CHAPTER XXI

LOADING

129. The Effect of Line Characteristics on Attenuation

In our branch of telephone service, we are ever concerned with the most practicable manner of satisfactorily transmitting voice currents over great distances. Due to the very length of the circuits, undesirable attenuation and distortion effects, which might not be serious in short circuits, become deciding factors in determining whether or not intelligible conversation is possible.

The total attenuation at a given frequency from a talking subscriber's station to a listening subscriber's station (where no telephone repeaters are used) depends upon the length of the circuit, the attenuation per unit length, the energy transfer at the two ends of the circuit or at any junctions of dissimilar sections, and the energy losses due to apparatus that may be associated with the circuit. If a circuit of any given type were long enough, we would naturally expect that the total loss would tend to become so great that the energy reaching the distant end would be insufficient to operate the telephone receiver. This is a transmission limitation from the energy standpoint which is peculiar to long circuits.

Moreover, it should be remembered that while transmission of the required volume of energy is essential, it is not the only consideration. Referring to Equation (87), it may be noted that α varies with the frequency f. In other words, currents of different frequencies may be attenuated unequally as they pass along the circuit. Thus we may easily imagine a long circuit on which a frequency of 500 cycles is transmitted satisfactorily while a frequency of 1500 cycles is not. Under such conditions we would have a distortion effect, and the longer the circuit the more serious this distortion would become. Taking all such factors into consideration we find that there is a very complex relation between the physical characteristics of a telephone line, which determines its efficiency for satisfactory telephone transmission. First of all, the length of the line is an important factor, as the overall attenuation varies directly with length. Second, the actual attenuation constant per unit length at any given frequency is a controlling factor in determining the extent of the loss. Third, the extent to which this attenuation varies with changes in frequency has a direct bearing upon the distortion of the voice currents, or the circuit's "quality". And fourth, apart from the circuit's efficiency as a transmitting medium, we are concerned from the standpoint of power transfer, telephone repeater operation, etc., with its characteristic impedance as given by Equation (82), that is—

$$Z_0 = \sqrt{\frac{z}{y}} = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \qquad (82)$$

One practice that has been helpful in solving the problem of long distance telephone transmission is the application of line loading. By such application we make certain improvements in the circuit's transmission efficiency through one or more of the following effects:

a. A reduction of the circuit's attenuation per unit length.

b. A more even attenuation of the various frequencies within the band of frequencies to be transmitted, thereby reducing distortion.

c. A more nearly constant characteristic impedance for the frequencies within the band to be transmitted, a consideration which is most important in the satisfactory operation of telephone repeaters, but of some importance in considerations having to do with the circuit's termination.

130. Loading as a Means of Reducing Attenuation

The theory of loading is by no means simple and loading results are somewhat difficult to analyze through any physical portrayal. Perhaps the best conception that can be had of loading is a more or less mathematical one that can be gained from studying the effect of line characteristics on attenuation.

The equation given below is the general attenuation constant equation given in the preceding chapter.

$$\alpha = \sqrt{\tfrac{1}{2}\sqrt{(R^2 + \omega^2 L^2)(G^2 + \omega^2 C^2)} + \tfrac{1}{2}(GR - \omega^2 LC)} \qquad (87)$$

Here α is the attenuation constant per unit length and R, G, L, and C are likewise for one unit length. The practice of loading is merely a means of increasing the inductance, or factor L per unit length of circuit, and was first used in the long distance plant to reduce the attenuation. To appreciate fully this particular application let us analyze Equation (87), assuming that we have a non-loaded cable circuit. The distance over which satisfactory transmission is possible with such a circuit is limited in practice to a few miles. If we desire to talk greater distances, the first solution that might suggest itself is the reduction of R in Equation (87). This means we must increase the size of the wires, thereby reducing the I^2R losses in the circuit. But certainly this would be an expensive way of obtaining our objective since it would necessitate the use of more copper per circuit and hence permit fewer circuits per cable.

If we next consider some change in the value of the leakage, G, we find that keeping the leakage to a minimum is advantageous because the lower the value of G, the lower the value of α, even with other line constants remaining unchanged. But with the value of G reduced to zero, or with a condition of perfect insulation, we still have too large a value for α due to the other constants of the line. Let us next consider the capacity per unit length appearing in the foregoing equation. Here we find that a reduction of C means a reduction of α, but on the other hand, a reduction in capacity can be secured only by a wider separation of the wires. This would greatly reduce the number of circuits that might be carried in a single cable and thereby increase the cost per circuit. As in the case of reduced resistance, we might find any tangible results prohibitive on account of this increased cost.

There remains only to consider what effect will be obtained by changing the inductance, L, at the same time not ignoring the influence of R, G, and C upon the attenuation. That is to say, we know that it is good maintenance practice to keep the insulation of our circuits as high as possible, and we know that other things being equal, large gage conductors permit us to talk over greater distances than small gage conductors because of their lower resistance. We know also that non-loaded cable, due to its high capacity, is a relatively poor talking medium. But in the case of both R and C we have already found practical limitations, while in the special case of the value, G, it may be so reduced through proper maintenance that it can be neglected in Equation (87).

With the leakage so low as to be neglected, or with $G = 0$, the attenuation equation becomes:

$$\alpha = \sqrt{\tfrac{1}{2}\omega C\left[\sqrt{R^2+\omega^2L^2} - \omega L\right]} \qquad (102)$$

A study of this expression shows that, within certain limits, an increase in L will result in a reduced α. The improvement that can be obtained by increasing L depends on the value of R. If R is small, but little decrease in α can be effected by increasing L; on the other hand, if R is relatively large, as it necessarily is in practical circuits, a substantial decrease in α can be effected. In other words, by "loading" the circuit with inductance we can reduce α, or expressing the same thing physically, we can reduce the energy loss in the circuit. However, inasmuch as any inductance we may add has resistance, we will by loading increase R, thus to some degree neutralizing our efforts to improve conditions. But with properly designed inductance units, the increase in L more than offsets the increase in R so that the attenuation constant is reduced.

Before proceeding further with our analysis, it will be well to consider the practicability of increasing the circuit inductance so as to obtain this desired reduction in energy loss. It may be remembered that Equations (82) and (87) were developed on the assumption that the circuit properties, resistance, inductance, capacity, and leakage, were uniformly distributed. Theoretically, therefore, in order to increase L we should find it necessary to increase the distributed inductance of the circuit. We could accomplish this by winding each conductor of the circuit with a spiral wrapping of magnetic (iron or permalloy) wire or tape, but the expense involved would be so great that only in special cases could this method be used practically.

In practice a solution is effected by supplying the loading inductance in the form of coils inserted in the circuit at regularly spaced intervals. We learned earlier that we may approximately simulate a circuit of distributed constants with a series of **T**-networks of lumped constants, and similarly the addition of inductance in "lumps" will produce the effect of increasing the distributed inductance, provided that the lumps are sufficiently close together. Thus it is that a loaded circuit usually has load coils, which are nothing more or less than lumps of inductance, inserted at periodic intervals along the circuit, the interval depending on a number of factors, but always being small enough to obtain the effect of increased distributed inductance with its accompanying reduction in attenuation.

131. Loading to Reduce Distortion

Although we have succeeded in finding a means to reduce the energy loss in the circuit, we still have to consider the distortion effects. At least, we must be sure that the distortion effects have not been so exaggerated by loading as to counteract its beneficial effects on the attenuation. As a matter of fact, we shall find that we can employ loading to reduce distortion as well as attenuation. To demonstrate this, we may set up certain simplified equations for the impedance and attenuation of a non-loaded circuit, and then compare them with similar expressions for the same circuit when loaded.

The leakage G, of cable circuits is so small that it may be assumed negligible and due to the very small separation between the wires, the inductance is likewise small enough to be neglected. For non-loaded cable, therefore, where $L = 0$, and $G = 0$, the expressions for impedance and attenuation become—

$$Z_0 = \sqrt{\frac{R}{j\omega C}} \qquad (103)$$

$$\alpha = \sqrt{\tfrac{1}{2}\omega RC} = \sqrt{\frac{\omega RC}{2}} \qquad (104)$$

From these two equations we see that both the attenuation and the impedance vary with frequency, and consequently there will be distortion effects, as mentioned earlier.

Now assume that we load this cable circuit. As before, G may be neglected but by increasing L we have made the reactance ωL very large, so large in fact that we may now consider R insignificant as compared with ωL. If we develop the impedance and attenuation equations on this basis, we will obtain—

$$Z_0 = \sqrt{\frac{L}{C}} \text{ (approximately)} \qquad (105)$$

$$\alpha = \sqrt{\frac{R^2 C}{4L}} \text{ (approximately)} \qquad (106)$$

Here we see that Z_0, since it contains neither inductive nor capacity reactance, has no angle, or what amounts to the same thing, the impedance is perfectly constant and independent of the frequency. Likewise α, the attenuation constant, is independent of frequency. Thus by loading the circuit we have not only reduced the energy loss but have further improved conditions by eliminating distortion effects through the frequency range in which we are interested.

The above expresses mathematically the results of loading. It will be more difficult to obtain a physical picture of these results, but let us first consider the effect of the increased inductance on the characteristic impedance. The load coils connected in series with the line wires naturally increase the impedance. At the same time they neutralize the effect of the capacity inherent to the circuit, but not so simply as in the case of a single inductance in series with a condenser. Nevertheless, we may consider the loaded circuit as made up of a number of sections consisting of series inductance and capacity, each section acted upon by the voltages set up across the capacity of the preceding section. This results in an increased voltage and a decreased current, as well as practically eliminating the phase angle of the characteristic impedance. Now due to the increase in the impedance of the circuit, or due to the power transmitted being in the form of higher voltage and less current, the I^2R losses are less and the total energy loss must inevitably be less. This is a means of higher voltaged energy transmission for the reduction of power losses, and is applicable to high-frequency transmission where the employment of step-up and step-down transformers, as described in Article 113, is not practicable. Viewed from an energy standpoint, it is obvious that if a loaded circuit receives the same amount of energy as a non-loaded circuit, the energy transmitted over the loaded circuit will be greater because its losses are less. Actually, the losses are so much reduced that notwithstanding the fact that the entering current may be less (due to $I = E/Z_0$), the received current on a loaded circuit is greater than that on a non-loaded circuit of the same length, providing that this length is of appreciable magnitude.

The next important aspect of loading which we might discuss is the spacing of the coils. It was stated above that these coils must be close enough together so that the effect of distributed inductance will be obtained. For a proper appreciation of this condition let us simulate a uniformly loaded line with a multisection uniform network. Now it must be remembered that the circuit properties, inductance and capacity, produce reactions which vary with frequency; capacity reactance decreases and inductive reactance increases as the frequency becomes greater While it is equally true that this same frequency effect takes place on the simulating network, the lumping of the inductance and capacity exaggerates the effect.

On the uniform line, current passes continuously from the positive voltaged wire to the negative voltaged wire through the small capacity elements distributed along the circuit, whereas in the network this passage of current can take place only at the middle of the T-sections. Thus, the inductive effect in the network will be somewhat greater than that in the uniform line due to the larger currents traversing the inductances, while the capacity effect will be correspondingly less. Now, of course, the greater the number of sections in the network, the less will be the difference in behavior of the uniform line and the network. But since any

Development of the Loading Coil

difference is increased with higher frequency, any network analysis of a uniform line really resolves itself into a determination of how many sections shall be taken in order that the simulation of the line by the multisection network may be satisfactory over a given frequency range. Similarly, on loaded circuits, the coil spacing (which in effect merely defines the length of each section of the multisection network), is so chosen that the change produced by frequency, over a given range, does not differ greatly from the effect that would be produced on a continuously loaded circuit of about the same average constants, R, L, G, and C. It should be noted, however, that either the value of the loading inductance will be a factor in determining the spacing since its inductance will be a part of the simulating network, or any chosen standard spacing will be a factor in determining the inductance values the coils may have.

Phantom Loading Coils for Toll Entrance Cable Assembled in Shielded Case to Prevent Carrier Crosstalk

132. The Cut-Off or Critical Frequency

We have learned that the band of frequencies between about 200 and 2700 cycles will transmit telephone conversations without any considerable distortion, and loaded circuits accordingly are designed with a view to transmitting at least this band. The lump loaded circuit simulates a smooth circuit, having a correspondingly larger series inductance, very closely over a considerable portion of the normal transmission band, but towards the upper range the simulation becomes less exact. In other words, whereas a uniformly loaded circuit would have an impedance and an attenuation constant which might vary but little with change in frequency, the network, i.e., the loaded circuit, has an impedance and an attenuation constant which generally increase with frequencies near the upper limit of the ordinary voice-frequency band. When necessary, however, it is quite possible to extend this range by changing the design of the loading, and the 16-gage B-22 loaded circuit is a most interesting illustration of what can be accomplished in this way. In any case, the circuit design is such that only the essential frequencies are transmitted, and a marked increase in both impedance and attenuation takes place above the desired transmission band. In fact, the attenuation rises so rapidly that only a few hundred cycles above the upper limit of the band, the amount of current that can be sent through is practically negligible and the circuit is said to "cut-off". This critical frequency at which cut-off occurs is dependent only on the inductance and capacity per loading section and is determined from the expression—

$$f_c = \frac{1000}{\pi\sqrt{L_0 C_0}} \tag{107}$$

where L_0 and C_0 are the inductance and capacity values of the equivalent network sections, and are approximately the actual inductance and capacity values of the loading section. They are expressed in henrys and microfarads, respectively.

133. The Effect of Loading Upon the Wave-Length Constant

Up to this time we have said little about the wave-length constant of a loaded circuit, but knowing from our analysis in the preceding chapter that increased inductance introduces a retarding effect on voltage and current, we would expect the wave-length, λ, to be decreased. That is, since—

$$\text{Velocity} = W = f\lambda \tag{95}$$

we can see that, with f constant, any reduction in W must be due to a decrease in λ. Now in a loaded circuit where we may assume R and G negligible, the value of β obtained by simplifying Equation (88) is—

$$\beta = \omega\sqrt{LC} \tag{108}$$

and since

$$\lambda = \frac{2\pi}{\beta} \tag{94}$$

we have—

$$W = f\lambda = \frac{2\pi f}{\beta} = \frac{1}{\sqrt{LC}} \tag{109}$$

Here C is in farads and W is in miles or "loads" per second accordingly as L and C are the values for one mile of circuit or one loading section. A casual glance at Equations (107) and (109) suggests that they are related, and such is actually the case.

134. Mechanical Analogy of Loading

Though the physical concept of loading apart from the mathematical analysis of the various formulas may be a difficult one, there is a mechanical analogy which

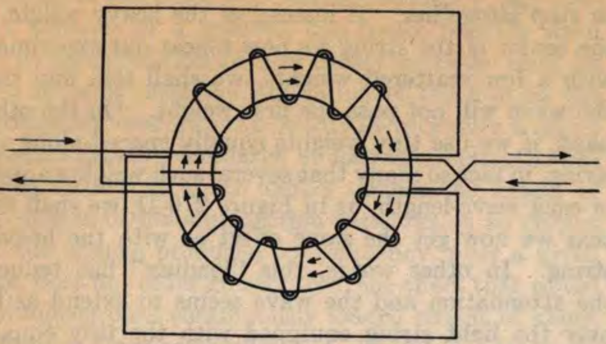


FIG. 271. WINDING OF SINGLE LOADING COIL

several coils are passed over in each wave-length (in practice, about nine coils). If the proper number is not encountered, the losses increase rapidly and "cut-off" follows.

135. Features in Loading Coil Design

In the design and manufacture of the various types of loading coils used on long distance lines, there are a number of requirements other than merely providing a specified inductance value. As stated previously, loading is effective in reducing attenuation only when the increase in the alternating-current resistance of the circuit is held within certain limits. A loading coil should, therefore, have minimum resistance in the winding and minimum losses in its core. Each coil's inductance must be accurately divided into two parts so that one-half of the inductance is inserted in one wire of the talking circuit and the other half is inserted in the other wire, thereby maintaining circuit balance. This requirement is a very exacting one and unless the two windings of each coil are identical in every respect, crosstalk or noise will result. Figure 271 illustrates the method of winding coils to give a high degree of balance.

Loading coil cores consist of toroids of permalloy or molybdenum-permalloy, the latter being standard for practically all new loading. In the manufacture of both types of cores, the magnetic material is first powdered, then mixed with shellac to insulate each particle, and finally pressed into solid rings. This process gives the coils a high degree of magnetic stability. Because of the remarkable magnetic properties of the permalloy cores, modern loading coils are very small. For field installation, they are usually installed in welded steel pots which are standardized in several sizes to hold from a few to a large number of coils. Where only a few circuits are to be lightly loaded, as in program circuit loading, however, the small coils may be placed in the sleeve of the loading splice.

Coils for phantom loading usually have lower inductance values than side-circuit coils but must, of course, have four windings. Figure 272-A illustrates the windings of a phantom loading coil and Figure 272-B shows the connections of a single loading point in a phantom group, where both the side circuits and the phantom are loaded.

136. Important Considerations in Loading Practices

There is a great deal to be said about the proper use, installation, and maintenance of loading coils in the plant which cannot be covered here but will be found in standard instructions. A few important considerations, however, are fundamental and should be remembered.

a. In connecting a loading coil, care must be exercised to prevent a reversal of one winding, thereby neutralizing the coil's inductance.
b. The inductance values of loading coils should be kept within 2% where the circuit is used in connection with telephone repeaters.
c. The loading coil spacing should be accurate to within 2% where the circuit is used in connection with telephone repeaters.
d. To prevent loading coil magnetization, the line current used for telegraph operation should not exceed the specified limits for the particular type of loading.

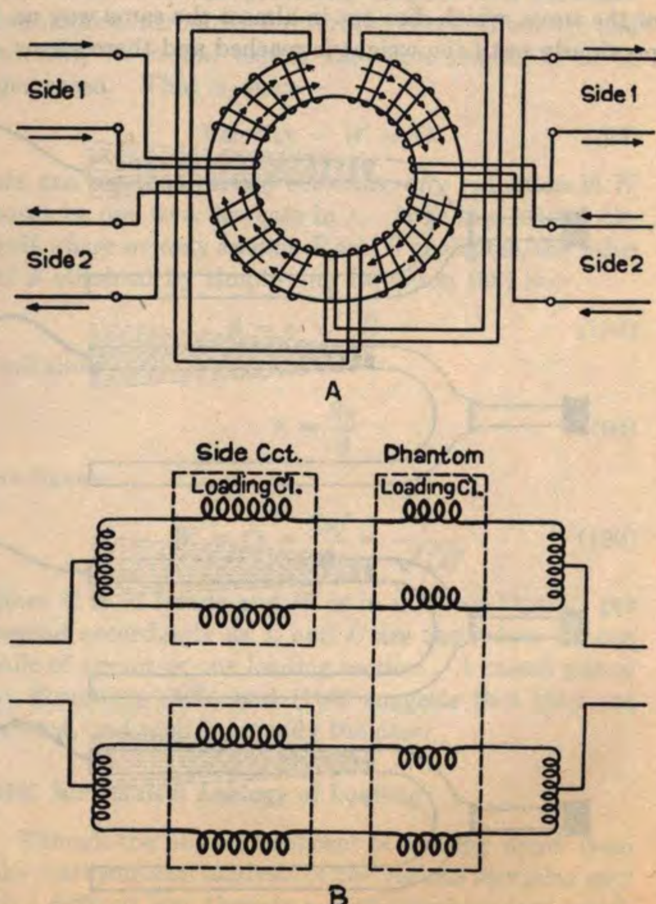


FIG. 272. PHANTOM LOADING COIL CONNECTIONS

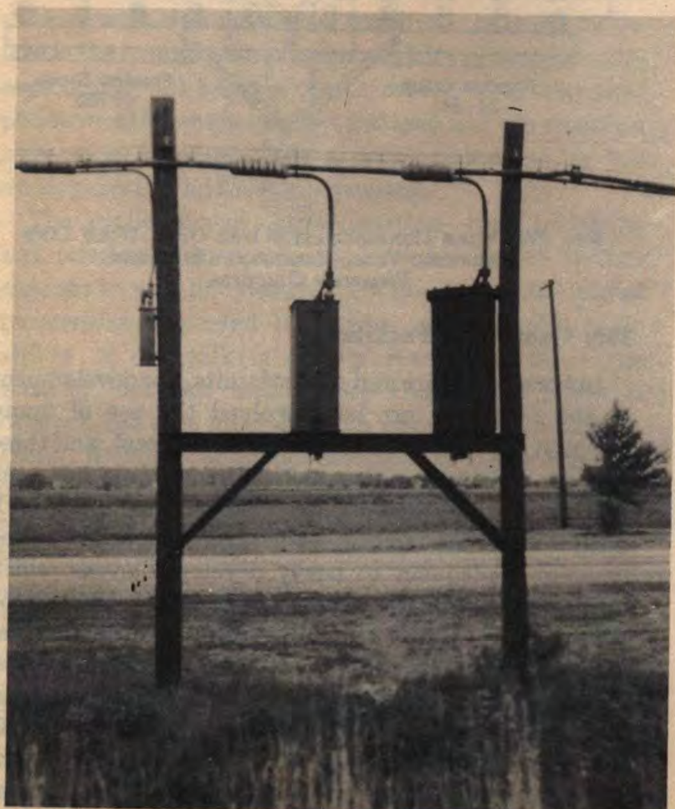
e. It is practicable to load the side circuits of a phantom group without loading the phantom or to load the phantom without loading the side circuits, but one side circuit cannot be loaded without the other.

f. The sending or receiving end impedance of a loaded circuit depends upon the termination of the circuit, i.e., whether terminated at a half-section point, .2 section point, mid-coil point, etc., and will not ordinarily be the same as the characteristic impedance. (This is covered more thoroughly in Chapter XXVIII.)

137. Building-out Short Loading Sections

On account of the actual conditions encountered in the field, it is not always feasible to effect uniform spacing of loading coils. Due to line changes, intermediate submarine cable, loops into stations, etc., loading sections may sometimes be too long or too short. This is usually corrected by "building-out" each short section by adding capacity at a convenient point in the section. In the case of a long section, it is necessary to create an additional loading point, and add capacity in the remaining short section.

The capacity value to be added is not exactly that obtained by multiplying the length of circuit by which



Aerial Cable Loading Showing New and Old Style Pots—Program Loading in Small Case on Pole at Left

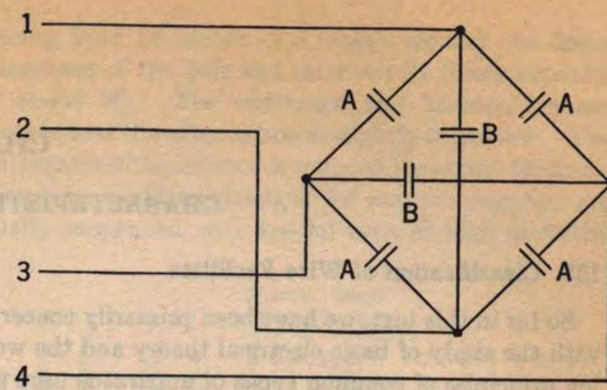


Fig. 273. Phantom Group Building-out Condenser

the section is short by the capacity per mile, but is given by the formula—

$$C_b = C_0 - lc \tag{110}$$

where C_b is the building-out capacity, l is the length of the short section, c is the capacity per mile, and C_0 is the average equivalent capacity per loading section or, in effect, the "lumped" capacity that would simulate the distributed capacity of a section.

One method of building-out is to use six condensers connected as shown in Figure 273. Their values can be calculated from the equations—

$$A = \tfrac{1}{4} C_{bp} \tag{111}$$

and

$$B = C_{bs} - A \tag{112}$$

where C_{bs} is the building-out capacity of the side circuit and C_{bp} is the building-out capacity of the phantom.

Because of difficulties in securing the precise values required for such building-out condensers, and certain maintenance problems, however, this method is generally less satisfactory than the use of "stub cables". These are short sections of cable, usually manufactured in such a manner as to have abnormally high capacity, which may be bridged across or connected in series with the cable section that needs to be built out. The exact capacity needed, as determined from Equation (110), is obtained by cutting the stub cable to the proper length. When the stub cable is bridged on the main cable, several pairs may be connected in parallel to give the required capacity, thus reducing the length of stub that must be used. In certain cases, where great precision of building-out is demanded, it is desirable to build-out the resistance of the short section as well as its capacity, and in such cases the stub cable may be connected in series with the main cable instead of being bridged. In either case, the conductors of the stub cable are carefully balanced against crosstalk in the same way as the conductors of the main cable.

CHAPTER XXII

CHARACTERISTICS OF CIRCUIT FACILITIES

138. Classification of Wire Facilities

So far in this text, we have been primarily concerned with the study of basic electrical theory and the working principles of common types of apparatus used generally in telephone and telegraph work. Beginning with this chapter, our attention will be directed more to the application of some of the principles that we have been considering to specific problems of long distance telephone transmission.

In this connection, we will naturally be interested at the beginning in the electrical characteristics of the various types of wire facilities used for carrying the telephone messages. Such facilities may be classified in several ways according to their uses, or on the basis of their physical or electrical peculiarities.

It is customary first to make a general distinction between facilities used for toll (long distance) and for exchange area transmission. The latter facilities include the greatest part of the total telephone plant since local or short haul service is naturally used much more frequently than long distance service. Accordingly, it is economical to design these facilities primarily on the basis of providing satisfactory transmission within the exchange area. For toll or long distance connections, of which local facilities necessarily form a part in every case, more costly types of facilities are used for the long distance links in order that the transmission shall remain satisfactory. This arrangement is in the interest of overall economy because the long distance facilities are relatively few as compared with the local facilities. It means in general that the latter facilities do not have to meet as exacting requirements as do the toll facilities with respect to attenuation per unit length, impedance regularity, or balance against noise and crosstalk. In exchange area cables, for example, wire conductors as fine as 22, 24, or 26-gage are widely used, whereas the minimum gage in long toll cables is 19. Moreover, it is not necessary to use the quadded construction employed in toll cables because the shorter distances involved in exchange area transmission make crosstalk problems relatively unimportant. Generally similar distinctions as between local and toll transmission apply in the case of open wire facilities. However, it may be noted that there is a certain middle ground where exchange area trunks are of such great length in some cases that their transmission requirements are not widely different from those of the shorter toll circuits. Loading is frequently applied to such trunks and in extreme cases it may even be necessary to use telephone repeaters.

The usual principal classification of toll or long distance facilities is as open wire or cable plant, although several other classifications are possible. Thus, such facilities might be classified according to the way in which they are used, as between those transmitting at voice frequencies and those transmitting at carrier frequencies. In the following articles, the characteristics of these facilities are considered under the three headings of "open wire", "toll cable" and "toll entrance cable".

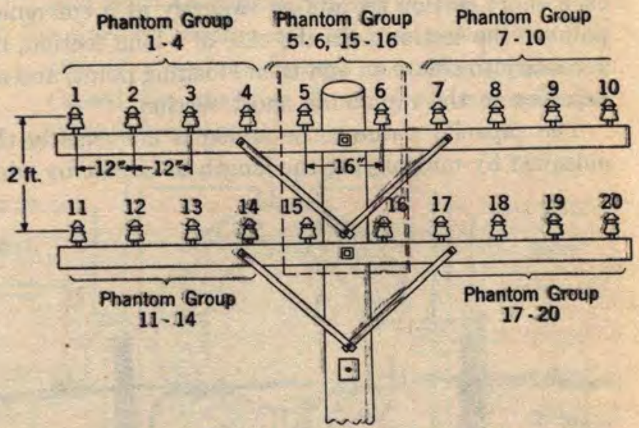


FIG. 274. WIRE CONFIGURATION FOR OPEN WIRE LINE CARRYING VOICE-FREQUENCY SIDE AND PHANTOM CIRCUITS

139. Open Wire Facilities

In both open wire and cable circuits, the development of the telephone art has involved the use of many different types of circuit facilities in the past, and these changes continue as new methods come into use. At any given time, accordingly, the working plant will include facilities ranging from types which are on the verge of inadequacy to newly developed types which are barely out of the experimental stage. Before the development of the telephone repeater (see Chapter XXVI), the great majority of long distance facilities were open wire and, in order to keep the attenuation down, practically all of this open wire was loaded with relatively high inductance coils spaced at intervals of about 8 miles. The conductors used were almost entirely 165, 128, or 104-gage and each group of four wires was arranged to carry a phantom circuit.

The wires were carried on crossarms in the manner indicated in Figure 274. Here each crossarm carries 10 wires which are numbered consecutively starting with the left-hand pin of the top crossarm when looking in the direction of the pole numbering of the line. The standard wire layout on two crossarms, shown in the figure, provides ten "side" and five phantom circuits. Phantoms are derived from wires 1–4, 7–10, 11–14, 17–20, and 5–6, 15–16. The last is called a "vertical" or "pole-pair" phantom and has somewhat different characteristics from the other phantoms because of the different spacing and configuration of the wires. Similarly the characteristics of the "non-pole-pair" side circuits such as 1–2 or 9–10, with 12 inch spacing between wires, are slightly different from those of the pole-pair circuits like 15–16, where the distance between wires is 18 inches. Many open wire lines, with an arrangement of wires on poles as shown in this figure, are still in use in the long distance plant.

Loading, however, is no longer used on open wire facilities. This is a result of the fact that the characteristics of open wire circuits—particularly the leakage and capacity—change markedly with varying weather conditions. In dry weather, open wire loading is effective in reducing the attenuation of the circuits considerably. But due principally to the increased leakage, loading may actually increase the attenuation of open wire circuits in wet weather. In order to increase the overall transmission stability of such circuits, accordingly, all loading was removed after the telephone repeater came into general use, and the resulting increase in attenuation was compensated for by the employment of additional repeaters.

The application of carrier systems—both telephone and telegraph—to open wire lines has led to further changes in pole line design. On account of the higher frequencies employed in carrier systems, the probability of crosstalk is greatly increased. Since the greatest crosstalk hazard (see Chapter XXXI) is between the side and phantom circuits of a phantom group, it has been found desirable in many cases to dispense with the phantom circuit altogether. Further reduction in crosstalk possibilities is effected by spacing the two wires of each pair closer together on the crossarm, and increasing the separation between pairs. Thus, Figure 275 shows a wire configuration widely used on lines carrying telegraph or Type-C telephone carrier systems, in which the non-pole-pairs have eight inch spacing between wires and the separation between the nearest wires of adjacent pairs is 16 inches.

This configuration includes a pole-pair phantom group which ordinarily would be used only for voice frequencies or carrier telegraph circuits. The change in spacing from 12 inches to 8 inches reduces the linear inductance of the pair and increases its linear capacity by about 8%. The resistance and leakage are not changed and the attenuation is slightly increased. The characteristic impedance is reduced by about 50 ohms. Open wire facilities arranged for carrier operation are usually suspended on a special type of high dielectric

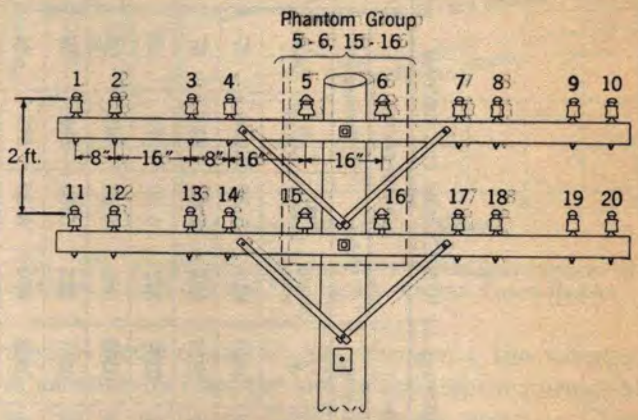


Fig. 275. Wire Configuration for Open Wire Line on which Type-C Carrier Systems are Superimposed

glass insulator coded CS, which gives somewhat more stable transmission characteristics than the ordinary glass (DP) insulator. Steel pins instead of wood for carrying the insulators are also used for the same reason.

Where open wire line facilities are designed to carry broad-band carrier systems (Type-J) 8 inch spacing between wires of a pair is employed, and the pole-pair

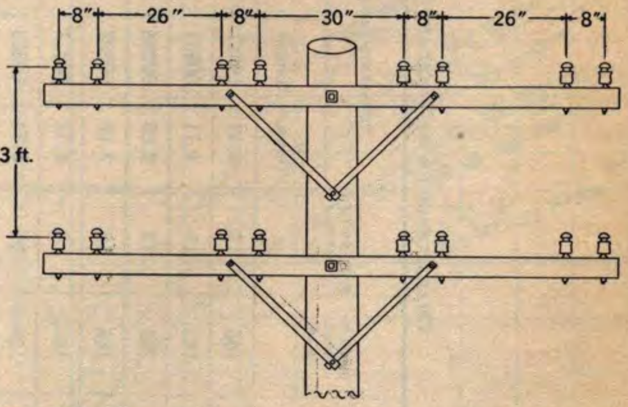


Fig. 276. Wire Configuration for Open Wire Line on which Type-J Carrier Systems are Superimposed

groups are dispensed with entirely. Each crossarm then carries 8 wires, with the spacing and configuration indicated in Figure 276, and no phantom circuits are provided for. In certain cases, also, wires carrying high-frequency systems may be spaced as closely as six inches, where average weather conditions are such that

TABLE XII

Characteristics of Standard Types of Open Wire Telephone Circuits at 1000 Cycles per Second

Type of Circuit	Gage of Wires (Mils)	Spacing of Wires (In.)	Constants per Loop Mile				Propagation Constant				Line Impedance				Wave-Length Miles	Velocity Miles per Second W	Trans-mission Equivalen DB per Mile
							Polar		Rectangular		Polar		Rectangular				
			R Ohms	L Henrys	C Mf.	G M.Mho.	Magni-tude	Angle Degrees +	α	β	Magni-tude	Angle Degrees −	R Ohms	X Ohms −			
Non-Pole Pair Side	165	12	4.11	.00337	.00915	.29	.0352	84.36	.00346	.0350	612	5.35	610	57	179.5	179,500	.030
Pole Pair Side	165	18	4.11	.00364	.00863	.29	.0355	84.75	.00325	.0353	653	5.00	651	57	178.0	178,000	.028
Non-Pole Pair Phan.	165	12	2.06	.00208	.01514	.58	.0355	85.34	.00288	.0354	373	4.30	372	28	177.5	177,500	.025
Pole Pair Phan.	165	18	2.06	.00207	.01563	.58	.0359	85.33	.00293	.0358	366	4.33	365	28	175.5	177,500	.025
Non-Pole Pair Phys.	165	8	4.11	.00311	.00996	.14	.0353	83.99	.00370	.0351	565	5.88	562	58	179.0	179,000	.032
Non-Pole Pair Side	128	12	6.74	.00353	.00871	.29	.0356	81.39	.00533	.0352	650	8.32	643	94	178.5	178,500	.046
Pole Pair Side	128	18	6.74	.00380	.00825	.29	.0358	81.95	.00502	.0355	693	7.72	686	93	177.0	177,000	.044
Non-Pole Pair Phan.	128	12	3.37	.00216	.01454	.58	.0357	82.84	.00445	.0355	401	6.73	398	47	177.0	177,000	.039
Pole Pair Phan.	128	18	3.37	.00215	.01501	.58	.0362	82.82	.00453	.0359	384	6.83	382	46	174.8	174,800	.039
Non-Pole Pair Phys.	128	8	6.74	.00327	.00944	.14	.0358	80.85	.00569	.0353	603	8.97	596	94	178.0	178,000	.049
Non-Pole Pair Side	104	12	10.15	.00366	.00837	.29	.0363	77.93	.00760	.0355	692	11.75	677	141	177.0	177,000	.066
Pole Pair Side	104	18	10.15	.00393	.00797	.29	.0365	78.66	.00718	.0358	730	10.97	717	139	175.5	175,500	.062
Non-Pole Pair Phan.	104	12	5.08	.00223	.01409	.58	.0363	79.84	.00640	.0357	421	9.70	415	71	176.0	176,000	.056
Pole Pair Phan.	104	18	5.08	.00222	.01454	.58	.0368	79.81	.00651	.0362	403	9.83	397	69	173.6	173,600	.056
Non-Pole Pair Phys.	104	8	10.15	.00340	.00905	.14	.0367	77.22	.00811	.0358	644	12.63	629	141	175.5	175,500	.070

Notes: 1. All values are for dry weather conditions.
2. All capacity values assume a line carrying 40 wires.
3. Resistance values are for temperature of 20° C. (68° F.).
4. DP Insulators assumed for all 12″ and 18″ spaced wires—CS Insulators assumed for all 8″ spaced wires.

