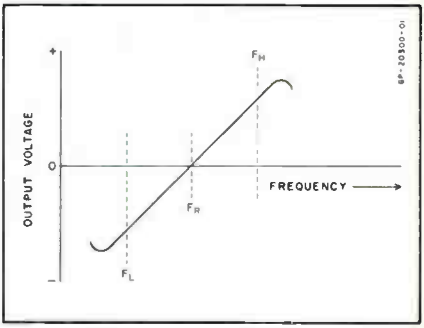


Fig. 9. Typical characteristic curve for a phase-shift discriminator.

not exist in ratio detectors. But, in general, the phase-shift discriminator has a more linear response which justifies the additional stage. Other forms of FM detectors using locked oscillators and pulse counters may also be used for specific applications. But the basic phase-shift discriminator or some modification of it is the most commonly used circuit for FM detection.

Conclusion

An FM wave is produced by using the changes in amplitude of the signal to produce changes in frequency of the modulated wave. For a given amplitude of the signal, the rate at which this frequency change occurs is determined by the frequency of the signal. Any form of FM detector performs the reverse process. This means converting the changes in frequency of the FM wave to changes in amplitude. The resulting wave is then rectified to recover the original modulating signal.

Almost every form of FM detector uses the properties of tuned circuits to convert the frequency changes to amplitude changes. The remaining step is similar to the detection process used in AM receivers.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6 NO. 12 DECEMBER 1957

6,000 MC RADIO SYSTEMS

Some Equipment and Operating Considerations

The 6,000 mc common carrier and industrial bands are valuable because they can accommodate very large numbers of voice channels. The characteristics of 6,000 mc radio waves, however, call for circuitry which differs in several important respects from conventional radio. Further, propagation is affected more by the atmosphere at this frequency than at lower frequencies.

This article describes two major differences between circuits for 6,000 mc and conventional radio. It also discusses the effect of the atmosphere on propagation at 6,000 mc.

Conventional radio circuits are used for communication at frequencies up to about 2,000 mc. Above this region they will not operate satisfactorily. Components such as standard radio tubes and two-wire or coaxial transmission lines either fail completely or have too much attenuation. In recent years, however, new components have been devised that make microwave communication in the region above 2,000 mc practical and economical.

One of these components, the klystron vacuum tube, is now widely used at frequencies up to 25,000 mc or greater. In some respects it performs more efficiently at microwave frequencies than an ordinary radio tube does at broadcast frequencies.

Another component, the waveguide, makes use of the very short wavelengths of microwaves to provide an effectively shielded transmission line having relatively low loss. The waveguide is not only capable of operating as a transmission line but also can be designed to operate as a capacitance, inductance, filter, or hybrid. When used with auxiliarly devices such as magnets and ferrites, it can be made to operate as an isolator, circulator, modulator, discriminator, or attenuator.

Besides having different terminal and repeater components, microwave radio systems are affected more by atmospheric conditions than broadcast or other low-frequency radio. Special consideration must be given to 6,000 mc equipment location to overcome the deeper and more frequent fades which occur as frequency increases in the microwave region. Additional importance is also given to the sensitivity and noise figure of receivers because of the deep fading.

Vacuum Tubes

In conventional vacuum tubes operating below microwave frequencies, the time required for an electron to travel from the cathode to plate is very small compared to the time required for a signal on the control grid to go through one cycle. However, if a microwave signal is applied to the grid, its cycle may be short compared to an electron's *transit time,* and will become shorter as the frequency increases. The flow of electrons in transit toward the grid will not be able to follow the signal variations exactly.

The plate-current wave will be very distorted and will be out of phase with the grid voltage. In addition, power will be dissipated at the control grid and, to a lesser extent, at the cathode. The result is distortion of the output signal and loss of gain.

To overcome this problem a klystron vacuum tube uses a stream of electrons to excite a resonant cavity within the tube. The tube output is taken from this cavity at its resonant frequency. Klystrons do not depend on a plate current to create an output voltage.

Klystrons used in microwave transmitters generate the carrier frequency which is modulated and fed to the sending antenna. In receivers they generate the radio frequency which is mixed with the incoming signal to produce an intermediate frequency for amplification and detection. Klystrons have now been developed to the point where they can be made about as reliable as the best standard industrial tubes. Fig. 1 is a schematic representation of the internal and external circuits of a klystron oscillator.

Waveguides

One of the characteristics of electromagnetic waves is that they can be confined in, and propagated along, hollow metal tubes. Such tubes may be circular or rectangular in cross-section, but a rectangular tube is generally the most practical because of its wide frequency range and ability to maintain wave polarization. The power loss in a waveguide is about one-third the loss in a comparable air-insulated coaxial line, and very small compared to the loss in a flexible coaxial cable with solid insulation (rubber, plastic, etc.). Waveguides can be made rigid or flexible and have an infinite life as long as they remain free of corrosion or dents.

One factor limits the use of waveguide. This is the physical dimension required for propagating given frequencies. The lowest frequency a waveguide can transmit is determined by its width. The wavelength at this lowest frequency, or *cut-off frequency,* is twice the width of the waveguide. This means, for example, that a waveguide designed to transmit a 30 mc signal must be at least 17 feet wide. Needless to say, this is impractical.

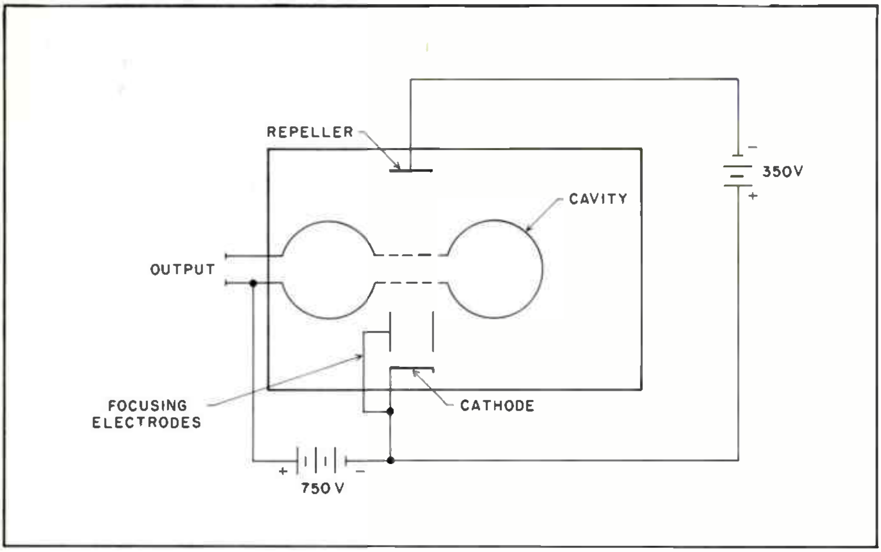


Fig. 1. Diagram of a reflex klystron. Electrons flow from the cathode across the cavity gap toward the repeller. The repeller returns the electrons in bunches to the cavity gap. These electron bunches induce voltages across the cavity at its resonant frequency and maintain a condition of oscillation.

At 2,000 mc, however, half a wavelength is about three inches, and at 6,000 mc only one inch. Waveguides can be built economically to handle these frequency ranges, and they lend themselves to convenient equipment arrangements. Fig. 2 shows the compact waveguide assembly used in Lenkurt Type 74A *Microtel* equipment.

Propagation

Radio signals travel from a transmitter to a receiver by three principal means: *ground waves, sky waves*, and *space waves*. Ground waves cannot be used at microwave frequencies because they are completely attenuated within a few feet of the transmitter. Sky waves are reflected, refracted, or scattered back to earth by ionized layers of the upper atmosphere (the ionosphere), but only to a small extent above 100 mc. At microwave frequencies they are usable only in very high-power expensive systems which must bridge great distances in a single hop. Space waves, which travel through the atmosphere immediately above the earth, are the most practical propagation means for microwaves.

To be usable, the space waves must arrive at the receiver with a certain minimum signal strength. Below this minimum, known as the *threshold level*, the signal is drowned out by receiver noise. When the received signal drops below this threshold in a common carrier radio system, telephone circuits connected to it through a dial exchange will disconnect. To restore service they

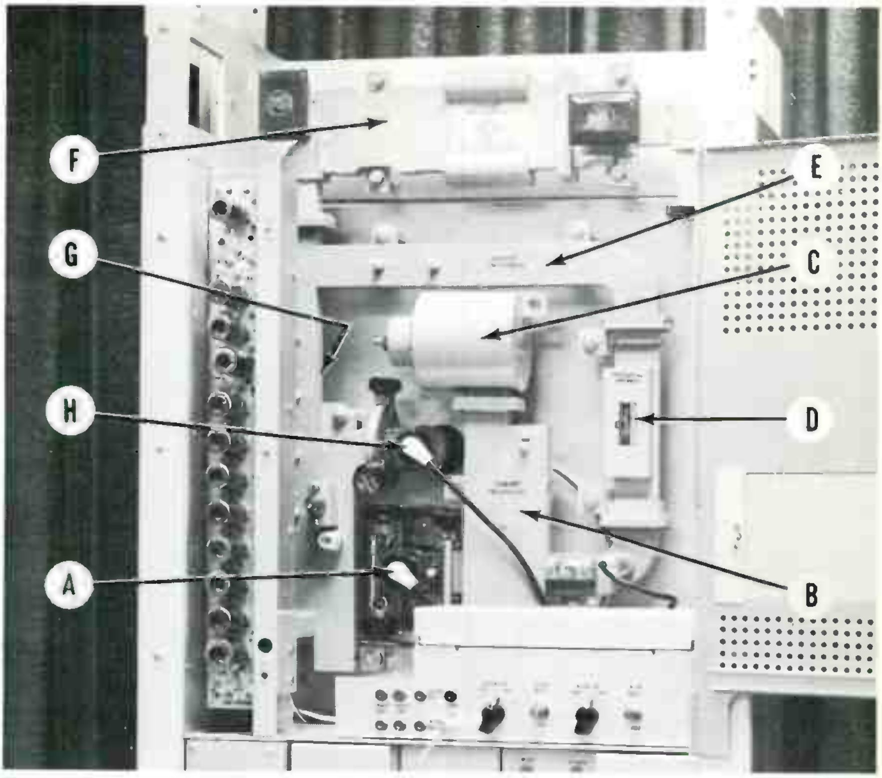


Fig. 2. Type 74A transmitter-receiver with waveguide assembly exposed. The principal microwave components shown are: (A) transmitting klystron; (B) waveguide discriminator; (C) reference cavity for controlling transmitter frequency; (D) isolator; (E) waveguide run to circulator panel; (F) circulator panel; (G) waveguide run to r-f mixer; and (H) local oscillator klystron.

must be redialed. Further, for toll quality communication, the signal must remain several decibels above the threshold level to maintain the desired signal-to-noise ratio.

The maximum distance allowable between transmitter and receiver for toll quality service is determined by transmitter power output, receiver threshold, and the sum of the losses between them. There is a relatively small loss from the transmitter or receiver to its antenna, an appreciable antenna gain, and a varying path loss. Fig. 3 shows gains and losses of a typical radio section.

The total path loss between antennas is made up of two parts: (1) path attenuation and (2) fading. These losses are determined by path length, path clearance above the earth, atmospheric conditions, and frequency.

Path Attenuation

If a space wave is radiated from a point (isotropic) antenna it spreads out equally to all directions in the shape of an ever-expanding sphere. As the surface of the sphere moves farther and farther from the point antenna, the radiated energy is spread over a larger area and the amount of energy per square-foot of wave front decreases. Mathematically speaking, the energy concentration at a point on a wave front is inversely proportional to the square of the distance from the antenna.

The power that can be extracted from a wave front by a similar point antenna is inversely proportional to the square of the frequency. Thus, the power received by a point antenna is inversely proportional to both the square of the distance from the source and the square of the frequency. The ratio of this power to the total power radiated is called *path attenuation*.

When the receiving antenna is something other than a point (a parabola-shaped dish, for example), the amount of power extracted from the wave front is greatly increased. The ratio of the amount of power received by a practical antenna to the amount extracted by a theoretical point antenna is called *antenna gain referred to an isotropic radiator*.

The gain of a parabolic antenna increases with antenna area. It also increases with operating frequency. So, for a given radio path with fixed-size antennas, the path attenuation increases with frequency. But so does the antenna gain. One tends to offset the other. Table 1 compares path attenuation and antenna gain for two radio sections operating at different frequencies over the same path length with antennas of the same size.

Fading

Fading of received signal strength is caused primarily by variations in atmospheric conditions. These variations

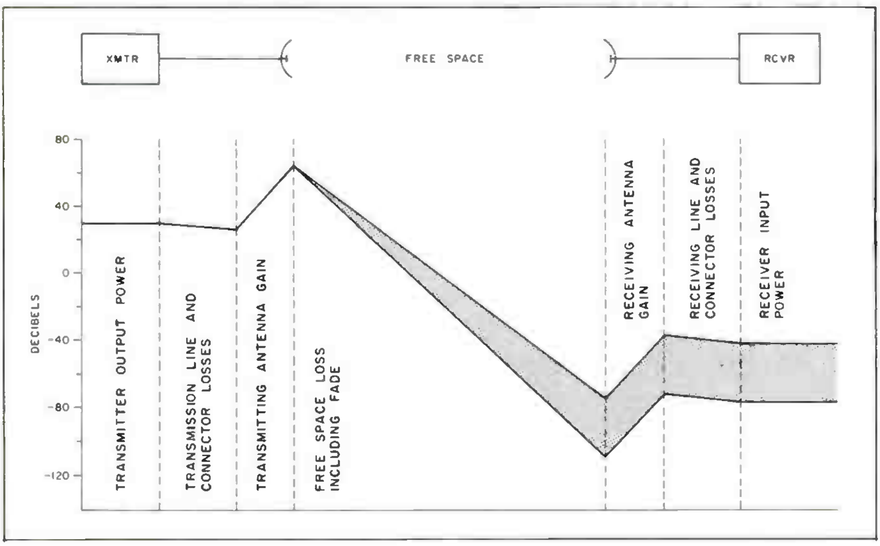


Fig. 3. Gains and losses in a typical radio section. The shaded area indicates the range of losses expected from fading during 99.9% of the time.

TABLE I

COMPARISON OF RADIO SECTION CHARACTERISTICS
AT 1,000 and 6,000 MC

	1,000 MC	6,000 MC	REMARKS
Path Length	25 mi	25 mi	
Antennas	6' Parabolic	6' Parabolic	
Free-Space Path Loss	124.5 db	140.0 db	Computed from: $L = 10 \log f^2d^2$
Antenna Gain (2 antennas)	46.0 db	77.0 db	Computed from: $G = 10 \log f^2 + 10 \log D^2 - 52.6$ for one antenna
Normal Transmitter Power	+37 dbm	+30 dbm	
Normal Misc. Losses (trans. lines, combining filters, circulators, etc.)	8 db	5.2 db	
Net Received Signal Power	-49.5 dbm	-49.5 dbm	

cause radio waves to bend away from their normal lines of propagation. The fades resulting under a given set of conditions occur in greater number and greater severity as signal frequency increases. The reasons for more frequent fading at higher frequencies can be explained by examining the mechanics of wave propagation.

A signal beamed toward a receiving antenna consists of a series of wave fronts whose centers are on the line of sight from trasmitting antenna to receiving antenna. The surface of each of these wave fronts consists of an infinite number of isotropic radiators sending signals in all directions away from the wave front. Thus, at any instant there are an infinite number of paths from a given wave front to its receiving antenna.

For example, in Fig. 4 if any two paths differ by one-half wavelength or any odd multiple of a half wavelength, the energies received over the paths will cancel. If they are the same length or differ by any whole number of wavelengths, they will reinforce each other. The paths AR and A′R from the wave front to the receiver R are one-half wavelength longer than the line of sight path OR. All secondary waves emanating from within the area defined by AOA′ as diameter will reinforce the direct wave, OR, at the receiver because they are less than one-half wavelength out of phase with OR. This area is known as the first Fresnel zone and provides one-quarter of the received field energy.

The path lengths BR and B′R are one wavelength longer than the direct path OR. All secondary waves emanating from the shaded area between the first Fresnel zone and the circle whose diameter is BOB′ will act at the receiver to partly cancel the waves from the first Fresnel zone because they are between one-half and one wavelength out of phase with OR. This shaded area is the second Fresnel zone. All odd-numbered Fresnel zones will reinforce the direct wave and all even-numbered Fresnel zones will cancel odd-zone energy.

The third Fresnel zone is defined by diameter COC′, and the fourth by DOD′. There are an unlimited number of Fresnel zones, with each succeeding one contributing less energy than the one before. The area of the Fresnel zones is determined by their distance from the transmitter and receiver, and the operating frequency. The higher the frequency, the smaller is the difference in path lengths which is equal to a half wavelength; hence, the smaller the first Fresnel zone and the others surrounding it. However, each Fresnel zone still contributes the same proportion of energy.

Fresnel zone sizes are important because they determine the effect of wave bending (refraction) on path clearance above the earth and on reflections from smooth earth surfaces. Smaller Fresnel zones cause obstacles in the radio path to obstruct a greater percentage of radiated energy. They also cause more severe and more frequent cancellations of energy between reflected and directly transmitted waves when the latter are bent or *refracted* by the earth's atmosphere.

Refraction

Refraction occurs when a wave changes velocity in passing from one medium into another. This occurs in air when two layers have different densities. As a radio wave travels upward at an angle through the atmosphere it normally encounters air of decreasing density. Since the top of the wave reaches the lighter air first, it increases

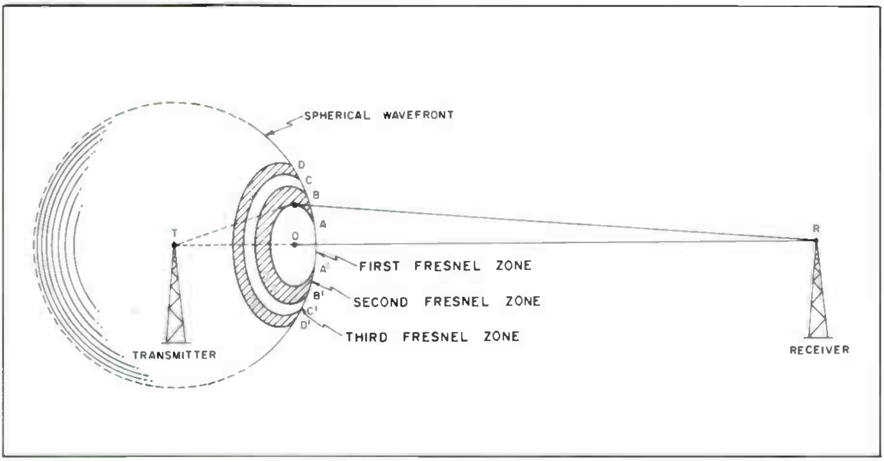


Fig. 4. Fresnel zones of a transmitted signal's wavefront at distance OR from a receiver. The difference in path length to receiver from edge of one zone and edge of adjacent zone is one-half wavelength. Odd-numbered zones reinforce the signal and even-numbered zones cancel it.

its speed first, and the wave bends back toward the earth.

In this way radio waves normally tend to follow the earth or, in effect, the earth appears to flatten and have a larger radius. Correspondingly, if inversion occurs and the air becomes heavier as altitude increases, the wave bends away from the earth.

When atmospheric conditions are such that the air density increases with height, the earth between transmitter and receiver appears to bulge up into the wave fronts. If this effective bulge reaches the line of sight between the transmitter and receiver, microwaves of any frequency will be attenuated about 20 db. However, if the refraction increases so that the earth bulge rises above the line of sight, attenuation will be even greater but no longer equal for all frequencies. Loss in this "shadow zone" increases rapidly with frequency.

For example, Figure 5 shows an effective earth bulge of 67 feet above the line-of-sight caused by atmospheric refraction. The first Fresnel zone radius for a 6,000 mc signal is 81 feet. The 67-foot obstruction results in a -0.83 clearance of the first Fresnel zone at 6,000 mc and causes a loss of 50 db over normal propagation conditions.

In addition to changing the clearance above obstacles, refraction causes fading by changing the relative path length of the direct and reflected waves. As the direct wave is bent above or below the line of sight, its path-length increases so that part of the time the two waves reinforce each other and part of the time they tend to cancel. This is shown in Fig. 6.

If the terrain between the transmitter and receiver is a good reflector, the

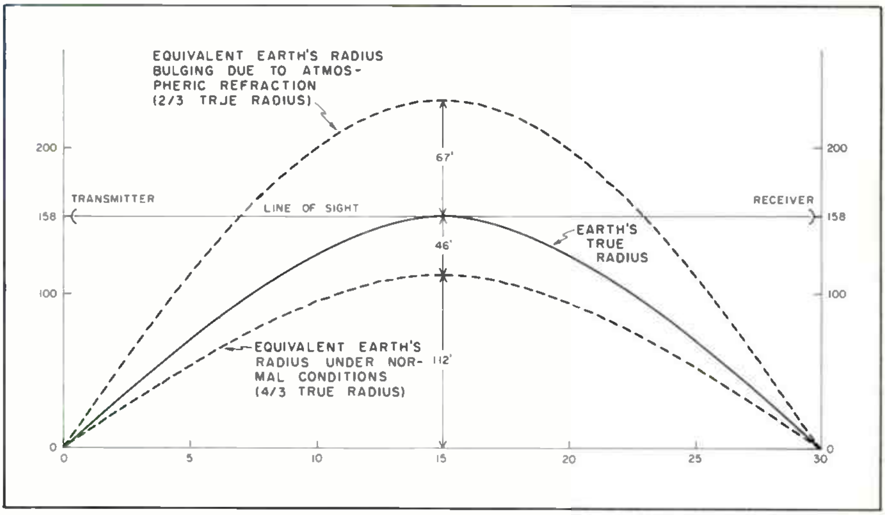


Fig. 5. In this example normal atmospheric refraction permits a signal clearance of 46 ft. above the earth. This is represented by an equivalent flattening of the earth's radius to 4/3 true radius. When atmospheric density increases with height the earth's surface appears to bulge upward. Here a bulge of 67 ft. above line-of-sight gives an equivalent earth's radius of 2/3 true radius.

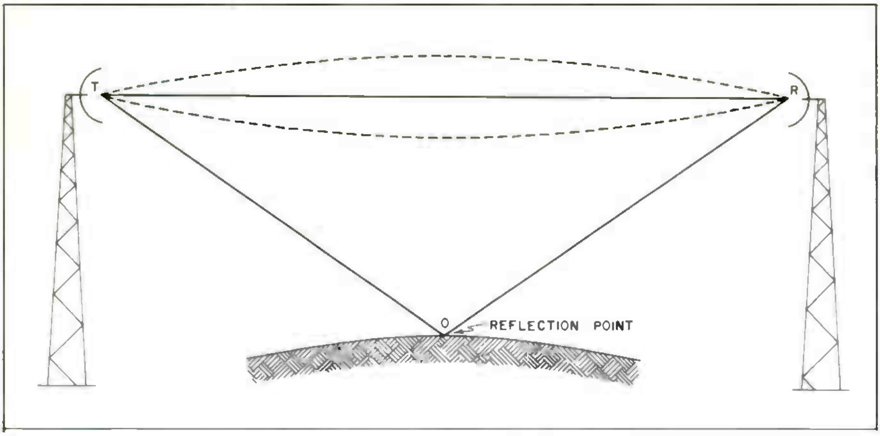


Fig. 6. Refraction of direct wave which results in partial cancellation by reflected wave. TR is the direct path and TOR is the reflected path from transmitter to receiver. Dotted lines show bending of direct wave due to refraction. This changes path length TR and causes alternate canceling and reinforcing by reflected wave TOR.

cancellations may be nearly complete and very deep fades will result. High-frequency waves will cancel each other more frequently than low-frequency waves because smaller changes in refraction are required to cause a difference in paths of one-half wavelength. Thus a 6,000 mc radio system will have more fades than a lower frequency system but they will be of shorter duration.

Fade Margin

To insure that a transmitted signal reaches a receiver with at least a minimum strength for a certain percentage of the time (for example, 99.9%), enough extra signal strength must be available during normal propagation to compensate for most fades. This is called *fade margin,* and is usually determined by actual field experience because there are no reliable formulas for predicting atmospheric conditions.

A fade-margin figure dictated by experience for a typical path at 6,000 mc is 30-40 db. The exact figure used depends on field studies of the particular location. When the signal path is over a good reflecting surface such as water, additional fade margin must normally be allowed.

Conclusions

The frequency bands in the 6,000 mc range available for common carrier and industrial use are valuable because of their high channel capacity. The use of these bands, however, involves special microwave circuitry and increased losses to the space wave. The necessary circuit elements—such as klystrons and waveguides—are readily available and present little difficulty in application. In fact, klystrons and waveguide circuits are more rugged and simpler than low-frequency tubes and coaxial cable. The additional path attenuation and fading at these frequencies can be overcome by shorter sections and antennas of larger gain.

HE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6, NO. 4 APRIL, 1957

WAVEGUIDES

For Microwave Systems

Over the years, the transmission frequencies of telephone message channels have gone higher and higher. Starting with the early open-wire line systems operating below 30-kc, transmitting frequencies have climbed steadily through the ranges suitable for multi-pair cable, coaxial cable and, most recently, microwave. Each increase in frequency has brought about new techniques and components for frequency generation, filtering and transmission.

The newest techniques are those associated with microwave transmission at frequencies above 1000 mc. This article discusses an important component of such systems—the waveguide.

Every radio system needs a method of conveying energy from the transmitter to the transmitting antenna and from the receiving antenna to the receiver. At the low radio-frequencies, a two-wire transmission line may be used. At frequencies above a few hundred megacycles, the losses of a two-wire line become too high and a coaxial cable is used. Above 1000 mc, even the losses of coaxial cable become too high. At these higher frequencies, a waveguide provides the most efficient path for electrical energy.

In a broad sense, any device which confines electrical energy to a specific path in space is a waveguide. Thus, two-wire lines and coaxial cables are forms of waveguides. But in a strict sense, a waveguide is a hollow tube. It receives energy at one end and delivers it to the other end.

Waveguide Operation

Radio waves in space tend to propagate outward in all directions. But if they are set up within a hollow tube of conducting material as shown in Fig. 1, they can be confined within the tube. What is more, if the tube has the proper dimensions with regard to the wavelength of the r-f energy, the waves will travel down the length of the tube with little loss. Such a tube is a waveguide. It may be rectangular, square or circular in cross-section. By far the most commonly used form is the rectangular-shaped cross-section.

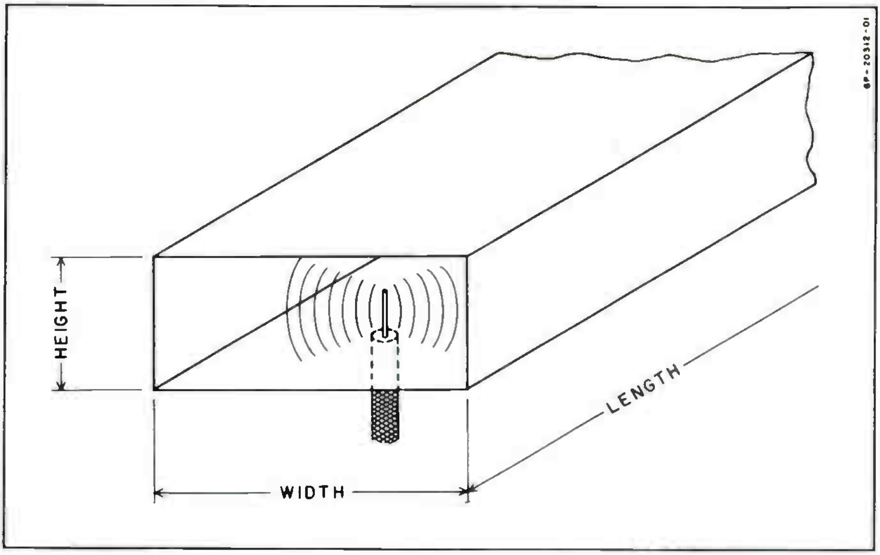


Fig. 1. Vertical probe radiating energy inside a rectangular waveguide. All sides of the guide are conductors.

Radio waves propagate within a waveguide by reflecting off the sidewalls of the guide. These reflections occur in a way that makes the electric and magnetic fields of the waves set up a definite pattern within the guide. The particular pattern depends on the wavelength of the exciting energy and the dimensions of the guide. Each such pattern is called a *mode*.

In addition, every mode has a *cutoff frequency*. This is the lowest frequency that a guide will transmit while operating in a particular mode. Energy at frequencies below this frequency will be attenuated instead of transmitted down the length of the guide. Above this frequency, energy will be transmitted with very little attenuation.

For a particular guide, the mode having the lowest cutoff frequency is called the *dominant mode*. This is the mode most often used in practice. For a rectangular waveguide operating in the dominant mode, the cutoff frequency is the frequency which has a wavelength equal to twice the width of the guide. At frequencies above the cutoff frequency, the electric and magnetic fields arrange themselves so that the energy follows a zigzag path through the guide. As the frequency is lowered, the energy rearranges itself so that the zigzag path is more compressed.

The energy paths of three different frequencies through the same waveguide are shown in Fig. 2. When the frequency reaches the cutoff frequency, the energy simply bounces back and forth between the sidewalls of the guide and has no forward motion through the guide. Thus energy at frequencies below the cutoff frequency will not appear at the receiving end of the guide.

Energy is usually fed into the guide by means of a probe which acts like a small antenna. Often the probe is simply the inner conductor of a coaxial cable as shown in Fig. 1. It is inserted through the bottom of the guide midway between the two sidewalls. At a distance of a quarter-wavelength back from the antenna, a conducting plate blocks off the end of the guide. The end plate acts as a reflector for the r-f energy and this reflected energy is in the proper phase to reinforce the original energy from the probe. The other end of the guide is open. It may feed to a resonant cavity or to another waveguide.

The construction of waveguides for use at the receiving end of a microwave system is the same, but the operation is reversed. The r-f energy enters the guide at its open end and travels through the guide to a probe which picks up the energy and feeds it to the receiver.