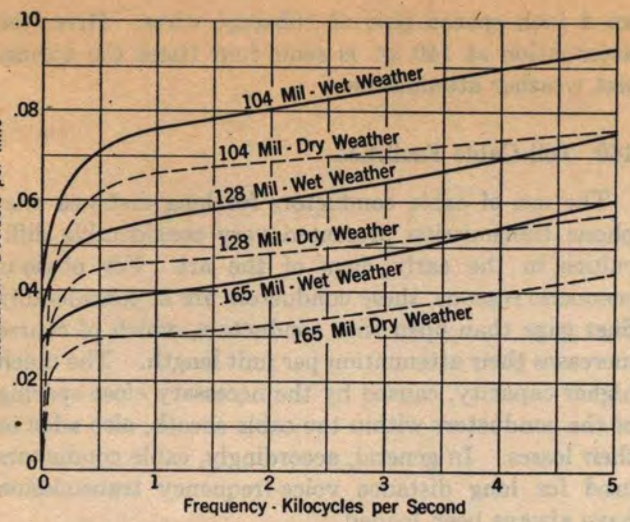


FIG. 277. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE SIDE CIRCUITS OVER THE VOICE RANGE

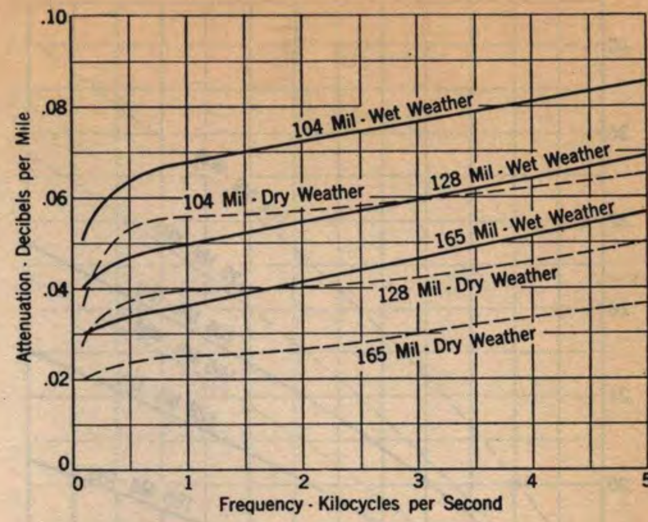


FIG. 278. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHANTOM CIRCUITS OVER THE VOICE RANGE

this close spacing will not result in excessive swinging together of the wires of the pair.

Table XII gives the more important physical and electrical constants of the commonly used types of open wire circuits. The values given are calculated for the single frequency of 1000 cycles and they apply only under more or less ideal conditions. Caution must therefore be used in applying them to practical problems. For example, the leakage of open wire conductors depends upon weather conditions. In wet weather the values for G given in the table may be very considerably increased, and the various constants dependent to a greater or lesser extent on this value, such as attenuation, wave-length, and characteristic impedance, would change accordingly.

The table of course does not give information regarding any variations of the circuit constants through the voice-frequency range. In practically all cases, however, the attenuation, as well as certain of the other circuit constants, changes somewhat with changing frequency. The magnitude of this attenuation change can be determined from curves in which attenuation is plotted against frequency through the working range. Figures 277 and 278 give representative attenuation-frequency curves for 104, 128, and 165 open wire, side and phantom circuits, having the wire spacing and configuration shown in Figure 274, over the frequency range from 0 to 5000 cycles. Separate curves are given for dry and wet weather conditions but the latter curves naturally represent merely an average situation since the "degree of wetness" of the weather is a rather variable quantity. From these curves, it will be noted that, in general, there is an increase of attenuation between 500 and 5000 cycles of somewhere in the order of 50%.

As would be expected, when open wire circuits are used as conductors for carrier systems, the variation in attenuation from the low- to the high-frequency end of the transmission band is much greater. Thus, Figure 279 gives curves for 8 inch spaced, physical circuits, transposed for Type-C carrier and equipped with CS insulators, through the frequency range up to 50,000 cycles. Here, in the band between 5000 and 50,000 cycles, it will be seen that the attenuation more than doubles. Similarly as shown in Figure 280, the

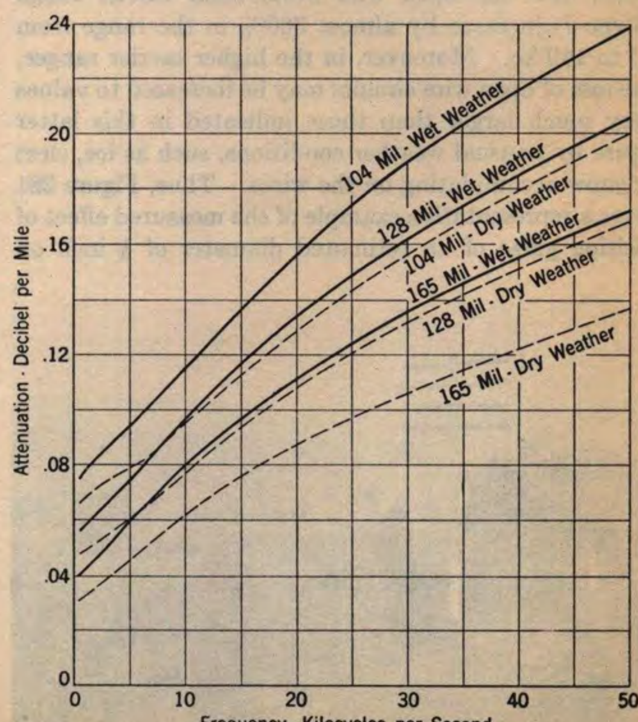


FIG. 279. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHYSICAL CIRCUITS OVER THE TYPE-C CARRIER RANGE

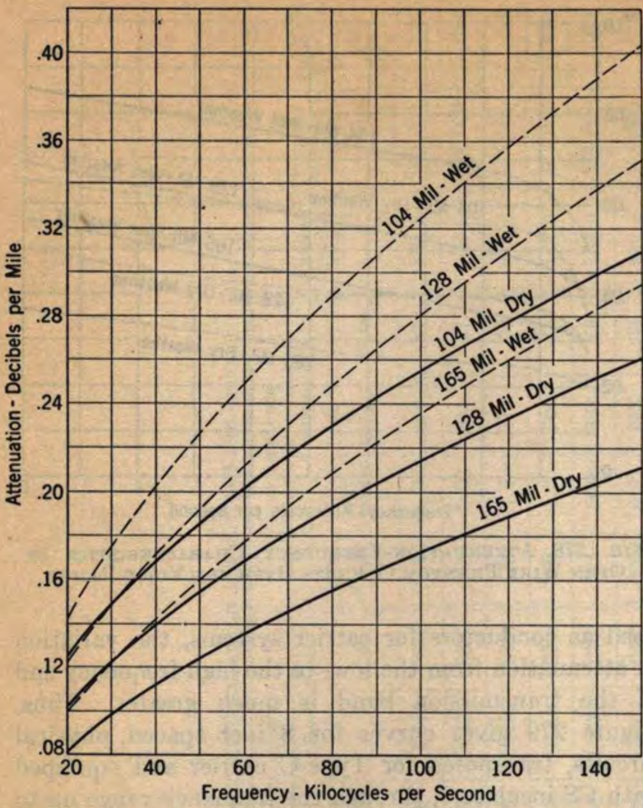


FIG. 280. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHYSICAL CIRCUITS OVER THE TYPE-J CARRIER RANGE

losses over the open wire broad-band carrier range (Type-J) increase by almost 300% in the range from 20 to 140 kc. Moreover, in the higher carrier ranges, the loss of open wire circuits may be increased to values very much larger than those indicated in this latter figure by unusual weather conditions, such as ice, sleet or snow accumulating on the wires. Thus, Figure 281 gives a representative example of the measured effect of melting glaze of an estimated diameter of $\frac{1}{2}$ inch on

TOLL CABLE IN BACKGROUND– 12-INCH SPACED OPEN WIRE LINE IN FOREGROUND

an 8 inch spaced pair of 165-gage wires. Here, the attenuation at 140 kc. is some four times the normal wet weather attenuation.

140. Toll Cable Facilities

The use of cable conductors for long distance telephone transmission presented very considerable difficulties in the early days of the art. For obvious economic reasons, these conductors are of considerably finer gage than open wire conductors, which of course increases their attenuation per unit length. The much higher capacity, caused by the necessary close spacing of the conductors within the cable sheath, also adds to their losses. In general, accordingly, cable conductors used for long distance voice-frequency transmission have always been loaded.

Before the development of the telephone repeater, toll cables were built with the largest gage conductors practicable—10, 13, and 16—and the loading was "heavy". That is to say, loading coils having inductances as high as .245 henry were inserted at intervals of 6000 to 9000 feet. As we noted in our discussion of loading in Chapter XXI, however, such heavy loading, while effective in reducing the attenuation, has two undesirable effects. In the first place, it reduces the velocity of propagation to relatively low values which may seriously interfere with effective transmission

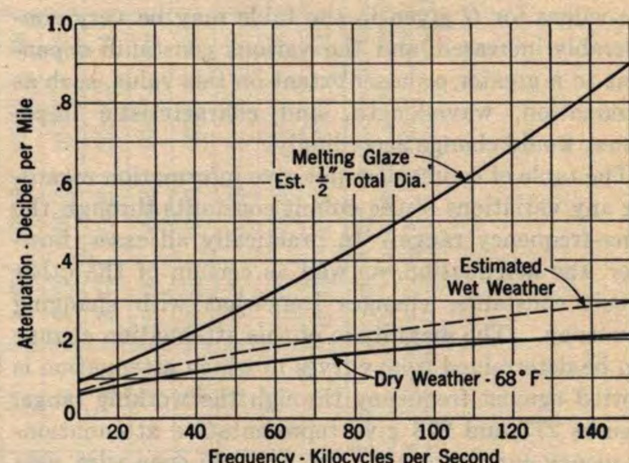


FIG. 281. CURVE SHOWING THE EFFECT OF SLEET DEPOSIT ON ATTENUATION OF OPEN WIRE CIRCUIT

over the longer circuits. Also, such a loaded circuit acts as a low-pass filter with a relatively low cut-off frequency—in the neighborhood of 2500 cycles. Requirements for high quality transmission at the present time demand that cut-off points be much higher than this.

Thus, it is now desirable that the ordinary cable telephone circuit transmit frequencies up to at least

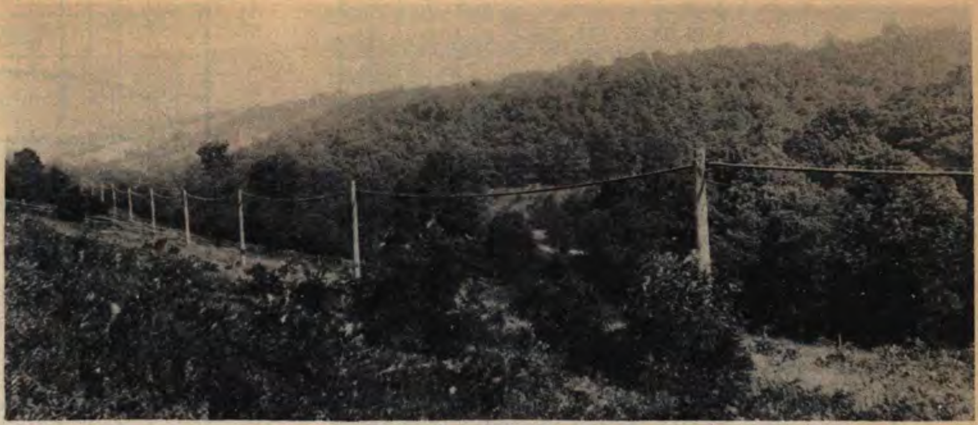
TABLE XIII

Characteristics of Standard Types of Paper Cable Telephone Circuits at 1000 Cycles per Second

Type of Circuit	Wire Gage A.W.G.	Type of Loading	Code No. of Loading Coils	Spacing of Load Coils Miles	Load Coil Constants per Load Section: R Ohms	Load Coil Constants per Load Section: L Henrys	Constants Assumed to be Distributed per Loop Mile: R Ohms	Constants Assumed to be Distributed per Loop Mile: L Henrys	Constants Assumed to be Distributed per Loop Mile: C Mf.	Constants Assumed to be Distributed per Loop Mile: G M. Mho.	Propagation Constant, Polar: Magnitude	Propagation Constant, Polar: Angle Degrees +	Propagation Constant, Rectangular: α	Propagation Constant, Rectangular: β	Line Impedance, Polar: Magnitude	Line Impedance, Polar: Angle Degrees −	Line Impedance, Rectangular: R Ohms	Line Impedance, Rectangular: X Ohms	Wave-Length Miles	Coils per Wave-Length	Velocity Loads per Second W	Velocity Miles per Second W	Cut-off Frequency f_c (Approx.)	Transmission Equivalent DB per Mile (Calculated)
Side	19	N.L.S.	—	—	—	—	85.8	.001	.062	1.5	.1830	46.98	.1249	.1338	470.1	42.80	344.9	319.37	46.93	—	—	46930	—	1.08
"	19	H-31-S	M-4	1.135	2.7	.031	88.2	.028	.062	1.5	.2769	76.58	.0643	.2693	710.0	13.20	691.2	162.17	23.33	20.55	20555	23331	6700	.56
"	19	H-44-S	M-2 & M-3	1.135	4.1	.043	89.4	.039	.062	1.5	.3188	79.87	.0561	.3138	818.0	9.91	805.8	140.80	20.02	17.64	17638	20022	5705	.49
"	19	H-88-S	M-11	1.135	7.3	.088	92.2	.078	.062	1.5	.4408	84.56	.0418	.4388	1131.0	5.22	1126.3	102.83	14.32	12.61	12615	14319	3997	.36
"	19	H-172-S	M-1	1.135	13.0	.170	97.3	.151	.062	1.5	.6095	86.96	.0323	.6085	1564.7	2.82	1562.8	76.90	10.33	9.10	9098	10326	2878	.28
"	19	H-174-S	D-99318	1.135	16.1	.171	100.0	.152	.062	1.5	.6116	86.90	.0331	.6107	1570.0	2.84	1568.0	78.83	10.29	9.59	9586	10288	2870	.29
"	19	H-245-S	M-7	1.135	24.5	.247	107.4	.219	.062	1.5	.7332	87.66	.0300	.7236	1882.0	2.12	1880.7	69.65	8.58	7.56	7556	8577	2389	.26
"	19	B-88-S	M-9	0.568	7.3	.088	98.7	.156	.062	1.5	.6195	87.01	.0322	.6186	1590.2	2.76	1588.3	76.70	10.16	17.88	17882	10157	5655	.28
"	16	N.L.S.	—	—	—	—	42.1	.001	.062	1.5	.1288	49.13	.0842	.0974	330.7	40.65	250.9	215.39	64.51	—	—	64506	—	.73
"	16	H-31-S	M-4	1.135	2.7	.031	44.5	.028	.062	1.5	.2659	82.79	.0334	.2638	682.5	6.99	677.4	83.02	23.82	20.99	20985	23818	6700	.29
"	16	H-44-S	M-2 & M-3	1.135	4.1	.043	45.7	.039	.062	1.5	.3148	84.61	.0296	.3134	808.0	5.17	804.7	72.83	20.05	17.66	17663	20048	5705	.26
"	16	H-88-S	M-11	1.135	7.3	.088	48.5	.078	.062	1.5	.4380	87.64	.0224	.4374	1124.0	2.71	1122.8	53.09	14.36	12.66	12656	14365	3997	.19
"	16	H-172-S	M-1	1.135	13.0	.170	53.6	.151	.062	1.5	.6084	88.27	.0183	.6082	1562.0	1.51	1561.5	41.06	10.33	9.10	9102	10331	2878	.16
"	16	H-174-S	D-99318	1.135	16.1	.171	56.3	.152	.062	1.5	.6105	88.20	.0191	.6102	1567.0	1.58	1566.4	43.11	10.30	9.06	9062	10297	2870	.17
"	16	H-245-S	M-7	1.135	24.5	.247	63.7	.219	.062	1.5	.7325	88.56	.0184	.7323	1880.0	1.22	1879.6	39.87	8.58	7.56	7559	8580	2389	.16
"	16	B-88-S	M-9	0.568	7.3	.088	54.9	.156	.062	1.5	.6184	88.29	.0185	.6181	1587.4	1.49	1586.9	41.35	10.17	17.90	17897	10165	5655	.16
Phantom	19	N.L.P.	—	—	—	—	42.9	.0007	.100	2.4	.1646	47.78	.1106	.1219	262.1	41.97	194.8	175.23	51.53	—	—	51525	—	.96
"	19	H-18-P	M-4	1.135	1.4	.018	44.1	.017	.100	2.4	.2695	78.67	.0529	.2642	428.8	11.11	420.8	82.61	23.78	20.95	20952	23781	6959	.46
"	19	H-25-P	M-2 & M-3	1.135	2.1	.025	44.7	.023	.100	2.4	.3082	81.30	.0466	.3047	490.7	8.48	485.3	72.39	20.62	18.16	18158	20621	5916	.40
"	19	H-50-P	M-11	1.135	3.7	.050	46.2	.045	.100	2.4	.4243	85.25	.0351	.4228	675.2	4.53	673.1	53.33	14.86	13.09	13093	14861	4193	.30
"	19	H-63-P	M-1	1.135	6.1	.063	48.3	.056	.100	2.4	.4724	85.98	.0331	.4712	751.8	3.80	750.1	49.81	13.33	11.75	11748	13334	3738	.29
"	19	H-106-P	D-99318	1.135	8.2	.107	50.1	.095	.100	2.4	.6135	87.49	.0269	.6129	976.4	2.29	975.6	38.98	10.25	9.03	9033	10252	2871	.23
"	19	H-155-P	M-7	1.135	12.5	.155	53.9	.137	.100	2.4	.7361	88.10	.0244	.7357	1171.6	1.68	1171.1	34.40	8.54	7.49	7492	8540	2386	.21
"	19	B-50-P	M-9	0.568	3.7	.050	49.4	.089	.100	2.4	.5939	87.37	.0273	.5933	945.2	2.41	944.4	39.83	10.59	18.64	18645	10590	5936	.24
"	16	N.L.P	—	—	—	—	21.0	.0007	.100	2.4	.1161	50.02	.0746	.0890	184.8	38.98	143.7	116.29	70.60	—	—	70604	—	.65
"	16	H-18-P	M-4	1.135	1.4	.018	22.2	.017	.100	2.4	.2618	84.02	.0273	.2604	416.7	5.76	414.6	41.83	24.13	21.26	21259	24129	6959	.24
"	16	H-25-P	M-2 & M-3	1.135	2.1	.025	22.8	.023	.100	2.4	.3032	85.41	.0243	.3022	482.5	4.37	481.1	36.80	20.79	18.32	18319	20792	5916	.21
"	16	H-50-P	M-11	1.135	3.7	.050	24.3	.045	.100	2.4	.4223	87.43	.0189	.4218	672.1	2.35	671.5	27.52	14.90	13.12	13124	14896	4193	.16
"	16	H-63-P	M-1	1.135	6.1	.063	26.4	.056	.100	2.4	.4709	87.74	.0185	.4705	749.4	2.04	748.9	26.63	13.35	11.77	11766	13354	3738	.16
"	16	H-106-P	D-99318	1.135	8.2	.107	28.2	.095	.100	2.4	.6128	88.54	.0156	.6126	975.2	1.24	975.0	21.13	10.26	9.04	9037	10257	2871	.14
"	16	H-155-P	M-7	1.135	12.5	.155	32.0	.137	.100	2.4	.7357	88.83	.0151	.7355	1170.9	0.95	1170.7	19.52	8.54	7.53	7527	8543	2386	.13
"	16	B-50-P	M-9	0.568	3.7	.050	27.5	.089	.100	2.4	.5931	88.48	.0157	.5929	943.9	1.30	943.7	21.39	10.60	18.66	18657	10597	5936	.14
Physical	16	B-22	616	0.568	1.25	.022	43.1	.040	.062	1.5	.3150	85.02	.0273	.3139	809.1	4.76	806.3	67.10	20.01	35.26	35260	20010	11276	.24

NOTE. The values for cut-off frequency and transmission equivalent per mile, as given in the last two columns, are calculated from the primary constants which are assumed as uniformly distributed. These values accordingly may not be identical with the measured values given in standard formal instructions. The values given in Bell System Practices should therefore be used for engineering work.

3000 cycles without appreciable distortion, and circuits used for program transmission work must handle frequencies much higher than this. There has been a continuing tendency, therefore, to use lighter and lighter loading in cable circuits—that is, to employ lower inductance coils and closer spacing between coils. At the same time, the general application of the telephone repeater has made it possible to use finer gage wire in toll cables, so that practically all conductors in modern cables are of either 16 or 19-gage.



Toll Cable

The computed constants at 1000 cycles of the types of circuits extensively used in toll cables are given in Table XIII. This table is similar in make-up to Table XII and the general comments made in the preceding article regarding the former table apply likewise to this. As Table XIII includes loading constants, however, it should be noted that the secondary constants in this case are computed on the assumption that the loading coil inductance and resistance are added directly to the corresponding basic wire constants and uniformly distributed. Certain of the resulting secondary constants, particularly the characteristic impedance, may therefore be expected to differ somewhat from values computed on the basis of "lumped" loading.

TABLE XIV

Loading Coil Spacing Code

CODE DESIGNATION	LOAD COIL SPACING IN FEET
A	700
B	3000
C	929
D	4500
E	5575
F	2787
H	6000
X	680
Y	2130

The loading designations given in the third column of the table make use of a standardized code. The first letter in the code indicates the spacing between coils as shown in Table XIV. In general, only the H and B spacings are used in toll cables proper. The other spacings listed in the table apply either to exchange area cables or to toll entrance cables, which are discussed in the next article. The number following the first letter in the code gives the inductance in milhenrys and the letter S, P or N following this indicates whether the circuit is side, phantom or physical (non-phantom), respectively. For example, H-25-P indicates a phantom circuit equipped with 25 milhenry coils spaced at intervals of 6000 feet. For convenience this code system is further extended so that phantom group loading may be indicated by a letter followed by two numbers. Thus, for example, B-88-50 indicates a phantom group in which the phantom circuit is loaded with coils of 50 milhenry inductance and the side circuits are loaded with coils of 88 milhenry inductance, both spaced at 3000-foot intervals. In phantom groups, loading is usually applied to the side and phantom circuits at the same point, but this is not always true. Where the spacing is different for the side and phantom loading, two letters are used in the first symbol of the code. Thus, BH-15-15 indicates a loaded phantom group where both side and phantom coils have an inductance of 15 milhenrys, but the side circuit coils are spaced at 3000-foot intervals and the phantom circuit coils at 6000-foot intervals.

In loaded cable circuits, the amount of variation of attenuation with frequency up to frequency values fairly close to the cut-off point is relatively small. This may be seen by referring to the curves of Figure 282. These curves also show how the cut-off frequency takes higher and higher values as lighter loading is employed. The lightest toll cable circuit loading is used for program circuits. As shown in Figure 282, such facilities (16-ga. B-22) are capable of transmitting frequencies up to some 8000 cycles without serious distortion. Where cable circuits are used for high-frequency carrier transmission—up to a maximum of 60 kc. in the case of Type-K carrier—loading is not practicable. The attenuation of the non-loaded conductors is of course very much higher, as shown by the curves of Figure 283, but this is compensated for by the use of closely spaced high-gain amplifiers.

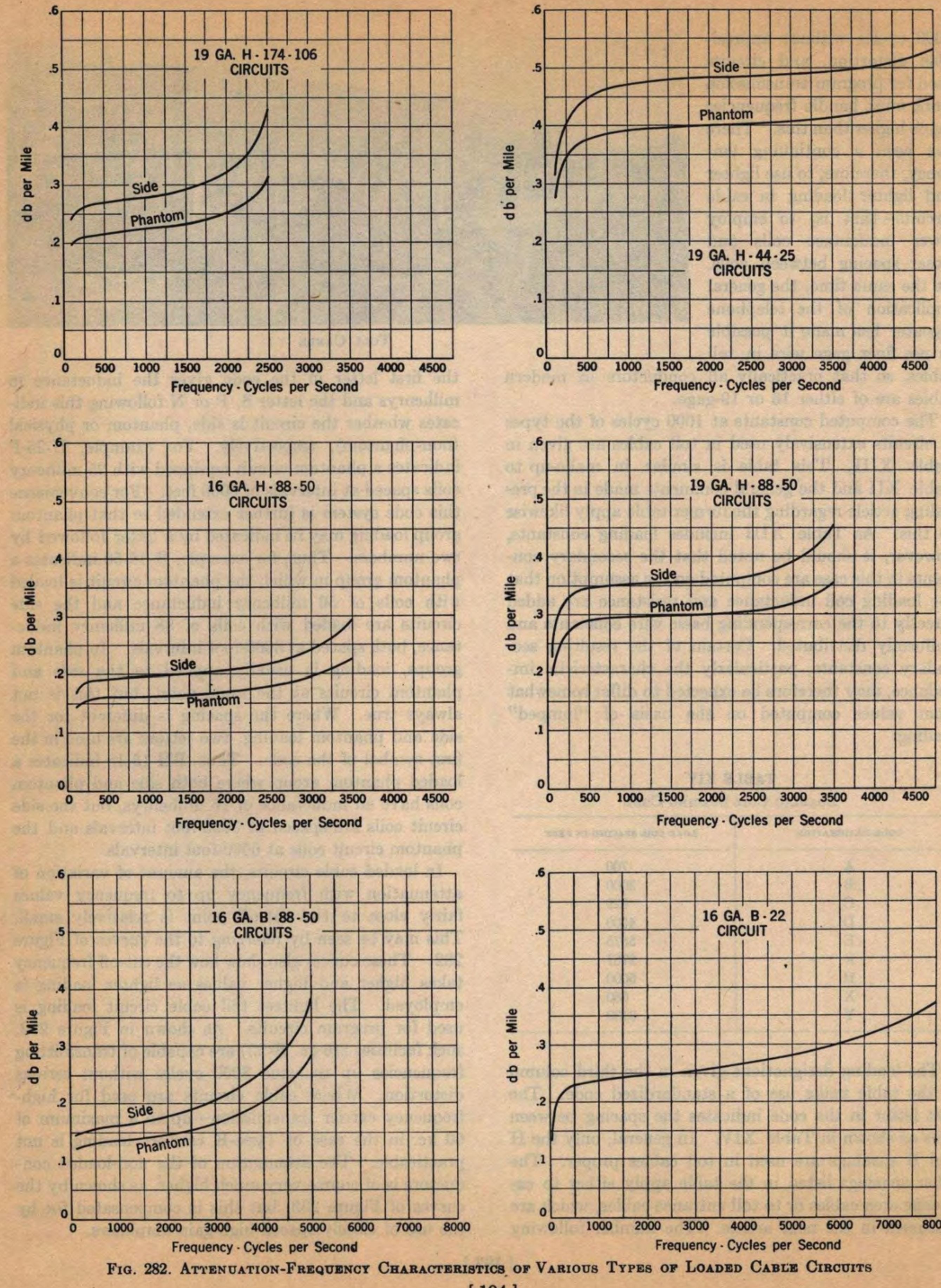


FIG. 282. ATTENUATION-FREQUENCY CHARACTERISTICS OF VARIOUS TYPES OF LOADED CABLE CIRCUITS

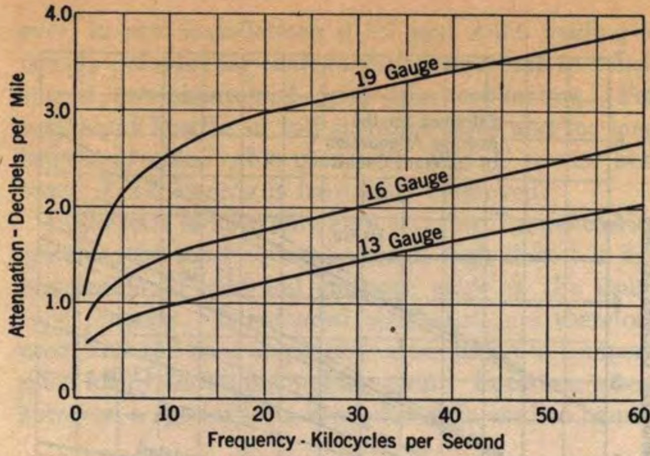


FIG. 283. ATTENUATION-FREQUENCY CHARACTERISTICS OF NON-LOADED CABLE CIRCUITS OVER THE TYPE-K CARRIER RANGE

In addition to the usual open wire and cable circuits that we have been considering, there is another important type of toll line facility which is usually classified under cable. This is the coaxial type of conductor. Here the conducting pair consists of a small central wire and a surrounding concentric tube insulated from the central wire by rubber or fiber discs. The transmission properties of such a conducting structure depend upon the gage of the central wire and the diameter of the tube which surrounds it, the loss per unit length naturally being smaller with larger structures. The chief virtue of the coaxial conductor as a transmission medium lies in the fact that the outer cylindrical conductor acts as a shield against crosstalk or external electrical interference. For this reason such a medium can be used effectively to transmit a frequency band of tremendous width—up to two million or more cycles per second. By employing carrier methods, accordingly, a single coaxial conductor may be used to transmit hundreds of separate telephone channels, or to transmit the very wide band of frequencies comprising a standard television signal.

COAXIAL CABLE

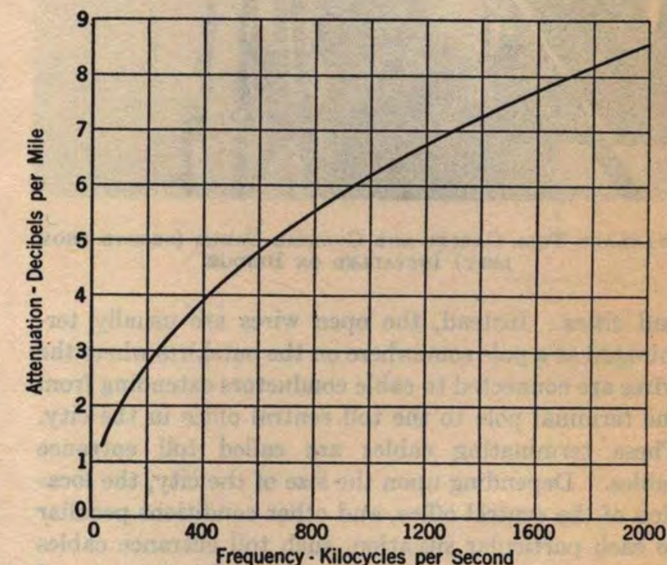


FIG. 284. ATTENUATION-FREQUENCY CHARACTERISTIC OF COAXIAL CIRCUIT

As might be expected the attenuation losses of any practicable conductor of this type are extremely large at the high frequencies with which the conductor is used. This is illustrated in Figure 284 which gives the attenuation per mile of the experimental coaxial cable installed between New York and Philadelphia in 1936. This conductor consists of a 13-gage central wire and a concentric tube slightly larger than $\frac{1}{4}$ inch in inside diameter, with insulation consisting of hard rubber discs spaced at $\frac{3}{4}$ inch intervals. Because of its high attenuation losses, the application of the coaxial type of line facility in practice necessarily involves the use of high-gain amplifiers spaced at very frequent intervals along the line.

141. Toll Entrance Cable Facilities

It is seldom practicable to extend open wire line facilities into the central sections of the larger towns

Standard Toll Cables and Coaxial Cable (second from left) Installed on Bridge

and cities. Instead, the open wires are usually terminated at a pole somewhere on the outskirts where the wires are connected to cable conductors extending from the terminal pole to the toll central office in the city. These terminating cables are called toll entrance cables. Depending upon the size of the city, the location of the central office, and other conditions peculiar to each particular situation, such toll entrance cables may vary in length from a few hundred feet to several miles.