Application

Many modes other than the dominant mode can exist in a waveguide. But the common practice is to design the guide to propagate the dominant mode and suppress all other modes. The usual rectangular guide has a width greater than one-half wavelength but less than one wavelength of the desired operating frequency. The height is then made about one-half the width. These dimensions give a cutoff frequency below the operating frequency and a cross-sectional area too small to allow higher modes to form.

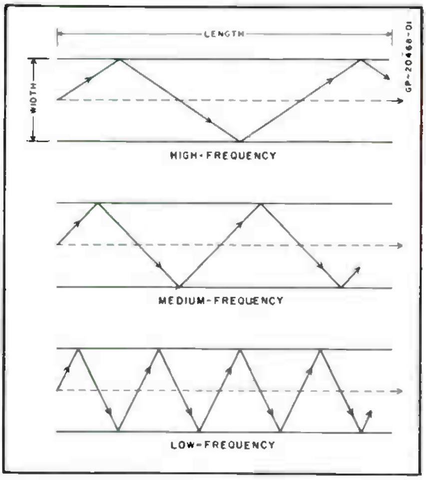


Fig. 2. Energy paths for three different frequencies in the same waveguide. Dotted line shows net direction of energy transmission.

The dominant mode is more commonly used because it gives the lowest cutoff frequency for a particular guide. In addition, it gives the simplest field pattern and is not as susceptible to impedance mismatches and reflections as the more complex modes.

The common conducting materials used in waveguides are brass, aluminum, gold, silver and copper. At the high operating frequencies of waveguides, the currents do not penetrate very deeply into the guide surface. Therefore, for short lengths, a thin inner plating of a high-conductivity material (gold, silver, aluminum) on a cheaper metal often provides a practical waveguide. Where weight is important, a guide may be constructed entirely of aluminum.

A new low-loss material coming into use is oxygen-free, high-conductivity (OFHC) copper. A typical attenuation-frequency curve for an OFHC copper guide for use with Lenkurt's new Type 74A microwave system is shown in Fig. 3.

When connecting two-wire transmission lines or coaxial cables, impedances have to be matched carefully to avoid reflections and losses. This is also true in waveguides. When two guides are connected together, there will be an impedance mismatch unless both have the same dimensions. If they differ, the losses can be reduced by using a tapered waveguide section to connect the two. By this means, the change in guide characteristics comes about gradually and reflection is kept to a minimum.

The physical path that a waveguide must take cannot always be a straight line. The over-all guide must be constructed to follow the curves of the physical path. Therefore, the guide path is usually made up of a combination of straight and curved sections of guide. The curved sections have the same cross-sectional dimensions as the straight sections. Two typical 90° curved sections are shown in Fig. 4.

Curved sections also tend to introduce reflections and power loss. These can be kept small by avoiding abrupt changes in direction. A common rule of thumb is to keep the radius of the curved section never less than two wavelengths of the signal to be transmitted.

A waveguide may also be used as an antenna. In this case, the open end of the guide is flared to make a gradual transition from the impedance of the guide to the impedance of space. Such an antenna is called an electromagnetic horn. It has characteristics similar to those of a directional antenna.

Horns may be shaped by flaring either the top and bottom of the guide or by flaring the sidewalls. To achieve maximum gain from a given length of horn, the top, bottom, and both sides are flared.

Waveguides offer many advantages at microwave frequencies. To begin with, they are of simpler construction than other means of transmission. But what is more important, they have very low losses. The conducting surface provides complete shielding to eliminate losses from radiation, and the losses in

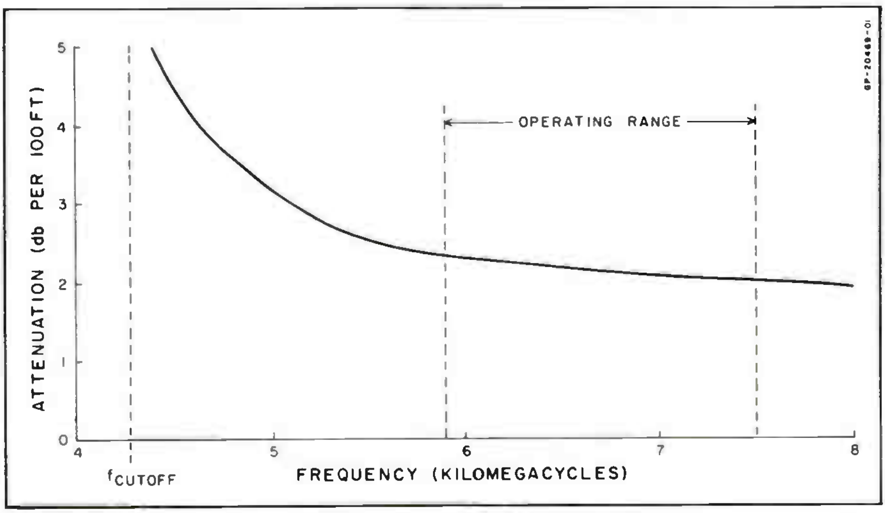


Fig. 3. Attenuation-frequency curve for typical waveguide used in Lenkurt Type 74A system. Operating range and cutoff frequency are shown by dotted lines.

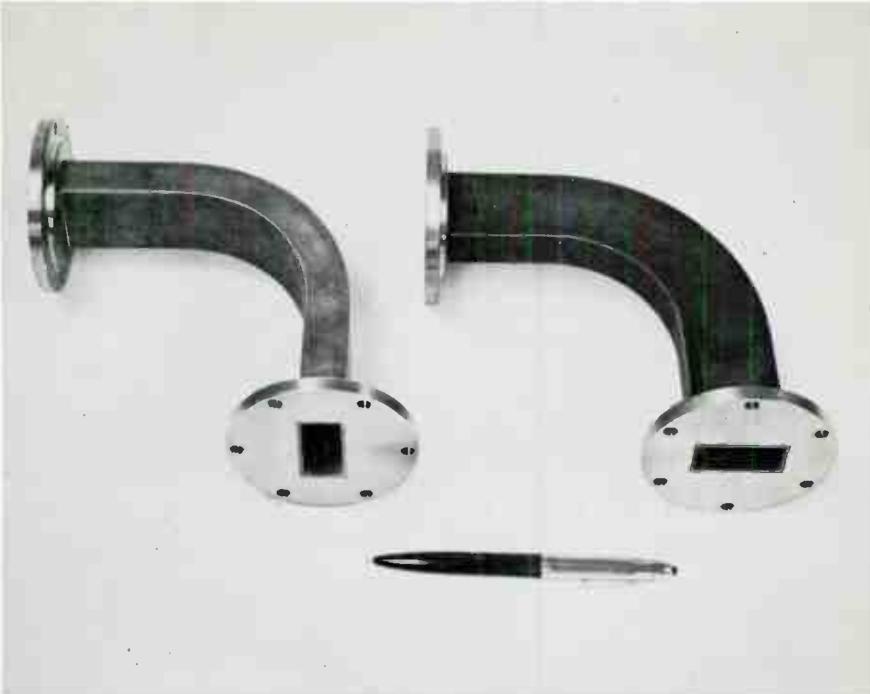


Fig. 4. Two forms of 90° bends for rectangular waveguides.

the conducting surface itself are small. In addition, a waveguide has more power-handling ability.

The main disadvantage of a waveguide lies in the limitation of its use to only the microwave frequencies. At lower frequencies, the guide dimensions become too large to be practical. Also, the installation of waveguides is more delicate than that of other methods. Solder beads, dents and bends will all tend to increase the attenuation and standing waves within the guide.

Circular waveguides are also used occasionally, especially where the guide must feed through a rotating joint. Their analysis differs from that of rectangular waveguides in several respects and, in general, is more complicated. They are seldom used unless the mechanical needs of the system call for them.

Conclusion

A waveguide is a device which isolates a specific path in space for the transmission of high-frequency energy. The energy, in the form of electromagnetic field patterns, travels along the guide from or to a probe enclosed in the guide. The size of a waveguide is inversely proportional to the lowest frequency it must handle. Therefore, the large size required at low frequencies makes them impractical at frequencies below the u-h-f bands. But at frequencies above 1000 mc, waveguides are the most efficient means of transmission and are used almost exclusively.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 6, NO. 6 JUNE, 1957

THE KLYSTRON

Most conventional vacuum tubes have little practical use at the higher microwave frequencies. Without some other means to generate and amplify these signals, all communications would be seriously limited to the already crowded lower bands of the spectrum.

One of the devices which opened up a vast new area of usable frequencies was the klystron tube. This article discusses the basic operating principles of klystrons and their applications.

A conventional vacuum tube conducts current by means of a stream of electrons which flows from a negative cathode to a positive plate. The electric field between cathode and plate controls the amount of electrons in the stream.

In a triode, a signal voltage on the grid changes this field and causes the electron stream to increase and decrease in density with the amplitude of the signal voltage. The resulting current flows through a load in the plate circuit to produce an output voltage whose amplitude is proportional to the signal and whose frequency is the same.

The distance between cathode and plate is short and the speed of the electrons is high. Therefore, the transit time between is very short compared to the time it takes for an audio-frequency or low radio-frequency signal to go through one cycle. But at frequencies in the microwave region, the transit time may be very long compared to the period of the signal. When this happens, the field between cathode and plate may reverse itself long before an individual electron has had time to go from cathode to plate. This upsets the phase relationship between plate voltage and current, limits the output power and frequency, and increases the plate dissipation.

Transit time effects are the most serious limitations of a conventional tube at microwave frequencies. These effects are counteracted in the klystron tube which actually makes use of transit time to amplify or generate microwave frequencies.

Klystron Operation

To understand the operation of a klystron, it is first necessary to know something of the basic mechanics of electron flow. First of all, the electron is a negatively charged particle with a definite mass. When it is in motion, it has a certain amount of kinetic energy. Increasing its velocity increases its energy. Decreasing its velocity decreases its energy.

A positive electric field attracts an electron and a negative field repels it. If an electron is moving toward a positive electrode, the attractive force will accelerate the electron. If it is moving away from a positive electrode, the force will exert a drag on the electron which decelerates it. A negative field has just the opposite effect.

But no form of energy comes into being or disappears spontaneously. So when an electron's speed is increased, the additional energy must come from somewhere. And when an electron's speed is decreased, the energy it gives up must go somewhere. In an electronic circuit, the energy supplied to or taken from the electron comes from or goes to the field.

Figure 1 shows the basic circuit of a klystron tube. A positive accelerator grid draws electrons from the cathode and shoots them in a high-velocity stream toward a pair of grids in a cavity resonator. These grids are called *buncher grids*. The cavity acts like a tuned LC circuit. A signal is fed into it through a waveguide, probe, or by a coupling loop as shown.

The signal voltage sets up an electric field between the buncher grids. On positive halves of the cycle, this field accelerates the approaching electrons and on negative halves it decelerates them. When the signal wave goes through zero, the field has no effect on the electrons. Therefore, the stream leaving the buncher grids consists of electrons moving at different speeds. Some are moving faster than when they entered the cavity, some are moving slower, and others are moving at the same speed.

The space between G_3 and G_4 in Fig. 1 is free of any fields and is called the *drift space*. In the absence of any accelerating field, the electrons will keep whatever speed they had on leaving the buncher gap. Since some are traveling faster than others, the faster ones will overtake the slower ones. Therefore, at some point in the drift space, the electrons will jam up and form bunches.

Another resonant cavity with a set of catcher grids (G_4 and G_5) is located at a point where the bunches of electrons form. The bunches induce an r-f voltage in the catcher cavity at its resonant frequency just as a current pulse excites a tuned LC circuit. With the proper phase relations between this voltage and the arriving bunches, the field will slow down the electrons as they pass through the catcher gap. The energy they give up goes into the resonator field and is taken out by another coupling loop. The slowed down electrons then go on to strike a collector plate which returns them to the cathode.

Since the electrons pass the buncher gap in a steady stream, the signal voltage speeds up as many electrons as it slows down. Therefore, neglecting losses, there is no net exchange of energy in the buncher cavity throughout a cycle of signal voltage.

But at the catcher, the electrons reach the gap in bunches when the cavity voltage is in its negative half of the cycle and very few electrons pass

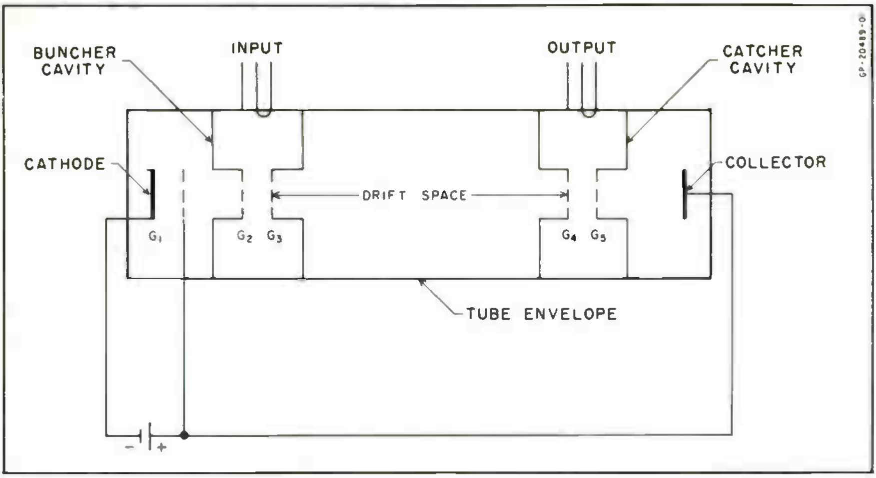


FIG. 1. Simplified diagram of a klystron amplifier. An actual klystron usually has a set of electrodes at the cathode which focuses into a narrow beam.

through on the positive half. As they pass through the gap, many more electrons are slowed down than are sped up. Thus, much more energy goes into the field than goes out. This extra energy represents an increase in power over the applied signal. In working terms, the klystron has amplified the signal.

Klystron amplifiers are often made with more than two cavities. These are called multi-cavity klystrons. They achieve greater efficiencies and higher outputs by cascading the power gain. Such amplifiers can supply output powers of the order of kilowatts.

Klystron tubes can also be used for oscillators and frequency-multipliers. The klystron oscillator simply feeds back some of its output power from the catcher cavity to the buncher cavity in the proper phase. In a frequency-multiplier, the catcher cavity is smaller than the buncher cavity and tuned to resonate at some multiple of the buncher cavity frequency.

The Reflex Klystron

Probably the most common form of klystron oscillator in use today is the reflex klystron. Its basic circuit is shown in Fig. 2. This form uses only one resonant cavity and one set of grids which act as both buncher and catcher for the electrons. In place of the collector of the two-cavity klystron, the reflex klystron has a repeller electrode. This has a negative voltage with respect to the cathode and turns the electrons back instead of returning them through the external circuit.

In operation, electrons from the cathode are drawn through the resonator gap and toward the repeller by the potential of the first grid. The positions of electrons in the beam are random in nature. Thus the beam will induce tiny noise voltages in the gap, and some of these will be in the resonant frequency range of the cavity. These voltages will velocity-modulate the electron beam just as the signal source does in the klystron amplifier of Fig. 1.

This means that some electrons will have greater speeds than others as they leave the gap and head toward the repeller. The repeller exerts a force similar to the force of gravity on an object thrown straight up in the air. At some point before an electron reaches the repeller, it will slow to a stop, reverse, and head back toward the gap. Electrons traveling at higher speeds will travel farther into the negative field of the repeller before reaching this turning point.