In order to meet the overall transmission requirements of the long distance circuits, it is of course desirable to keep the attenuation of the toll entrance cable conductors to as low a value as practicable. It is even more important in most cases that the cable conductors should be so designed that their impedance matches the impedance of the open wire facilities to which they are connected. Loading of the proper weight is used to obtain both of these results. Toll entrance cables usually contain three gages of wire—namely, 13, 16, and 19, and the larger gages are connected to the larger gages of open wire. That is, 165-gage open wire will be connected to a 13-gage pair in the toll entrance cable while 104-gage will be connected to a 19-gage pair. For voice-frequency open wire circuits, the standard loading for toll entrance cables is H-31-18. Loading may be applied to cable lengths as short as 2000 feet, but short or irregular lengths (end sections) of cable may require artificial building-out in order that the loading section shall be

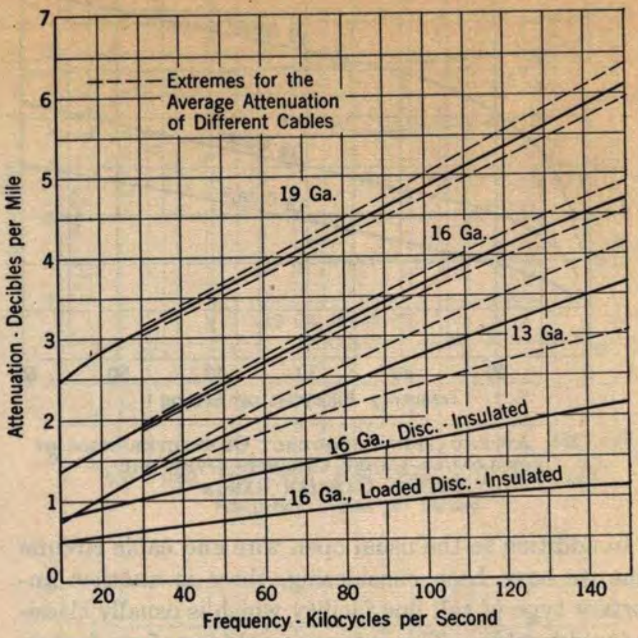
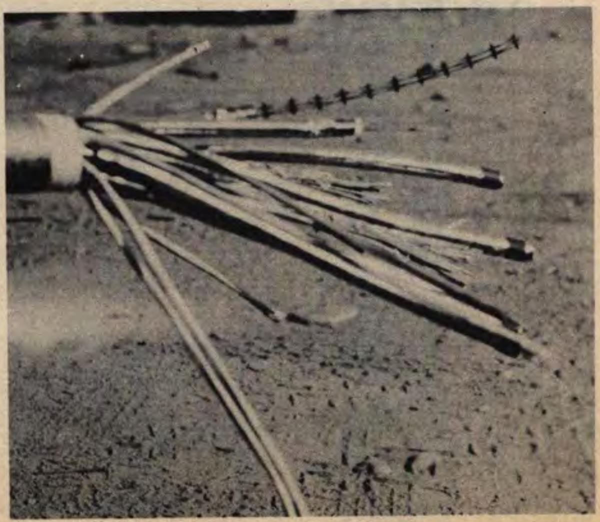


Fig. 285. Attenuation-Frequency Characteristics of Disc-Insulated Toll Entrance Cable Circuits Compared with Non-loaded Paper Insulated Circuits

of the proper length to give the desired characteristic impedance.

Where open wire lines carry Type-C carrier systems, the toll entrance cable loading must be very light in order to transmit frequencies up to 30,000 cycles. For 12 inch spaced open wires, the corresponding entrance cable loading generally used is C-4.1 and C-4.8, the former being used for the larger gage wires and the latter for the smaller. This loading may be modified by means of capacitance and resistance building-out to be satisfactory with 8 inch spaced open wire. How-



Spiral-Four, Disc-Insulated Toll Entrance Cable for Type-J Carrier Systems ever, in new installations A-2.7 and A-3.0 loading is usually provided for cable facilities connected to 8 inch spaced non-phantomed open wire conductors. For very short lengths of toll entrance cable and for long lengths of office cable used with Type-C carrier systems, X-2.5 loading is frequently employed.

In the case of open wire lines carrying Type-J carrier systems, the top frequencies are so high that it is impracticable to load toll entrance cable of the usual type properly. Non-loaded conductors are therefore used, and the resulting higher attenuation is compensated for by additional repeater gain. In certain cases, however, a special type of conductor is used to handle these high frequencies. This consists of a cable made up of individually shielded 16-gage disc-insulated ("spiral-four") quads. Each such quad consists of four wires placed at the corners of a square, the two wires at the diagonals of the square forming a pair and having a separation of .302 inch. The capacity of each pair is about .025 microfarad per mile and the attenuation (non-loaded) is about 2 db per mile at 140 kc. This may be compared with the attenuation of ordinary non-loaded cable pairs at comparable frequencies by referring to Figure 285. These disc-insulated quads may also be loaded to improve still further their attenuation and impedance characteristics.

CHAPTER XXIII

ATTENUATING, EQUALIZING AND TIME-DELAY CORRECTING NETWORKS

142. Pads or Attenuators

In the operation of various telephone circuits, it is frequently necessary artificially to reduce the currents and voltages at various points within the circuits. To accomplish this result, attenuating networks are inserted at the required points. To attenuate all currents of the different frequencies the same amount, the attenuating network obviously must be made up of

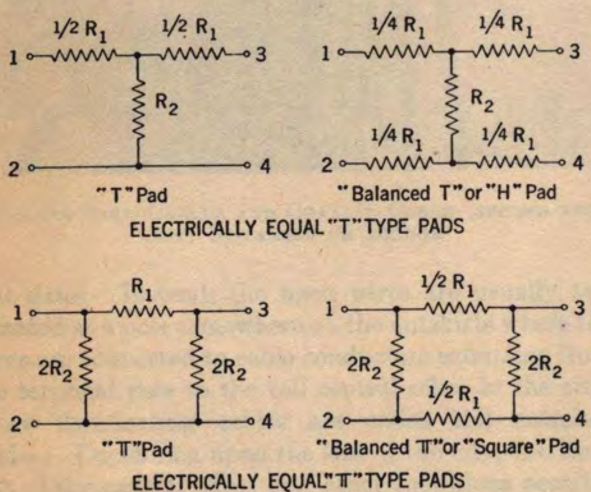
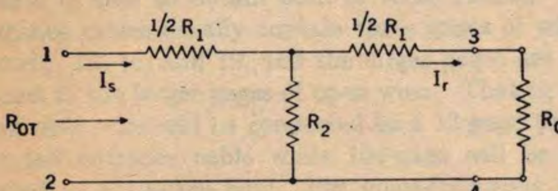


Figure 286

resistances. By arranging appropriate resistances in a network of series and shunt paths, any specified value of attenuation may be obtained without introducing any impedance irregularities in the circuit in which the network is connected. Such resistance networks are usually called **pads** and the most common of these are the **"T"** and **"π"** types illustrated in Figure 286.

These same pads may be made up in **"H"** and **"Square"** networks where the series resistances in



$$R_{0T} = \sqrt{1/4\, R_1^2 + R_1 R_2}$$

$$\frac{I_s}{I_r} = \frac{1/2\, R_1 + R_2 + R_{0T}}{R_2}$$

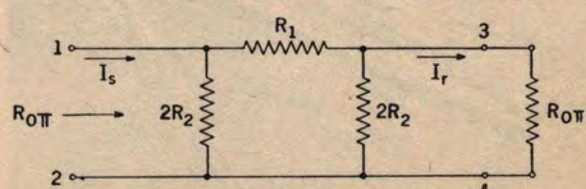
Figure 287

wires 1–3 and 2–4 are equal. This balances the two sides of the circuit without changing the electrical characteristics of the pad. Thus, in Figure 286, the **T-** and **H-pads** are electrically identical; that is, both pads have the same impedance (resistance) and provide exactly the same attenuation. The same applies to the π and **Square** pads illustrated.

In most cases pads are symmetrical; that is, their impedance, as seen from either terminals 1–2 or 3–4, is the same. This is the case for the four pads illustrated in Figure 286. It is possible, however, for a pad to have a different impedance as seen from either terminal. Under these conditions the pad may be used to match two unequal impedances and at the same time produce the desired attenuation.

The two basic facts required for designing a pad are, first, the impedance of the circuit in which the pad is to operate because this must match the impedance of the pad to prevent reflection loss; second, the amount of attenuation the pad is to produce. The characteristic impedance of most circuits in which pads are connected in practice is approximately a pure resistance. It is this resistance the pad is designed to match in order to prevent reflection loss.

When a symmetrical pad is terminated in its characteristic resistance, the resistance it presents at the other terminal is still, of course, its characteristic resistance. In view of this it is a simple matter to solve for the characteristic resistance of the pad in terms of its series and shunt resistances. The loss, or attenuation, of a pad is measured by the relation of the received current, I_r, to the sent current, I_s. This loss expressed in db is $20 \log_{10} \frac{I_s}{I_r}$. The value of this current ratio, $\frac{I_s}{I_r}$, can also be obtained in terms of the series, shunt, and characteristic resistances of the pad. These relations for



$$R_{0\pi} = \frac{R_1 R_2}{\sqrt{1/4\, R_1^2 + R_1 R_2}}$$

$$\frac{I_s}{I_r} = 1 + \frac{R_{0\pi}}{R_2} + \frac{R_1}{2R_2} + \frac{R_1 R_{0\pi}}{4R_2^2}$$

Figure 288

both the characteristic resistance and the current ratio are given for the T and π type pads in Figures 287 and 288, respectively.

If the same values of series (R_1) and shunt (R_2) resistances are used in T- and π-pads, their characteristic resistances and db losses will be different. However, by using proper resistances of different values, two such pads can be made electrically identical. In Article 122 of Chapter XIX, Equations (72), (73) and (74), show how any network can be replaced by a simple **T**-network, after measuring or calculating three specific resistance values. By applying these equations to the π-pad in Figure 289, the equivalent **T**-pad can

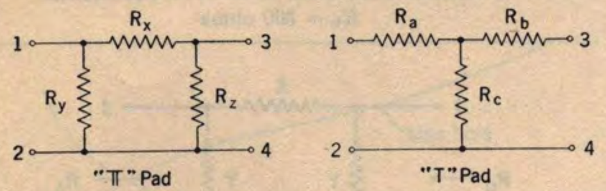


FIGURE 289

be calculated. The resistances R_1, R_2, and R_3 in Chapter XIX are as follows:

R_1 = Resistance across terminals 1–2 with terminals 3–4 open.

R_2 = Resistance across terminals 1–2 with terminals 3–4 shorted.

R_3 = Resistance across terminals 3–4 with terminals 1–2 open.

In terms of R_x, R_y, and R_z in Figure 289, these resistances now become:

$$R_1 = \frac{(R_x + R_z)R_y}{R_x + R_y + R_z} = \frac{R_xR_y + R_yR_z}{R_x + R_y + R_z} \qquad (113)$$

$$R_2 = \frac{R_xR_y}{R_x + R_y} \qquad (114)$$

$$R_3 = \frac{(R_x + R_y)R_z}{R_x + R_y + R_z} = \frac{R_xR_z + R_yR_z}{R_x + R_y + R_z} \qquad (115)$$

By substituting these values in Equations (72), (73) and (74), and solving for the resistances of the **T**-network, we get—

$$R_a = \frac{R_xR_y}{R_x + R_y + R_z} \qquad (116)$$

$$R_b = \frac{R_xR_z}{R_x + R_y + R_z} \qquad (117)$$

$$R_c = \frac{R_yR_z}{R_x + R_y + R_z} \qquad (118)$$

Where the π-pad is symmetrical ($R_y = R_z$), the values of R_a and R_b will, of course, be equal.

To convert a T-pad to a π type, we can make use of three similar equations which can be developed from Equations (116), (117) and (118) above. This involves obtaining values of R_x, R_y, and R_z in terms of R_a, R_b, and R_c. This can be done by first obtaining the sum of the products of Equations (116) and (117), (116) and (118), and (117) and (118) which gives—

$$R_aR_b + R_aR_c + R_bR_c = \frac{R_x^2R_yR_z + R_xR_y^2R_z + R_xR_yR_z^2}{(R_x + R_y + R_z)^2}$$

This becomes—

$$R_aR_b + R_aR_c + R_bR_c = \frac{R_xR_yR_z}{R_x + R_y + R_z}$$

Then by dividing this equation by each of Equations (118), (117), and (116), we get—

$$R_x = \frac{R_aR_b + R_aR_c + R_bR_c}{R_c} \qquad (119)$$

$$R_y = \frac{R_aR_b + R_aR_c + R_bR_c}{R_b} \qquad (120)$$

$$R_z = \frac{R_aR_b + R_aR_c + R_bR_c}{R_a} \qquad (121)$$

When the **T**-pad is symmetrical ($R_a = R_b$), the values of R_y and R_z will be equal.

In designing a pad, it is only necessary to calculate the resistance values for either a **T** or π type and from these values, each of the types illustrated in Figure 286 can be obtained. Probably the simplest method is to first calculate the **T**-pad, and if any of the other types are desired, they can be obtained from the **T**.

Example: Determine the resistance values for a symmetrical 600-ohm, 10 db H-pad.

Solution: From Figure 287, $600 = \sqrt{\frac{R_1^2}{4} + R_1R_2}$

or

$$\frac{R_1^2}{4} + R_1R_2 = 360{,}000$$

$$20 \log_{10} \frac{I_s}{I_r} = 10 \text{ db.}$$

$$\log_{10} \frac{I_s}{I_r} = \frac{10}{20} = .50$$

$$\frac{I_s}{I_r} = 3.16$$

Also from Figure 287

$$\frac{I_s}{I_r} = \frac{\frac{R_1}{2} + R_2 + R_0}{R_2} = 3.16$$

$$\frac{R_1}{2} + R_2 + 600 = 3.16\,R_2$$

$$R_2 = \frac{\frac{R_1}{2} + 600}{2.16} = .231\,R_1 + 277.8$$

Substituting R_2 in the first equation above—

$$\frac{R_1^2}{4} + R_1(.231\,R_1 + 277.8) - 360{,}000 = 0$$

$$.481\,R_1^2 + 277.8\,R_1 - 360{,}000 = 0$$

(This may be solved by the formula for a quadratic equation where $ax^2 + bx + c = 0$, in which

$$x = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}\Big)$$

From this

$$R_1 = \frac{-277.8 + \sqrt{77{,}170 + 692{,}640}}{.962}$$

$$= \frac{-277.8 + 877.3}{.962} = 623.2 \text{ ohms}$$

and

$$R_2 = .231\,R_1 + 277.8$$

$$= 144 + 277.8 = 421.8 \text{ ohms.}$$

The **H**-pad will have a shunt resistance (R_2) of 421.8 ohms, and each of the four series arm resistances $\left(\frac{R_1}{4}\right)$ will be 155.8 ohms.

A common use of pads is as an auxiliary method to control the net gain of 22-type telephone repeaters. As may be seen from Figure 336, these pads are of the square type and, of course, designed to match the 300-ohm impedance seen in both directions at this point in the circuit. Table XV gives the series and shunt resistance values for such 300-ohm Square pads in the range from 2.5 db to 25.0 db. Similar tables can be prepared for other types of pads used for various purposes in the telephone plant.

In making certain tests, it is often desirable to use a variable pad which has a fairly wide range of loss values. Such a pad is called a **variable attenuator** and usually consists of both fixed and variable **H** type units. The variable units are adjusted by dials while the fixed **H** units can be cut in or out of the attenuating circuit by a switching key. This provides a loss that can be varied over the entire range of the attenuator.

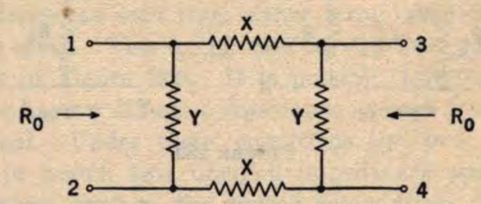
143. Attenuation Equalizers

One of the factors tending to decrease the intelligibility of telephone conversations is unequal attenuation of the currents of different frequencies as they pass over the circuits. For example, the attenuation of a non-loaded open wire circuit is greater for the higher frequencies than for the lower, and this difference in attenuation is directly proportional to the length of line. Therefore, when long circuits are employed it is frequently necessary to make use of **attenuation equalizers** to correct for the unequal attenuation of the line. These equalizers are usually associated with the repeaters which must be included in the circuit to assure a satisfactory volume of sound at the receiving end.

TABLE XV

SERIES (X) AND SHUNT (Y) RESISTANCE VALUES OF BALANCED π OR SQUARE PADS USED IN 22-TYPE REPEATERS

$R_0 = 300$ ohms



Resistance Values for Square Pad

DB LOSS	RESISTANCE VALUES	
	Series, X	Shunt, Y
0	0	Infinite
2.5	44	2099
5.0	91	1071
7.5	146	738
10.0	213	577
12.5	298	487
15.0	408	430
17.5	553	392
20.0	742	367
25.0	1330	336

Attenuation equalizers are networks consisting of retardation coils, condensers, and resistances, which are so proportioned and arranged that their attenuation-frequency characteristics are complementary to the line characteristics that produce the distortion. In brief, the total loss of the line plus that produced by the equalizer will be substantially the same for all frequencies in the transmitted band. This principle is shown in Figure 290.

One of the simplest types of equalizers, shown schematically in Figure 291, is bridged directly across the circuit to be corrected. Obviously, the impedance of such a bridged equalizer must be low enough at certain frequencies to allow sufficient current to flow through it to produce the required losses at these frequencies. Accordingly, the equalizer circuit naturally changes the circuit impedance, particularly at the frequencies where the equalizer is to provide a substantial loss. This introduces an impedance irregularity into the circuit of a sizeable value. The use of bridged equalizers of this type therefore has definite limitations in practice.

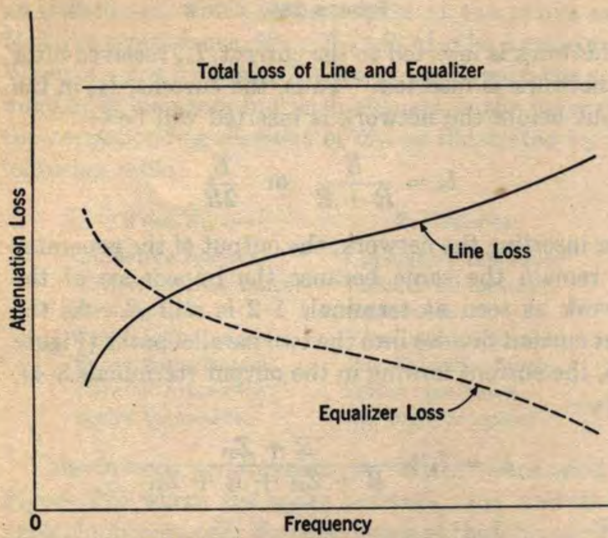


Fig. 290. Principle of Attenuation Equalizer

In long circuits equipped with telephone repeaters, the desired equalizing effects can be obtained without introducing an appreciable impedance irregularity by inserting equalizing networks at the mid-point of the primary sides of the repeater input transformers. Such applications are illustrated in Figures 336 and 337. Instead of changing the net loss of the line, however,

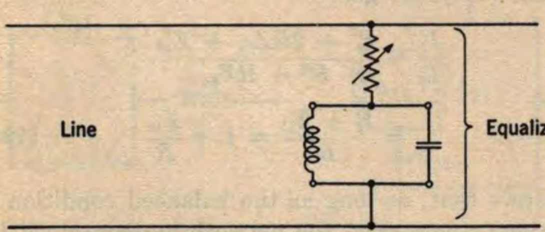


Fig. 291. Simple Bridged Equalizer

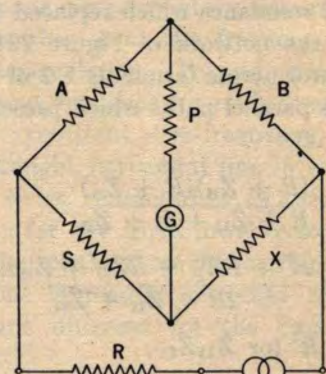
this arrangement changes the overall gain-frequency characteristic of the repeater to match reasonably closely the loss-frequency characteristic of the line. That is, for the frequencies where the line loss is high the repeater gain is also high and vice versa. The overall loss-frequency characteristic of the line and repeater together is then reasonably uniform over the transmitted frequency band.

144. Bridged T-Equalizer

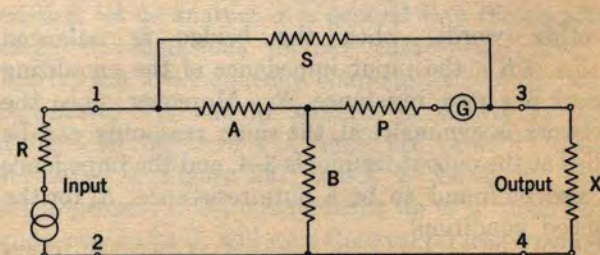
Both of the above methods of equalization give satisfactory results where the amount of attenuation distortion to be corrected is relatively small. To use either of these methods to correct a large attenuation distortion, might result in an impedance irregularity of such a magnitude as to more than offset the benefits obtained by equalizing. To equalize for these relatively large amounts of attenuation distortion, a somewhat more complex equalizing network, in the form of a bridged T-structure, may be used. This equalizer is designed to have a constant impedance over the entire frequency band transmitted.

As its name implies, the bridged T-equalizer is built in the general form of a T-network, but it has an additional impedance path bridged across its series elements. This latter path controls the loss of the equalizer. The elements of the bridged T-equalizer are connected in a Wheatstone bridge arrangement, and the principle of its operation may be best grasped by first referring to the ordinary Wheatstone bridge circuit illustrated in Figure 292-A. Here a generator is connected to the two opposite points of the bridge through the impedance R, and a galvanometer, G, is connected across the other two points through an impedance P. The bridge is balanced and no current flows through the impedance P when the following proportion holds true:

$$\frac{A}{B}=\frac{S}{X} \quad \text{or} \quad SB = XA$$



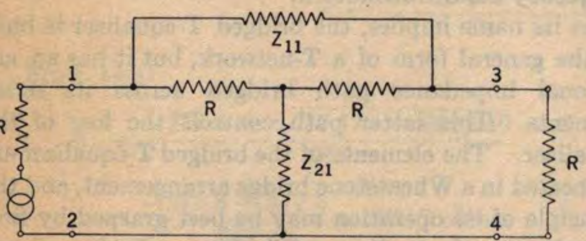
A. WHEATSTONE BRIDGE CIRCUIT



B. WHEATSTONE BRIDGE IN THE FORM OF A "T" NETWORK

Figure 292

Now let us rearrange this bridge circuit in the form of a T-network where the series elements are bridged by the impedance S, as illustrated in Figure 292-B. The T-network proper is formed by A, P, and B, with S as the bridging impedance while R and X now become the input and output impedances, respectively. Next, let us change the impedances R, A, P, and X to resistances of equal value, which may then all be designated as R. For reasons to be explained later, we shall also redesignate the impedances S and B as Z_{11} and Z_{21}, respectively. Then, as illustrated in Figure 293, we still have the same bridge which was



CONSTANT RESISTANCE "T" NETWORK WHEN $Z_{11}\,Z_{21}=R^2$

FIG. 293. BRIDGED T-EQUALIZER

balanced when $SB = XA$, and is now balanced when—

$$Z_{11}Z_{21} = R \times R \quad \text{or} \quad R^2 \tag{122}$$

When the bridge is balanced no current flows in the impedance, P (Figure 292-A), and for purposes of analysis we may therefore simplify the network by removing the R resistance which replaced P in Figure 293, giving us the network of Figure 294. Looking from the generator across terminals 1–2 of this circuit, we now see two parallel paths which present an input impedance Z_{in} of—

$$Z_{in} = \frac{(R + Z_{11})(R + Z_{21})}{R + Z_{11} + R + Z_{21}}$$

$$= \frac{R^2 + RZ_{11} + RZ_{21} + Z_{11}Z_{21}}{2R + Z_{11} + Z_{21}}$$

or substituting R^2 for $Z_{11}Z_{21}$,

$$Z_{in} = \frac{R(2R + Z_{11} + Z_{21})}{2R + Z_{11} + Z_{21}} = R \tag{123}$$

In other words, when the bridge is balanced ($Z_{11}Z_{21} = R^2$), the input impedance of the equalizing network is a pure resistance, R. Moreover, since the T-network is symmetrical, the same reasoning can be applied at the output terminals 3–4, and the impedance will also be found to be a pure resistance, R, for the balanced condition.

As in the case of any other circuit, the loss produced by this network may be determined by the ratio of the current, I_b, received in the output impedance before

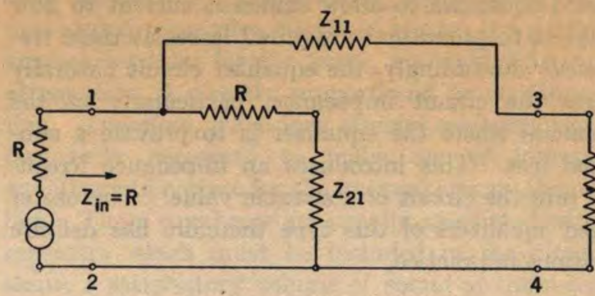


FIGURE 294

the network is inserted to the current, I_a, received after the network is inserted. Thus, the current, I_b, in the output before the network is inserted will be—

$$I_b = \frac{E}{R + R} \quad \text{or} \quad \frac{E}{2R}$$

After inserting the network, the output of the generator will remain the same because the impedance of the network as seen at terminals 1–2 is still R. As the input current divides into the two parallel paths (Figure 294), the current flowing in the output (terminals 3–4), I_a, is—

$$I_a = I_b \times \frac{R + Z_{21}}{R + Z_{21} + R + Z_{11}}$$

Then

$$\frac{I_a}{I_b} = \frac{R + Z_{21}}{2R + Z_{11} + Z_{21}}$$

or

$$\frac{I_b}{I_a} = \frac{2R + Z_{11} + Z_{21}}{R + Z_{21}} \tag{124}$$

Since we are considering the balanced condition where $Z_{11}Z_{21} = R^2$, then $Z_{21} = \dfrac{R^2}{Z_{11}}$. Substituting this in Equation (124), we get—

$$\frac{I_b}{I_a} = \frac{R^2 + 2RZ_{11} + Z_{11}^2}{R^2 + RZ_{11}}$$

$$= \frac{R + Z_{11}}{R} = 1 + \frac{Z_{11}}{R} \tag{125}$$

This shows that, as long as the balanced condition is maintained, the loss of the network is determined by Z_{11}. This is also apparent from an inspection of Figure 294 because this impedance is in series with the receiving circuit, and any value of loss may be secured without affecting the input or output impedances, providing the balanced condition is maintained.

To summarize, **the bridged T-equalizer has a constant impedance, as seen from both terminals, equal to R when Z_{21} is the inverse of Z_{11} ($Z_{21}Z_{11} = R^2$), and its overall loss-frequency characteristic is determined by the bridged series impedance network, Z_{11}.**

Both Z_{11} and Z_{21} represent generalized impedances which may be resistances, capacitances, inductances, or any combination of them. The one and only requirement is that established by the balanced condition ($Z_{11}Z_{21} = R^2$), which means there must always be an inverse relationship between Z_{11} and Z_{21}. If Z_{11} is a pure inductive reactance represented by $j\omega L$, then Z_{21} must be $R^2/j\omega L$ or $-jR^2/\omega L$, which represents a capacitive reactance, $-j/\omega C$, where $C = L/R^2$. On the other hand, if Z_{11} is a capacitance, Z_{21} must be an inductance, which is the reverse of the above case. If Z_{11} is a resistance, then Z_{21} will also be a resistance. When Z_{11} is a network, Z_{21} is a network with the same number of elements but each element is the inverse of the corresponding element of Z_{11} as illustrated by the following table:

When Z_{11} is:	*Z_{21} becomes:*
Inductive reactance.	Capacitive reactance.
Capacitive reactance.	Inductive reactance.
Resistance.	Resistance.
Series inductance.	Parallel capacitance.
Series capacitance.	Parallel inductance.
Parallel resonance.	Series resonance.
Series resonance.	Parallel resonance.

This inverse relationship is further illustrated in Figure 295 where the series network, Z_{11}, and its inverse shunt network, Z_{21}, are shown at the left and right, respectively. Here the advantages of using the two-digit subscript for Z become more evident. The first digit of the subscript indicates whether the element belongs to the series or shunt impedance, while the second digit designates the corresponding inverse elements of the two networks. Therefore, in Figure 295, C_{21} is the inverse of L_{11}; C_{22} is the inverse of L_{12}; L_{21} is the inverse of C_{11}; and R_{21} is the inverse of R_{11}.

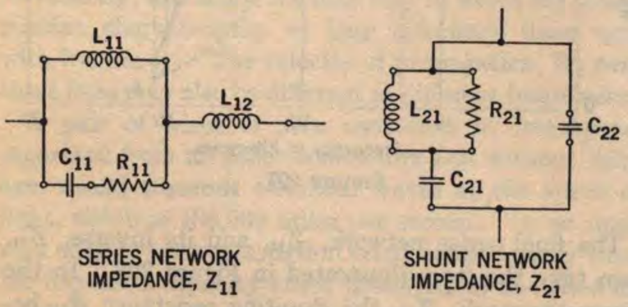


FIG. 295. INVERSE SERIES AND SHUNT NETWORKS

In designing a bridged T-equalizer for a specific use, the attenuation-frequency characteristic of the Z_{11} network must be complementary to the attenuation-frequency characteristic of the circuit to be corrected. This is true because, as we have seen, the loss-frequency characteristic of the bridged T-equalizer is controlled by the series impedance network, Z_{11}.

As an example of the general problem involved in the design of the Z_{11} network, let us consider a representative application of its use in a Type-C carrier system. In this, as in other carrier systems, separate frequency bands are employed for transmission in the two directions. For example, in the Type-CS system, transmission in the East to West direction occupies the frequency range from about 6 to 16 kc., while transmission in the West to East direction is in the range from about 18 to 28 kc. At the terminals and intermediate repeater points, the entire frequency band used in transmitting in each direction, which in the C systems includes three separate voice channels, is amplified by a single amplifier. The frequency bands transmitting in opposite directions are separated by means of so-called "directional filters".

The attenuation of the line facilities varies very considerably over the wide band of frequencies used. The directional filters also introduce appreciable distortion near their cut-off frequencies. In order to maintain uniform transmission, therefore, it is necessary to employ equalizers to counteract both of these attenuation distortion factors. This situation is illustrated in Figure 296. Here the loss produced by the line, filters, and their combined total, are indicated by the heavy lines so designated. (The frequency positions of the three voice channels in each direction of transmission are indicated by the vertical dashed lines.) The required loss-frequency characteristic of the equalizers is shown by the two upper curves A and B, each of which is made complementary (inverse) to the total line and filter loss over the frequency band for its direction of transmission. By adding the losses of the line, filters, and equalizer for each direction of transmission, the resultant loss-frequency characteristic becomes a straight horizontal line in each case. Because of the rising characteristic of the line, however, the total loss for the three lower voice channels, L_1, is less than that of the three higher voice channels, L_2. This difference is readily corrected by making the amplifier gains different for the two directions of transmission.

Now that we have noted the factors which give these equalization curves (A and B) their particular characteristics, let us analyze in a general way the equalizer design considerations for one curve—say curve A. Clearly, the loss-frequency characteristic of the series impedance, Z_{11}, should conform as closely as practicable with the curve A of Figure 296, or with the solid line curve of Figure 297, which is the same. As a first approach, a Z_{11} circuit made up of a single series condenser, as in B, will give the general loss-frequency characteristic indicated by curve *b*. This, of course, is due to the fact that the current through a condenser increases with frequency; consequently, its loss decreases. However, it will be noted that curve *b* diverges widely from the desired characteristic at the higher frequencies. To reduce the loss at f_1, and thus bring the curves closer together, we can add in series with the condenser an inductance, L_{11}, of such a value that series resonance occurs at approximately the frequency f_1. This fails to solve the problem, however, because below the resonant frequency this series circuit produces a loss that increases with decreasing frequency, as indicated by curve *c*. Because of the inductance, the curve has now become too low over most of the frequency range but yet fairly close to the desired value at the two extremities.

Apparently what is needed is an inductance that is considerably smaller than that of L_{11} over most of the frequency range but equal to it at f_1. This can be simulated by a parallel resonant circuit which has a resonant frequency somewhat above f_1, as indicated at D, because up to the anti-resonant frequency the inductive reactance of a parallel resonant circuit increases with frequency. On this basis, L_{11} can be selected so that it is small enough to approximate the desired loss at the lower and mid frequencies. Then by shunting a condenser around it, forming a parallel resonant circuit, the effective inductance of the parallel combination at f_1 can be made equal to that of the former L_{11}. In this way the low impedance, and hence low loss, is preserved at f_1 and the loss is still increased at lower frequencies. The net effect is the characteristic shown by curve *d*. This comes very close to the desired characteristic, but even greater precision can be obtained by adding the shunt resistance R_{11}, as shown by E. This introduces a small increase in the loss over most of the frequency range and modifies the curve as shown by *e*.

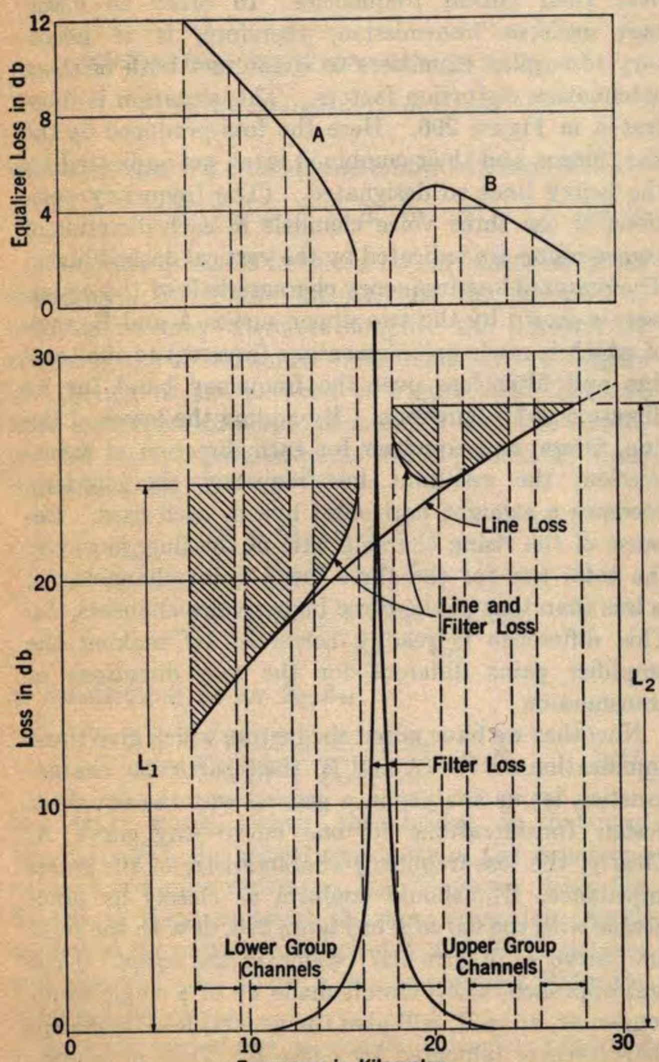


Fig. 296. Attenuation Equalization for Type-C Carrier System

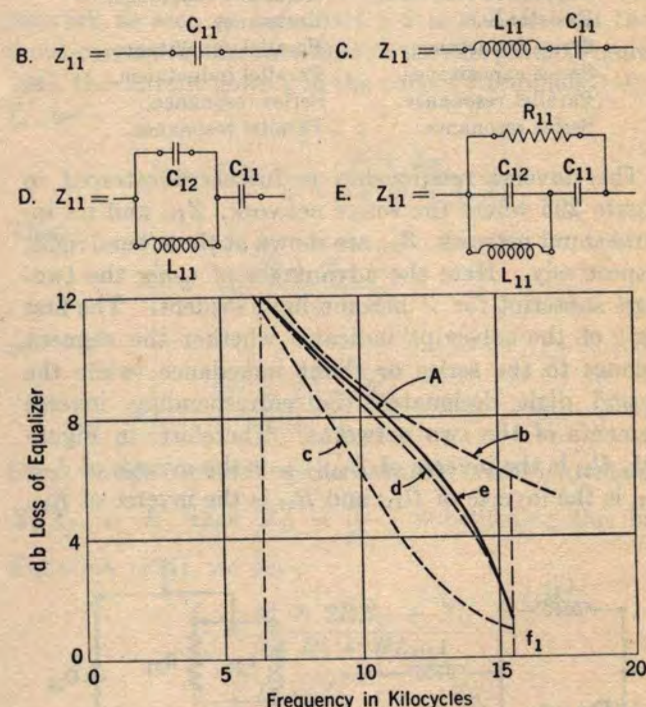


Figure 297

The final series network, Z_{11}, and its inverse, Z_{21}, then take the form illustrated in Figure 298. In the inverse network, Z_{21}, the shunting resistance R_{11} becomes a series resistance R_{21}; the series condenser C_{11} becomes a shunt inductance L_{21}; while the parallel resonant circuit C_{12} and L_{11} becomes a series resonant circuit L_{22} and C_{21}. The degree of perfection with which a given loss-frequency characteristic can be matched by such an equalizer depends upon the number of coils, condensers, or resistances it is considered economical to use.