On the way back, the electrons will form bunches. This bunching process takes place in much the same manner as a number of balls thrown straight up in the air, one at a time, can be made to strike the ground at the same time. If each one is thrown up with a little less force than the one before and the time interval between throws is spaced properly, the balls will all reach the ground in a bunch.

In the reflex klystron, the electron bunches pass back through the gap toward the cathode and induce voltages in the gap which reinforce the original voltage. The reinforced voltage speeds up or slows down other electrons coming from the cathode and the cycle repeats itself. At some point, conditions will reach an equilibrium state where the amount of energy coming back just balances the circuit losses. Then the klystron oscillates at the frequency of a resonant mode of the cavity.

Both the frequency and power output of a reflex klystron depend on the time it takes for the electrons to leave the gap and return in a bunch. This time depends on the distance between gap and repeller and the supply voltage on the grids and repeller. When the dis-

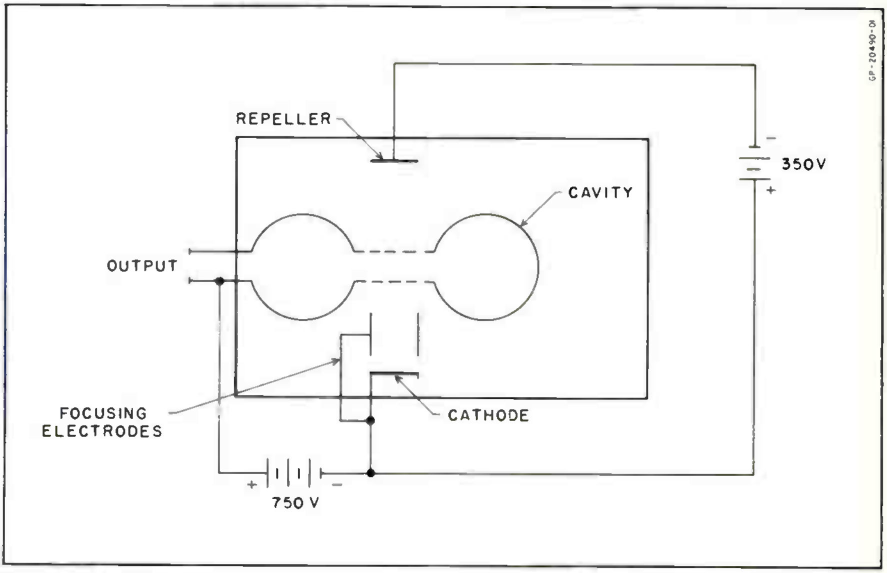


FIG. 2. Diagram of a reflex klystron. The output may be through a waveguide, coaxial cable or probe.

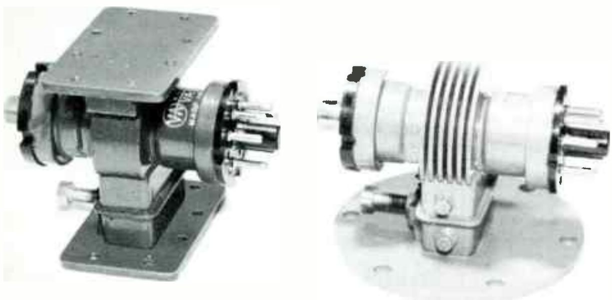


FIG. 3. Two typical reflex klystrons. The one at left is rated at one watt, and the one at right, at 35 milliwatts.

tance and the grid voltage are fixed, both frequency and power can be controlled by the repeller voltage.

To transfer power to the cavity, the r-f voltage across the gap must decelerate the electron bunches as they pass through. This happens when the repeller voltage is the proper value to get the bunches back to the gap at a time when the r-f voltage is going through a negative part of its cycle with respect to the arriving bunches. The energy given up by the electrons then transfers to the field.

Since the repeller voltage controls the time necessary for the bunches to get back to the gap, it also controls the frequency of oscillation. With a constant supply voltage on the repeller, the frequency will be the resonant frequency of the cavity. A more negative voltage will stop and turn back the electrons at a point farther from the repeller. Their travel time will then be shorter and the frequency will increase. A less negative voltage will have the opposite effect.

This ability to convert variations in voltage amplitude to variations in frequency makes the reflex klystron a very convenient oscillator for frequency-modulation systems. For the circuit of Fig. 2, the modulating signal can simply be placed in series with the repeller supply voltage. As the signal voltage varies in amplitude, the repeller voltage also varies and causes the output frequency of the oscillator to vary.

The reflex klystron has a low efficiency and so is not used to produce very high power outputs. But in many applications, not much power is needed and the reflex klystron makes a very effective oscillator. Two reflex klystrons are used in the new Lenkurt Type 74A 6,000-mc microwave radio system. One supplies the radio carrier frequency in the transmitter and another acts as a local oscillator in the receiver. Two typical low-power reflex klystrons are shown in Fig. 3.

THE

Lenkurt Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 3 NO. 6 JUNE, 1954

Some Factors Affecting

THE PROPAGATION OF MICROWAVES

Over Point-to-Point Radio Systems

Reliable communications can be obtained over point-to-point radio systems just as easily as they can be obtained over conventional wire and cable lines. Just as a wire-line system is made reliable by engineering the system to compensate for predictable variations in line losses, a radio system can be made reliable by engineering the system to compensate for predictable variations in propagation losses.

In this article, the important factors which affect propagation of radio waves are discussed and some of the methods of utilizing them or compensating for them are described.

In the outside telephone plant, wire and cable have long been the standard transmission facilities for toll and exchange routes. Until after World War II, no extensive use of radio was made in the telephone industry. It was normally used only where land lines or submarine cables were impractical. This situation has now changed. Radio equipment designed specifically for telephone toll plant usage is presently available. Operation and maintenance of this equipment is fully compatible with normal telephone practices. Because of its many advantages, radio is finding wide application for expansion and replacement of existing outside plant wire lines and cables.

While much public attention has been given to national microwave networks for the transmission of hundreds of telephone channels and several television channels, the recent development of radio and channelizing equipment for light to medium traffic routes has made the use of point-to-point radio economically practical for expansion and extension of toll and exchange facilities.

Many telephone companies are now finding that microwave has a definite place in their outside plant. Numerous installations already made have demonstrated that microwave systems can be engineered to be equally or more reliable than conventional wire lines or cables.

To engineer a wire line system

requires a knowledge of the transmission characteristics of wire lines at the frequencies used. In the same manner, to engineer a microwave system requires a knowledge of the propagation characteristics of radio waves at microwave frequencies.

Fading, a phenomenon encountered in radio links, is comparable to increased attenuation under severe weather conditions, a basic factor in wire line engineering. Fading is caused by the effect of air and terrain on radio wave propagation.

Radio waves at microwave frequencies and light waves have many of the same characteristics. Since the behavior of light waves is well known through the science of optics, and since radio waves and light waves have many of the same properties, certain optical principles are useful in describing radio wave propagation. The most important of these are reflection, refraction and diffraction.

Optical Properties

Because they behave like light, radio waves can be reflected from smooth conducting surfaces and focused by reflectors or lenses. When radio waves pass from one medium to another (such as from dry air to moist air) they are bent or refracted in the same manner as light waves are bent by a lens or prism. Radio waves tend to bend around large obstacles in their path by a process known as diffraction. They also are scattered by small particles such as rain and snow.

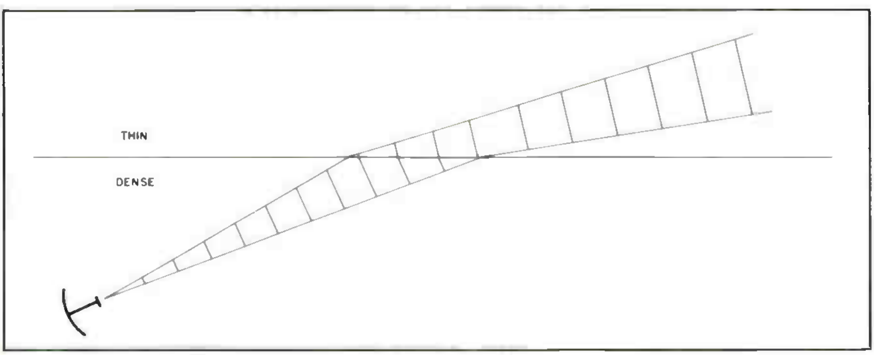
Each of these properties can cause variations in the received signal strength. They must be considered and allowances made for their effects when engineering a radio system.

Reflection

Very short radio waves are usually focused by dish-shaped metal reflectors. Such reflectors concentrate all the energy into a relatively narrow beam that can be directed like a light beam of a searchlight. This concentration of radio energy allows transmission over longer paths with much less power than would otherwise be required with non-directional antennas. (See Demodulator, Vol. 1, May, 1952.)

While the ability to reflect radio waves is very useful for focusing them into a beam, reflection is also a primary source of received signal variation. Reflections occur when radio waves strike a smooth surface such as water or smooth earth. If both reflected and direct waves reach the receiving antenna, it is possible for the two waves to cancel each other and reduce the received signal strength.

FIGURE 1. Refraction at a boundary between air at different densities. The speed of radio waves is slower in the denser medium.



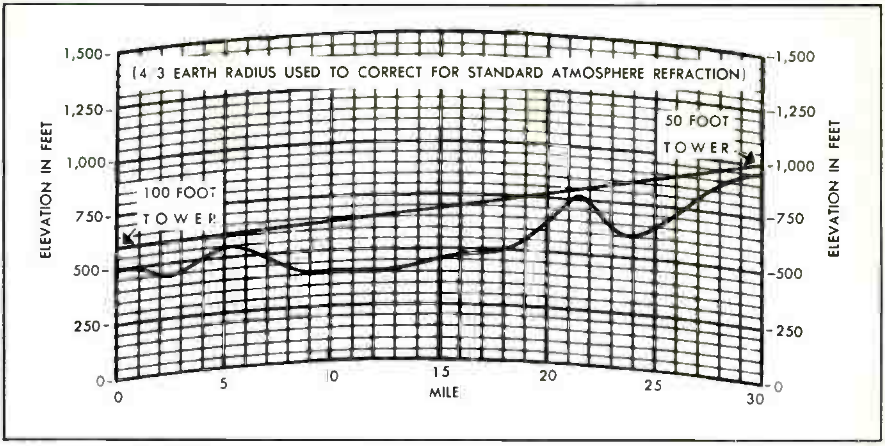


FIGURE 2. Profile charts are often prepared with 4/3 true Earth's radius to allow for normal atmospheric refraction. Charts prepared with true Earth's radius are also widely used. True Earth's radius provides a more conservative method of system engineering.

Depending on the length of the reflected path compared to the direct path, the reflected wave may arrive at the receiving antenna either in phase, out of phase, or partially out of phase with the direct wave. Under conditions where the reflecting surface is very smooth and the reflected wave and direct wave are exactly out of phase at the receiver, the reflected wave may temporarily almost completely cancel the direct wave and cause a very deep fade in received signal strength. Cancellation is worst when the reflecting surface is a calm body of water, smooth moist earth or the thin layer of hot air that lays just above the surface of desert sand in the daytime.

In general, reflected waves are undesirable. Changes in the refractive qualities of the air cause the point of reflection to shift and the reflected and direct waves pass in and out of phase with each other causing wide variations in received signal strength.

Rough terrain, such as a rocky or wooded area, is generally a very poor reflector of radio waves. Such terrain either absorbs much of the radio energy or scatters it so that little reflected energy reaches the receiving antenna. For this reason, radio paths with reflection points in rough terrain have very little interference from reflected waves.

Refraction

Refraction occurs because radio waves travel with different speeds in different media. In free space (a vacuum) the speed is maximum. In any other medium, however, radio waves travel slower. As shown in Figure 1, when radio waves pass from dense air to thin air, their direction is changed. As the upper part of a wave front enters the thinner air it starts traveling faster than the lower portion which is still in the dense air. The result is that the path of the waves is bent or refracted.

When considering refraction of radio waves through the Earth's atmosphere, it is usually assumed that under normal conditions the atmosphere is densest at the Earth's surface and becomes thinner at higher altitudes. This variation in air density above the Earth causes radio waves near the surface of the Earth to travel more slowly

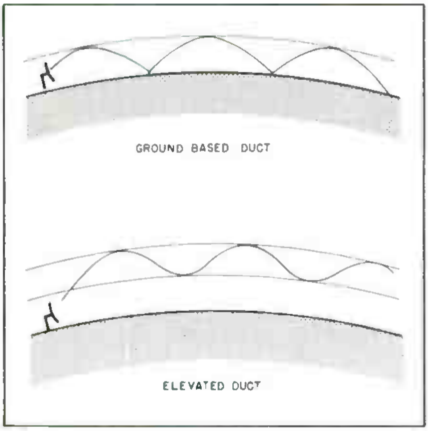


FIGURE 3. Ducts formed by stratification of the Earth's atmosphere. Ducts tend to trap radio waves and guide them around the Earth's Surface.

than those considerably above the surface. The result of these different velocities is a bending of the direction of wave travel which causes the waves to tend to follow the Earth's surface.

Because curved paths through the atmosphere are difficult to represent graphically, it is customary to draw profile charts of the Earth's surface with the Earth's radius represented as 4/3 actual size. The use of this fictitious radius approximately compensates (under average conditions) for the bending of the waves by the Earth's atmosphere and permits the illustration of radio paths on a profile chart as straight lines. An example of a profile chart with 4/3 Earth's radius is shown in Figure 2.

Simple refraction causes no great difficulty in the engineering of radio routes. Occasionally, however, refraction effects can seriously disturb the transmission of signals over line-of-sight paths. Under unusual conditions, the atmosphere becomes stratified with definite boundaries between layers of different densities. This causes the path of the radio waves to be bent first down and then up so that the waves become trapped in a layer of dense air. As a result the waves are guided along the air layer in much the same manner as microwaves travel in a wave guide. Layers of dense air that trap radio waves are commonly referred to as ducts.

The existence of a duct may either increase or decrease the received signal strength depending on whether the duct guides the waves toward or away from the receiving antenna. Ducts are more frequent near or over large bodies of water and in climates subject to frequent temperature inversions (stratification of air). While ducts may cause fading, their most important effect is that they sometimes guide radio waves well beyond the optical line-of-sight so that they are detected by distant repeaters of the same system. This type of interference can be avoided by changing transmitted frequencies at repeaters and locating repeater stations along a zig-zag path. Examples of two common types of ducts are shown in Figure 3.

Diffraction and Fresnel Zones

Ordinarily, radio paths are selected so that there is a direct line-of-sight between the transmitting and receiving antennas. However, a direct path between transmitting and receiving antennas is not necessarily a sufficient condition for good radio transmission. If a radio wave passes near an obstacle such as a hilltop or a large building, part of the wave front will be obstructed, and the amount of energy received will differ from that received if no obstacle were there. The cause of this difference is known as diffraction.