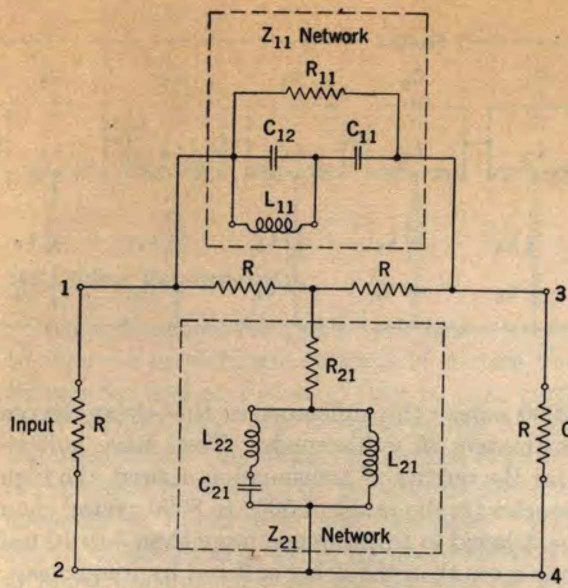


FIGURE 298

Another general application of the bridged T-equalizer is in equalizing lines for program transmission, which use 16-gage B-22 loaded cable circuits. The loss on these facilities varies with frequency as may be noted from Figure 282, and in view of the wide voice-frequency band transmitted, it is apparent that attenuation equalizers must be used to provide a uniform loss-frequency characteristic over the frequency band transmitted. The principles involved are, of course, the same as those we have just been considering although the details of design may be somewhat different.

145. Time-Delay or Phase Equalizers

The equalizers discussed in the preceding article take care of the variation of line attenuation. Unfortunately, this is not the only way in which the transmission characteristics of long telephone lines vary with frequency. The velocity of propagation, W, over these lines may also be different at different frequencies.

A pair of wires of zero resistance in free space, separated from all other conductors and without leakage, would transmit electrical waves at the speed of light, which is 186,000 miles per second. In an open wire circuit, what retardation exists comes largely from the increased capacity effect produced by the glass insulators, and the resistance of the wires. In cable circuits, there is a still further retardation due in part to the greater capacity between the wires, and even more to the inductance of the loading coils which are inserted to decrease the attenuation. In any case there is a finite time interval, or time-delay, between the sending end of any circuit and its receiving end.

This delay is greater with some types of facilities than others and, of course, increases in direct proportion with the length of circuit in every case. Moreover, the delay is not the same for all frequencies in the transmitted band because the line reactance components which determine the wave-length constant, β, and thereby the velocity of propagation, W, vary with frequency. This difference in velocity at the various frequencies means that the phase relationships of the currents at the receiving end are not the same as at the sending end. In long circuits of certain types, this produces a form of distortion sometimes called "phase distortion". This phase or time-delay distortion may be equalized by inserting in series with the circuit a network having the inverse time-delay=frequency characteristic. The total time-delay produced by the circuit, added to that of the time-delay equalizer, will then be approximately the same over the frequency band transmitted. These **time-delay equalizers** are sometimes referred to as "phase correctors".

The time-delay (phase shift) in a circuit is determined by the electrical characteristics of the circuit, and varies in proportion to the circuit's length. When all the frequencies in the band transmitted are delayed the same amount, there is obviously no delay distortion because

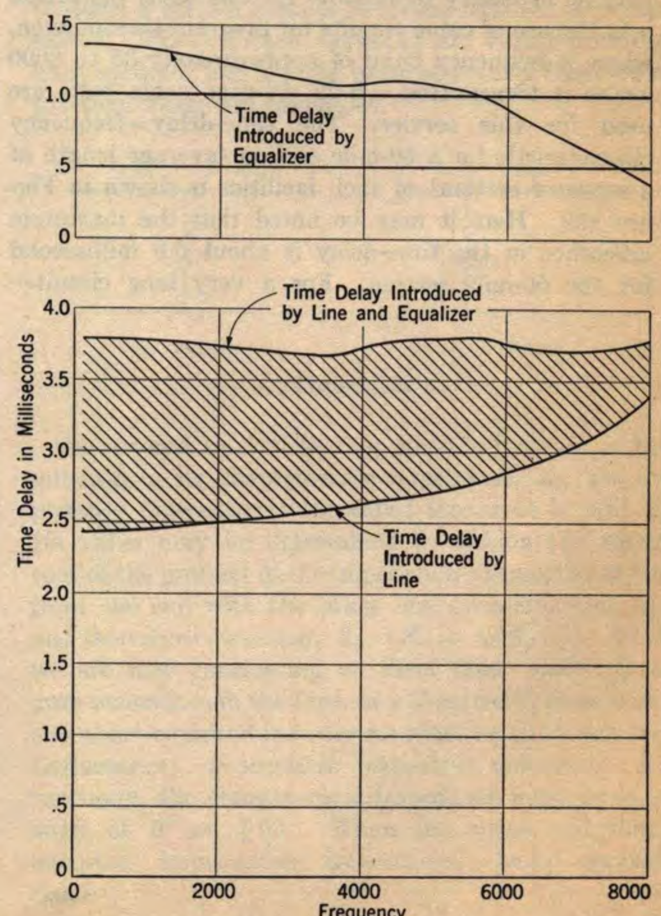


FIG. 299. PRINCIPLE OF TIME-DELAY OR PHASE EQUALIZING NETWORK

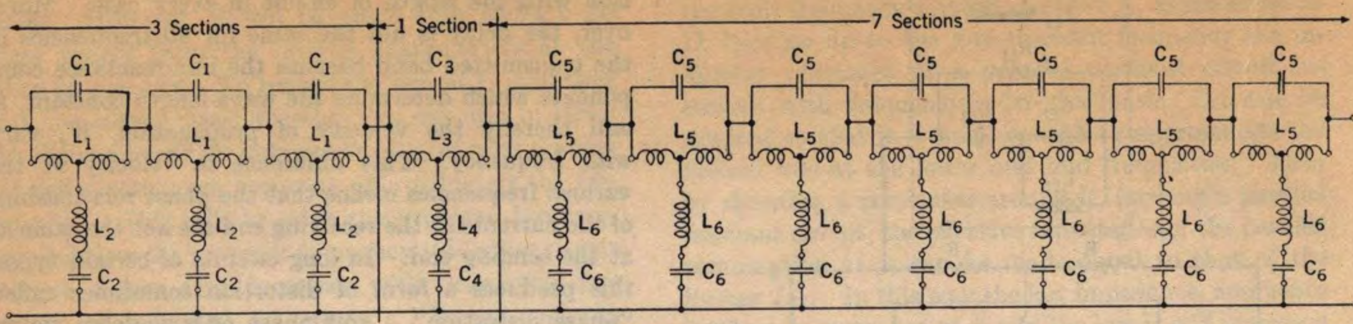


FIGURE 300

it is the differences in the time-delays of the different frequencies that produce the time-delay, or phase distortion. The allowable time-delay (phase shift) on a circuit depends, of course, upon the quality of transmission desired. The higher the quality the lower the allowable time-delay, and vice versa. For most telephone circuits, designed to transmit a voice band of about 2500 cycles, it is not necessary to equalize for time-delay distortion unless the circuit is exceptionally long.

A representative illustration of a situation where it may be necessary to equalize for this delay distortion is in the use of cable circuits for program transmission, where a frequency band of approximately 35 to 8000 cycles is transmitted. B-22, 16-gage cable pairs are used for this service. The time-delay=frequency characteristic for a 50-mile section (average length of a repeater section) of such facilities is shown in Figure 299. Here it may be noted that the maximum difference in the time-delay is about 0.9 millisecond for the 50-mile section. For a very long circuit—say 2000 miles—this difference in time-delay becomes approximately 36 milliseconds. Tests have indicated that for the quality of transmission desired, the higher frequencies (in the range of 5000 to 8000 cycles) should not be delayed in transmission more than 5 to 10 milliseconds more than the delay suffered by frequencies in the neighborhood of 1000 cycles.

Time-delay equalizers, designed to provide an inverse time-delay=frequency characteristic from that of the line, as illustrated in Figure 299, are inserted at the input of the program amplifiers as indicated in Figure 338. These time-delay equalizers are built in the form of a series of bridged T-networks as schematically illustrated in Figure 300. It will be noted that eleven of these sections, seven of one kind, three of another, and one of a third, are needed to make up the network normally used with each 50-mile section of a program cable circuit. The detailed analysis of the problems involved in the design of such time-delay correcting networks is somewhat involved and is accordingly left to the more technical literature.

CHAPTER XXIV

FILTERS

146. Filter Requirements

In telephone and telegraph work, it is often desirable to suppress or eliminate currents of certain undesired frequencies and at the same time to pass currents of other frequencies. This is accomplished by means of **electrical filters.** As would be expected from our study of resonance in Article 111, these filters are essentially networks of inductances and capacities. While the details of design of some of the more elaborate types of filters are somewhat beyond the scope of this text, we may consider here the general principles of the three major types commonly used. These are known respectively as (1) "low-pass" filters which transmit, with very little attenuation, currents of all frequencies from zero up to some designated cut-off frequency and offer very high attenuation to all higher frequencies; (2) "high-pass" filters which perform the reverse of this action and attenuate up to the cut-off value but readily transmit currents of all higher frequencies; (3) "band-pass" filters which have both an upper and a lower cut-off point, and which permit the transmission of only those frequencies lying between the two cut-off frequencies.

At the beginning it is convenient to consider the requirements of an ideal filter, even though the ideal cannot be fully obtained in practice. In such a filter, (1) frequencies lying within the pass bands would be transmitted without hindrance; in other words, over these bands the attenuation would be zero and no power would be dissipated by the filter; (2) the frequencies outside of the pass bands would be completely suppressed and the attenuation would be infinite; (3) the frequency intervals between the transmitted and attenuated frequency bands would be very small; in other words, the change from passing to suppressing or vice versa, would occur in a very narrow transition band; (4) throughout the transmitted bands, the characteristic impedance at the filter terminals would match the impedance of the terminating apparatus to prevent reflection losses.

From our study of resonance and the effect of resistance on the attenuation-frequency curves of resonant circuits, it is apparent that an ideal filter must be constructed entirely of pure reactances because the presence of resistance would produce attenuation in the transmitted bands. If there are no limitations as to the complexity of the reactance arms used in forming filters, or in the configuration in which these arms may be arranged, then there are an infinite number of possible types of filters. Naturally, however, filters are designed to meet the technical requirements using the simplest networks practicable.

147. Low-and High-Pass Filter Sections

The simplest arrangements of elementary filter networks are the "T" and "π" sections shown in Figure 301, where Z_1 and Z_2 represent the series and shunt impedances, respectively. (Note: the values of Z_1 and Z_2 are not necessarily the same in the two drawings.) For simplicity our discussion will be confined to the **T** type of network, but as we learned in the preceding chapter, these two networks may be interchanged if certain definite electrical relations are maintained. These simple networks are called **prototype** filter sections, and are the basic structures from which practical filters are developed.

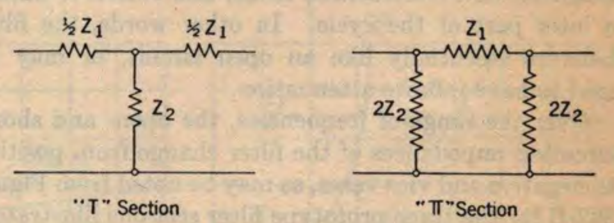


FIGURE 301

As covered in Article 124, when a network is terminated in its characteristic impedance, Z_0, the impedance presented at the input terminals is still Z_0. Its value may be determined by taking the square root of the product of the impedances (geometric mean) from one end with the other end open-circuited, Z_{oc}, and then short-circuited, Z_{sc}, ($Z_0 = \sqrt{Z_{oc}Z_{sc}}$). Since we are now considering an ideal filter made up of pure reactances in the form of a **T**-network, these open- and short-circuited impedances must be either positive (inductance), or negative (capacity) reactance. Accordingly, the characteristic impedance must have an angle of 0° or $\pm 90°$. When the open- and short-circuited impedances (reactances) have opposite signs—

$$Z_0 = \sqrt{(\pm jX_{oc})(\mp jX_{sc})}$$
$$= \sqrt{-j^2(X_{oc}X_{sc})}$$

or

$$Z_0 = \sqrt{X_{oc}X_{sc}} \tag{126}$$

This is an impedance with a zero angle, or effectively a pure resistance. If, however, the signs are alike—

$$Z_0 = \sqrt{(\pm jX_{oc})(\pm jX_{sc})}$$
$$= \sqrt{j^2(X_{oc}X_{sc})}$$

or

$$Z_0 = \pm j\sqrt{X_{oc}X_{sc}} \tag{127}$$

and the impedance has an angle of $\pm 90°$ which, of course, represents a pure reactance.

When, as in the first case, the characteristic impedance is a resistance equal to the impedance of the terminating device, the network (consisting of filter and termination) is capable of absorbing power (energy) from any source connected to the input. Since none of this power can be dissipated in the reactances of the filter, it must be passed on to the receiving device (termination). In other words, the termination absorbs all the power and the attenuation in the filter itself is zero. On the other hand, when the characteristic impedance is a pure reactance, no power can be absorbed by the network. Under this condition the filter network would merely take energy from the source during part of a cycle, store it in the electromagnetic and electrostatic fields, and return it during a later part of the cycle. In other words, the filter behaves essentially like an open circuit, or may be said to have infinite attenuation.

Over the range of frequencies, the open- and short-circuited impedances of the filter change from positive to negative and vice versa, as may be noted from Figure 302-B for the three prototype filter sections illustrated. At the same time, the characteristic impedance changes from resistance to reactance and vice versa. **When the characteristic impedance is a resistance, the filter transmits; when the characteristic impedance is a reactance, the filter attenuates.** This is also shown by the solid line curve of Figure 302-D.

As we have already seen, the characteristic impedance, Z_0 is

$$Z_0 = \sqrt{Z_{oc}Z_{sc}} \tag{128}$$

but from inspection of the T-section of Figure 301, it is evident that—

$$Z_{oc} = \frac{Z_1}{2} + Z_2$$

and

$$Z_{sc} = \frac{\dfrac{Z_1Z_2}{2}}{\dfrac{Z_1}{2} + Z_2} + \frac{Z_1}{2}$$

which, simplified, is—

$$Z_{sc} = \frac{\dfrac{Z_1^2}{4} + Z_1Z_2}{\dfrac{Z_1}{2} + Z_2}$$

Then by substituting these values in Equation (128) we get—

$$Z_0 = \sqrt{\frac{Z_1^2}{4} + Z_1Z_2} = \sqrt{Z_1\left(\frac{Z_1}{4} + Z_2\right)} \tag{129}$$

When the reactance represented by Z_1 is opposite in sign to the reactance represented by the quantity $\left(\frac{Z_1}{4} + Z_2\right)$, their product is positive and the characteristic impedance is a resistance. This can be seen by substituting reactances in Equation (129) as follows:

$$Z_0 = \sqrt{\pm jX_1\left(\frac{\pm jX_1}{4} \pm jX_2\right)}$$

Next, let the combined value of $\frac{\pm jX_1}{4} \pm jX_2$ be designated jX_3 which will be either negative or positive depending upon the relative values of X_1 and X_2. In the case when the sign of jX_3 is opposite to that of jX_1—

$$Z_0 = \sqrt{\pm jX_1(\mp jX_3)} = \sqrt{-j^2X_1X_3} = \sqrt{X_1X_3}$$

In this case the characteristic impedance, Z_0, has a zero angle, which means it is effectively a pure resistance. In the other case, when the sign of jX_3 is the same as that of jX_1, the characteristic impedance will be—

$$Z_0 = \sqrt{\pm jX_1(\pm jX_3)} = \sqrt{j^2X_1X_3} = j\sqrt{X_1X_3}$$

which is a pure reactance. In the first case, the filter transmits; in the second case, the filter attenuates.

The passed and attenuated frequencies may also be determined from the reactance curves for Z_1 and $\left(\frac{Z_1}{4} + Z_2\right)$ which are shown in Figure 302-C. When these two curves have opposite signs, the characteristic impedance of the filter is a resistance and the filter transmits, but when the signs are alike, the characteristic impedance is a reactance and the filter attenuates. This is illustrated by the solid curve of Figure 302-D. At the critical point where the $\left(\frac{Z_1}{4} + Z_2\right)$ curve crosses the zero axis, the characteristic impedance, Z_0, becomes zero as is evident from Equation (129). This is the frequency at which the filter is said to cut off. On one side of this point is the pass band and on the

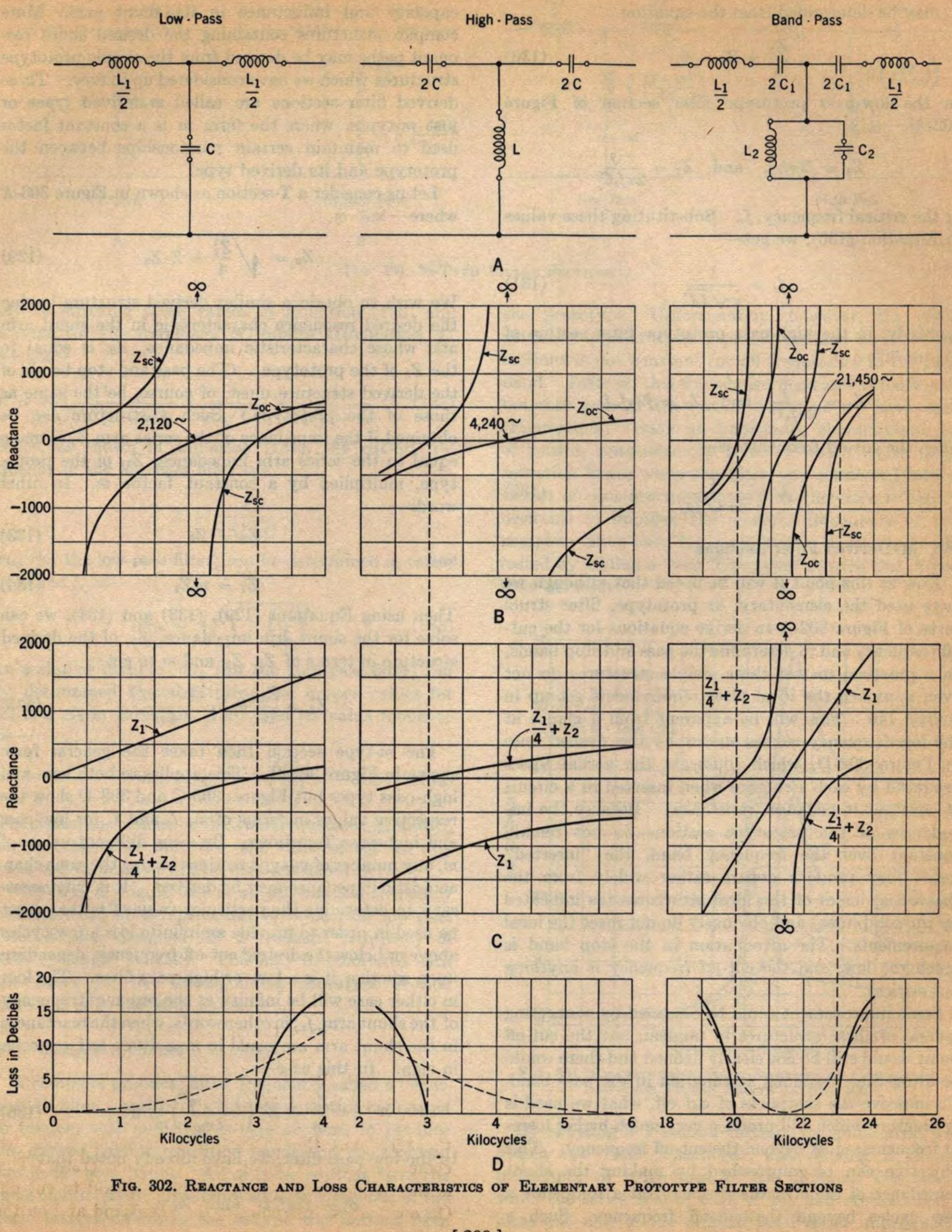


FIG. 302. REACTANCE AND LOSS CHARACTERISTICS OF ELEMENTARY PROTOTYPE FILTER SECTIONS

other side the filter attenuates. This cut-off frequency, f_c, may be determined from the equation

$$\frac{Z_1}{4} + Z_2 = 0 \qquad (130)$$

In the low-pass prototype filter section of Figure 302-A—

$$Z_1 = j2\pi f_c L \quad \text{and} \quad Z_2 = \frac{-j}{2\pi f_c C}$$

at the critical frequency, f_c. Substituting these values in Equation (130), we get—

$$f_c = \frac{1}{\pi\sqrt{LC}} \qquad (131)$$

Similarly, in the high-pass prototype filter section of Figure 302-A—

$$Z_1 = \frac{-j}{2\pi f_c C} \quad \text{and} \quad Z_2 = j2\pi f_c L$$

Then the cut-off frequency is—

$$f_c = \frac{1}{4\pi\sqrt{LC}} \qquad (132)$$

148. M-Derived Filter Sections

Now at this point it will be noted that although we have used the elementary, or prototype, filter structures of Figure 302-A to derive equations for the cut-off frequency and to determine the pass and stop bands, as a practical matter these simple structures do not even approach the ideal filter requirements set up in Article 146. This will be apparent from a glance at the loss-frequency curves shown by the dashed lines in Figure 302-D, which illustrate the actual losses produced by each structure when inserted in a circuit of constant impedance (resistance). Because the impedances of the prototype sections do not remain constant over the frequency band, the "inserted" losses they produce depart rather widely from the theoretical losses of the ideal structures (as indicated by the solid lines) and obviously do not meet the ideal requirements. The attenuation in the stop band is much too low, and the cut-off frequency is anything but critical.

Some improvement could be obtained by connecting several of these structures in tandem, but the cut-off point would still be not clearly defined and there would be altogether too much attenuation in the pass band. To improve the sharpness of cut off, what we need is a structure which will produce very much higher losses at frequencies just beyond the cut-off frequency. This objective can be approached by making the shunt impedance of the structure resonant at a frequency a few cycles beyond the cut-off frequency. Such a structure would, of course, have to contain series capacity and inductance in its shunt arm. More complex structures containing the desired shunt resonant paths may be derived from the simple prototype structures which we have considered up to now. These derived filter sections are called **m-derived types** or just m-types, where the term m is a constant factor used to maintain certain relationships between the prototype and its derived type.

Let us consider a **T**-section as shown in Figure 303-A where—

$$Z_0 = \sqrt{\frac{Z_1^2}{4} + Z_1 Z_2} \qquad (129)$$

We wish to obtain a similar derived structure having the desired resonance characteristic in the shunt arm and whose characteristic impedance, Z_0', is equal to the Z_0 of the prototype. (The pass and stop bands of the derived structure must, of course, be the same as those of the prototype.) Such a structure can be obtained if the impedance of the series arm Z_1' is made equal to the series arm impedance, Z_1, of the prototype, multiplied by a constant factor m. In other words—

$$Z_0' = Z_0 \qquad (133)$$

and

$$Z_1' = mZ_1 \qquad (134)$$

Then using Equations (129), (133) and (134), we can solve for the shunt arm impedance, Z_2', of the derived structure in terms of Z_1, Z_2, and m to get—

$$Z_2' = \frac{Z_2}{m} + Z_1\frac{(1 - m^2)}{4m} \qquad (135)$$

The m-type section then takes the general form shown in Figure 303-B. This applies to both low- and high-pass types but Figures 303-C and 303-D show the respective values in terms of m, L and C for low-pass and high-pass **T**-sections. By using different values of m, any number of m-type sections having the same characteristic impedance may be derived. It is only necessary to determine the particular value of m that must be used in order to provide an infinite loss a few cycles above or below the desired cut-off frequency, depending upon whether it is a low- or high-pass filter. The loss in either case will be infinity at the resonant frequency of the shunt arm, f_r; in other words, when the reactances in the shunt arm are equal in magnitude but opposite in sign. In this case—

$$\frac{Z_2}{m} + Z_1\left(\frac{1 - m^2}{4m}\right) = 0 \qquad (136)$$

In the low-pass filter, we have already noted that—

$$Z_1 = j2\pi f_r L \quad \text{and} \quad Z_2 = \frac{-j}{2\pi f_r C}$$

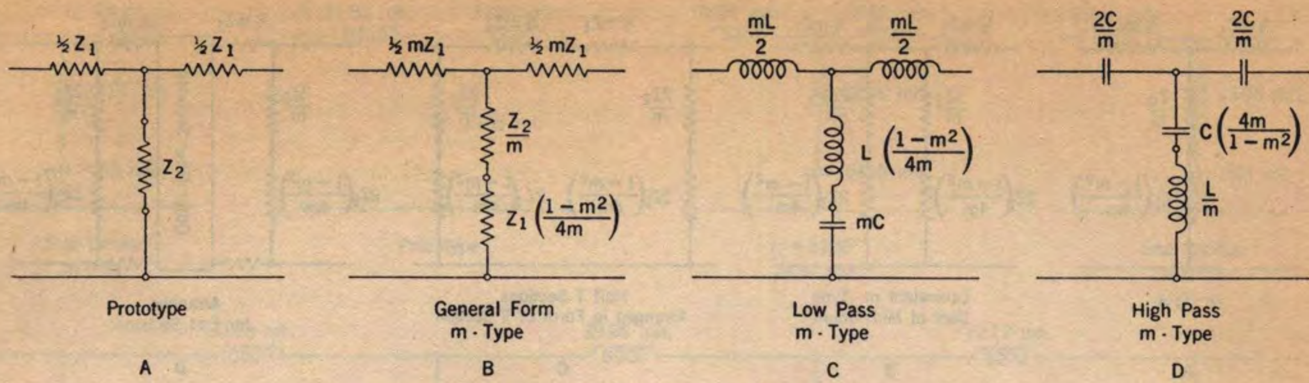


FIG. 303. M-TYPE FILTER SECTIONS

By substituting these values in Equation (136) and solving for the frequency of infinite attenuation, f_r, we get—

$$f_r = \frac{1}{\pi\sqrt{LC(1-m^2)}} = \frac{1}{\pi\sqrt{LC}} \times \frac{1}{\sqrt{1-m^2}} \qquad (137)$$

Since, as shown in Equation (131), the cut-off frequency in the low-pass filter is—

$$f_c = \frac{1}{\pi\sqrt{LC}}$$

m_{lp} (for the low-pass filter) can be determined in terms of f_c and f_r as—

$$m_{lp} = \sqrt{1-\left(\frac{f_c}{f_r}\right)^2} \qquad (138)$$

In a similar manner, m_{hp} (for the high-pass filter) can be determined by substituting the proper values for Z_1 and Z_2 in Equation (136), and its value found to be—

$$m_{hp} = \sqrt{1-\left(\frac{f_r}{f_c}\right)^2} \qquad (139)$$

The closer the values of the two frequencies, f_c and f_r, are to each other the lower the value of m. In any case, the value of m will be between 0 and +1.

Curve B of Figure 304 is an attenuation-frequency characteristic of a representative m-derived low-pass filter section designed for a resonant frequency of 3100 cycles (m = .252). This may be compared with the curve for the corresponding prototype section shown as A in the same figure.

149. Composite Filters

A complete practical filter, frequently called a "composite" filter, consists of a prototype section connected in tandem with sufficient m-type sections to produce the desired narrow transition bands and the required loss in the stop bands. All of these sections have the same characteristic impedance since this was one of the basic factors on which the m-type was derived from the prototype. Unfortunately, however, the characteristic impedance of T-prototypes and their m-type sections is not constant for all frequencies in the pass band. Instead, this impedance decreases rapidly and becomes zero at the cut-off value as we have seen (Figure 302). Such an impedance characteristic is, of course, unsatisfactory since it would cause large reflection losses when the filter was connected into a circuit of constant impedance. It therefore is highly desirable to improve the terminal impedance of the complete filter over the pass band. This can be effected by adding a "half T-section" at each end of the filter structure.

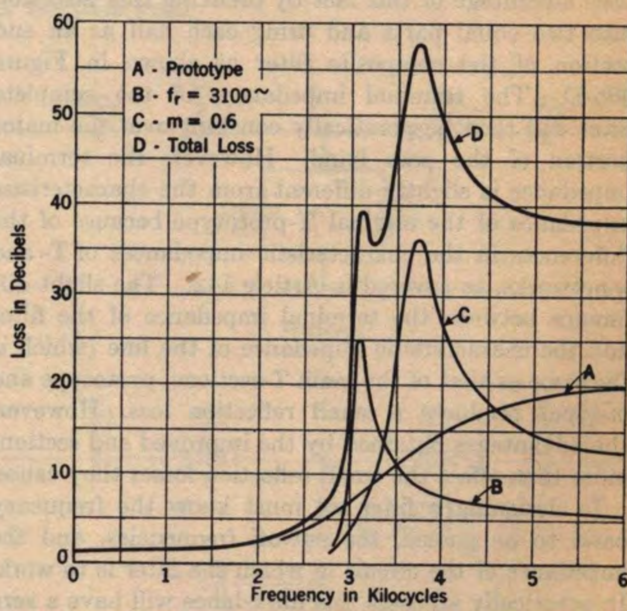


FIG. 304. CHARACTERISTICS OF SECTIONS OF COMPOSITE LOW-PASS FILTER SHOWN IN FIG. 306

By splitting a T-section through its shunt arm, we obtain two "half T-sections", each having a shunt impedance of twice that of the original T-section, while the series arm of each of the half sections contains one half of the original total series impedance.

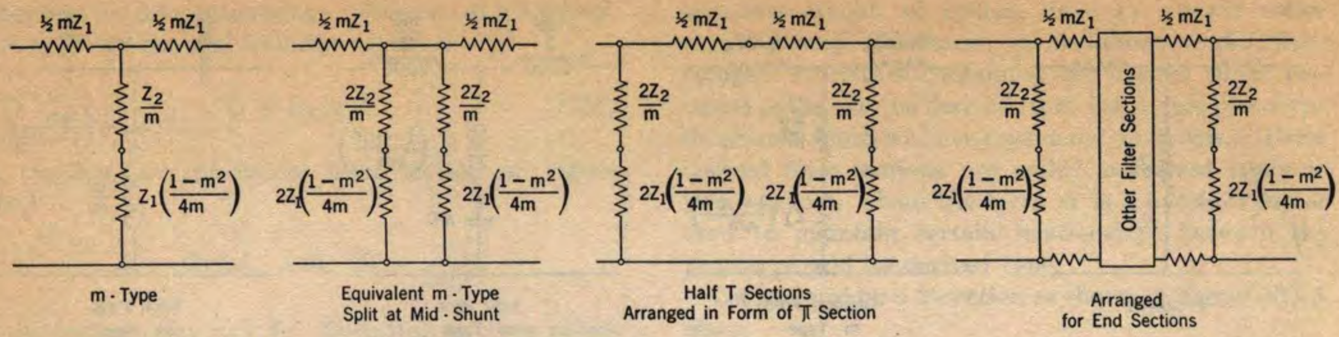


FIG. 305. HALF T M-TYPE FILTER END SECTIONS

This splitting arrangement is illustrated in Figure 305-B. We may then rearrange the series and shunt arms so that the network takes the form of a π as in Figure 305-C. It should be noted, however, that we are not converting the **T**-section into an electrically equivalent π-section, but simply rearranging the positions of the series and shunt arms without changing their respective values. The new structure will have the same loss as it had in its original **T** form but its impedance will be changed because we are now viewing the network "mid-shunt" instead of "mid-series".

Now, it can be shown that the impedance of such a structure is practically uniform over the major portion of the pass band, if m is made equal to 0.6. We can take advantage of this fact by breaking this π-section into two equal parts and using each half as an end section of the composite filter as shown in Figure 305-D. The terminal impedances of the complete filter will then be practically constant over the major portion of the pass band. However, the terminal impedance is slightly different from the characteristic impedance of the original **T**-prototype because of the differences in the characteristic impedances of **T**- and π-networks, as covered in Article 142. The slight difference between the terminal impedance of the filter and the characteristic impedance of the line (which is the same as that of the main **T**-sections, prototype and m-type) produces a small reflection loss. However, the advantages obtained by the improved end sections more than offset the small reflection losses they cause.

In designing a filter we must know the frequency band to be passed, the cut-off frequencies, and the impedance of the circuit in which the filter is to work. In practically all cases this impedance will have a zero angle, and it may therefore be considered as a pure resistance, R. It can be shown that in such a case the matching impedance of the filter is—

$$R = \sqrt{\frac{L}{C}} \tag{140}$$

for both the low- and high-pass filters. Combining this relation with the value of the cut-off frequencies as given in Equations (131) and (132), we have for the low-pass filter—

$$L = \frac{R}{\pi f_c} \tag{141}$$

and

$$C = \frac{1}{\pi f_c R} \tag{142}$$

and for the high-pass filter—

$$L = \frac{R}{4\pi f_c} \tag{143}$$

and

$$C = \frac{1}{4\pi f_c R} \tag{144}$$

These are the values of the inductance and capacity in the prototype sections. Constants of the other sections are developed from these prototype constants.