A simplified physical explanation of diffraction is shown in Figures 4 and 5. In Figure 4 a succession of unobstructed radio wave fronts are shown progressing from the transmitter to the receiver. The whole surface of each individual wave front contributes energy to the receiving antenna. However, energy from some portions of the wave front tends to cancel energy from other portions because of differences in the total distances traveled. The shaded areas in Figure 4 show the paths of energy that cancel some of the energy transmitted by the paths shown unshaded. The cancellation is such that half of the energy reaching the receiver is cancelled out. Most of the energy that is received is contributed by the large unshaded central area of that portion of the wave front that is closest to the receiver. If an obstacle is now raised in front of the wave so that all of the wave front below the line-of-sight is obstructed, (this is shown in Figure 5) half of the broad central area is obstructed and a greater loss of energy occurs. Under this condition the radiated power reaching the receiver is reduced to one fourth normal or by 6 db. If the obstruction is lowered (or the receiving antenna raised) so that all of the central zone is exposed, the power received by the receiving antenna is even greater than it would be if the obstacle were not there. This is shown in Figure 6.

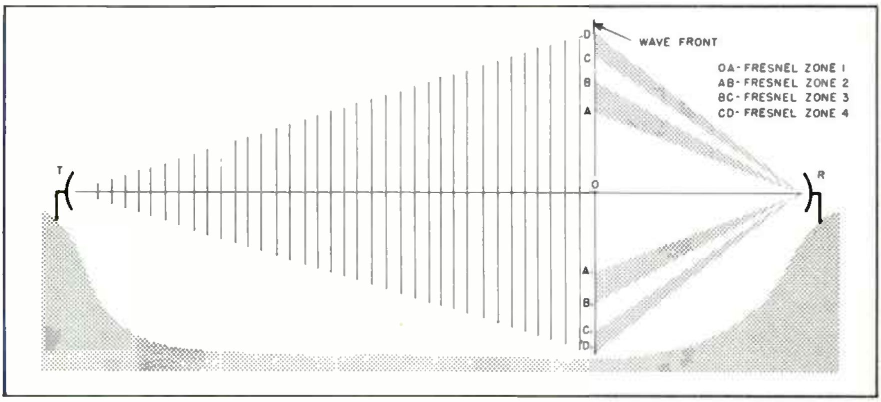
The various zones of the wave front that contribute either in-phase or out-of-phase energy are called Fresnel zones after their discoverer Augustin Jean Fresnel (1788-1827). The large central zone is called the first Fresnel zone and zones farther removed from the line of sight are called the second, third, fourth zones, etc. If the obstruction is such that the first zone is above the obstruction, the radio path is said to have first Fresnel zone clearance. In general, first Fresnel zone clearance is considered to be very desirable, although clearance of only one half the first zone is adequate. Of course, clearance greater than first Fresnel zone is also adequate.

Paths without adequate clearance are not desirable because refraction by the atmosphere may change. If a path just clears an obstacle under normal conditions, a change in refraction may cause the path to be obstructed. Careful construction of a profile chart and visual examination of the proposed route should show whether adequate path clearance is available.

Absorption and Scattering

Because no transmitting media other than free space is perfect, some radio energy is absorbed

FIGURE 4. Energy contribution from a wave front to a receiver. The dark areas are out of phase with the light areas.



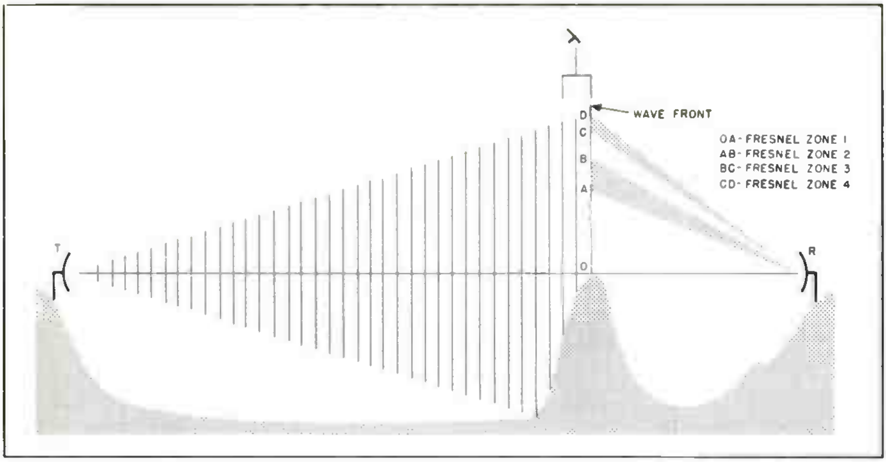


FIGURE 5. When an obstacle is raised in front of a receiver to the line of sight, half of the wave front is obstructed and the received signal strength is reduced by 6 db.

from a wave when traveling through the Earth's atmosphere. In clear weather, absorption is very slight for radio waves of less than 10,000 megacycles frequency. Rain, snow, and fog, however, can absorb or scatter large amounts of radio energy, especially at the higher microwave frequencies. Below 1000 megacycles, however, reduction of received signal strength by scattering and absorption by fog or precipitation is not a serious problem over paths of the usual length.

Radio Route Considerations

Each of the factors that affect radio propagation must be taken into consideration when planning a radio route. In many cases, visual examination of the route topography and a knowledge of weather conditions along the route are sufficient to determine the feasibility of proposed transmitter, repeater, and receiver locations. Where there is a doubt, profile charts can be used to determine more precisely the transmission conditions to be expected. In exceptional cases it may be desirable to make propagation tests.

Where visual examination indicates that the radio path is completely or partially over smooth terrain or water, the point of reflection from the terrain of the transmitted wave should be determined from a profile chart. The reflection point is located where the angle the transmitted wave makes with the Earth's surface is equal to the angle of the reflected wave.

If the reflection point is found to be on a smooth surface such as a flat field or water, relocation or a change in height of the transmitting or receiving antenna may be desirable. Often one of the antennas can be located so that it is masked from smooth ground or water reflection but still in line-of-sight to the other antenna. An example of using nearby terrain to mask unwanted reflections is shown in Figure 6.

How Bad is Fading?

Fading can only be determined absolutely over a particular path on the basis of experience just as the ice or frost attenuation of an open wire line must be learned absolutely by experience. It is the combined effects of all the various factors described above that can cause variation in received radio

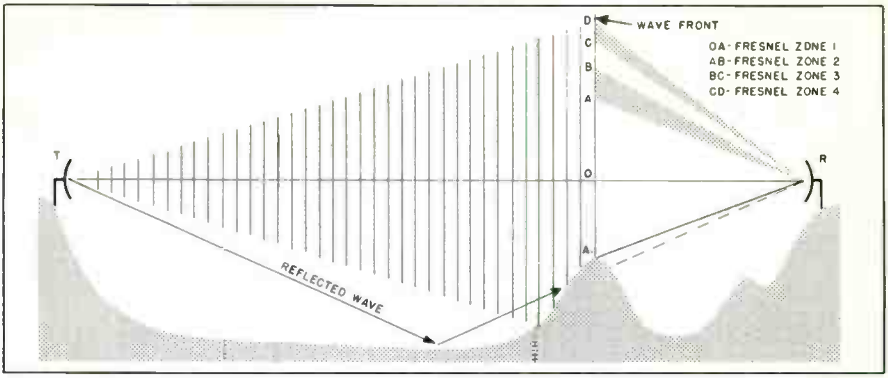


FIGURE 6. When the obstruction is lowered (or the receiving antenna raised) until all of the first Fresnel zone is exposed, the received signal strength is greater than if the obstruction did not exist.

frequency signal strength. Observations of fading over practical systems operating at frequencies below 1000 megacycles have shown that fade margins above normal space losses of 0.5 to 1.5 db per mile (depending on the terrain) will provide satisfactory transmission for 99.9 percent of the time. The 0.1 percent of time that transmission quality is below standards of such systems amounts to only 9 hours per year; which compares favorably with the performance obtained from many wire line and cable systems. In most cases a well planned system can be expected to be completely out of service due to excessive fades for less than 1 hour per year.

Conclusions

The factors affecting radio propagation over line-of-sight paths, while differing greatly in details, have much the same effect on transmission quality and reliability as do weather and temperature changes on open-wire lines and cables. The problems created by these factors are as amenable to solution as are the transmission problems of conventional wire and cable.

Each of the factors affecting the propagation of radio waves contributes somewhat to variations in received signal strength. However, by understanding the way in which each factor affects propagation, by engineering the radio system to minimize their effects, and finally by allowing a sufficient margin for unavoidable fading, very high quality circuits can be obtained with maximum reliability and reasonable cost.

What are Roof Filters?

Low-pass filters when used for certain applications in telephone communications are called 'roof filters.'

The principle application of 'roof filters' is in the repeater and terminal equipment of carrier systems where it is desirable or necessary to limit the frequency response of the equipment assembly to only those frequencies needed for normal transmission.

Roof filters, when used in this manner, reduce unwanted higher frequencies induced into the circuit from external sources. This action prevents overloading of amplifiers, improves run-around crosstalk suppression, and lessens the possibility of high frequency singing.

THE Lenkurt Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 4 NO. 1 JANUARY 1955

DIFFRACTION OF MICROWAVES

In the June 1954 issue of the Lenkurt Demodulator, an explanation was given for the change observed in received signal strength when an obstacle was in or near the line of sight of a microwave system. The June article attributed the change to a phenomenon called diffraction in which it is considered that the whole surface of a wave front contributes energy to a receiving antenna. According to diffraction theory, when part of the wave front is obstructed, the amount of energy contributed to the receiving antenna is different than when the wave front is unobstructed. Not explained in the earlier article, however, was how energy could be received from a portion of a wave front that was not traveling directly towards the receiving antenna.

The purpose of this article is to explain this apparent discrepancy in microwave propagation by giving a brief description of the mechanics of wave travel.

The manner in which waves travel is described by a fundamental principle of optics known as "Huygens Principle". This principle, though originally applied only to light, applies equally well to microwaves. Huygen, in attempting to explain some of the characteristics of light, proposed that light traveled as waves in an "ether" and that the orderly progression of the waves through the ether could be explained if every point of an advancing wave front were considered to be a secondary source of radiation, and that each point of a wave front be considered the source of secondary wavelets emanating in the general direction of wave travel.

Huygen's principle is illustrated by the sketch shown in figure 1. In this sketch, a mask is placed in the path of a train of microwaves completely blocking wave travel except for a small hole. If a screen is placed behind the mask, it will be found that the whole screen will be illuminated by the pinhole source with the strongest illumination occuring at the point (P) that is in line of sight with the pinhole and original source of radiation. If a second pinhole is placed in the mask, it will be found that this second pinhole will also contribute some radiant energy to point P.

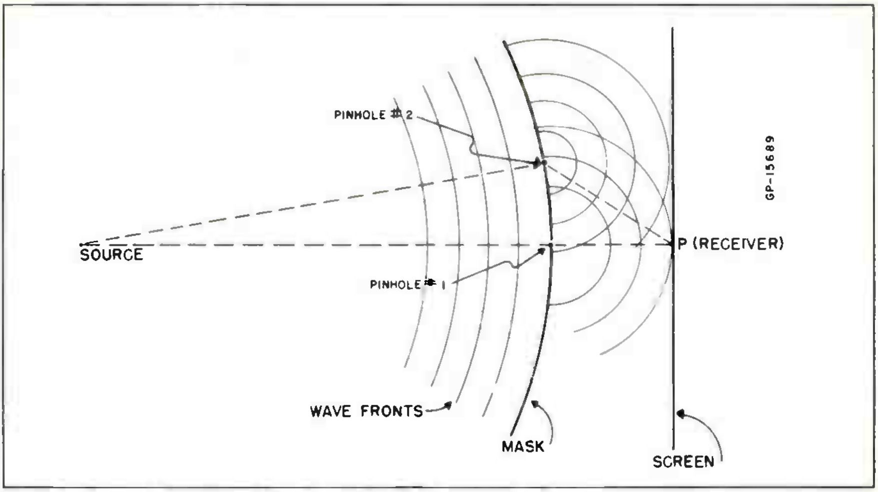


FIGURE 1. Illustration of Huygen's Principle. The pinholes in the mask act as secondary point sources of radio energy.

Thus, if a very large number of pinholes are made in the mask (or the mask is removed), energy will be received at point P from all parts of the wave front.

In the example given in figure 1, energy arrives at point P over two different paths; one path being longer than the other. If the difference in path length is some odd multiple (1, 3, 5, etc.) of a half-wave length, the two sources of illumination will tend to cancel each other; if the difference is an even multiple of a half-wave length they will tend to reinforce each other. When the mask is removed, the combined result of all cancellations and reinforcements is such that the energy received at P appears to come from the source in a straight line. If, instead of a mask, an obstacle such as a hill or tall building, is placed in or near the path of the radio wave front, the balance between the cancelling wavelets and the reinforcing wavelets is upset and the amount of energy received at point P will be greater or less than the energy received without the obstacles' presence. Even if the receiver lies in the geometrical shadow of the obstacle (for example, a microwave receiver masked by a hilltop), some energy will be received.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 2 NO. 1 JANUARY, 1953

How to Prepare and Use PROFILE CHARTS OF RADIO LINK ROUTES

With the help of an accurate profile chart of a radio link route, competent engineers often can predict performance closely enough to make actual propagation tests unnecessary.

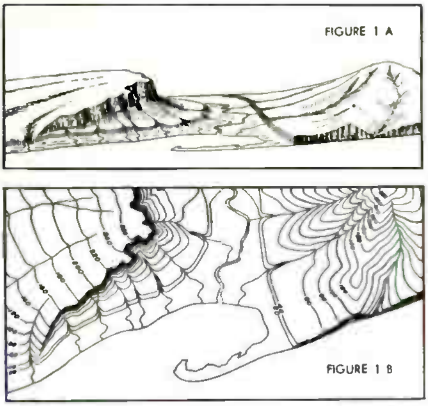
In this article a few aspects of preparing and using profile charts are discussed to give the reader a practical acquaintance with some of the problems encountered when installing a radio link.

Radio waves in the higher frequency ranges used for point-to-point radio links (commonly called microwaves) exhibit many of the properties of light. They travel in relatively straight lines, and they are bent (refracted) by the atmosphere, reflected by solid objects or surfaces, and diffracted by physical objects in or near the transmission path. To predict the effect of these properties upon the propagation of energy between two antennas, the nature of the terrain between antennas must be considered.

The first step in estimating the propagation characteristics between two antenna sites is to assemble elevation data about the intervening terrain. Using this data a profile chart is prepared to show the elevation of all hills, ridges, tall buildings, or other obstacles that might interfere with line-of-sight transmission of radio waves. A satisfactory transmission path can then be intelligently selected by analyzing the information on the profile chart along with any other pertinent data.

Sources of Data

Several different sources can provide the data required for preparing a profile chart. In many cases it can be obtained from topographic maps with contour lines showing elevation of land at convenient intervals. Information on maps of this type, prepared by the United States Department of the Interior, is available from the Director, United States Geological Survey, Washington, D.C. A section of a typical map of this type is shown in Figure 1B. Figure 1A is a sketch of the area which this map represents.



COURTESY U. S. DEPT. OF THE INTERIOR

FIGURE 1. *Contour maps provide an excellent data source for preparing profile charts. Figure 1B is a contour map of the area sketched in Figure 1A.*

For locations where these topographic maps are unavailable, local county surveyors can often provide the required data.

If no previously prepared maps can be obtained, a special survey may be required although, in many cases, sufficient data can be obtained through one or more "common sense" procedures. In one such method an altimeter is used to determine the relative heights of land along a proposed transmission path. In some cases a spotlight (or sun reflecting mirror) can be used to determine if a line-of-sight path exists and, if conditions permit, the light source can be moved vertically to determine path clearance.

If data obtained from topographic maps shows that only marginal clearance exists, the elevations of high points should be checked by survey or altimeter to insure their accuracy.

Consulting engineering services are ordinarily available to make either ground or aerial surveys of proposed radio link routes.

Preparing the Profile Chart

After tentative antenna sites have been selected, and the relative elevation of land between these sites has been determined, a profile chart can be prepared. In some cases a complete profile such as those shown in the examples will be necessary; in other cases only certain hills or ridges need be indicated to be sure adequate path clearance exists.

An important factor influences the shape of a profile chart. Although the surface of the earth is curved, microwaves tend to travel in straight lines. However, they are bent (refracted) a small amount by the atmosphere. The amount of bending (refraction) varies with atmospheric conditions. The effect of refraction is such that if a profile chart is prepared on the basis of four-thirds (4/3) of true earth radius, a straight line between antenna sites will indicate clearance between the actual transmission path and the earth. However this factor of 4/3, which would increase the permissible distance between antennas, is not always accurate. Under some atmospheric conditions the refraction caused by the atmosphere will diminish and the actual transmission path will approach true line-of-sight conditions.

Because reliability and continuity of service are very important for a multichannel radio link, many radio engineers prefer to be conservative and base propagation predictions on the basis of path clearance shown on a profile chart prepared with true earth radius.

The effect of using 4/3 and true earth radius for the same path is shown in the two sketches of Figure 2. Figure 2B shows that when the amount of atmospheric bending is normal, a clear path is indicated when planning is done with a profile chart drawn with 4/3 true earth radius. Under abnormal conditions, however, if a true line-of-sight path exists, the transmission path between antennas is interfered with by the ridge shown on the chart in Figure 2A, drawn with true earth radius.

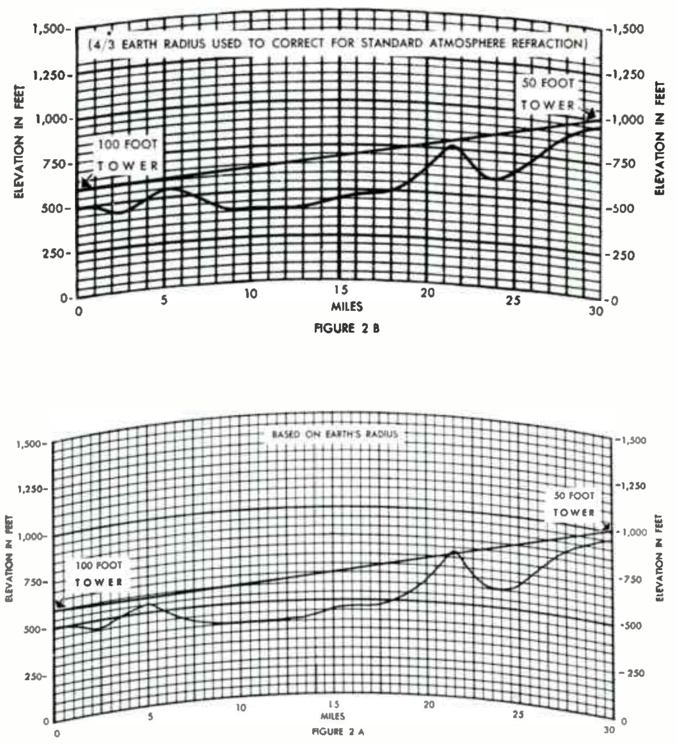
Since the choice of earth radius varies with topography and climate and is influenced by the amount of fading allowable, the advice of a competent radio engineer should be obtained when a radio link is being installed over a path where some question about clearance or other factors exists.

Printed forms are available for plotting profile charts. A form used by Lenkurt's engineers has been used for the examples in this article. This form, which uses true earth radius, is based on the relationship between the height of an observer and the distance to the horizon where, if "h" is in feet and "d" is in miles, $h = 2/3d^2$. If 4/3 earth radius is used, this relationship becomes $h = 1/2d^2$. The scale of either chart can be changed if desired by doubling the horizontal interval and quadrupling the vertical interval (or by dividing the horizontal interval by 2 and the vertical interval by 4).

Using the Profile Chart

An accurately drawn profile chart will show whether or not

FIGURE 2. *Profile charts prepared with true earth radius provide a more conservative estimate of propagation conditions than those prepared with 4/3 true earth radius. The clear path indicated in Figure 2B (4/3 true earth radius) is interfered with when true earth radius is used (Figure 2A).*



adequate path clearance exists for the transmission path between antennas. The chart can also be used to determine the "reflection point" as shown in Figure 3.

The path clearance desired varies with frequency and with distance from the transmitting antenna. Lenkurt's engineers usually consider that about 75 feet minimum clearance is acceptable for a system operating at 900 megacycles.

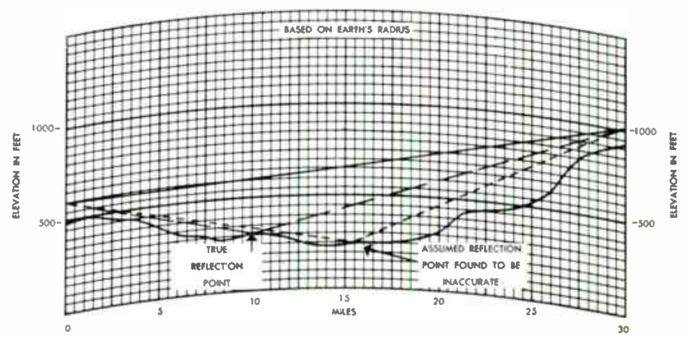
The effect that a ray reflected by the earth will have depends to a great extent upon the character of the surface at the reflection point. A strong reflection will be caused by a smooth body of water or by smooth earth while a weaker reflection will come from wooded terrain. In general, a strong reflected wave is undesirable because it can cause fading and distortion.

The reflection point can be found from a profile chart by using a "cut and try" method illustrated in Figure 3. By inspection, an assumed reflection point is selected and straight lines are drawn between this point and the two antennas. The assumed reflection point is the true point if the two lines rise the same number of feet in going an equal number of miles to the right and left of it. In Figure 3 the dotted line indicates an incorrect reflection point because the lines do not rise equal amounts in equal distances to the right and left of the assumed point. The dashed lines, however, indicate the true reflection point.

Conclusions

An experienced radio engineer often can predict the effects of available path clearance and reflection with sufficient accuracy to determine whether or not proposed antenna sites can be used. In many cases, the losses caused by unfavorable topography can be overcome by using higher gain antennas, higher transmitter power or lower loss cables between radio equipment and antennas. Sometimes different antenna sites or higher antenna mountings might be required. In any case an accurate profile chart is an invaluable tool for the experienced engineer planning radio links.

FIGURE 3. *Reflection point is easily determined from a profile chart by using a "cut and try" method.*



THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 5 NO. 12 DECEMBER, 1956

Selection of

TOWERS AND ANTENNAS

for Microwave System Use

In microwave installations, tower and antenna equipment should be selected and located so as to assure maximum transmission quality and reliability. Both technical and economic factors are involved. In many cases, the final choice is a compromise between them.

This article discusses some of the technical and economic factors that enter into the selection of towers and antennas.

Microwave systems in the United States operate above 890 megacycles. Because they behave somewhat like light waves, microwaves can be reflected from smooth conducting surfaces and focused by reflectors or lenses. When radio waves pass from one medium to another (as from dry air to moist air) they are bent or refracted in the same manner as light waves are bent by lenses or prisms. Radio waves bend around large obstacles in their path by a process known as diffraction. And at frequencies above 4000 mc, they are scattered by small particles such as rain and snow.

Each of these effects can cause variations in the received signal strength. Consequently, it is necessary that they be considered and allowances made for them when engineering a radio system. The proper selection and location of towers and antennas is an important part of that engineering.

Towers and antennas are generally chosen after surveys have been made of the area where they are to be located. The microwave propagation path, topography of surrounding area, weather conditions, and economic factors are all considered. They determine which of the many available types of towers and antennas will best meet system requirements.

Propagation Path Considerations

Adequate clearance for the microwave propagation path and its length are among the most important consid-

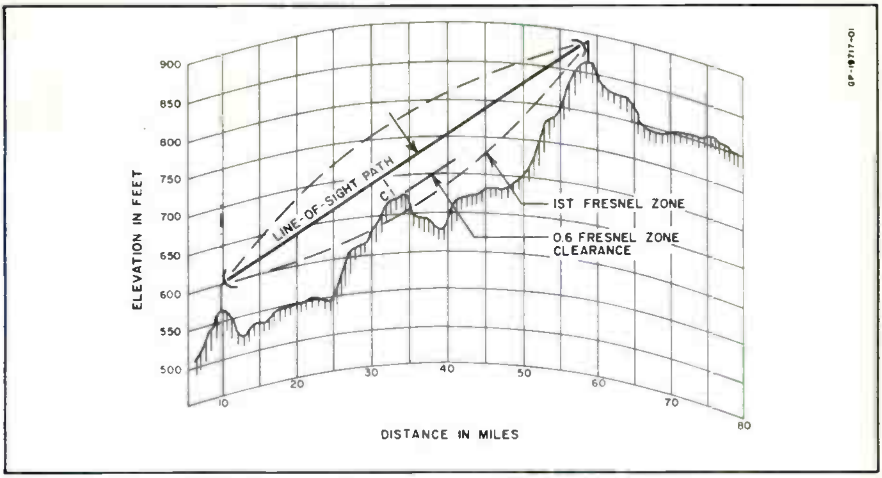


FIG. 1. Profile charts are prepared from contour maps to obtain elevations of points along the propagation path. Adequate path clearance and antenna heights are determined from these charts.

erations in system engineering. They determine the number of repeater stations, antenna reflector sizes, and tower heights.

Propagation path clearance and antenna heights are usually established through the use of profile charts. A chart, such as the one shown in Fig. 1, is plotted from contour maps of the proposed area. As trees and buildings are not shown on ordinary contour maps, a visual survey of the area is often made to verify the path. Altimeter readings are also taken to check the accuracy of the profile chart or to plot the chart if contour maps are not available.

After the profile has been checked, the path clearance is determined. On the chart in Fig. 1, point C is the worst obstacle in the system's path. To achieve propagation essentially the same as free space, the line-of-sight path should pass over the obstacle with a clearance of 0.6 of the first Fresnel zone.[1] With this clearance established, point C can then be used as a "leverage point" to determine the most suitable antenna heights.

Tower and Antenna Locations

Desirable tower locations will provide a clear, unobstructed path with the necessary clearances. Mountain tops, hills, buildings, and other tall structures can reduce the height requirements for towers. Other factors to be considered are land costs and availability, accessibility for post-installation maintenance and the provision for interconnection with other facilities.

Although sometimes impractical, paths are often planned to bypass certain areas that may later disturb system operation. Large bodies of water, large flat land areas, power plants, high voltage power lines, chemical plants and areas that flood should be avoided if possible.

[1] See Lenkurt Demodulator for June, 1954 and January, 1955.

Towers located near airfields will be subject to stricter requirements. Civil Aeronautics Administration regulations require that all towers within a specified distance of an airfield be equipped with warning lights and a light failure alarm. And in addition, the towers must be painted red and white.

Tower sites are often located in remote or sparsely settled areas. This is especially true of repeater installations. When selecting such a site, it is important to evaluate the expense of building access roads and the obtaining of power to operate equipment. Locations near existing roads and power facilities will reduce both the expense of installation and subsequent maintenance.

Equipment Strength and Rigidity

When selecting microwave system towers, a primary factor to consider is the tower's ability to safely withstand years of wear under varying weather conditions. The possibility of high winds, heavy ice formations and corrosion should be anticipated.

Present-day tower construction methods follow safety and construction standards agreed to by the industry. The standards recommended and published by the Radio-Electronics-Television Manufacturers Association (RETMA) include loading specifications, unit stress allowances, foundation requirements and quality of materials and workmanship. These are minimum requirements. Most installations will require a greater safety margin.

According to standard specifications, all towers higher than 600 feet and those of any height located within a city's limits must be capable of withstanding 30 pounds per square foot of horizontal wind pressure (equal to an actual wind velocity of about 85 miles per hour). For all other towers, the minimum requirement is 20 pounds per square foot. These requirements apply to antennas as well.

Under certain conditions, towers and

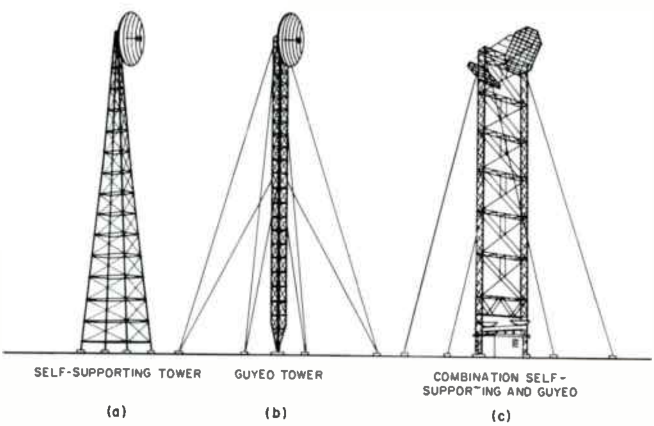


FIG. 2. Three types of steel towers are available for microwave system use. Towers (a) and (b) are shown with parabolic reflectors. Tower (c) supports two passive reflectors with two parabolic reflectors mounted on the roof of the building.

antennas will need to be much stronger. In some southern hurricane areas, for instance, local laws require that they be able to withstand 50 pounds per square foot of wind pressure. Local restrictions should be checked when determining tower strength. In all cases, it is advisable to allow a wide safety margin.

Winds present still another problem—torsional twist of the upper portion of the tower. For maximum reception of the transmitted energy, the transmitting and receiving antennas must be aimed directly at each other. In actual practice, however, some torsional twist is unavoidable.

Field tests have determined limits for the maximum torsional twist that can be allowed. For systems operating in the 900 mc region, ± 3 degrees is allowed. Since the twist is more noticeable at higher frequencies, systems operating in the 6000 mc region have a limit of $\pm \frac{1}{2}$ of a degree. In general, twists of about $\frac{1}{3}$ of beam width are permissible.

Equipment Types and Applications

Supporting structures for microwave antennas vary from tubular frame brackets mounted on the side of a building to large steel towers over 300 feet high. In some cases, the desired antenna height may be low enough to use the radio equipment building itself, or poles such as those used for wire-line facilities, or any other structure strong enough to support antennas and reflectors.

For high antenna elevations where no other suitable structures are available, steel towers are needed. They are available in three general types. These types, shown in Fig. 2, are (a) self-supporting; (b) guyed; and (c) a combination of self-supporting and guyed. The structures vary somewhat from manufacturer to manufacturer, but most conform to RETMA and government standards for safety, wind loading, and quality of construction.