Example: Design a low-pass filter with a cut-off frequency of 3000 cycles to work into an impedance of $600\underline{/0°}$. To provide a sharp cut-off, f_r should be 3100 cycles. The end sections should be designed for $m = 0.6$.

Solutions: From Equations (141) and (142), the inductance and capacity for the prototype will be—

$$L = \frac{R}{\pi f_c} = \frac{600}{\pi 3000} = .06366 \text{ henry or } 63.66 \text{ mil hen.}$$

$$C = \frac{1}{\pi f_c R} = \frac{1}{\pi 3000 \times 600}$$

$$= .1768 \times 10^{-6} \text{ farad or } .1768 \text{ mf.}$$

From Equation (138) and Figure 303-C, the m-type section for $f_r = 3100$ will be—

$$m_{lp} = \sqrt{1 - \left(\frac{f_c}{f_r}\right)^2} = \sqrt{1 - \left(\frac{3000}{3100}\right)^2} = 0.252$$

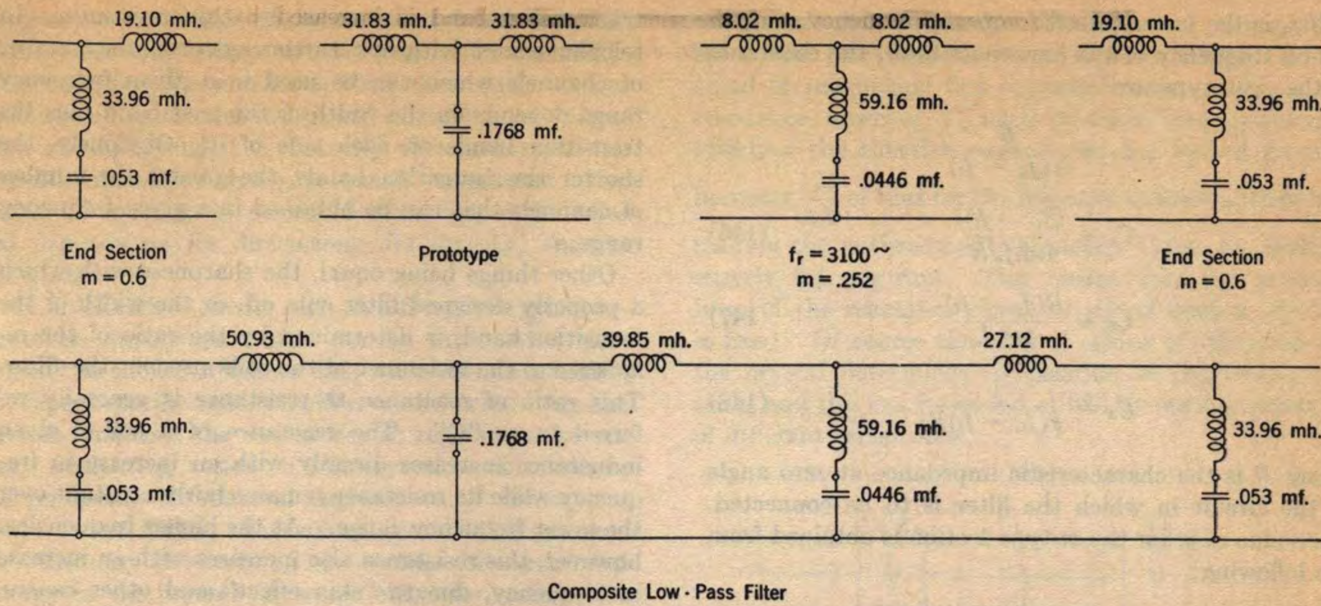


FIG. 306. MAKE-UP OF TYPICAL COMPOSITE LOW-PASS FILTER

$$\frac{mL}{2} = \frac{.252 \times 63.66}{2} = 8.02 \text{ mil hen.}$$

$$L\left(\frac{1-m^2}{4m}\right) = 63.66\left(\frac{1-(.252)^2}{4 \times .252}\right)$$

$$= 59.16 \text{ mil hen.}$$

$$mC = .252 \times .1768 = .04455 \text{ mf.}$$

In a similar manner the end sections, where $m = 0.6$, will be—

$$\frac{mL}{2} = \frac{.6 \times 63.66}{2} = 19.10 \text{ mil hen.}$$

$$L\left(\frac{1-m^2}{4m}\right) = 63.66\left(\frac{1-(.6)^2}{4 \times .6}\right) = 16.98 \text{ mil hen.}$$

$$mC = .6 \times .1768 = .106 \text{ mf.}$$

The composite filter thus takes the form illustrated in Figure 306. The attenuation-frequency characteristics for this composite filter and each of its component structures are illustrated in Figure 304.

150. Band-Pass Filters

Band-pass filters may be obtained by connecting a low-pass and a high-pass filter in series, with their cut-off frequencies so arranged as to pass the desired band. In practice, however, these filters are designed and built as a single structure having two cut-off frequencies, using the same principles as already discussed for the low- and high-pass types. The band-pass filter shown in Figure 302 is one of several prototype forms. An m-type section can be derived for each of the various prototypes, and the one in question is illustrated in Figure 307. This m-type is derived in the same general manner as those for the low- and high-pass filters. In the band-pass filter, just as with low-pass and high-pass filters, it is desirable to use an m-derived type T-section, where $m = 0.6$, and to split this to form half T-sections for the end terminations, thus providing uniform terminal impedances over the major portion of the pass band.

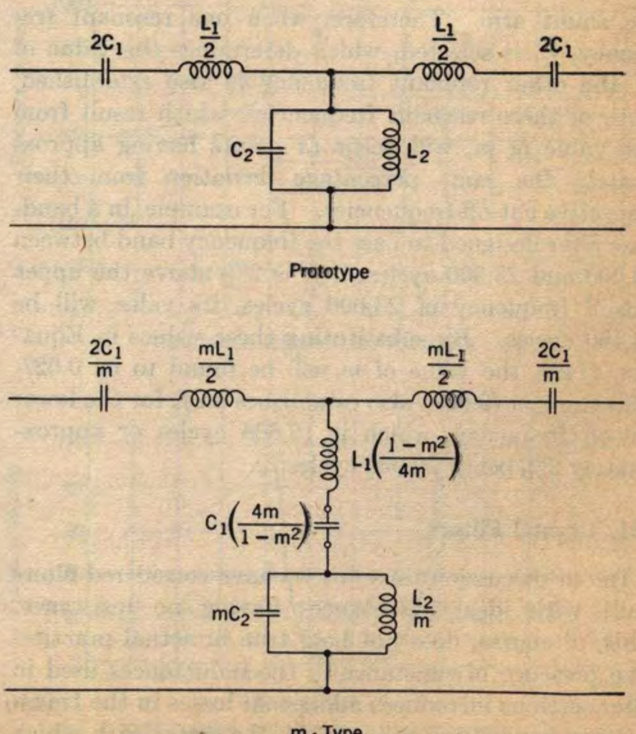


FIG. 307. BAND-PASS FILTER SECTIONS

If f_l is the lower cut-off frequency, and f_u the upper cut-off frequency of the band-pass filter, the constants of the prototype are—

$$L_1 = \frac{R}{\pi(f_u - f_l)} \tag{145}$$

$$C_1 = \frac{(f_u - f_l)}{4\pi f_l f_u R} \tag{146}$$

$$L_2 = \frac{R(f_u - f_l)}{4\pi f_l f_u} \tag{147}$$

$$C_2 = \frac{1}{\pi(f_u - f_l)R} \tag{148}$$

where R is the characteristic impedance at zero angle of the circuit in which the filter is to be connected. The value of m for the m-type section is obtained from the following:

$$m = \sqrt{1 - \left[\frac{f_r(f_u - f_l)}{f_r^2 - f_u f_l}\right]^2} \tag{149}$$

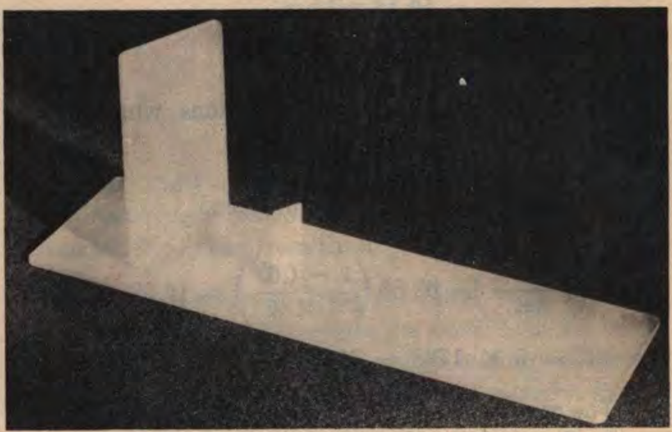
where f_r is the resonant frequency of infinite attenuation.

Since there are two cut-off frequencies, f_u and f_l, there should be two resonant frequencies to provide infinite attenuation both above the upper cut-off frequency, f_u, and below the lower cut-off frequency, f_l. It may be noted, however, from an inspection of the m-type section of Figure 307, that for each value of m there will be two resonant frequencies in the overall network of the shunt arm. Therefore, when one resonant frequency, f_r, is selected, which determines the value of m, the other resonant frequency is also established. Both of these resonant frequencies, which result from one value of m, will occur at points having approximately the same percentage deviation from their respective cut-off frequencies. For example, in a band-pass filter designed to pass the frequency band between 20,000 and 23,000 cycles, if f_r is 2% above the upper cut-off frequency of 23,000 cycles, its value will be 23,460 cycles. By substituting these values in Equation (149), the value of m will be found to be 0.627. This same m (0.627) also establishes an f_r for the lower cut-off frequency, which is 19,608 cycles or approximately 2% below 20,000 cycles.

151. Crystal Filters

In our discussion thus far, we have considered filters built with ideal inductances having no resistance. This, of course, does not hold true in actual practice. The presence of resistance in the inductances used in filter sections introduces additional losses in the transmitting bands, and reduces the sharpness with which the filter cuts off. In other words, the width of the transition band is increased by this resistance. In telephone and telegraph carrier systems, the number of channels which can be used in a given frequency range depends on the width of the pass band plus the transition bands on each side of it. Obviously, the shorter the transition bands, the greater the number of channels that can be obtained in a given frequency range.

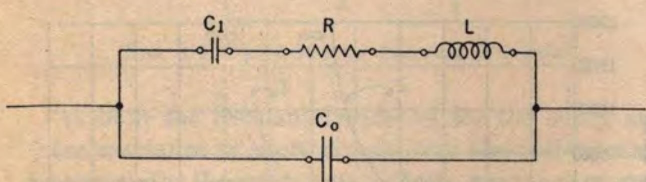
Other things being equal, the sharpness with which a properly designed filter cuts off, or the width of the transition band, is determined by the ratio of the reactance to the resistance of the coils used in the filter. This ratio of reactance to resistance is generally referred to as "Q". The reactance of a coil of given inductance increases directly with an increase in frequency while its resistance remains fairly constant over the lower frequency range. At the higher frequencies, however, this resistance also increases with an increase in frequency, due to "skin effect" and other causes. As a result the Q of a coil seldom exceeds 400 in practice. In the frequency range up to about 30 kc., this value of Q has been found high enough to provide satisfactory filters for carrier operation, but for frequencies above this value filter elements having higher Q's are desirable.



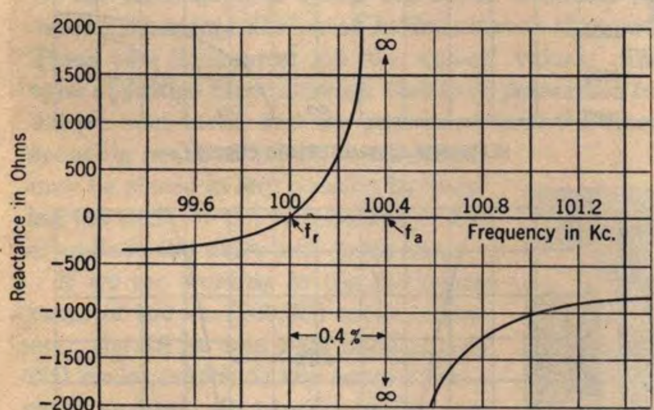
Quartz Crystals

One of the most practicable ways to obtain such high Q elements is by the use of mechanical vibrating systems, such as the "piezo-electric crystal", which possesses a natural mechanical-electrical relationship. Of the many substances that exhibit this piezo-electric effect, the one that has been found most satisfactory for these purposes is crystalline quartz. When a piece of this crystalline quartz is strained mechanically, it sets up an electric field in its neighborhood, introducing electric potentials on conductors in the field. Conversely, when a piece of crystal is placed in an electric field, its shape changes very slightly. This not only applies to a crystal in its natural form, but thin slabs of the crystalline material cut in certain ways will exhibit the same characteristics even more markedly. When the frequency of an applied alternating voltage is the same as the natural period of mechanical vibration of the crystal, the intensity of vibration of the crystal will reach a sharp "resonant" maximum. The natural frequency, or period of vibration, of the crystal depends on its dimensions, its density, and its elasticity.

In an electrical circuit such as a filter, a crystal presents an impedance which can be represented electrically as shown in Figure 308, where the inductance, L, represents the mass reaction of the crystal against motion (inertia); the resistance, R, represents the energy dissipating action in the crystal as it vibrates; C_0 represents the natural capacitance of the crystal when at rest (static characteristic); and C_1 represents the elasticity determining the storage of mechanical energy in the crystal (dynamic characteristic).



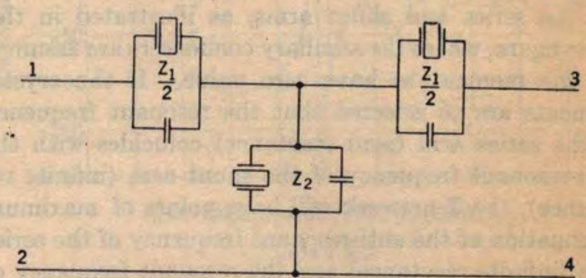
EQUIVALENT ELECTRICAL CIRCUIT OF A QUARTZ CRYSTAL



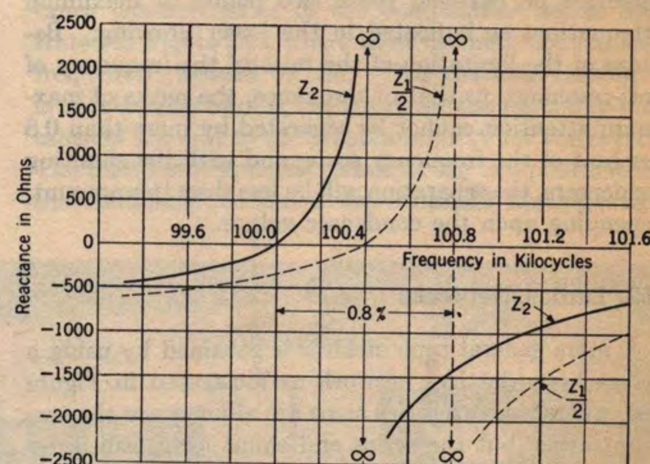
REACTANCE CHARACTERISTIC OF A QUARTZ CRYSTAL

FIGURE 308

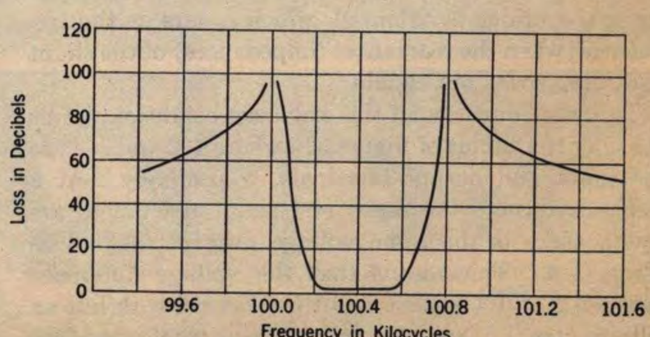
The ratio of C_0 to C_1 is a constant for any given crystal material. For quartz it is 125 to 1. The Q for quartz crystals is of the order of 20,000 or more. The general reactance characteristics of such a crystal are also shown in Figure 308 where the resonant frequency is f_r, and the anti-resonant frequency is f_a. Since there is a fixed ratio of C_0 to C_1 (125 to 1) for quartz, the anti-resonant frequency, f_a, is always 0.4 per cent higher than the resonant frequency, f_r. This may be more easily understood if we keep in mind that resonance occurs when the mass reaction represented by L and the elastic reaction represented by C_1 are equal in magnitude but opposite in sign. For anti-resonance, however, C_0 must be taken into consideration and the effective capacity in the looped circuit becomes $\frac{125}{126}$ of that for the resonant condition, thereby making the anti-resonant frequency higher by almost exactly 0.4 per cent. This means that the general form of the reactance characteristic of such a crystal is fixed. Of course the natural period of vibration of the crystal determines its position in the frequency scale, and this can be varied at will by cutting crystals of different dimensions.



"T" FILTER SECTION USING CRYSTALS



REACTANCE CHARACTERISTIC CURVES



ATTENUATION · FREQUENCY CHARACTERISTIC

FIGURE 309

If an auxiliary condenser is placed in parallel with the crystal, however, the effective capacity of C_0 is increased, and as a result the anti-resonant frequency can be made less than 0.4 per cent above the resonant frequency. This can be done without detracting from the favorable characteristics of the crystal because the Q of the added condenser will normally compare favorably with the Q of the crystal.

As we have seen, the simplest form of a filter circuit is the T-network. Now if crystals are placed in the series and shunt arms of such a network, we have the circuit shown in Figure 309, where condensers are also added in parallel with the crystals to permit control of the band width. The characteristics of such a filter circuit can be analyzed by drawing the reactance curves for the series and shunt arms, as illustrated in this same figure, where the auxiliary condensers are assumed for the moment to have zero value. If the crystal elements are so selected that the resonant frequency of the series arm (zero reactance) coincides with the anti-resonant frequency of the shunt arm (infinite reactance), the T-network will have points of maximum attenuation at the anti-resonant frequency of the series arm (infinite reactance) and the resonant frequency of the shunt arm (zero reactance). The pass band must therefore lie between these two points of maximum attenuation, as indicated in the lower drawing. Because of the limitation of the ratio of the frequency of anti-resonance to that of resonance, the peaks of maximum attention cannot be separated by more than 0.8 per cent of the frequency scale; and with the shunting condensers, the separation will be less than this amount, depending upon the condenser values.

152. Lattice Networks

A more general type of filter is obtained by using a bridge type (lattice) network as illustrated in Figure 310, where the two series arms are alike as are the two shunt arms, but the series and shunt arms still differ from each other. This type of network is equivalent to a bridge circuit where no current flows in the output when the bridge is balanced, which occurs at the frequencies when the reactances (impedances) of the shunt and series arms are equal.

To better understand this balanced condition, let us consider the circuit of Figure 310 where 1–2 and 3–4 are the input and output terminals, respectively. At a frequency where the series and shunt reactances are equal, there is the same voltage drop across 1–3 as across 1–4. This means that the voltage difference between 3 and 4 is zero, and the filter network has an infinite loss. When the frequency is such that the reactances in the series and shunt arms are of opposite sign, on the other hand, the voltage difference between the output terminals 3–4 is equal to the input voltage across 1–2, and current of this frequency is passed with zero attenuation (considering no energy dissipated in the crystals).

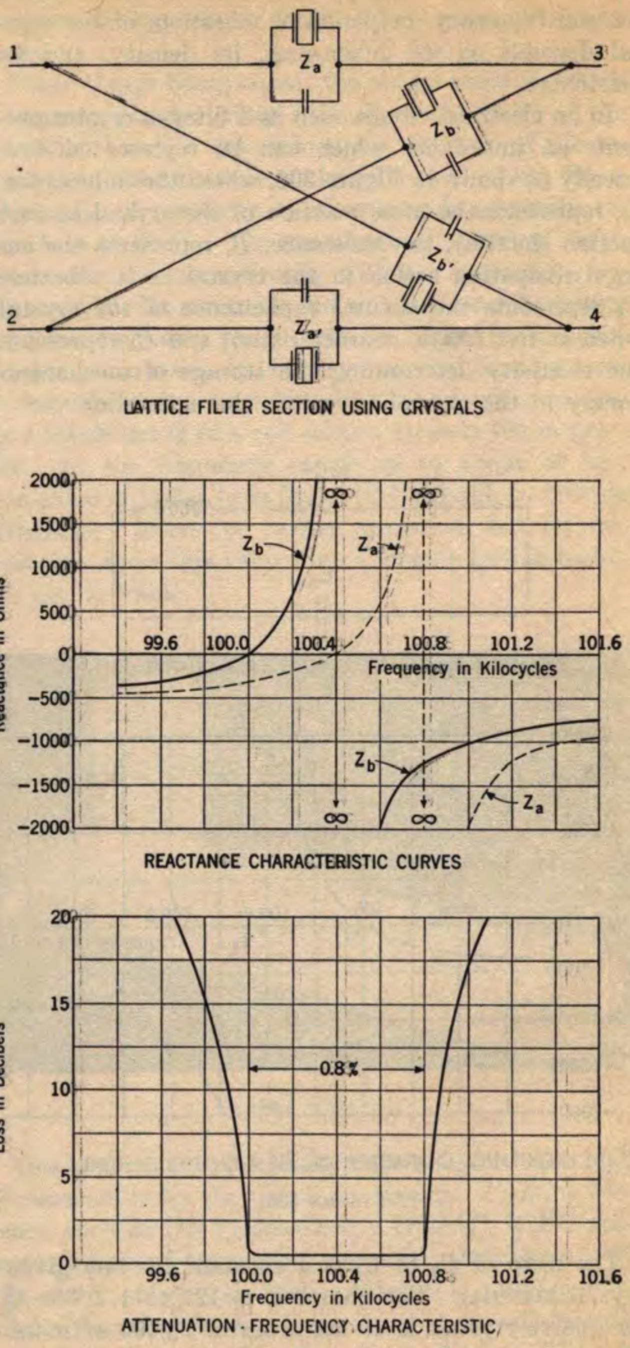


FIGURE 310

This may be more easily seen from examination of this network rearranged in the more conventional balanced bridge form of Figure 311. Here the series arms, A and A', and the shunt arms, B and B', may be either positive (inductive) or negative (capacity) reactance, depending upon the frequency of the input. For example, if an input of 5 volts produces a voltage drop across A and A' of $+7$ volts and across B and B' of -2 volts, there will be the same voltage in the output (3–4) as at the input. On the other hand, when the frequency is such that the reactances in these two arms are of the same sign, the voltage across the output terminals 3–4 becomes less than that across the input. The loss of the network then takes some definite value between zero and infinity.

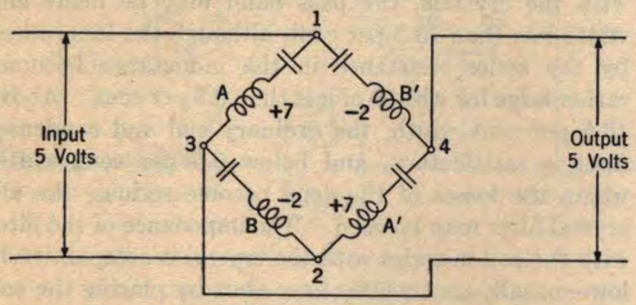
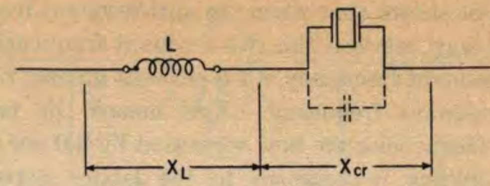


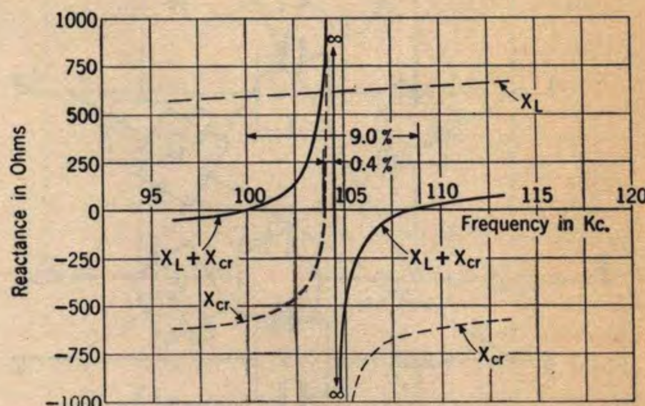
Fig. 311. Principle of Lattice Network

At both the resonant frequency for the shunt arm (zero reactance or short circuit) and the anti-resonant frequency for the series arm (infinite reactance or open circuit), the voltage across the output terminals will be the same as that across the input terminals (assuming no energy dissipated in the network elements). These two frequencies are the cut-off values. This type of lattice filter network therefore passes the full 0.8 per cent band, and the pass band (and the corresponding peaks of infinite attenuation) may be placed in any position by varying the ratio of the impedances of the crystals in the series and shunt arms.

If we are working in the frequency range of 100 kc. (100,000 cycles), however, the 0.8 per cent pass band is only 800 cycles, which is too narrow for a voice channel. In practice, wider pass bands are required. They may be obtained by adding inductance coils in the network of crystals. As we have seen such coils have a relatively low Q, and they will be satisfactory, therefore, only if they can be used in such a manner that the loss (energy dissipated) they introduce does not overcome the beneficial effect of the low loss of the crystals. As a practical matter, it has been found that within certain limits the Q of the crystals has sufficient margin to make the use of such coils possible without serious detrimental effects.



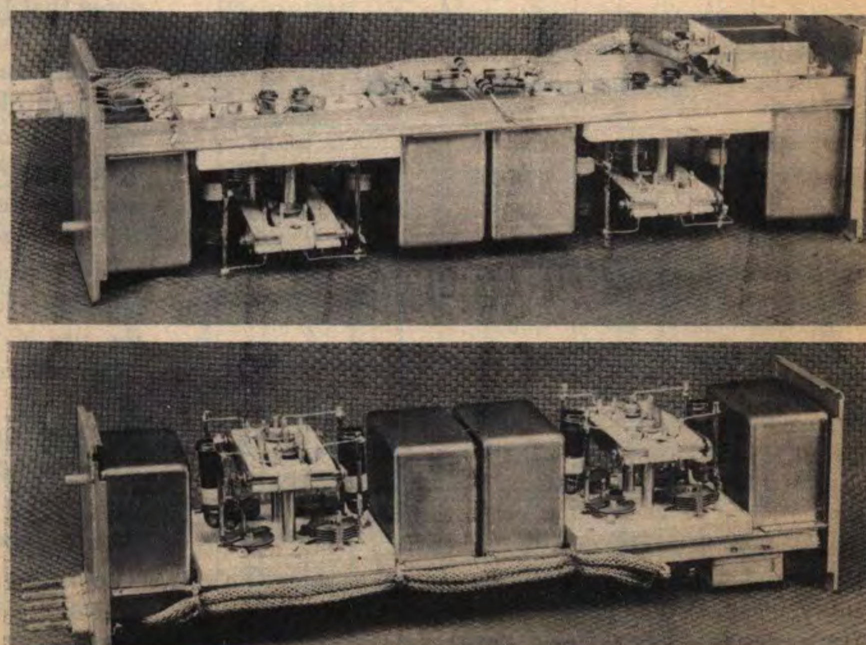
INDUCTANCE IN SERIES WITH CRYSTAL



REACTANCE CHARACTERISTIC CURVES

Figure 312

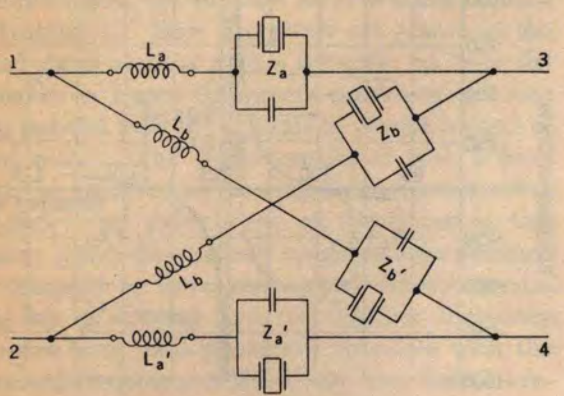
If a coil is placed in series with a crystal, the reactance characteristics of the combination are as illustrated in Figure 312, where there are now two resonant frequencies instead of one. This characteristic curve is obtained by adding the curve for the inductive reactance of the coil to the reactance curve for the crystal. The resonant frequencies occur at the two points where the combined reactance curve crosses the zero axis.



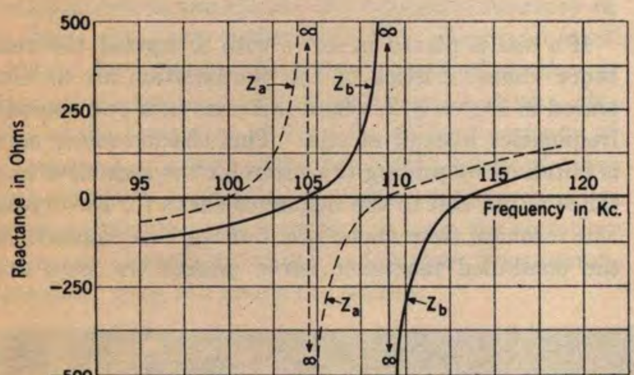
Channel Filter for Broad-Band Carrier Systems

It can be shown that when the anti-resonant frequency is half-way between the two resonant frequencies, the anti-resonant frequency is 4.5 per cent greater than the first resonant frequency. This means the two resonant frequencies are now separated by 9.0 per cent.

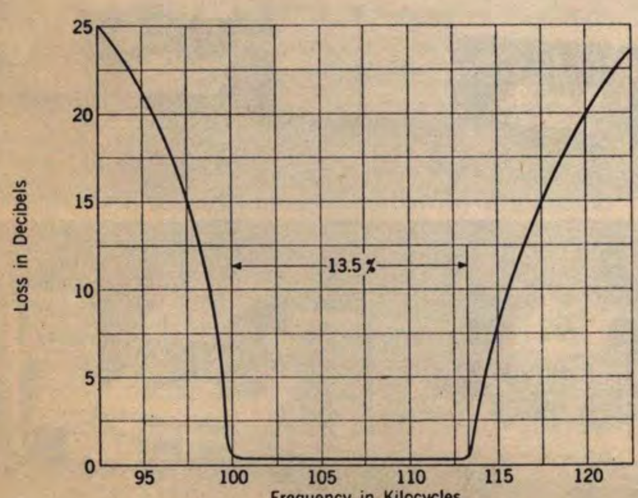
By adding inductances to the lattice network of Figure 310 we have the filter illustrated in Figure 313,



LATTICE FILTER SECTION WITH SERIES INDUCTANCE



REACTANCE CHARACTERISTIC CURVES



ATTENUATION - FREQUENCY CHARACTERISTIC

Figure 313

which passes the band of frequencies from the first resonant frequency of the series arm to the second resonant frequency of the shunt arm. The characteristic curves show that this filter arrangement now has a pass band of 3 x 4.5 or 13.5 per cent. In the frequency range of 100 kc. this means a pass band of 13,500 cycles. The frequencies of infinite attenuation, of course, occur where the impedances of the two arms are equal and of the same sign, which is another way of saying the bridge arrangement is balanced.

By varying the size of the condensers in parallel with the crystals, the pass band may be made any width less than 13.5 per cent, although the loss caused by the series resistance in the inductance becomes rather large for widths of less than 0.5 per cent. Above 13.5 per cent width, the ordinary coil and condenser filter is satisfactory, and below 0.5 per cent width, where the losses of the coils become serious, the all-crystal filter may be used. The impedance of the filter with the coil in series with the crystal is comparatively low—usually under 600 ohms—but by placing the coil in parallel with the crystal (which will have the same general effect on the width of the pass band as the series connection), this impedance may be made as high as 400,000 ohms for the narrower band widths. Such filters are suitable for connecting high impedance screen-grid vacuum tubes without the use of transformers.

Crystal filters have a wide field of application—especially where "broad-band" carrier is employed. A representative band-pass crystal filter, such as is used in the high-frequency cable carrier system (Type-K), is shown schematically in Figure 314. It may be noted that this filter contains two lattice sections. In each section, the crystal elements in the series arms are identical and those in the lattice (shunt) arms are identical. However, the elements of one section are not the same as those of the other section. The inductances are now cut in half and placed outside of the lattice. The shunting resistance between the two sections, together with the resistance components of the adjacent coils, form a resistance pad to match the impedance of the two lattice sections. By varying the size of this shunting resistance, the loss may be adjusted over a small range to obtain the same loss in each filter.

Electrically there are four crystal elements in each lattice section but for reasons of economy and for convenience in handling and adjusting the crystals, those in corresponding arms are "mechanically one". This means actually that there are but two physical crystals in each lattice section, or four for the whole filter shown in Figure 314. This is accomplished by plating the surfaces of the crystals used with a thin

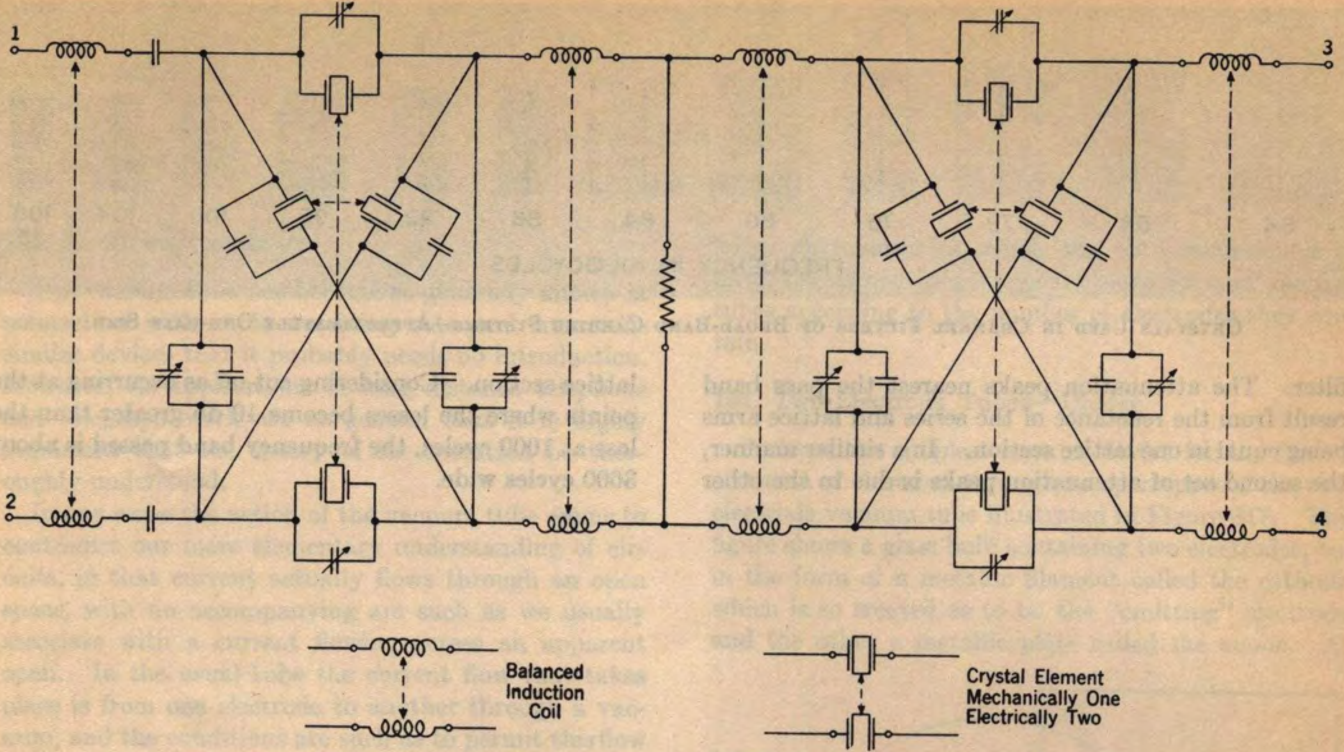


Fig. 314. Typical Band-Pass Crystal Filter

layer of aluminum which is divided along the center line lengthwise of the crystal to form two electrically independent crystal units from a single crystal. Since the crystals vibrate longitudinally with a node across the middle, they are clamped at this node in mounting as illustrated in Figure 315. The four crystals used in the lowest frequency channel (64 kc.) of the Type-K carrier system range in length from about 1.65 to 1.58 inches, while those in the highest frequency channel (108 kc.) range from about 0.96 to 0.94 inch. These crystals vary in thickness from about 0.0248 inch to 0.0433 inch, while the widths are in the order of 50 to 80 per cent of their lengths.