The choice of a tower is largely dependent upon the land area available and the cost of transmission lines from radio equipment to the antenna. For example, self-supporting towers are generally used atop buildings, in cities, or any other area where space is limited. When towers are put on buildings, extensive reinforcement of the building may be necessary to support the structure.

Guyed towers are generally less expensive than the self-supporting types. They are used in areas where space is not a limiting factor or when the cost of the transmission line from the central office is less than the cost of a self-supporting installation.

Combination self-supporting and guyed tower installations have many of the better features of the two basic types. As shown in Fig. 2(c), combination towers can straddle-mount the radio equipment building to utilize the advantages of passive reflectors. The radiating and receiving antennas, located on top of the equipment building, eliminate the need for costly high-frequency transmission lines to the top of the tower. Other types of antennas and reflectors can also be used.

Antenna selection is determined by the operating frequency, desired gain and directivity, and ability to meet wind loading requirements. Antenna gain and directivity increase with an increase

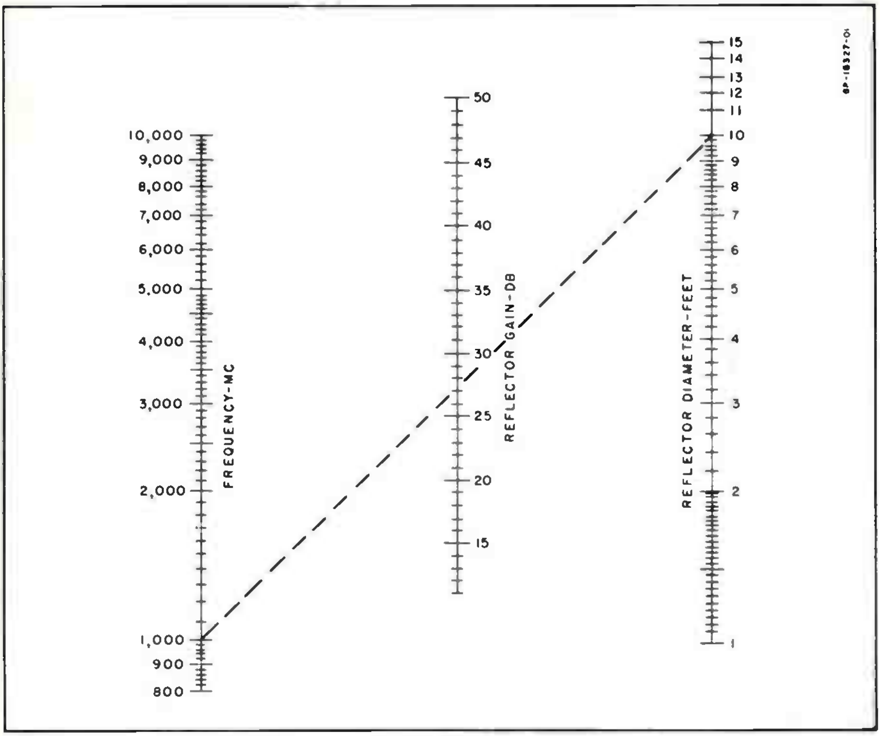


FIG. 3. Antenna gain for one-half wavelength dipole antenna with a parabolic reflector. The gain is with reference to an isotropic antenna with an assumed efficiency of 65 per cent.

in reflector size or operating frequency. Thus the selection of antennas is usually dependent upon system transmission requirements.

High antenna gain and directivity may be obtained by locating a large parabolic reflector behind the radiating element. The approximate gain of such an arrangement can be determined from Fig. 3. For example, a 10-foot reflector would have a gain of 28 db at 1000 megacycles.

Using the nomograph, antenna gain for a one-half wavelength dipole antenna with a parabolic reflector can be determined for given values of frequency and reflector diameters. The resulting gain is with reference to an isotropic (omnidirectionally radiating) antenna with an assumed efficiency of 65 per cent.

Conclusion

The factors involving tower and antenna selection are many and varied. Propagation path clearance, reflection, wind loading, and antenna directivity are but a few. However, much experience in system engineering has been accumulated by the manufacturers of microwave equipment over past years. This experience, properly applied, assures the microwave system buyer many years of reliable communications.

HE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 1 NO. 3 MAY, 1952

How to Determine

TRANSMISSION LOSSES IN RADIO LINKS

Point-to-point radio equipment is now available for economically transmitting light or medium traffic loads over short or medium distances. Application of this equipment presents telephone engineers, accustomed to working with open wire and cable, with the problem of planning radio links as integral parts of existing plant facilities. This article is presented to acquaint telephone men with the basic problems involved in determining the transmission losses in a radio system. Terms are defined, an outline of one method which can be used for determining losses in a radio link is included, and references are given to technical literature which covers the subject in greater detail.

In recent years telephone companies and industrial organizations have made numerous installations of radio communication systems. These installations have shown that in many cases radio links are the most economical means of obtaining dependable point-to-point communication channels.

At the present time point-to-point radio systems often use frequencies in the range of 900 megacycles and higher. The choice of frequency is influenced by the allocation available (to prevent conflict with other radio applications), the number of channels required, the power requirements, and the required reliability.

In general, higher antenna gains are obtained and more channels can be transmitted over a single radio link as the frequency increases; at the same time, however, the equipment becomes more specialized and variable factors affecting transmission losses have more effect.

Equipment for medium traffic installations is currently available in the 900 megacycle range. The Lenkurt Type 72A radio system, for example, can transmit up to 36 toll quality carrier derived voice channels with repeater spacing of 25 to 35 miles. Conventional low frequency design practices can usually be applied to 900 megacycle radio equipment, and the problem of fading (irregularities in the received signal strength) is less severe than at higher frequencies.

The principles discussed here can be applied, of course, to systems operating at either higher or lower frequencies.

A Typical Radio Link

A profile map of a typical radio link is shown in Figure 1. This type of map is often used when planning radio circuits to determine what the minimum path clearance will be with various antenna heights and locations. A line-of-sight path with adequate clearance between the path and the highest point between antennas is necessary because radio waves at frequencies above about 50 megacycles travel in essentially straight lines. Any obstacles between the transmitting and receiving antennas will tend to reflect or absorb the signals.

A certain amount of bending is introduced into the line of sight path by atmospheric refraction. To correct for this, the curvature of the earth in a profile map is drawn with a radius to indicate four-thirds times the true radius of the earth.

Levels in a Radio Link

For planning purposes a radio link can be treated in the same manner as any other telephone circuit. Levels at any point in the radio circuit can be determined, and the entire system can be designed to provide the desired signal strength at the receiving terminal.

Figure 2 shows a block diagram and a level diagram of a typical radio link operating at 900 megacycles with the transmitting and receiving antennas separated by 25 miles.

Operating levels are based on the specifications for the Lenkurt Type 72A radio system designed for the 900 megacycle band. This system has a transmitter output of +37 dbm (5 watts). Operation for toll quality multi-channel systems requires a receiver input level of -63 dbm (160 microvolts) to provide a noise level at the receiver output for the worst channel of less than +29 dba at the -9 level. The antenna gain is 27 db referred to an isotropic radiator at each end; the loss in the lines to the antennas is 1 db at each end; and the propagation loss between antennas is approximately 123 db.

With these theoretical losses the receiver input is -34 dbm which allows an operating margin of 29 db. Therefore the total losses can increase by 29 db before the noise level rises to a value of +29 dba, equivalent to a weighted signal to noise ratio of about 52 db.

The manner in which these various gains and losses were determined will be described below.

Total Losses

Total attenuation between the transmitter output and the input to the receiver is the sum of the losses and gains shown in the level diagram of Figure 2. Sometimes, however, specifications for radio systems and published methods of calculating the losses group two or more of the losses together. The antenna gain, for example, is often included in formulas for calculating propagation losses.

Manufacturers' specifications and published data for calculating losses should always be examined carefully to determine exactly which losses are included in any particular figure or formula.

Power Levels

Since telephone and radio engineers developed their present-day methods independently, each group

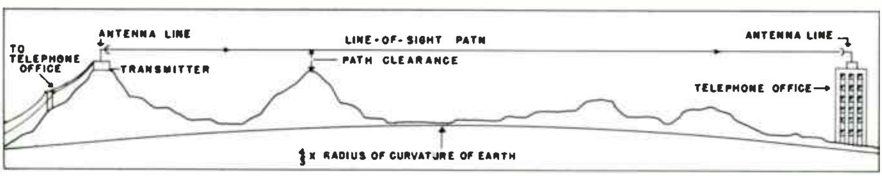


Figure 1. Profile map of a typical radio link.

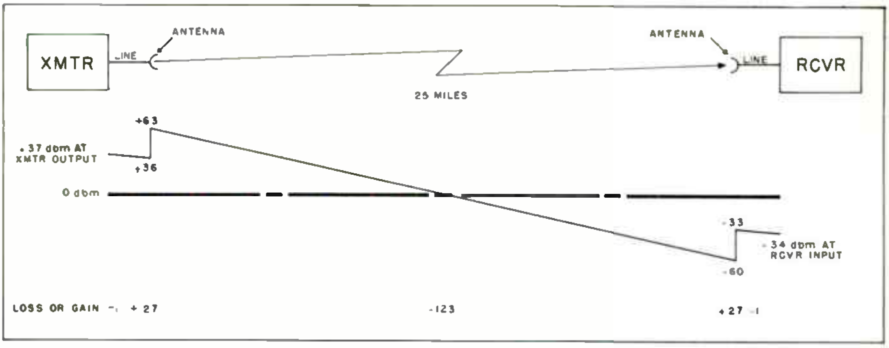


Figure 2. Block diagram and level diagram of a typical radio link.

began stating power levels in terms best suited to their own requirements.

In the telephone industry the dbm (decibels above or below one milliwatt) has been accepted as the standard term for expressing power levels. One reason for this selection is that all new telephone facilities must have transmission characteristics similar to those of existing plant facilities. The determining factor is usually the allowable noise level, and this is most conveniently expressed in db.

In the radio industry it has been common practice to express transmitter output power in watts, while the receiver input level is often stated in microvolts (μv) to give an indication of receiver sensitivity.

When planning a radio circuit for telephone applications it is usually most convenient to convert watts and microvolts to dbm. This can be done easily by using the following formulas.

The basic expression for power in decibels is:

$$\text{db} = 10 \log \frac{P_1}{P_2}$$

Since dbm refers to db above or below a reference level of one milliwatt (.001 watt) any power expressed in watts can be converted to dbm by:

(a) If P is more than one milliwatt:

$$+\text{dbm} = 10 \log \frac{P}{.001}$$

(b) If P is less than one milliwatt:

$$-\text{dbm} = 10 \log \frac{.001}{P}$$

Thus, if the output of a radio transmitter is given as five watts, it can be expressed as 10 log 5000, or +37 dbm.

To convert microvolts to dbm, it is first necessary to know the impedance across which the voltage is measured. This is almost always 50 ohms in radio receivers operating at frequencies above 100 megacycles. When the impedance is known, microvolts can be converted to watts, and then expressed in dbm.

For example, an input level of 160 microvolts is first expressed in watts:

$$P = \frac{E^2}{R} = \frac{(160 \times 10^{-6})^2}{50} = 5.12 \times 10^{-10} \text{ watts}$$

and then in dbm:

$$-\text{dbm} = 10 \log \frac{.001}{5.12 \times 10^{-10}} = -63 \text{ dbm}$$

Loss in Antenna Lines

Attenuation in the lines to the transmitting antenna and from the receiving antenna will depend upon the type and length of line used.

Various types of coaxial cable or concentric transmission line are available, and the particular one used will depend upon the location of the line and the amount of attenuation which can be tolerated. Information on attenuation per foot of various types of line can be obtained from standard handbooks or from manufacturers' literature. The attenuation of several common types is shown in Table 1.

Antenna Gain

The term 'antenna gain' refers only to an apparent gain rather than to an actual increase in power between the input to an antenna and its output. This apparent gain is produced by directing the radiated energy in the required direction.

If the output of a radio transmitter were applied to an 'isotropic' antenna (one which would theoretically radiate energy equally in all directions) there would be no antenna gain. It is only when the radiated energy is concentrated in a particular direction that the concept of gain is introduced.

A typical radiation pattern for an isotropic antenna is shown in Figure 3a. The theoretically perfect antenna would be a small sphere, and the field pattern would then be a concentric sphere.

The simplest type of antenna commonly used is a half wave dipole which produces a radiation pattern similar to that of Figure 3b. By concentrating the radiated energy in one plane, a gain of 2. 15 db is obtained (as referred to an isotropic antenna).

In a microwave radio link the energy radiated from a transmitting antenna is wasted unless it is sent towards the receiving antenna. To obtain this directivity an antenna array can be constructed, or a reflecting screen or parabolic reflector can be used to focus the radio waves in the same manner that a searchlight focuses a beam of light. A radiation pattern similar to that shown in Figure 3c is obtained with a parabolic reflector antenna.

Parabolic reflectors are quite commonly used at frequencies of 900 megacycles and higher, with the diameter of the reflector for a given gain decreasing as the frequency increases. The gain of this type of antenna system increases with both the diameter of the reflector and the frequency of the signal. Gains of 20 to 27 db are commonly obtained at 900 megacycles.

Antenna gain can be referred to either an isotropic antenna or to a dipole. If manufacturers' specifications use a dipole as reference, 2. 15 db should be added to obtain the absolute gain as referred to an

	Description of Line	Attenuation per Hundred Feet
1.	3/8 in. solid dielectric, flexible coaxial cable, Type RG8/U	8.0 db
2.	7/8 in. solid dielectric, flexible coaxial cable, Type RG17/U	3. 7 db
3.	7/8 in. air dielectric, semi-flexible, soft drawn copper coaxial cable	1. 19 db

Table 1. Attenuation per hundred feet at 900 megacycles for three common types of transmission lines.

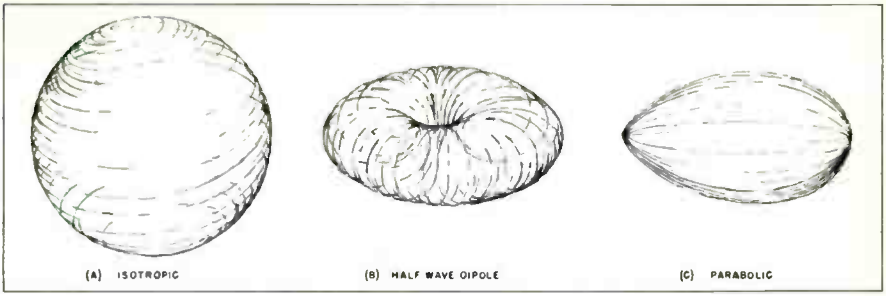


Figure 3. Typical radiation patterns of isotropic, dipole, and parabolic reflector antennas.

isotropic antenna.

Propagation Loss

The propagation loss of a radio system is, as the name implies, the difference between the power leaving the transmitting antenna and the power entering the receiving antenna. The ideal propagation loss, and the one which all radio systems attempt to equal, is the free space loss which would occur if all variable factors could be disregarded. Free space loss depends only upon the frequency and the distance between antennas.