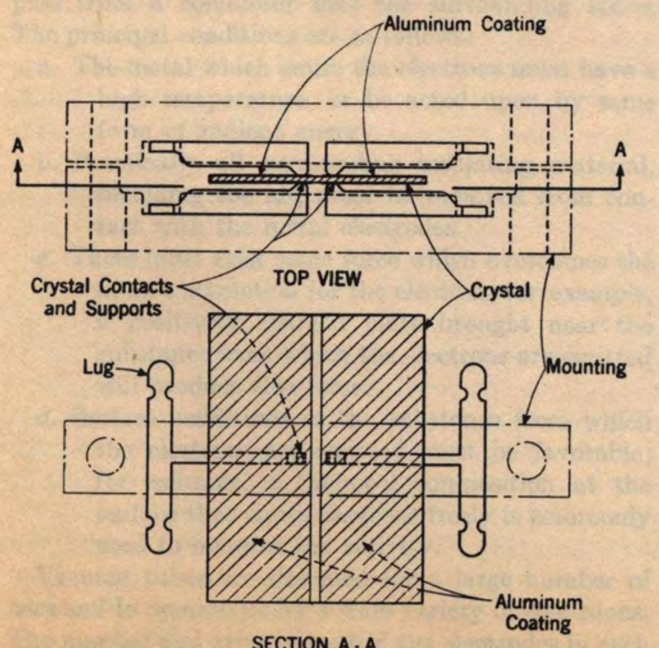


Fig. 315. Method of Mounting Quartz Crystal

Figure 316 is a representative loss-frequency characteristic of a two-lattice section crystal band-pass

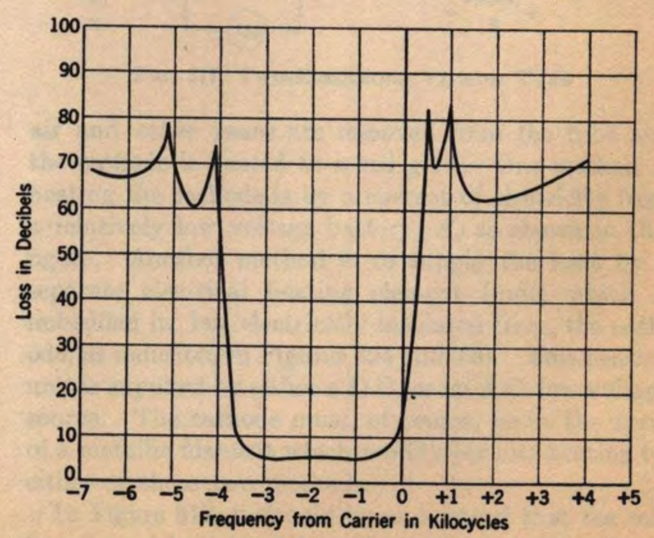
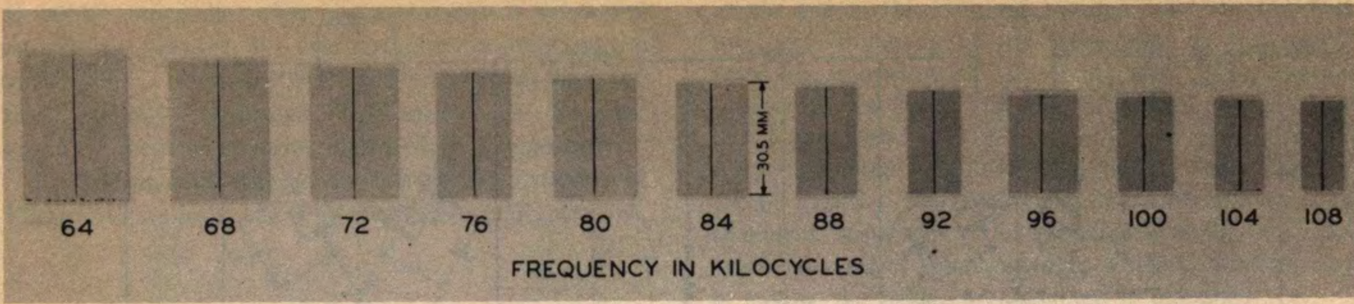


Fig. 316. Loss-frequency Characteristic of Filter Shown in Fig. 314



CRYSTALS USED IN CHANNEL FILTERS OF BROAD-BAND CARRIER SYSTEMS—APPROXIMATELY ONE-HALF SIZE

filter. The attenuation peaks nearest the pass band result from the reactance of the series and lattice arms being equal in one lattice section. In a similar manner, the second set of attenuation peaks is due to the other lattice section. Considering cut-off as occurring at the points where the losses become 10 db greater than the loss at 1000 cycles, the frequency band passed is about 3600 cycles wide.

CHAPTER XXV

VACUUM TUBES

153. Thermionic Emission

The vacuum tube has become so generally known in connection with radio receiving sets and numerous similar devices that it probably needs no introduction. However, its applications in long distance telephone and telegraph work are so general that it is highly important that the theory of its operation be thoroughly understood.

In one sense the action of the vacuum tube seems to contradict our more elementary understanding of circuits, in that current actually flows through an open space, with no accompanying arc such as we usually associate with a current flowing across an apparent open. In the usual tube the current flow that takes place is from one electrode to another through a vacuum, and the conditions are such as to permit this flow without the application of extremely high potentials.

Any fundamental conception of the tube's action is based on the electron theory, which was outlined briefly in Chapter I. But we have not heretofore dealt with the passage of electrons from one conducting substance to another, unless the two substances were actually in contact. Even in the case of the arc, the air or surrounding gas becomes a conductor. Yet, under a certain combination of physical conditions, electrons will pass from a conductor into the surrounding space. The principal conditions are as follows:

a. The metal which emits the electrons must have a high temperature, or be acted upon by some form of radiant energy.

b. Practically all surrounding insulating material, including the air, must be removed from contact with the metal electrodes.

c. There must exist some force which overcomes the atom's attraction for the electron; for example, a positively charged plate brought near the substance from which the electrons are emitted will produce this force.

d. Surface conditions of the substance from which the electrons are emitted must be favorable; for example, a chemical composition at the surface that emits electrons freely is commonly used to increase the activity.

Vacuum tubes are designed for a large number of uses and to operate under a wide variety of conditions. The number and arrangement of the electrodes in each type of vacuum tube is determined largely by the electrical characteristics which the tube is designed to have. This provides a logical classification of vacuum tubes according to the number of electrodes they contain.

154. Two-Electrode Tubes—Diodes

Perhaps the simplest form of device for the flow of a stream of electrons through an open space is the two-electrode vacuum tube illustrated in Figure 317. This figure shows a glass bulb containing two electrodes, one in the form of a metallic filament called the cathode, which is so treated as to be the "emitting" electrode, and the other a metallic plate called the anode. All

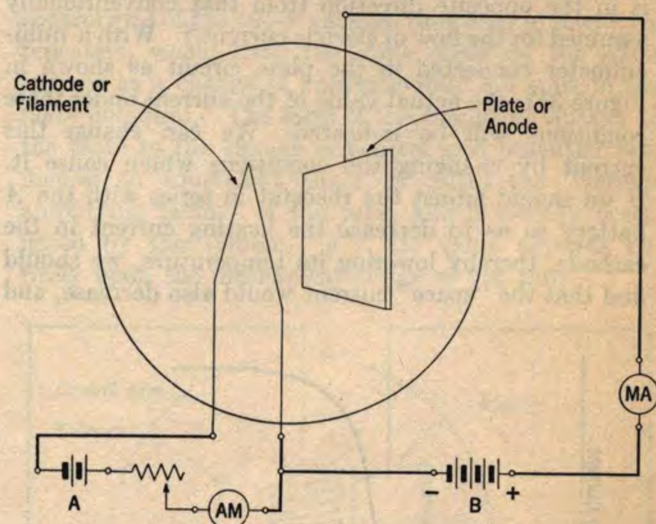


FIG. 317. TWO-ELECTRODE VACUUM TUBE

air and other gases are removed from the tube and the cathode is heated to a red glow. One method of heating the cathode is by a current of electricity from a relatively low voltage battery, *A*, as shown in this figure. Another method is to supply the heat by a separate electrical heating element (unit) which is imbedded in, but electrically insulated from, the cathode, as indicated in Figures 324 and 326. This heating unit is supplied by either a D.C. or an A.C. low voltage source. The cathode must, of course, be in the form of a metallic filament which readily permits heating by either of these two methods.

In Figure 317, it should be understood that the sole function of battery *A* is to heat the cathode; certain secondary effects which may be traced to this battery are explained later. Assuming that the filament is either of the proper chemical substance, or is coated with some chemical substance which will promote the emission of electrons, we have provided all conditions necessary for electron flow with the exception of (c) above. If now, a battery is connected between the cathode (filament) and the metallic plate (anode), and is so poled as to give the plate a positive charge, this charge will exert a force of attraction on any electrons that may be emitted into the space surrounding the heated cathode. Electrons escaping from the cathode will be drawn to the plate by the force set up by its positive charge, and a continuous flow of electrons from cathode to plate will result. The speed with which the electrons cross the gap is determined by the potential of the plate with respect to the cathode.

This flow of electrons is simply an electrical current, and the battery *B* will sustain this current in the same way that a battery sustains a current when it is connected to any closed electrical circuit. (It may be noted that the flow of electrons from cathode to plate is in the opposite direction from that conventionally assumed for the flow of electric current.) With a milliammeter connected in the plate circuit as shown in Figure 317, the actual value of the current under these conditions will be indicated. We can change this current by changing the conditions which cause it. If we should adjust the rheostat in series with the *A* battery so as to decrease the heating current in the cathode, thereby lowering its temperature, we should find that the "space" current would also decrease, and

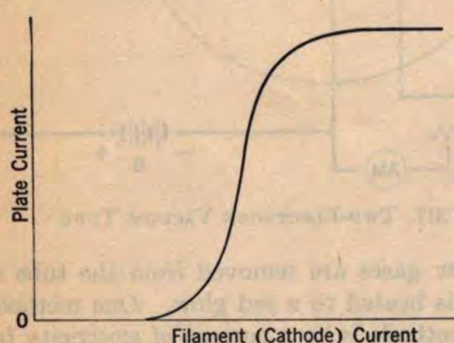


Fig. 318. Cathode Current vs. Plate Current Characteristic

with the cathode reduced to room temperature, the milliammeter in the plate circuit would read zero, indicating zero space current. On the other hand, if we should increase the voltage of the *A* battery, thereby increasing the cathode heating current and in turn increasing its temperature, the space current would be increased. This general relation between cathode current and space current for a representative vacuum tube is illustrated in Figure 318.

There is a limit, however, to the increase in space current that can be obtained by increasing the cathode temperature. This is due to the fact that electrons repel each other because they are all negatively charged, and free electrons in the space tend to keep new electrons from leaving the cathode. In other words, the electrons themselves, when emitted, tend to counteract further emission of other electrons, or to exert a repelling force on electrons within the cathode. This is called the "space-charge effect". When the cathode

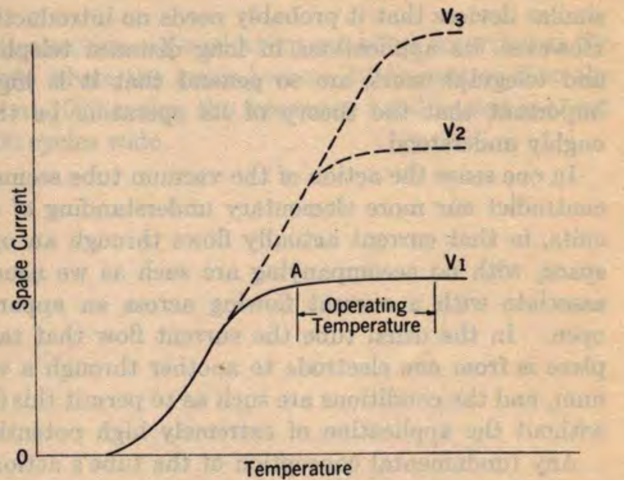


Fig. 319. Cathode Temperature vs. Plate Current Characteristic

reaches a certain temperature, there will be so many electrons in the space that their repelling effect prevents any further increase in the number leaving the cathode. The space current then becomes constant regardless of further increase of temperature. This accounts for the bending over of the curves in Figures 318 and 319. When conditions are as shown by point *A* in Figure 319, the tube is said to have reached the **temperature saturation** point. In practice this is the operating temperature for the cathode, since a slight change in the *A* battery voltage or heating current will not appreciably affect the tube's action.

Having disposed of the effects of changing the temperature of the cathode, let us next consider the effects of a change in the voltage of the plate. We have said that the speed of the electrons in proceeding from the cathode to the plate depends upon the plate potential. We should expect an increase in voltage to give two effects; first, the space current would be increased inasmuch as electrons are transferred more rapidly; second, since the number of electrons in the space surrounding the cathode is reduced, the space-charge effect will be lessened and the saturation point will be reached only at a higher temperature. The curves shown in Figure 319 illustrate such results— V_1, V_2, and V_3 represent three plate voltage values and it is seen that increasing the voltage increases the space current, extending the curve upward until a new temperature saturation point is reached. We find, therefore, that the cathode heating current which will give stable tube operation, i.e., the value giving saturation, depends on the plate voltage, and any change in plate potential will affect the stability of the tube unless a corresponding change is made in the cathode heating current. There is always a practical limit, however, to the cathode heating current that may be used, because the cathode will either burn out or its life be greatly shortened if the current rises above a certain value.

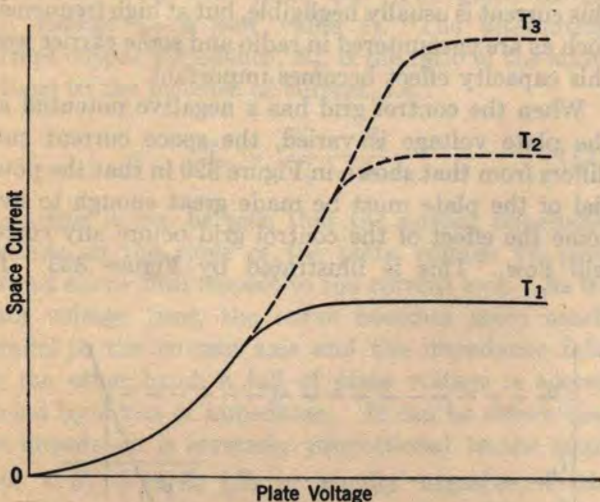


FIG. 320. PLATE VOLTAGE VS. PLATE CURRENT CHARACTERISTIC

Figure 320 is a curve showing the relation between space current and plate voltage for three different values of cathode temperature. Here we find that as the plate voltage is increased from zero, there is an increase in the space current until a saturation point for the given temperature is reached. The failure of the space current to continue its increase with increasing plate voltage is now due to the fact that the cathode is emitting the maximum number of electrons possible for the particular temperature. If the cathode temperature is increased, the **voltage saturation** point will increase correspondingly, as shown by T_2 and T_3 in the figure. In practice, the vacuum tube is operated with a plate voltage well below the voltage saturation value.

The two-electrode vacuum tube has practical use to some extent in radio receiving, electrical measuring, etc., on account of its rectifying property, i.e., the unidirectional flow of current between the cathode and the plate. If an alternating E.M.F. is substituted for the battery B in Figure 317, a space current varying in value for one-half cycle of voltage, but completely cut off for that half of the cycle which gives the plate a negative charge instead of a positive charge will result. In other words there will be a series of "pulses" of current, always in the same direction. The original tube for such use was called the "Fleming" valve.

155. Three-Electrode Tubes—Triodes

The type of vacuum tube commonly employed in telephone repeaters and similar apparatus, differs from the two-electrode tube in that a third electrode or "control grid" is interposed between the cathode and the plate, as indicated in Figure 321. In this device, the electrons which leave the cathode must pass through the meshes of the control grid to reach the plate. Their passage, therefore, is influenced by any force that may be set up by a charge on this control grid. Due to the relative positions of the control grid and plate with respect to the cathode, a change of potential of the control grid has a greater effect on the space current than an equal change in the potential of the plate. For example, a change of one volt in the potential of the control grid of the 101-F vacuum tube used in telephone repeaters, would have the same effect on the space current as a change of approximately 6.5 volts in the plate potential. The ratio of the change of voltage on the plate to the change in control grid potential producing an equivalent effect, is called the "voltage amplifying factor" of the tube, and is usually designated by the symbol, "μ". Its value depends entirely upon the mechanical design of the tube.

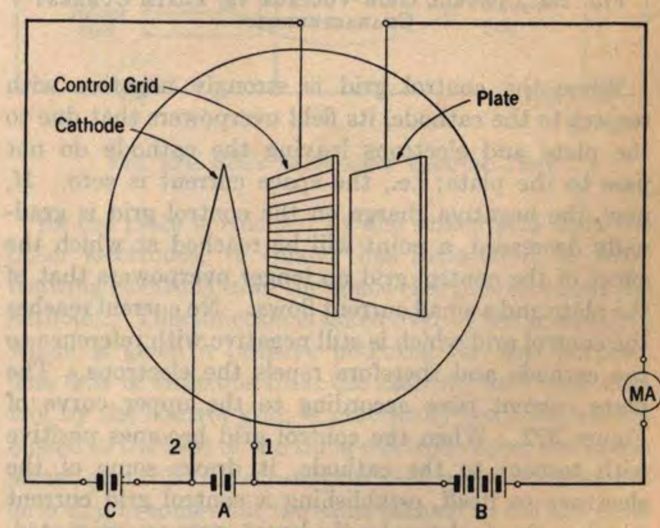


FIG. 321. THREE-ELECTRODE VACUUM TUBE

The utility of the vacuum tube in communication circuits is primarily due to the sensitive response in the plate circuit to small impressed potentials on the control grid. In this connection the control grid in its control over the current in another circuit, is analogous to the valve of a water faucet. It decreases or increases the current in the plate circuit, and the force necessary to thus regulate it is independent of the value of the current or the amount of energy that may exist in the plate circuit. To best illustrate the relation between the control grid voltage and the current in the plate circuit, a curve is employed which is known as the characteristic operating curve of the particular type of tube. Figure 322 illustrates such a curve. Here any voltage that is impressed on the control grid, either positive or negative, is laid off to the right or left of the zero point, respectively, and any vertical ordinate shows the corresponding plate current value.

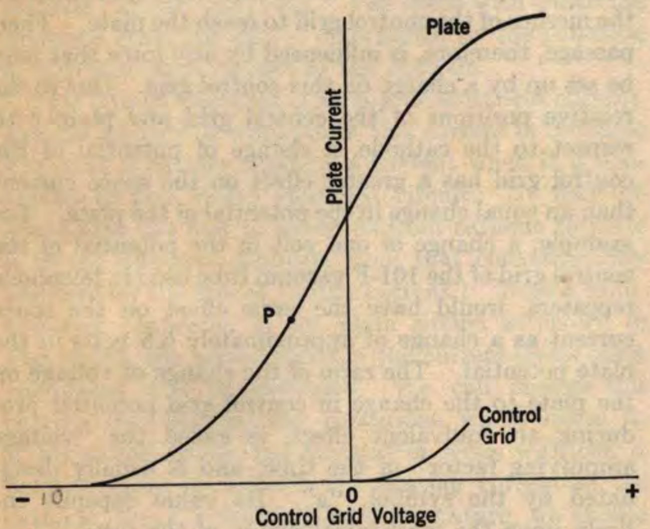


FIG. 322. CONTROL GRID VOLTAGE VS. PLATE CURRENT CHARACTERISTIC

When the control grid is strongly negative with respect to the cathode, its field overpowers that due to the plate and electrons leaving the cathode do not pass to the plate; i.e., the space current is zero. If, now, the negative charge on the control grid is gradually decreased, a point will be reached at which the effect of the control grid no longer overpowers that of the plate and a small current flows. No current reaches the control grid which is still negative with reference to the cathode and therefore repels the electrons. The plate current rises according to the upper curve of Figure 322. When the control grid becomes positive with respect to the cathode, it draws some of the electrons to itself, establishing a control grid current which varies as shown by the lower curve so designated. The sum of the control grid and plate currents is limited by the ability of the cathode to emit electrons; consequently, as the control grid becomes more and more positive, the plate current curve bends towards a horizonal direction at its upper end and may even fall again due to the control grid taking a larger share of electrons. The point at which this flattening takes place depends on the temperature of the cathode, as pointed out in connection with Figure 320. As previously noted, repeater vacuum tubes are usually so worked that the limiting effects of cathode emission are not encountered, but when the activity of the cathode, i.e., its ability to emit electrons, is reduced through age or low cathode current, the effect is manifested by reduced space current.

The circuit between the control grid and cathode is substantially open at telephone frequencies when the control grid is negative since no current flows through the space by transfer of electrons, but a very small charging current flows due to the electrostatic capacity between these electrodes. At telephonic frequencies this current is usually negligible, but at high frequencies, such as are encountered in radio and some carrier work, this capacity effect becomes important.

When the control grid has a negative potential and the plate voltage is varied, the space current curve differs from that shown in Figure 320 in that the potential of the plate must be made great enough to overcome the effect of the control grid before any current will flow. This is illustrated by Figure 323. The

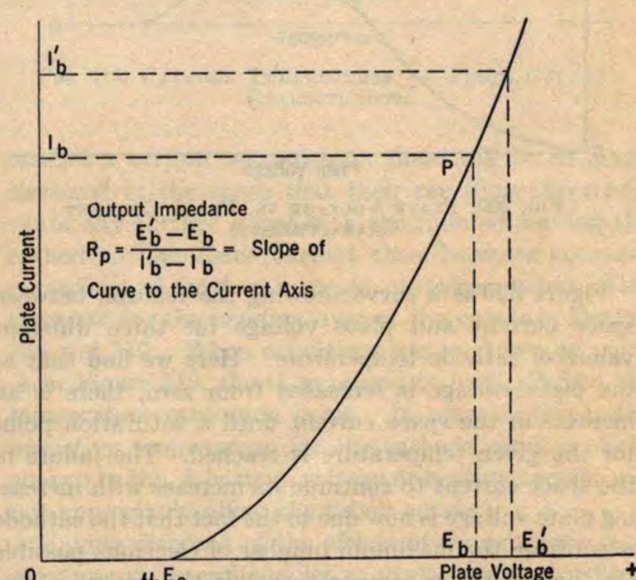


FIG. 323. PLATE VOLTAGE VS. PLATE CURRENT CHARACTERISTIC OF THREE-ELECTRODE TUBE

potential at which current begins to flow is μE_c, where E_c is the voltage of the control grid or C battery. Above this value the current varies as shown by the curve in Figure 323, which is similar to the curve for a two-electrode tube. Some point P on this curve corresponds to the working plate voltage E_b as determined by the plate battery. The corresponding space current is I_b. The direct-current resistance, i.e., the resistance that the tube offers to the direct current from the battery, is given by the expression—

$$R_{d.c.} = \frac{E_b}{I_b} \qquad (150)$$

which is the ordinary form of Ohm's Law. It should be remembered that $R_{d.c.}$ is not constant but varies with both control grid and plate voltages.

The alternating-current output resistance of the tube is quite different from this direct-current resistance and should not be confused with it. The alternating output voltage and current are superimposed on the direct plate voltage and current referred to above. Imagine that the plate voltage (Figure 323) is changed somewhat to a new value E_b'. The plate current will change to a new value I_b'. The alternating-current output impedance, R_p, is the ratio of the added voltage to the increase of current, i.e.:

$$R_p = \frac{E_b' - E_b}{I_b' - I_b} \qquad (151)$$

From this it can be seen that the output impedance depends on the **slope** of the plate voltage vs. plate current curve with respect to the current axis. As the plate voltage rises, the curve becomes more nearly parallel to the current axis and the impedance falls. On the other hand, a fall of plate voltage is accompanied by a rise of impedance. It can be shown that the impedance is inversely proportional to the quantity, $\sqrt{E_b + \mu E_c}$ (E_c is usually negative, so the quantity under the radical is less than E_b). From this fact and the value of the impedance under standard conditions, the value of the output impedance for any values of E_b and E_c may be estimated. The output impedance can also be found from the control grid voltage vs. plate current curve of Figure 322 by taking the slope at the point P, corresponding to the steady control grid potential, with respect to the current axis, and multiplying this by μ, because a change in the control grid potential has the same effect as a change μ times as great in the plate voltage.

In the foregoing the alternating-current resistance of the "space" and the output **impedance** of the tube are considered the same since the impedance does not contain a reactive component. This is permissible if we ignore the capacity between the plate and the control grid, which like the capacity between the cathode and control grid, is negligible at voice frequencies.

156. Four-Electrode Tubes—Tetrodes

At relatively low frequencies the amplification factor, μ, of a three-electrode vacuum tube can be made to have almost any desired value by properly spacing and proportioning the three electrodes. Thus, within reasonable limits, μ can be increased by locating the plate closer to the cathode and decreasing the size of the open spaces in the control grid mesh. However, closer spacing naturally increases the electrostatic capacity between the electrodes. When tubes are used with higher frequencies, such as are encountered in radio and high-frequency carrier systems, the effect of this capacity becomes increasingly important. This is particularly true of the capacity between the plate and control grid circuits, where its coupling effect at the higher frequencies tends to limit the value of the amplifying factor, μ. This plate-control grid capacitance effect can be practically eliminated by the addition of an electrostatic shield between the control grid and plate as illustrated in Figure 324. This shield is in the form of an additional grid, known as a "screen grid", and the four-electrode tube is known as the **screen grid tube.**

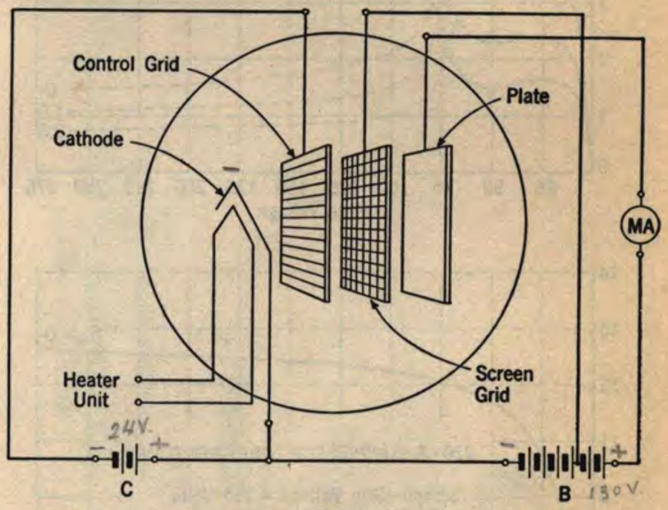


FIG. 324. FOUR-ELECTRODE VACUUM TUBE

As the plate is shielded by the screen grid from the other electrodes, it (plate) has little effect in withdrawing electrons from the space charge area about the cathode. This function is taken over by the screen grid which is given a positive potential for this purpose. The flow of electrons from the cathode, and their control by the control grid, is practically the same as discussed in the case of the three-electrode tube, but in the screen grid tube the screen grid itself may be considered as acting in somewhat the same manner as did the plate of the three-electrode tube. However, the electrons constituting the space current, on arriving in the area of the screen grid have acquired such a velocity that most of them pass through the openings in the screen grid and, attracted by the still higher positive voltage of the plate, continue on to the plate. A small portion of the electrons is, of course, intercepted by the screen grid and does not reach the plate. This is illustrated

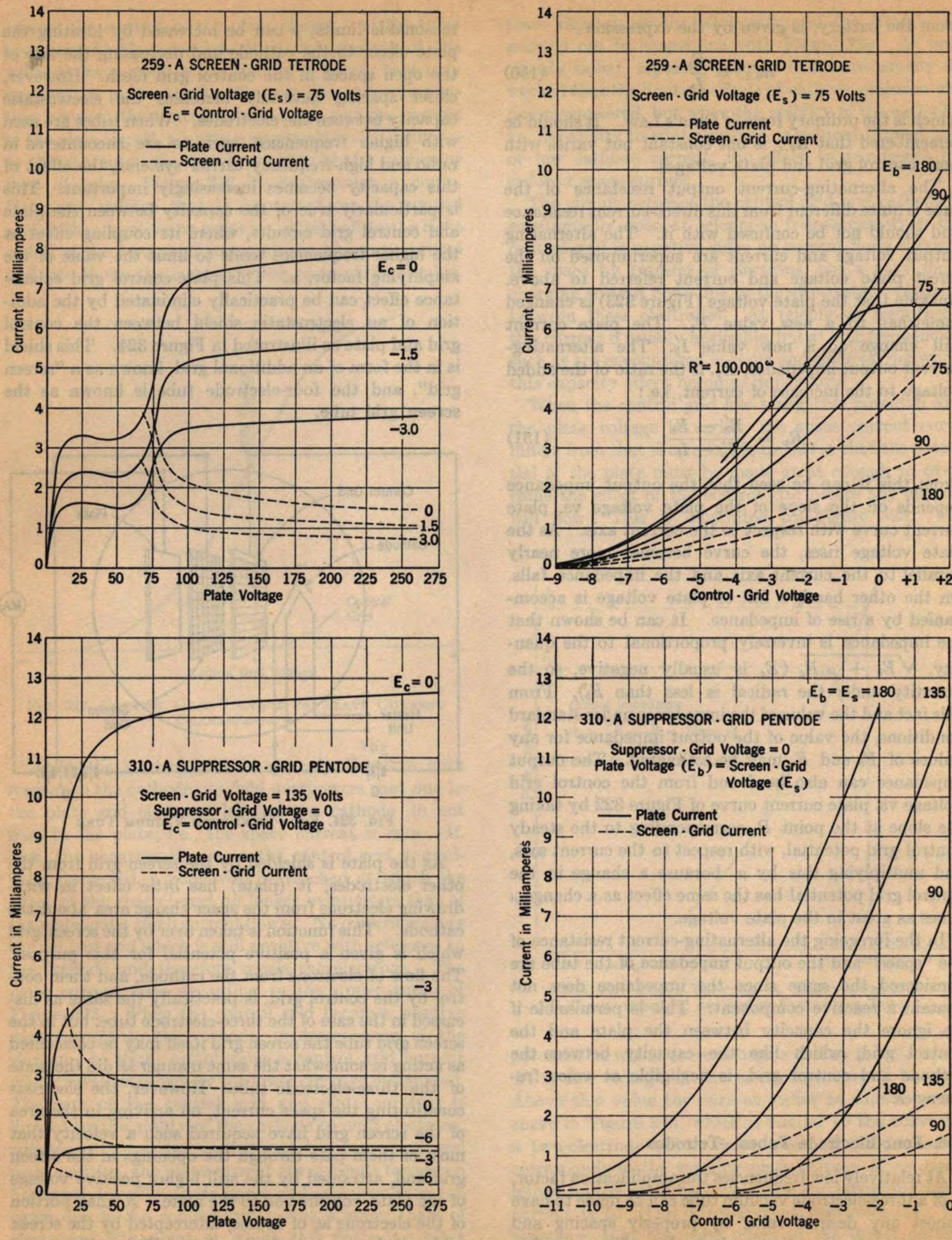


Fig. 325. Characteristics of Standard Four- and Five-Electrode Vacuum Tubes

by the plate voltage vs. plate and screen grid current curves in Figure 325 for a representative Screen-Grid Tetrode.

Here it may be noted that when the plate voltage is decreased to values in the neighborhood of the screen grid potential, the plate current takes a sudden drop while the screen grid current shows a corresponding increase. The sum of these two currents is approximately constant showing that the total number of electrons leaving the cathode remains fairly constant for a given cathode temperature. There are two reasons for the sudden drop in plate current and increase in screen grid current. First, the screen grid, by virtue of its closer location to the cathode, naturally provides a greater attraction for the electrons than a similar potential on the plate, but many of the electrons nevertheless pass through its meshes to the plate. Second, these electrons that pass through the openings of the screen grid may have sufficient velocity when they strike the plate to "knock out secondary electrons", i.e., to force certain electrons to bounce away from the surface of the plate. Some of these secondary electrons may pass over to the screen grid. This condition of secondary emission becomes pronounced if the screen grid is at a higher positive potential than that of the plate but is present to some extent in any case. Thus, there is a space current from the plate to the screen grid which is in the opposite direction to the space current represented by the electrons arriving from the cathode. This secondary current, subtracted from the plate current and added to the screen grid current, accounts in part for the characteristics noted above.

In the working range of plate voltages (150 to 200 volts) for the screen grid tube of Figure 325, it may be noted that the plate current change is very small for a sizeable change in plate voltage. This is another way of saying that the direct-current resistance (Equation 150) and the output impedance (Equation 151) of the tube are very high, which in fact is a desirable characteristic of this type of tube. The output impedance for these tubes is usually in the order of several hundred thousand ohms.

The plate voltage vs. plate and screen grid current curves also indicate the limits of the working output voltage of the tube. The alternating-current voltage at the plate terminal must not be so great that the negative peaks will depress the net instantaneous potential of the plate to a value lower than the potential of the screen grid, as this would cause considerable distortion.