Free space attenuation at 900 megacycles between isotropic radiators is shown for various distances in Figure 4. Note that the free space loss is increased by only 6 db when the distance between antennas is doubled; this is contrasted to open wire or cable where the loss per mile in db is a constant so the total loss in db is doubled if the distance is doubled.

The curve of Figure 4 is based on the formula:[2]

$$L = 10 \log 4543 f^2 + 10 \log d^2$$

where L is free space loss in db, f is in megacycles and d is in miles.

Average propagation loss for normal repeater spacings with adequate path clearance will be approximately equal to the theoretical free space loss.

Fading (irregularities in received signal strength) occurs when signals are reflected by obstacles in the path or refracted by abnormal atmospheric conditions. Since fading will have some effect on the propagation loss, an allowance (operating margin) should be made for it when a system is planned.

Operating margin requirements depend on the reliability required and upon the location of the radio system. Experience to date has indicated that, on an average, fades of 30 db are experienced about 0.1 percent of the time on 25-mile radio links (the received signals will be 30 db or more below normal for less than 9 hours per year); 20 db fades are experienced about 1 percent of the time (received signals will be 20 db or more below

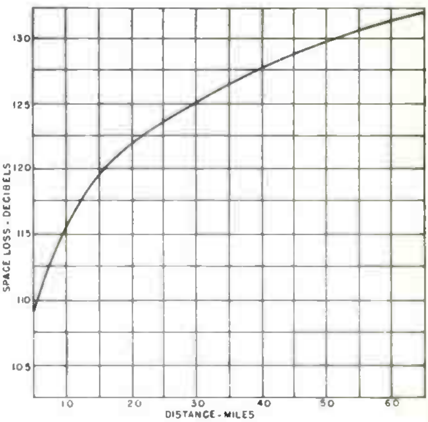


Figure 4. Free space attenuation between isotropic radiators at 900 megacycles.

normal for less than 90 hours per year).

Radio systems now in operation commonly have operating margins of from 1/2 to 2 db per mile.

System Calculations

For any given radio link consisting of a transmitter, receiver, and antenna system, an approximation of the overall performance can be obtained by comparing the maximum attenuation permissible between the transmitter output and the receiver input with the total calculated theoretical loss between the same points.

Consider a Lenkurt Type 72A System as an example. The transmitter output is +37 dbm. A receiver input of -63 dbm is required to obtain a noise level for the worst channel less than +29 dba at the -9 level (approximately 52 db weighted signal to noise ratio). Therefore, the maximum permissible loss between the transmitter output and the receiver input for toll quality circuits is 100 db, and the sum of the line losses, antenna gains, and propagation loss should not exceed this figure.

The loss in the transmission lines to and from the antenna will depend upon the type of cable used and the distance. A loss of 1 db at each end would be typical with air dielectric coaxial cable. If ten foot parabolic reflector antennas are used for both transmitting and receiving, the gain of each would be approximately 27 db referred to isotropic antennas.

The free space loss between isotropic antennas separated by 25 miles is approximately 123 db (from Figure 4).

Then the sum of free space loss, transmission line loss, and antenna gain gives the total attenuation between the output of the transmitter and the input of the receiver. Since this total is 71 db while the maximum permissible attenuation is 100 db, a margin of 29 db is available to provide for fading or other propagation irregularities.

The amount of margin required will depend upon the factors which affect propagation loss and upon the allowable deviation from the nominal maximum channel noise level.

SYSTEM RATINGS

(1)	Transmitter Power.	+ 37 dbm
(2)	Minimum received signal level to maintain noise level below +29 dba in worst channel	— 63 dbm
(3)	Maximum permissable loss between transmitter output and receiver input.	100 db

TRANSMISSION CALCULATIONS

(4)	Transmitting antenna gain.	+ 27 db
(5)	Receiving antenna gain.	+ 27 db
(6)	Allowance for losses in antenna lines at both ends.	— 2 db
(7)	Free space loss.	— 123 db
(8)	Total losses between transmitter output and receiver input.	71 db

OPERATING MARGIN

(9)	Allowance for fading and other additional system losses. (3) - (8)	29 db

Conclusions

When construction of a radio link is contemplated, the selection of basic radio equipment, antennas, and antenna sites can be made from the results of preliminary calculations of the type outlined above.

A thorough study should be made of the factors which affect propagation loss and an operating margin should be established which will provide the desired degree of reliability.

THE Lenkurt® Demodulator

NEWS FROM LENKURT ELECTRIC

VOL. 5 NO. 10 OCTOBER 1956

ESTIMATING THE NOISE PERFORMANCE

of FM Microwave Systems

The purpose of microwave system engineering is to provide a facility that has the desired quality and reliability at minimum cost. Although many different factors affect over-all circuit quality, the amount of noise introduced into a system is one of the most important.

In this article, system noise is divided into five distinct categories. Each is discussed and its importance to the over-all noise performance of the system is evaluated.

When a microwave system is to be used for transmission of voice channels, the primary performance characteristics which must be considered are frequency response, level stability, frequency stability, and system noise.

The first three of these characteristics can be controlled by the radio and channelizing equipment manufacturers through engineering and production techniques. The user is most interested in them when selecting equipment that will provide a transmission facility with the desired quality and bandwidth. The fourth characteristic, system noise, can be controlled only partially by equipment manufacturers because it is affected by such factors as the type of channelizing, system length, terrain, and path engineering as well as equipment design.

The total noise in a microwave system can be separated into five major categories: channelizing equipment noise, transmitter noise, receiver noise, intermodulation noise, and path distortion noise. The magnitude and character of the noise in each category are important factors in determining system details

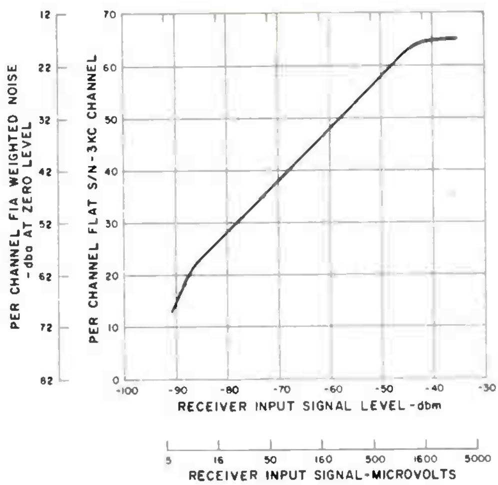


FIG. 1. Typical thermal noise characteristics showing noise with respect to receiver input power.

such as section length, antenna heights, and number of repeaters.

Noise Objective

One often-stated design objective for the noise performance of microwave systems is based on the noise objective used in the enginering of individual toll trunk links for the nation-wide toll dialing system. The objective for nation-wide toll dialing is the maintenance of average noise below 31 dba[1] per link for more than 50% of the time and below 52 dba for 99.9% of the time. The 31 dba noise objective is for an entire link, whether the link consists of a single type of facility or a combination of facilities placed in tandem.

If the link consists of two or more separate facilities in tandem, the permissible noise of any one of them depends upon the noise characteristics of the others. For example, if a cable section producing 29 dba of noise is placed in tandem with a microwave section to form a link, the microwave section noise would have to be below 26.8 dba to stay within the 31 dba objective. If the link consisted of microwave only, the full 31 dba objective could be applied.

Channelizing Equipment Noise

The amount of noise generated by the microwave channelizing equipment can normally be determined from equipment specifications furnished by the manufacturer. For example, the specified maximum noise level of the noisiest channel of Lenkurt Type 45BX channelizing equipment is from 20 to 23 dba. Since channelizing equipment noise occurs only in the two terminals, this noise source will be of the same

[1] As used here, dba refers to the interfering effect of noise in a voice channel measured with a Western Electric 2B noise measuring set, F1A weighted, at a 0 db transmission level point.

magnitude regardless of the number or length of microwave sections in the system.

Transmitter and Intermodulation Noise

Transmitter noise is caused by random current variations in the modulator stage of a transmitter and is present whether or not a signal is being applied to the system. The amount of transmitter noise generated in a system varies with the equipment design and number of transmitters in the signal path. In Lenkurt Type 72 systems, it is normally between 14 and 18 dba per transmitter.

Intermodulation noise, resulting from intermodulation distortion, is a variable quantity since its magnitude at any instant is affected by such factors as the number of channels in use and the speech levels being transmitted. It results when high energy speech peaks cause excessive frequency swings of the FM transmitter or cause any of the amplifier stages to be overloaded.

Although transmitter noise and intermodulation noise are generated by completely different processes, their interfering effects are mutually dependent upon the transmission level of the modulating signal. That is, if the modulating signal is strong, the interfering effect of transmitter noise will be less than if the signal is weak. A strong modulating signal, however, tends to cause more intermodulation noise.

To provide optimum noise performance, the manufacturer normally specifies the particular modulating level that will prevent the intermodulation noise from exceeding the transmitter noise for more than 1 per cent of the busy hour. When adjusted in this manner, idle circuit noise will be low and intermodulation noise will be within acceptable limits even during heavy loading conditions.

Receiver Noise

Receiver noise is caused by random current fluctuations in the "front end" of the receiver—primarily in the mixer diode. The interfering effect of noise generated in the receiver depends upon the magnitude of the received r-f signals, the receiver bandwidth, and the ratio of carrier frequency deviation to the highest modulating frequency (deviation ratio).

Receiver noise is of greatest importance during fading. A signal fade results in a corresponding rise in the receiver-originated background noise on the telephone channels. However, fades deep enough to bring receiver noise up to noticeable audibility are few on a properly engineered system.

Also, when repeatered systems are considered, the likelihood of several radio sections falling into a deep fade at any one time is remote. The total fade time should be expected to amount to only minutes per month.

Signal-to-noise ratios as poor as about 30 db are acceptable as the bottom limit for a deep fade (the signal-to-noise ratio referred to here compares a zero dbm test tone in a channel to the noise in a 3 kc bandwidth). This corresponds to per-channel noise of about 52 dba at a zero transmission level point. Such noise value is acceptable with the limitation that it should not be attained more than about 0.1% of the time, or about 40 minutes per month.

This noise limit is set at about the point where dialing circuits may be

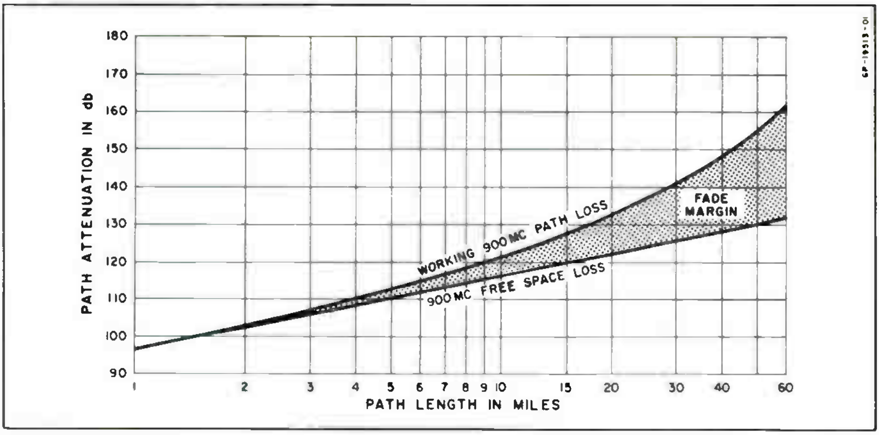


FIG. 2. Attenuation curve for 900-mc path showing fade margin required to maintain thermal noise within acceptable limits for 99.9 per cent of the time.

affected by noise hits, or, where under certain conditions, supervision may be interrupted long enough to cause a disconnect. Thermal noise during normal propagation periods is negligible for all practical purposes. Receiver noise data such as illustrated by Fig. 1 can often be obtained from the equipment manufacturer.

Exact predictions of fades to be expected for a given repeater section are difficult to make. However, it has proven practical to approximate the outside limit of the fade and engineer the system to these requirements. Figure 2 shows the fade margin required at 900 mc to maintain thermal noise within acceptable limits for 99.9% of the time.

Path Distortion Noise

Path distortion noise appears in the system as intermodulation noise and results when reflected waves in the propagation path distort directly transmitted waves. Reflection in the path between antennas may occur from hills, water surfaces or layers of air having different densities. Reflections serious enough to cause path distortion noise may also occur in long antenna transmission feed lines that have sizable voltage standing wave ratios (VSWR is a measure of line reflections resulting from improper terminations of the transmission feed lines).

The effect of this type of noise is minimized when the r-f propagation path is engineered for minimum multipath transmission and when the antenna transmission lines are terminated in their characteristic impedances.

Estimating Noise Performance

With good equipment design and proper system layout, path distortion noise, intermodulation noise, and receiver noise can be reduced to such low values that they will be a very small part of the total noise of a system. Transmitter noise and channelizing equipment noise, however, cannot be reduced beyond a certain point and are important factors in determining whether or not the noise performance of a system will meet design objectives.

If a microwave system consists of only one section, the total average noise interference can be determined from the sum of the channelizing noise power and the transmitter noise power. In multi-section systems, the total transmitter noise interference for the system also can be determined by adding the noise powers generated by the various transmitters. Since it is cumbersome (sometimes impossible) to combine the interfering effects of two different noises by converting each of them to noise powers, adding them and then reconverting them to dba, a table or graph such as shown in Fig. 3 is used to speed the process.

When using this chart, the decibel difference between the two known noise values is located on the horizontal axis. The point on the vertical axis corresponding to this difference is the number of decibels to be added to the larger quantity to get the combined noise.

For example, if a microwave section with 18 dba of transmitter noise is channelized with a Lenkurt Type 45BX System producing 23 dba of noise, the chart can be used to determine that the total noise of the system will be slightly more than 24 dba. This is well within the usual single link objective of 31 dba.

If the link were to consist of a microwave system connected in tandem with a cable carrier system, the same noise computing method could be used. For example, if the cable carrier system produced 29 dba of noise, the total noise for the microwave system should not be more than 26.8 dba if the 31 dba objective is to be met.

Since the total microwave channelizing equipment noise is 23 dba and the transmitter noise is 18 dba per section, it can be determined from Fig. 3 that a microwave system with four sections will have 26.6 dba of noise and a system with five sections will have 27.2 dba of noise. Hence, to be within the 26.8 dba noise limitation for the microwave system, up to four sections could be used. If only microwave were to be used in the link, 15 sections could be connected in tandem before exceeding the noise objective.

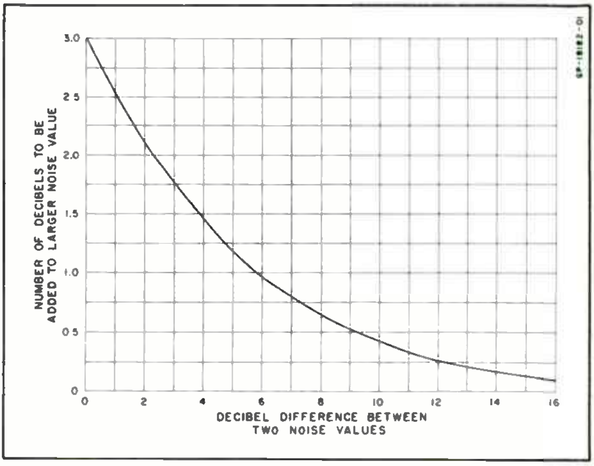


FIG. 3. Graph for determining combined values of noise.

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