Figure 325 also gives for the same tube the control grid voltage vs. plate and screen grid current characteristics for three different values of plate voltage. The instantaneous net values of the plate current when the tube is connected to a load of 100,000 ohms and the instantaneous applied control grid voltage is varied from -4 to $+2$ volts is indicated by the curve marked "R = 100,000 ohms". This load impedance is a representative value for this tube, as it is the practice to operate screen grid tubes into load impedances which are considerably lower than the output impedance of the tube. The instantaneous potentials on the plate are indicated by the points of intersection of this load characteristic curve with the various static curves for different plate potentials. It may be noted that the load characteristic becomes nearly horizontal for net values of instantaneous plate potential in the neighborhood of and below the potential of the screen grid. The output voltage must be confined to values which will not swing to this part of the curve if distortion is to be avoided.

Radio Transmitting Tubes

The amplification constant, μ, of the screen grid tube is usually in the order of 100 to 600. In the screen grid tube, μ is not as constant as in the case of the three-electrode tubes since its value depends upon the voltages applied to the plate and the two grids (control grid and screen grid). It is, therefore, necessary to know the particular conditions under which the screen grid tube is to operate before the value of μ can be determined.

Another form of the four-electrode tube is that where the additional grid (screen), having a positive potential, is placed between the cathode and control grid. The function of the screen grid in this location is to neutralize partially the negative space charge about the cathode and so facilitate the escape of the electrons from the cathode to the plate. This tube offers advantages where it is desired to operate at low plate voltages. When the electrodes are arranged in this manner the tube is known as the **Space-Charge Grid Tube.**

In still another form of the four-electrode tube, the two grids are so constructed that all of the grid wires are in the same plane, alternate wires being associated with each grid. One grid is used as a control grid to which the input is connected and is operated at a negative potential, while the other is given a positive potential. A greater output power is obtained in this tube for a given dissipation of D.C. power in the plate circuit. Only a moderate plate potential is required as compared with a three-electrode tube. A four-electrode tube of this type is known as a **Coplanar Tube.**

157. Five-Electrode Tubes—Pentodes

In the five-electrode tube, or pentode, there are two auxiliary grids in addition to the usual control grid, plate, and cathode. One of these auxiliary grids is a screen grid having the same function as the screen grid of the tubes discussed above. There are two general types of pentodes, and the location of the second auxiliary grid with respect to the other electrodes determines the classification.

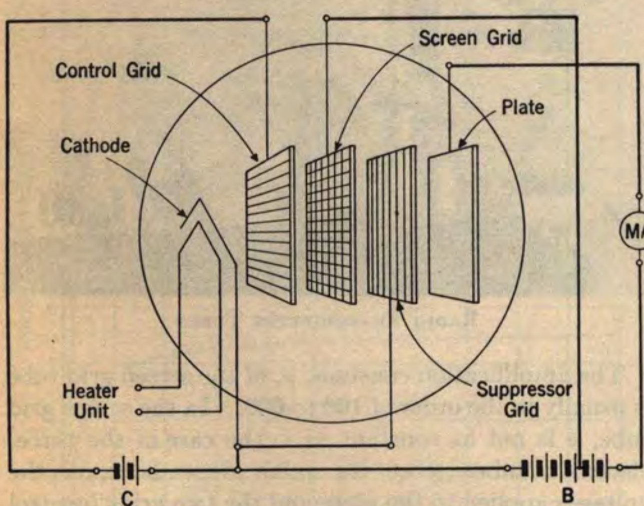
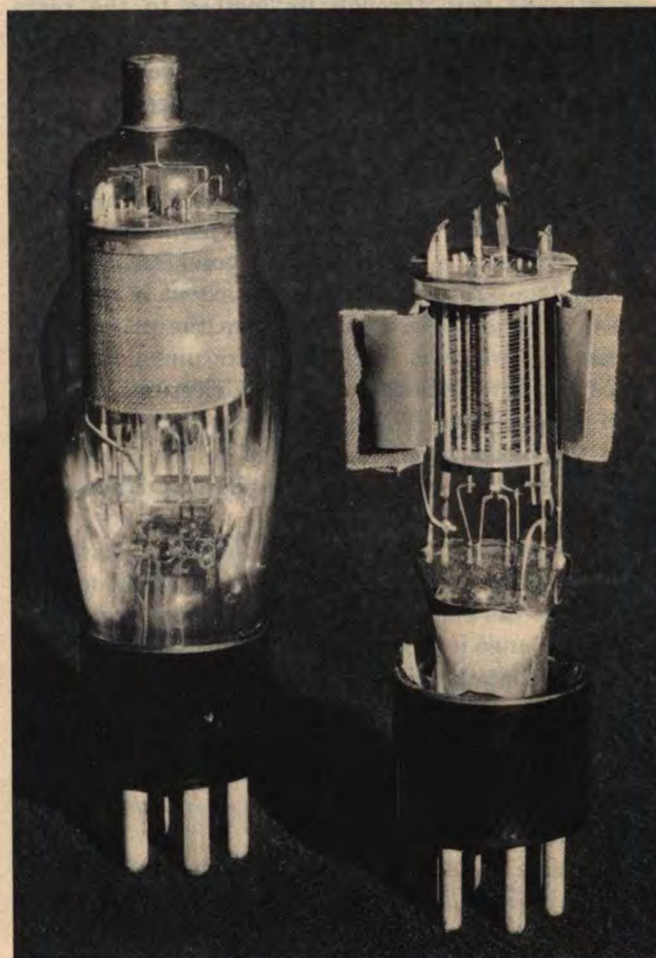


Fig. 326. Five-Electrode Vacuum Tube

When this second auxiliary grid is placed between the screen grid and the plate, as illustrated in Figure 326, and is given a potential at or near that of the cathode, the tube is classified as a **power pentode.** The second auxiliary grid is known as a "suppressor grid" and its purpose is to prevent the low velocity secondary electrons from escaping from the plate. With this arrangement the limitation due to secondary electrons, encountered in the screen grid tube, is removed and practically the full swing of the plate potential can be utilized. This may be seen from the plate and screen grid current vs. plate voltage curves for this type of tube, as shown in Figure 325. This is true even when the plate potential is decreased so as to be equal to that of the screen grid. Both high amplification and large power outputs can be obtained. The plate and screen grid current vs. control grid voltage characteristic curves for a power pentode used generally in high-frequency carrier systems are also shown in Figure 325.



310-A Pentode Vacuum Tube

In the other type of pentode the second auxiliary grid is a space charge grid placed between the cathode and control grid. This tube is known as a **Space Charge Grid Pentode** and is a combination of the space charge grid and screen grid four-electrode tubes previously discussed.

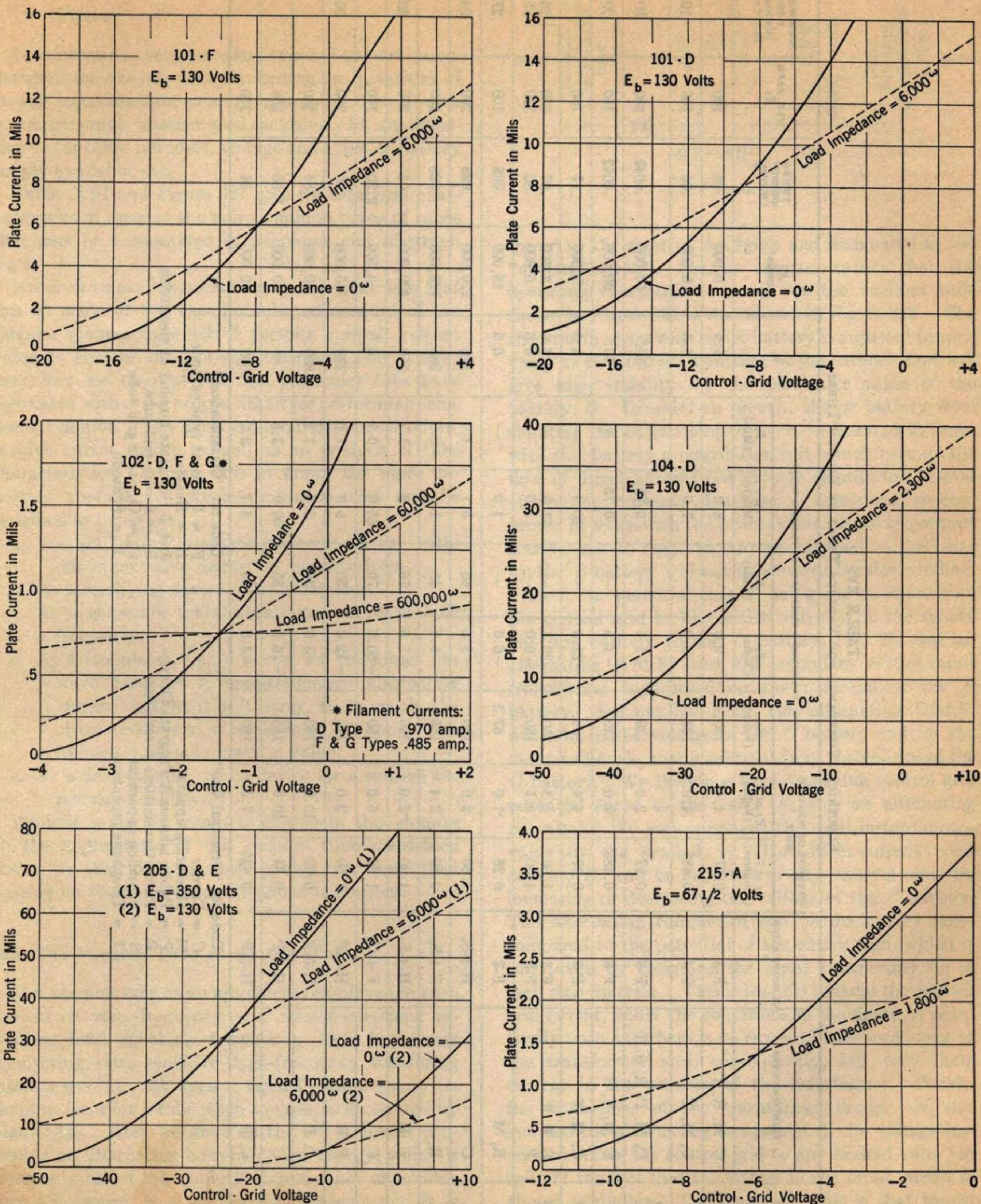


Fig. 327. Characteristics of Standard Three-Electrode Vacuum Tubes

TABLE XVI
Electrical Constants for Various Types of Vacuum Tubes

CODE NO.	PRINCIPAL USES	CLASSIFICATION	CATHODE OR HEATING UNIT		VOLTAGES & CURRENTS OF PLATE & GRIDS					μ	R_p (OHMS)	POWER OUTPUT (WATTS)	MAX. SAFE E_b	MICROPHONIC NOISE (DB BELOW 10^{-6} WATTS)
			I_f or I_h (Amperes)	E_f or E_h (Volts)	E_b (Volts)	E_c (Volts)	E_s (Volts)	I_b (Mils)	I_s (Mils)					
101-D	C, M, R, T	F-3	0.97	4.4	130.	−9.0	—	8.0	—	5.9	6,000	.06	160.	
101-F	C, M, T	F-3	0.485	4.0	130.	−8.0	—	7.0	—	6.5	6,000	.06	160.	19
102-D or G	C, M, P, R, S, T	F-3	0.97	2.0	130.	−1.5	—	0.75	—	30.	60,000	.0042	160.	30
102-F	C, M, T, S	F-3	0.485	2.0	130.	−1.5	—	0.75	—	30.	60,000	.0042	160.	20
104-D	C, M, T	F-3	0.97	4.4	130.	−22.5	—	20.0	—	2.5	2,300	.17	160.	
205-D	C, P, R	F-3	1.60	4.5	350.	−22.5	—	33.0	—	7.0	3,500	.07	350.	25
205-E	P	F-3	1.60	4.5	130.	−9.0	—	5.0	—	6.8	6,600	.89	350.	25
215-A	R, M	F-3	0.25	1.0	67.5	−6.0	—	1.0	—	6.0	18,000	.008	100.	27
245-A	C	H-4-S_c	1.60	2.0	180.	−1.5	45.	5.1	1.4	170.	220,000	.025	200.	29
246-A	R	F-4-S_c	0.10	3.4	135.	−1.5	45.	1.5	0.2	285.	725,000	.028	180.	
259-A	C, R	H-4-S_c	1.60	2.0	180.	−1.5	75.	5.5	1.2	550.	400,000	.05	200.	20
281-A	C	F-4-C_o	1.60	5.0	150.	−60.0	70.	35.0	0.7	5.0	3,500	2.20	170.	
285-A	P, R	H-5-P	1.60	2.0	180.	−12.0	150.	9.0	1.6	130.	145,000	.65	200.	20
310-A	C, P	H-5-P	0.32	10.0	180.	−3.0	135.	5.4	1.2	1800.	40,000	.34	250.	
311-A	C	H-5-P	0.64	10.0	155.	−16.0	155.	40.0	3.5	85.	30,000	1.80	250.	
328-A	C, P	H-5-P	0.425	7.5	180.	−3.0	135.	5.4	2.2	1800.	40,000	.34	250.	

Code for "Principal Use"

C = Carrier
M = Measuring apparatus
P = Public address & Program Pickup
R = Radio transmitters & receivers
S = Signaling apparatus
T = Telephone repeaters

Code for "Classification"

F = Filament type cathode
H = Heater " "
3, 4, 5 Designate number of electrodes
S_c = Screen grid tube
C_o = Coplanar grid tube
P = Pentode

158. Characteristics of Various Standard Vacuum Tubes

Commercially there are many types of vacuum tubes designed for various uses, depending on quantities of energy to be handled, amount of amplification desired in a single stage, whether used exclusively for amplification or for other purposes, voltage and types of battery supply available, etc.

Table XVI and Figure 327 give the electrical characteristics of some of the more common types of tubes that may be encountered in telephone and telegraph work.

Needless to say, there are many practical uses that can be made of the vacuum tube on account of the singular manner in which it permits a small voltage value to control an appreciable current flow. However, any use requires that the particular tube have associated with it the proper circuit for performing some useful function. The practical applications, therefore, involve circuit theory as well as an analysis of tube characteristics. Bearing this in mind, the more important applications in communication work may be classified as follows:

a. For amplifying alternating-current energy without appreciable distortion in wave form.
b. As a rectifying device or as a detector.
c. As a generator for alternating currents of high frequency.
d. As a modulator, i.e., a device for "molding" the wave form of a high-frequency alternating current so that it will carry, so to speak, the characteristics of a wave form of some other frequency, usually lower in value.
e. As a demodulator, i.e., a device for a process the reverse of *d* above.

While it is not the intent to discuss in this chapter all the applications of the vacuum tube mentioned above, we shall take up those having the most direct bearing on the more usual telephone operations.

159. The Vacuum Tube as an Amplifier

The vacuum tube as an amplifier is employed in such devices as voice-frequency and carrier-telephone repeaters, loud speaking telephones, high impedance monitoring sets, sensitive high-frequency measuring instruments, etc. A typical circuit connection in its simplest form for a tube when so used, is illustrated by Figure 328. Here we have on the left a circuit containing an alternating current of low energy, which is commonly called the "input" circuit. Let us assume that this energy is a feeble voice current, and it is desired to amplify it many times, and reproduce it without appreciable distortion in another circuit, which is

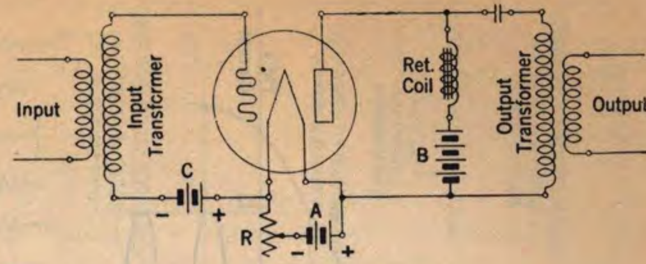


FIGURE 328

shown at the right of the figure and designated as the "output" circuit. Let us assume further that the operating characteristic curve for the vacuum tube shown in Figure 328 is that shown in Figure 329. The rheostat in series with the *A* battery is adjusted to such value as to establish a current in the cathode that will give high stability for the particular value of the battery *B*. In order to prevent the *B* battery from shunting the output circuit, we have inserted in series with this battery a retardation coil, which permits the flow of direct current but greatly retards the flow of alternating current. Likewise a condenser is connected in series with the primary winding of the output transformer so that the current supplied to the plate by the *B* battery will not be shunted by this winding.

Now, the potential that we are going to impress on the control grid in Figure 328 will not be the steady one due entirely to the *C* battery, i.e., E_c, but an alternating E.M.F. from the secondary of the input transformer superposed on the potential of the *C* battery. For one half-cycle, this alternating E.M.F. will add to the voltage of the *C* battery and for the other half-cycle, it will subtract from the voltage of the *C* battery. We therefore have a varying control grid potential equal to the value E_c plus an alternating potential. We may represent the total instantaneous value by the symbol, e_c. The direct-current component, E_c, can be adjusted to any value desired by increasing or decreasing the voltage of the *C* battery. The alternating component can be made very large compared to the potential of the circuit from which it was taken by designing the input transformer for a high step-up ratio. This is feasible because the control grid circuit, under the conditions in the figure, is practically open and there is no current in the secondary of the transformer and, consequently, but very little current in the primary of the transformer. Within the limitations of the transformer design, we can increase the alternating component of the voltage impressed across the control grid to any desired value, in spite of the fact that the energy in the input circuit is almost negligible. The reason for this is that we do not theoretically "draw from" this energy because if we did, we should require control grid current as well

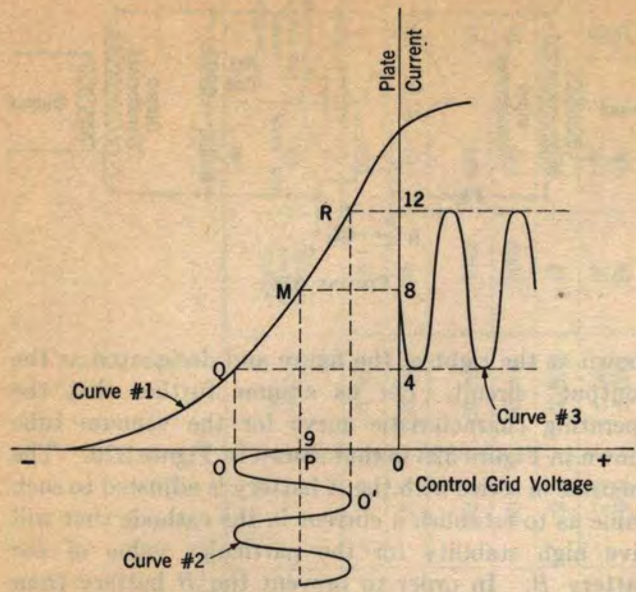


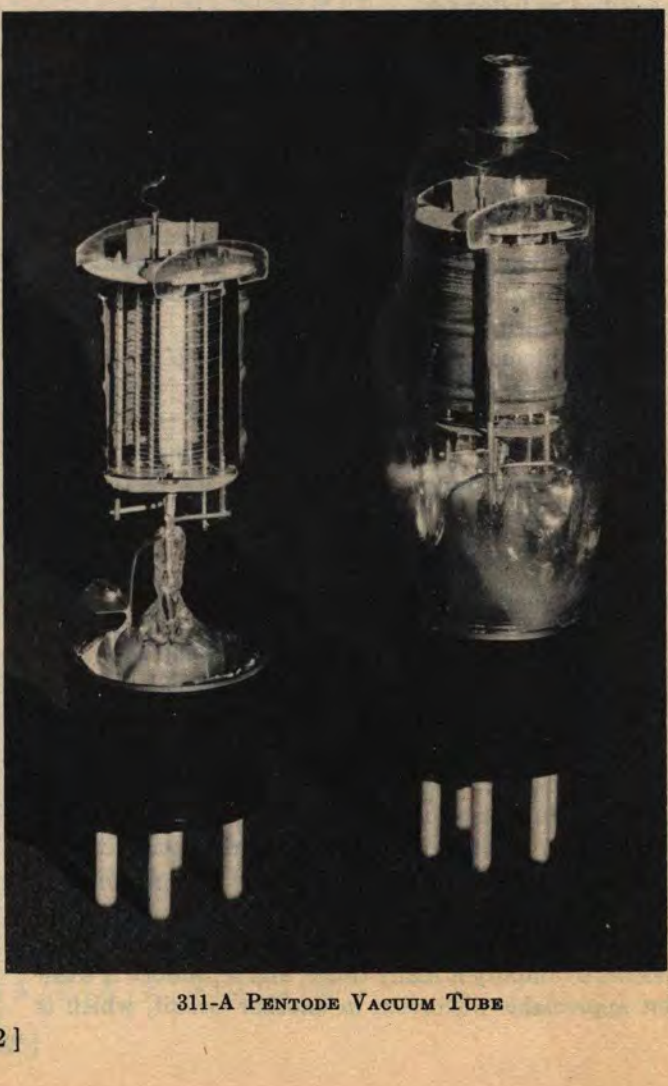
Fig. 329. Vacuum Tube as Amplifier

as control grid voltage. But with no current, or with an open circuit, any value of voltage we might name would represent zero energy.

Referring again to Figure 329, let us assume that we have an appreciable alternating E.M.F. impressed on the control grid, which tends to increase and decrease the voltage of the control grid alternately for each cycle. This we can represent by curve #2, which for convenience is charted downward. Now let us follow this curve beginning at the point *P*. Here we have a control grid E.M.F. created by the *C* battery alone, which is 9 volts in value and which fixes the value of the current in the plate circuit at 8 milliamperes. This we find to be the case by projecting upward to the characteristic curve, point *M*, and projecting across to the plate current scale. With 9 volts fixed control grid potential, we have 8 milliamperes fixed plate current, but now when the first half-cycle of the alternating component, beginning at point *P* and reaching a peak at point *O*, is added to the *C* battery voltage and projected upward to the characteristic curve, we have the point *Q* which corresponds to a plate current of 4 milliamperes. Now, going from the point *O* to the point *O′*, which is the peak of the other half-cycle, and projecting from the point *O′* to the characteristic curve, we have the point *R* which corresponds to 12 milliamperes. The value of the plate current is changing, therefore, as determined by the factor μe_c, and we have in the plate circuit an alternating component of current, in the same way that we have in the control grid circuit an alternating component of voltage impressed on the direct voltage of the *C* battery. In the plate circuit, this alternating component of current is not permitted to flow through the battery on account of the retardation coil in series with the battery, but is forced to flow through the primary of the output transformer.

Referring again to our characteristic curve, if the portion of it between the points *Q* and *R* had been a straight line, every point in curve #3 measured from a neutral axis would be proportional to corresponding points on curve #2 measured from its neutral axis, and we could say that curve #3 was identical in wave form to curve #2. To illustrate, if curve #2 were a sine wave, curve #3 would be a sine wave; if curve #2 were a complex wave representing some vowel of the voice, curve #3 would be a complex wave representing the same vowel of the voice.

In the above action we have accomplished considerable amplification of energy. The current in the input circuit was very feeble, being merely that required to maintain magnetization of the transformer. In the output circuit, on the other hand, we had a current of several milliamperes, which represents a large amplification of the energy impressed on the input circuit. Further, the amplified energy had the same frequency



311-A Pentode Vacuum Tube

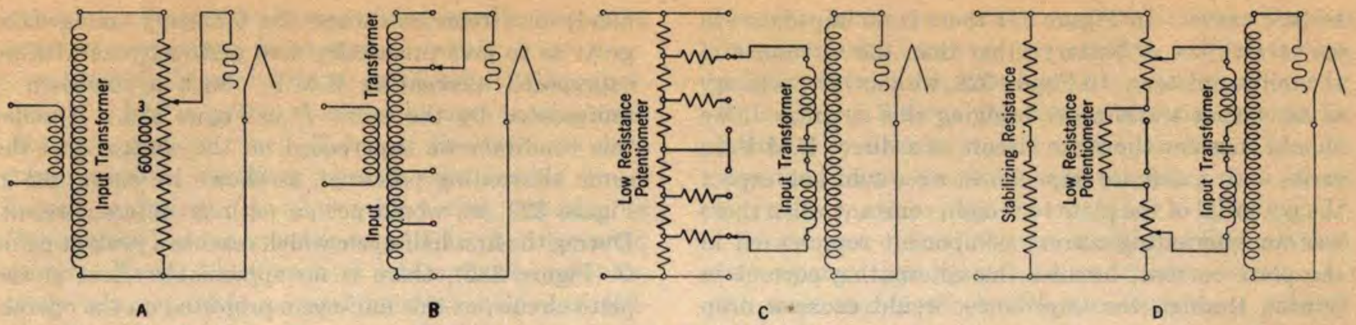


Fig. 330. Methods for Varying Amplifier Gain

and wave form as the input energy. The amount of amplification for a device of this kind depends on three factors: first, the slope of the straight line portion of the characteristic curve, or the amplification constant of the tube, which is the same thing expressed in other terms; second, the voltage step-up ratio of the input transformer; and third, the losses in the circuit, which must, of course, be subtracted.

There are a number of additional circuit features not shown in Figure 328 that are required to meet conditions in practice. One of the most important of these is the arrangement for adjusting the amount of amplification or "gain". There are five practical methods of doing this, one of which is to place some form of network in the output circuit to absorb some portion of the amplified energy. The other four are all schemes for regulating the potential impressed across the control grid, and these are the ones most generally used.

Figure 330 illustrates the four devices. Figure 330-A is the oldest scheme and employs a very high resistance potentiometer (600,000ω) between the secondary winding of the input transformer and the control grid. This, of course, draws a certain amount of current from the secondary of the transformer, which represents a certain amount of energy supplied to the input circuit. A more common practice is to employ a transformer having numerous taps on the secondary winding. This arrangement is illustrated in Figure 330-B. There is also in general use a gain regulating device which consists of a potentiometer on the primary side of the input coil, as represented by Figure 330-C. Here the potentiometer has much lower resistance inasmuch as it is on the low side of the transformer. It requires duplicate contacts, however, in order not to throw an unbalance on the connecting line due to the lack of symmetry in the circuit. The circuit illustrated in Figure 330-D is similar to that of C and is the arrangement used in 22-type telephone repeaters. The impedance presented by the parallel combination of the stabilizing resistance and low resistance potentiometer is usually made equal to the characteristic impedance of the circuit to which it is connected.

Having a picture of the circumstances under which the tube operates in the ordinary circuit, we may now deal with certain adjustments that must be made in the values of E_b and E_c and in the characteristic curve between these values, i.e., the control grid voltage vs. plate current curve which we discussed in a preceding article and which is represented by Figure 322. In the first place, the battery A, although intended primarily to heat the cathode, affects the values E_b and E_c. This can be understood by referring again to Figure 321. Here we have represented the voltage between the plate and cathode, E_b, by the battery B, but this E.M.F. is impressed between point 1 of the cathode and the plate, while the E.M.F. impressed between point 2 of the cathode and the plate is equal to the voltage of the battery B plus the I_aR drop due to the current the battery A furnishes through the resistance of the cathode. The average value for E_b, then, is the voltage between the plate and the middle of the cathode, which is equal to the voltage of the battery B plus one-half of the I_aR drop in the cathode.

In the same way that the I_aR drop may slightly affect the voltage of the B battery, it may more appreciably affect the voltage of the C battery since this is usually small. It should be remembered, therefore, that while the function of the A battery is primarily to heat the cathode, it tends to increase the effective value of both the B and C batteries when connected as shown in Figure 321 and would decrease their effective values if connected with its polarity reversed. For reasons of economy it should always be connected as shown, thereby permitting the use of B and C batteries of less voltage and consequently less cost. Ordinarily, in plotting characteristic curves of vacuum tubes, it is understood that the A battery is poled so as to add to a negative control grid and positive plate, and the characteristic curves in this chapter are plotted on that basis. This permits the use of actual voltage values as ordinates, instead of corrected values.

Another very important consideration coming from the actual conditions under which the tube is operated, is the effect of external plate circuit impedance on the $E_c - I_b$, or control grid voltage vs. plate current characteristic curve. In Figure 321 there is no impedance in series with the B battery other than the resistance of the milliammeter. In Figure 328, we show the primary of an output transformer bridging this circuit. If we should consider the plate circuit as a direct E.M.F. in series with a definite impedance, we would not expect the potential of the plate to remain constant when there was an alternating-current component represented in the plate current, because this alternating current in flowing through the impedance, would cause a drop which for the instant would seriously affect the plate voltage. It is, therefore, necessary to take this into consideration in the characteristic curve, and the effect of doing so is to flatten the curve as illustrated by Figure 331. The dotted curves in Figure 327 are corrected operating characteristics for external impedance conditions as given.

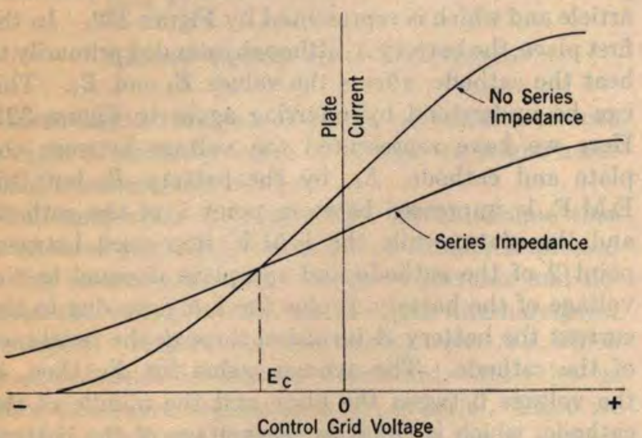


FIG. 331. EFFECT OF OUTPUT IMPEDANCE ON TUBE CHARACTERISTICS

The same general principles as discussed above for the three-electrode tube, when used as an amplifier, apply generally to the higher gain tetrodes and pentodes. The voltage applied to the screen grid is usually obtained from a tap on the battery which supplies the plate voltage. The control grid biasing voltage may be obtained from a separate battery tap or from the potential drop across a resistance inserted in the cathode circuit.

160. The Vacuum Tube as a Rectifying Device

We have seen that for amplification without distortion, the straight line portion of the characteristic curve must be employed, as shown in Figure 329. If a curved portion of the curve were employed, distortion would result. Within certain limits this can be controlled by the C battery which in Figure 329 had a value of 9 volts. This was a case of the chosen value restricting the amplifying operation to the straight line portion of curve #1. Let us consider, on the other hand, an extreme case where the C battery voltage is so great as to give practically zero plate current with no superposed alternating E.M.F. Such a condition is represented by the point P in Figure 332. If under this condition we superposed on the control grid the same alternating potential, as shown by curve #2 in Figure 329, we would get an entirely different result. During the first half-cycle which reaches a peak at point O (Figure 332), there is no appreciable effect in the plate circuit, as this half-cycle projected on the operating curve, falls on the zero line. The other half-cycle, however, subtracts from the E_c value and projects on a portion of the characteristic curve which has appreciable slope, though somewhat curved. This establishes a plate current in the form of unidirectional pulses for each half-cycle of the impressed E.M.F. that subtracts from the C battery voltage. The tube's response in this case is a rectifying action that is similar to that for which the two-electrode tube is sometimes employed, but there is a certain amount of amplification at the same time, which is not given by the two-electrode tube.

Although in Figure 332 we have chosen a value of E_c that gives very nearly zero value for I_b, there would be some rectifying effect on any curved portion of the characteristic curve. It is, therefore important that operation for amplification be restricted to the straight line portion, as any degree of rectification will distort the wave form and thereby impair the quality.

Three-electrode vacuum tubes so operated are widely used as detectors in radio receiving circuits and as modulators and demodulators in radio and carrier systems. They are also used in various types of

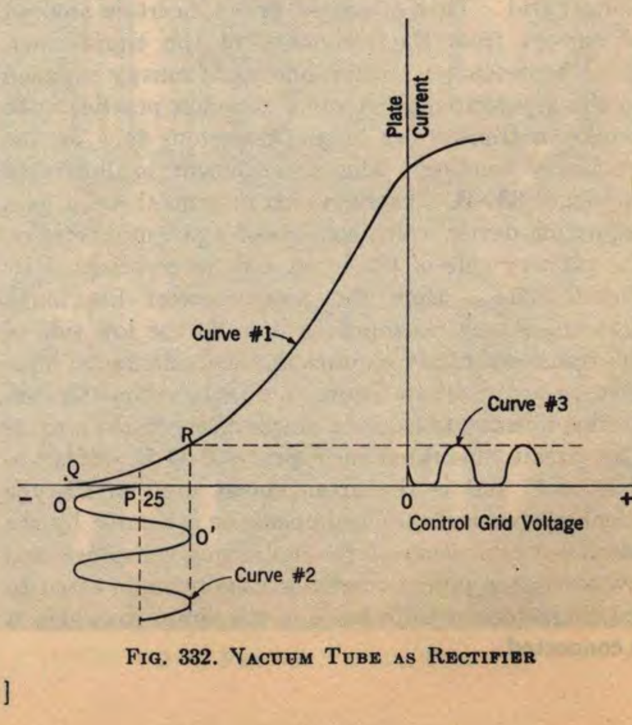


FIG. 332. VACUUM TUBE AS RECTIFIER

measuring apparatus for rectifying small alternating currents to operate direct-current meters.

As previously pointed out, rectification of an alternating E.M.F. can be obtained by means of a two-electrode vacuum tube. Tubes of this type are commonly used as rectifiers when a considerable power output is desired. In this respect they function as "converters" of alternating-current to direct-current power. Figure 333 illustrates schematically two-electrode vacuum tubes as used for both half- and full-wave rectification, together with their associated circuits. In the particular tubes illustrated, the cathodes are of the filament type and are heated by a low voltage alternating current.

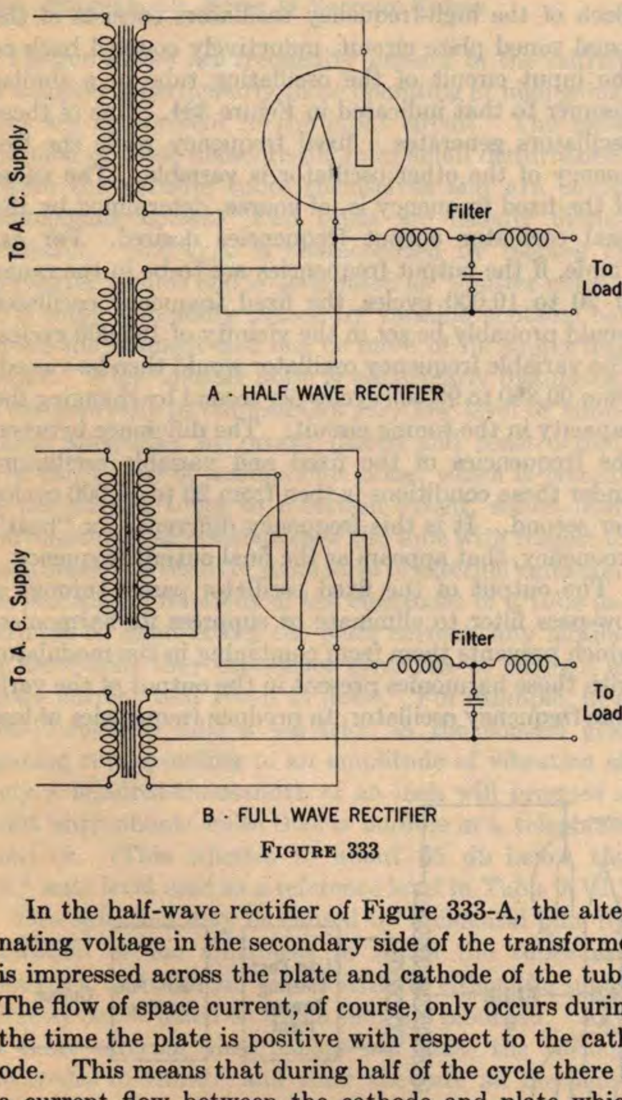


FIGURE 333

In the half-wave rectifier of Figure 333-A, the alternating voltage in the secondary side of the transformer is impressed across the plate and cathode of the tube. The flow of space current, of course, only occurs during the time the plate is positive with respect to the cathode. This means that during half of the cycle there is a current flow between the cathode and plate which gives the effect of closing the circuit or connecting the filter and load to the transformer. During the other half of the cycle (plate is negative with respect to the cathode) there is no space current and the tube may be considered as opening the circuit. The output current from the half-wave rectifier flows in one direction, but its magnitude varies as the positive half of the impressed wave and is therefore of a pulsating character. By adding a filter in the output containing series inductance and shunt capacity, this pulsating current is smoothed out somewhat into a more even direct current.

In the case of the full-wave rectifier of Figure 333-B, two separate plates are used, to which are connected the terminals of the transformer secondary winding. It can be seen that there will always be current between the cathode and one of the plates, because one of the plates will always be positive with respect to the cathode. This means there will be current flowing in the same direction in the output (filter and load) during both halves of the cycle or full-wave. As the full-wave rectifier uses both halves of the cycle, its power output is approximately twice that of the half-wave rectifier, with other factors being equal.

Very efficient rectifiers may be secured by admitting a small amount of certain gases at controlled pressure into the vacuum of the tube. In this case the flow of electrons between the cathode and plate ionizes the gas by the electrons colliding with the gas molecules. The collision between an electron and a gas molecule knocks some electrons out of the molecule, thereby separating it into a positive ion and one or more negative electrons. The electrons, being negative, are attracted to the positive plate, and the ions being positive travel to the cathode. The positive ions neutralize the negative space charge that would otherwise exist near the cathode, thus greatly facilitating the escape of further electrons. The net result is that the opposition to the space current flow is reduced, which permits it (space current) to increase to a value limited only by the external resistance in the plate circuit. In some tubes a mercury gas vapor, supplied by mercury within the tube, is used. Other gases such as argon have also been found to give very satisfactory results.

161. The Vacuum Tube as a Generator (Oscillator)

Any vacuum tube containing a control grid (three or more electrodes) can be used as an alternating-current generator under the following conditions:

a. There must be some connection or coupling, between the output and input circuit whereby a part of the output energy will be fed back into the input circuit.

b. The amount of energy that is fed back from the output to the input must be at least as great as the reciprocal of the energy amplification, e.g., if the circuit amplifies the energy 300 times, at least 1/300 must be fed back into the input circuit.

c. Either the input or output circuit must have adjusted capacity and inductance to establish resonance, thereby determining the frequency generated.

d. The current coming from the output circuit and reaching the input circuit through the feedback connection, should be "synchronized" (added in phase) to the existing current in the input circuit.

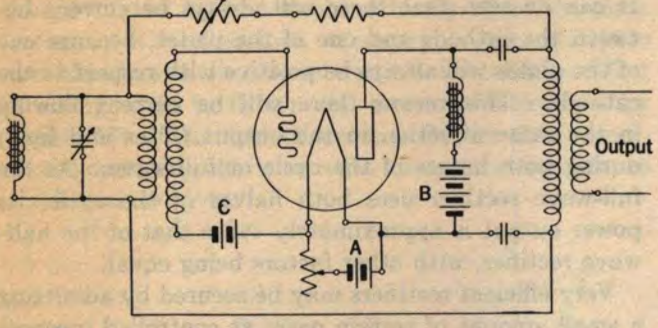


Fig. 334. Vacuum Tube Oscillator Circuit

Figure 334 represents a simple vacuum tube circuit arranged for generating an alternating current. Here we can see that the operation will be maintained indefinitely by input energy being fed back from the output circuit. Such a circuit will gradually build up until the full capacity of the tube is reached. Under these conditions, however, the operation cannot be restricted to the straight portion of the curve and the wave form of the generated current will not be a pure sine wave but will contain harmonics. These are usually minimized by filtering devices and by circuit modifications consisting of IR drops introduced at various places by the insertion of pure resistance. An alternating-current generating device of this kind is called a "vacuum tube oscillator".

There are various types of vacuum tube oscillators and their electrical design is determined by their uses. The "heterodyne" or "beat frequency" type of vacuum tube oscillator, in which the output frequencies from two oscillators operate into the same circuit and the final output frequency is their difference (the so-called "beat" frequency), has been found to be quite applicable in telephone work. This oscillator can be designed to have a fairly wide frequency range that extends to both very low and very high frequencies. Its wave shape is good because the distortion in the output is kept to a minimum. An oscillator of this type is schematically illustrated in Figure 335. It consists essentially of two separate high-frequency oscillators, a balanced modulator, and a "push-pull" amplifier. Each of the high-frequency oscillators consists of the usual tuned plate circuit, inductively coupled back to the input circuit of the oscillating tube in a similar manner to that indicated in Figure 334. One of these oscillators generates a fixed frequency while the frequency of the other oscillator is variable. The value of the fixed frequency is, of course, determined by the final oscillator output frequencies desired. For example, if the output frequencies are to be in the range of 20 to 10,000 cycles, the fixed frequency oscillator would probably be set in the vicinity of 100,000 cycles. The variable frequency oscillator would then be varied, from 99,890 to 90,000 cycles per second by changing the capacity in the tuning circuit. The difference between the frequencies of the fixed and variable oscillators under these conditions is then from 20 to 10,000 cycles per second. It is this frequency difference, or "beat" frequency, that appears as the final output frequency.

The output of the fixed oscillator passes through a low-pass filter to eliminate or suppress its harmonics, which prevents them from combining in the modulator with those harmonics present in the output of the variable frequency oscillator, to produce frequencies of less

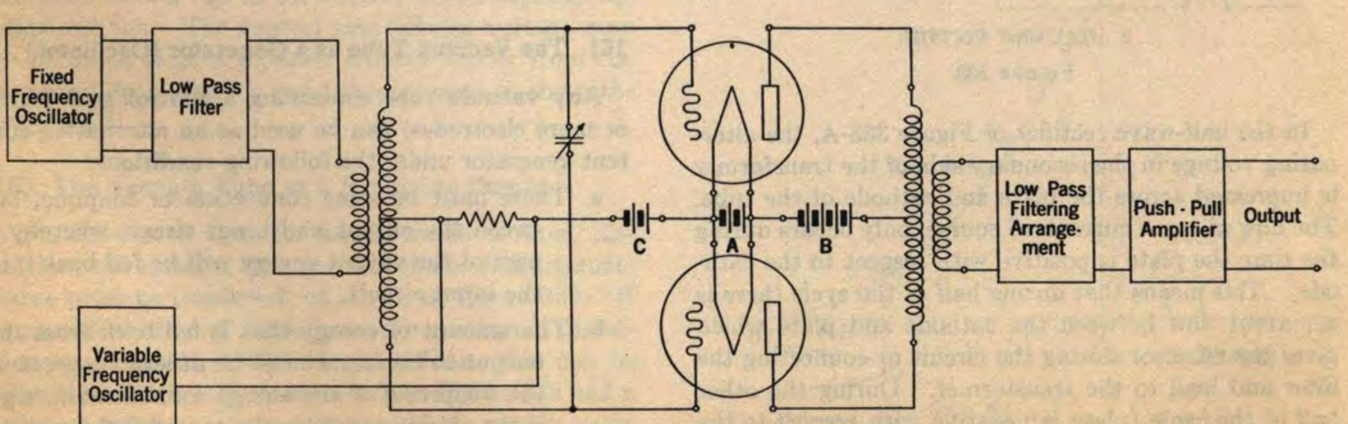


Fig. 335. Beat-Frequency Vacuum Tube Oscillator Circuit

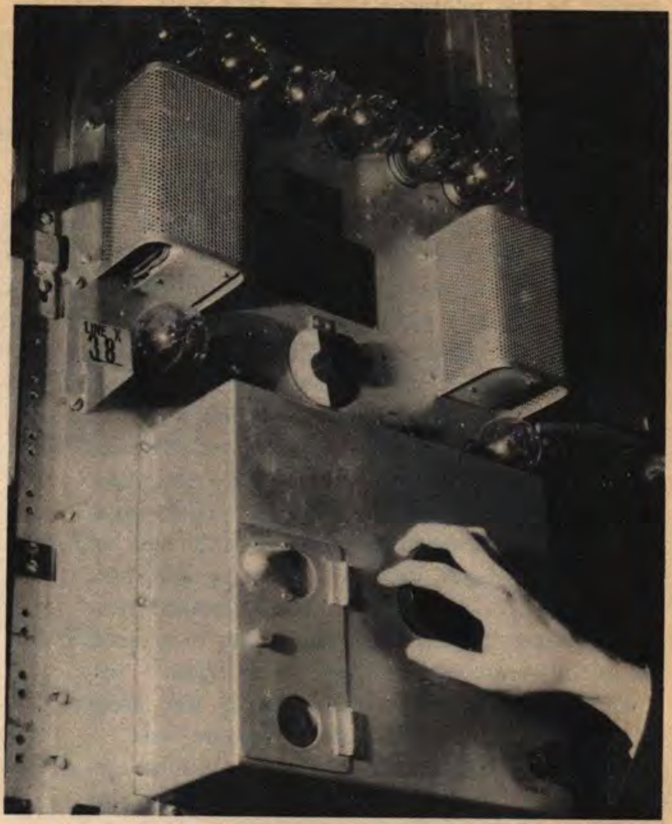
than 100,000 cycles per second. In addition, a low-pass filtering arrangement is inserted between the modulator and the push-pull amplifier which prevents the fundamental frequencies of the two oscillators, and any other high-frequency products of modulation, from passing into the amplifier. Therefore, the input to the amplifier is reasonably free from distortion and the wave shape is practically a pure sine wave for the frequency desired. The amplifier is of the push-pull design (see Article 167) so that the second harmonic inherently generated in each amplifier tube is balanced out, and thus the harmonics in the final output are kept to a minimum.

162. Microphonic Noise in Vacuum Tubes

Certain noises are frequently present in the output of amplifiers, particularly those operating at high gains, that are not present in the input circuit. The most common of these noises result from small disturbances within the vacuum tubes themselves and are called **microphonic noises.** Two types are usually distinguishable. One type is characterized by rasping or sputtering sounds and is designated as "sputter noise". It is most commonly caused by variable contacts between the filament and one or more of its supporting hooks. While sputter noise is particularly disturbing, it can generally be reduced to a very low level or eliminated altogether by proper design and construction. The other type of microphonic noise, which is present in all vacuum tubes to a certain extent, comes from movements of the electrodes of the tube with respect to each other as a result of some form of external agitation. As the spacial relations of the electrodes of a tube determine to some extent the plate current, any motion or movement they experience gives rise to changes in space current that result in noise. For example, in a 102-F vacuum tube a variation in the control grid spacing corresponding to an amplitude of vibration of only a hundred-thousandth of an inch will produce a faint microphonic noise that is audible in a telephone receiver. (This equates to about 55 db below the 10^{-5} watt level used as a reference level in Table XVI.)

This agitation may be caused by accidental jars, by vibration of the apparatus in which the tubes are mounted, or by sound waves striking either the tubes directly or the panel on which they are mounted. Any of these external disturbances will cause the various electrodes to vibrate, and since they are all mechanically coupled by means of their supports, a large number of vibration peaks are possible.

The microphonic noise output of a vacuum tube will, in general, include a wide range of frequencies depending on the various peaks of vibration of the electrodes of the tube. The disturbance to the listener will, on the other hand, depend on the characteristics of the amplifier and the ear. Certain frequency ranges will contribute much more to the general level of the disturbance than others of equal intensity.



13-A Beat-Frequency Variable Oscillator—20 to 10,000 Cycles

With the increasing requirements in vacuum tube performance, it becomes necessary to reduce as far as practicable these internal effects which result in microphonic noises. This is especially true where amplifiers are used on high grade program circuits. Obviously, there must be some means provided to measure the microphonic noise output of a vacuum tube for a given amount of external agitation. Such a testing arrangement is used by the manufacturer in checking the noise characteristics of the various tubes manufactured. The amount of microphonic noise a vacuum tube produces is, of course, determined to a certain extent by the severity of the external agitation to which it is subjected. Therefore, there must be some standard agitation arrangement for the tubes tested before any weight can be given the amount of microphonic noise measured, and then the measurements are primarily for comparison. It is, however, a simple matter to establish a limit for the amount of permissible microphonic noise in various types of amplifiers for a given amount of agitation. By using a fixed reference point this noise can be measured in terms of decibels with

respect to the reference level of 10^{-5} watt. Since no ordinary vacuum tube produces a microphonic noise output as high as this level, for the given standard amount of agitation used in the manufacturer's test, the measurements are in db below this reference level. The agitation the vacuum tube is subjected to in this test is, however, much more severe than would ordinarily occur in service. This mechanical agitation would have to be reduced in the order of 40 db to duplicate approximately the maximum agitation that a vacuum tube is likely to be subjected to in a telephone repeater station.

CHAPTER XXVI

TELEPHONE REPEATERS AND AMPLIFIER CIRCUITS

163. Uses of Telephone Repeaters and Amplifiers

For satisfactory telephone communication, there must be sufficient energy transmitted over the line to provide adequate sound volume at the receiving end. The loss of a 19-gage H-44 side cable circuit 1000 miles long is approximately 480 db. If we attempted to transmit energy over such a circuit without any means of boosting the transmitted power along the line, an input power of one milliwatt at the sending end would be attenuated to 10^{-51} watt at the receiving end. This reduction is so great that if one milliampere of current, which is 6.28×10^{15} electrons per second, was sent into this line, the current at the receiving end would be approximately one electron every four or five years. If we were to attempt to increase the power received to a value equal to that of the power sent (one milliwatt) by means of a single amplifier inserted anywhere in the circuit, we would have to use a device capable of amplifying power by 10^{48}. Such an amplifier is, of course, a practical absurdity. Neither would it be possible to accomplish the desired results if only two or three amplifiers were inserted. However, by placing amplifiers at more frequent intervals—usually about 50 miles apart on cable circuits—the power may be increased by each amplifier in steps of practicable size.

From the above it may be seen that telephone repeaters, or amplifiers, are an essential factor in our present system of long distance telephone communication. It is possible, of course, to talk over considerable distances without their use if large conductors and heavily loaded lines are employed. However, the use of repeaters and less expensive lines, even over comparatively short distances, is favorable on an overall economy basis. Vacuum tubes, which were covered in the preceding chapter, are universally used as the amplifying elements of these repeaters. Since the vacuum tube is a one-way device, it can operate in only one direction of transmission. For two-way transmission, therefore, it is necessary either to use 4-wire circuits with a pair of conductors carrying the conversation in each direction, or to use 2-wire circuits with a double amplifying, or repeating, device which operates independently in both directions of transmission.

164. 22-Type Repeaters

The type of amplifying arrangement commonly used in 2-wire voice-frequency telephone circuits is the "22-type" telephone repeater. Here the significance of "22" is "two-element, two-way", meaning that there are two distinct one-way amplifiers employed, and that the repeater is arranged for use with the ordinary two-way telephone circuit. In general, 2-wire circuit telephone repeater practices involve the following:

a. The use of one-way amplifier circuits designed to give the required amplification or transmission gain and equipped with regulating devices for adjusting the gain to meet operating conditions.

b. The use of special transformers, called hybrid coils, for adapting one-way amplifiers to two-way transmission.

c. Provision of proper network balancing equipment for closely approximating the impedances of each line circuit and its associated apparatus in the frequency band transmitted, thereby maintaining the degree of balance required by the hybrid coil for its proper functioning. Here proper functioning means that energy at voice-current frequencies from the output of one amplifier must be prevented from reaching the input of the other, which would cause impairment in the quality of transmission or even "singing", as was explained in Article 117.

d. The use of filters for eliminating any energy not at the frequencies essential for the required quality in the voice transmission.

e. The use of miscellaneous apparatus and circuit features for adapting the telephone repeater circuit to the standard operating practices.

Figure 336 illustrates a simplified 22-type telephone repeater with all the features outlined above. Its operation is briefly as follows:

We may assume that the subscriber at the east end of the connection is talking and that the greatly attenuated voice current from his station reaches the telephone repeater circuit at the hybrid coil associated with the east side of the repeater. Half of this energy is transmitted from points 3T and 8T of the coil through the pad, which will further attenuate the current depending upon the relative values of X and Y (see Table XV of Chapter XXIII), to the East-West potentiometer which is bridged across the input circuit. The adjustable element of this potentiometer consists of a double slide-wire with a 200-ohm resistance inserted between its halves. A shunt resistance of 1212 ohms, grounded at its mid-point, is placed ahead of the

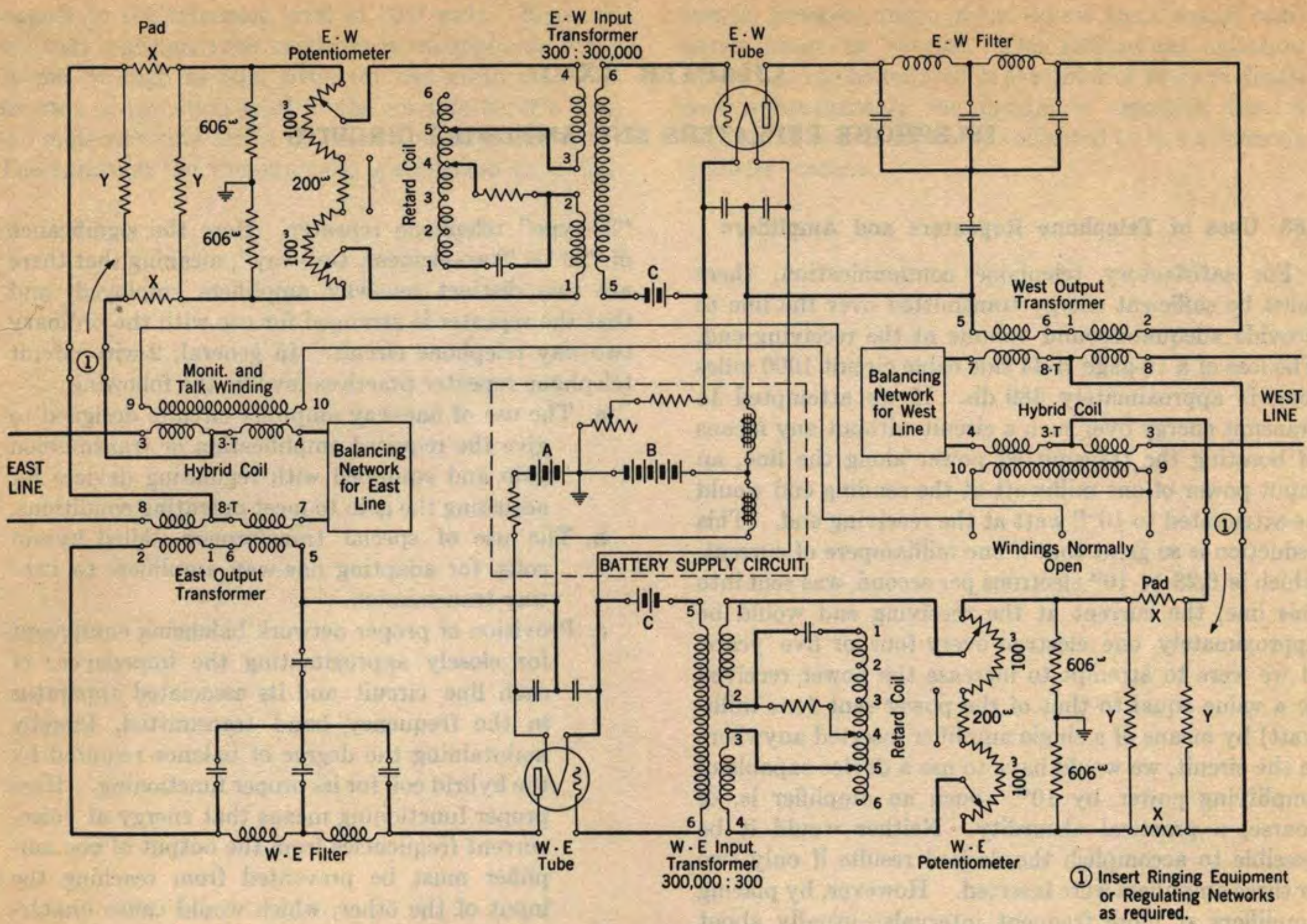


FIG. 336. 22-TYPE TELEPHONE REPEATER CIRCUIT

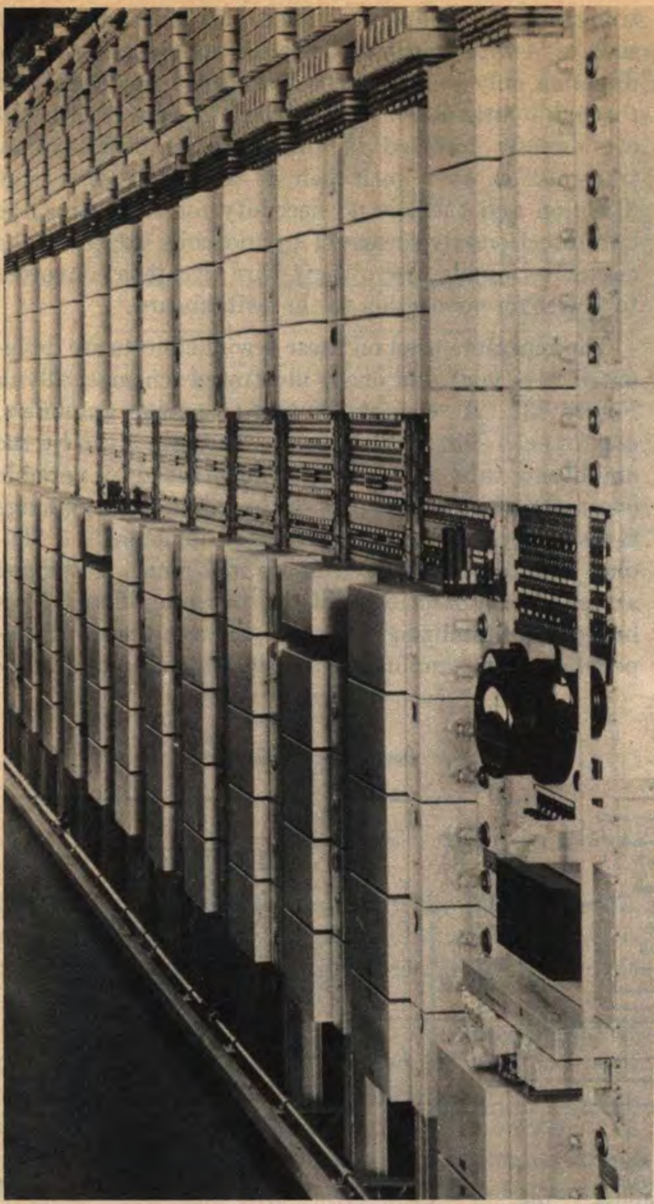
potentiometer, thus giving the circuit an impedance of approximately 300 ohms as seen from the hybrid coil side. The internal balance of the repeater is improved by grounding the shunt resistance at its mid-point. By means of the double slide-wire contacts, which are used in part to control the gain of the amplifier, from half to the full voltage drop across the potentiometer is "picked off" and impressed across the low impedance winding of the input transformer. In the event half of this voltage drop is greater than that required to provide the desired overall gain, a pad having a greater loss may be inserted in the input circuit. This picked off voltage is stepped up some thirty times by the transformer and impressed on the control grid of the vacuum tube, where it acts to control the current in the plate circuit. The plate circuit energy, which while having the same characteristics, is of much greater magnitude than that reaching the potentiometer on the input side, passes through a filter to windings 2–5 of the west output transformer (hybrid coil).

As we learned in our study of the theory of the hybrid coil (Article 117), half of this energy is transmitted to the west line while the other half is lost in the balancing network. If this network balances the line exactly, no part of the energy reaches the input of the West-East amplifying circuit. On the other hand, if the balance is not perfect, a part of the energy proportional to the degree of unbalance, will "cross" the hybrid coil and be amplified and returned to the east line. Furthermore, if there is also some unbalance in the network associated with the east line, some part of this returned energy will likewise cross the east transformer and return amplified to the west end of the circuit. If these unbalances are sufficiently large, the repeater circuit will act like the generator described in Article 161, and becomes inoperative as a repeater because of "singing". As the repeater circuit is perfectly symmetrical, its operation for transmission in the opposite direction may be followed through in exactly the same manner as described above.

To compensate, or equalize, for the difference in transmission characteristics of various lines, equalizing networks are connected in series at the mid-point of the low impedance side of the input transformers, as (sometimes referred to as "nets") balance the line at the voice frequencies, and are not required to balance it at these higher frequencies when the filter is employed.

The 22-type repeater is designed to operate with any of the three standard ringing frequencies of 20, 135, or 1000 cycles but the necessary circuit arrangements are not included in Figure 336. When either 20 or 135-cycle signaling is employed, the ringer connections are made between the bridge points of the hybrid coils and the pads as indicated. Signals at 1000 cycles fall within the voice range and are transmitted by the repeater itself and hence require no special provisions.

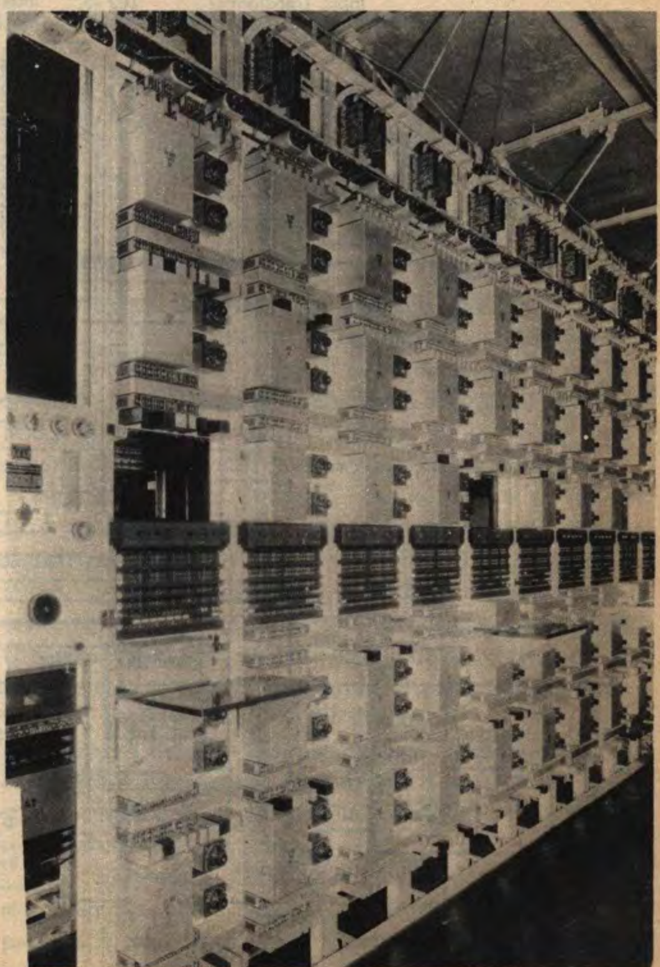
In order to make this repeater readily adaptable to various types of line circuits, the filters, balancing networks, and signaling apparatus are designed as separate units so as to be easily changed. The repeater proper, therefore, consists of only the input and output transformers, the potentiometers, the tubes, and the input equalizing apparatus. The repeater is designed to have an impedance of approximately 600 ohms. This impedance, and that of the various types of lines to which it may be connected, is matched by using



22-Type Telephone Repeaters

pointed out in Article 143. Resistance and capacitance in parallel provides the equalizing arrangement for the low frequencies, while the adjustable inductance gives the necessary equalization at the higher frequencies.

It will be noted that in each amplifier, between the plate circuit of the tube and points 2 and 5 of the hybrid coil, is a network of series inductance and shunt capacitance which forms a low-pass filter. This filter is designed to prevent the passage of high frequencies which are not essential for the successful transmission of the voice. The reason for this elimination is that it is difficult to design a balancing network that will exactly balance the ordinary telephone line at these frequencies, and at the same time balance it for the essential voice frequencies. The balancing networks



44-Type Telephone Repeaters

repeating coils (transformers) of the proper impedance ratio.

The maximum overall gain of the 22-type repeater is approximately 19.5 db, but by means of the slide-wire potentiometer and resistance pads, this overall gain can be adjusted to any lower value. It is well to remember, however, that one-half of the energy is lost each time it passes through a hybrid coil circuit. This means that the actual gain of each amplifying element must be at least 6 db greater than the overall gain required. This is compensated for in the calibration of the repeater potentiometers.

165. 44-Type Repeaters

It is customary to think of a telephone circuit as consisting of but a single pair of wires, although with phantomed or carrier circuits not even the exclusive use of a single pair of wires is required. However, in the case of long cable circuits, requiring several repeaters (four or more), two pairs of wires usually provide the most satisfactory and economical arrangement from a transmission standpoint. Each pair of conductors of such a "4-wire" circuit carries the conversation in one direction only so that the repeaters are not connected through hybrid coils as in the case of 2-wire circuits but are directly inserted in the line. In other words, transmission over each pair is always in the same direction and there is no necessity for separating the two directions by means of hybrid coils except at the circuit terminals where the 4-wire circuit is converted to 2-wire for connection to the switchboard.

The repeaters used on these 4-wire circuits are designated "44-type" and one is illustrated schematically in Figure 337. It consists of two two-stage amplifiers, one for each side of the circuit. The gain given by the amplifying tube of each stage is controlled by variable contacts on the secondaries of the input and interstage transformers, rather than by potentiometers. A 700-ohm resistance, grounded at its mid-point, is bridged across the input of each of the two amplifiers. This bridged or stabilizing resistance reduces the input impedance of the amplifier to a value which can be readily

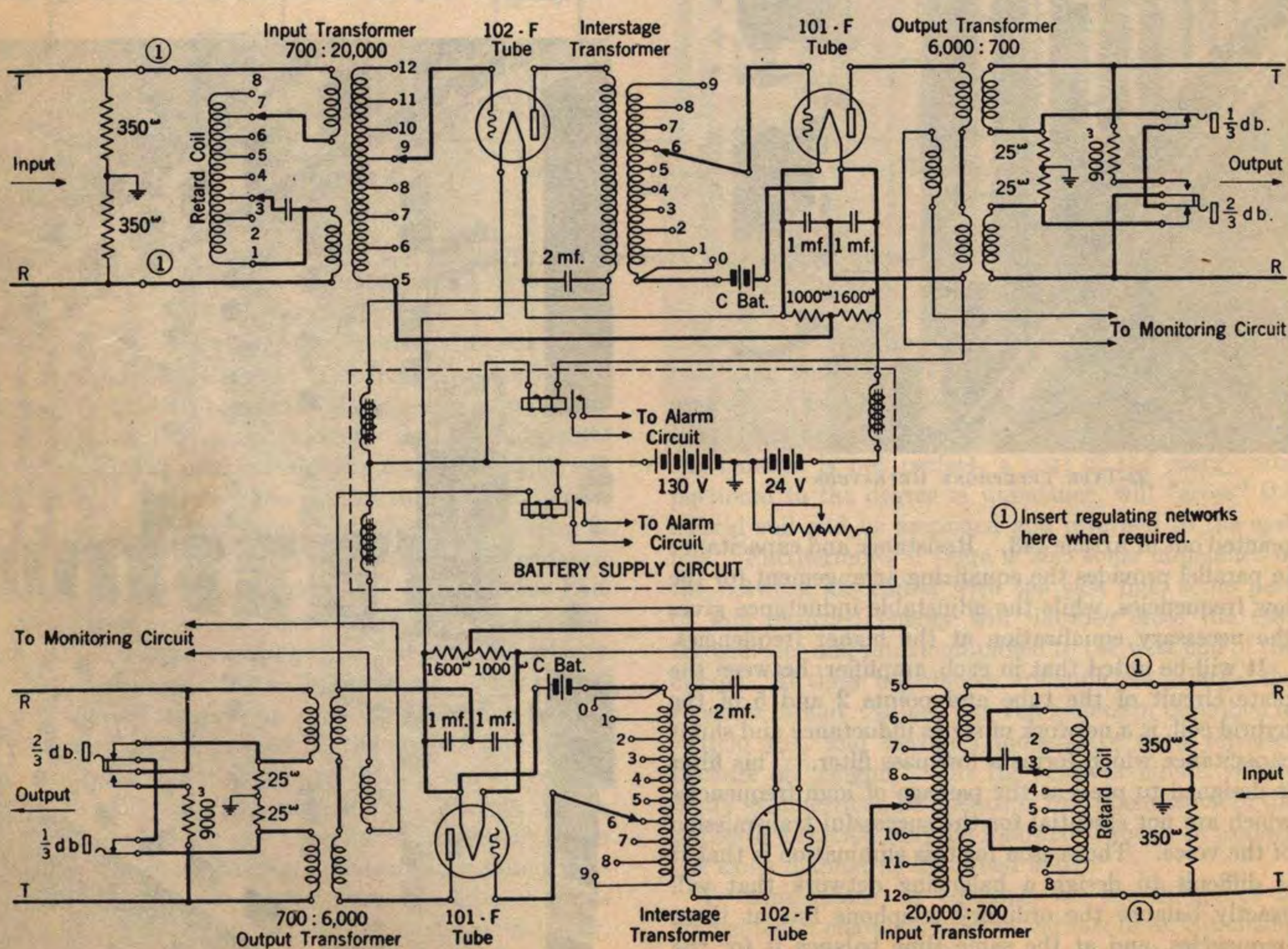


FIG. 337. 44-TYPE TELEPHONE REPEATER CIRCUIT

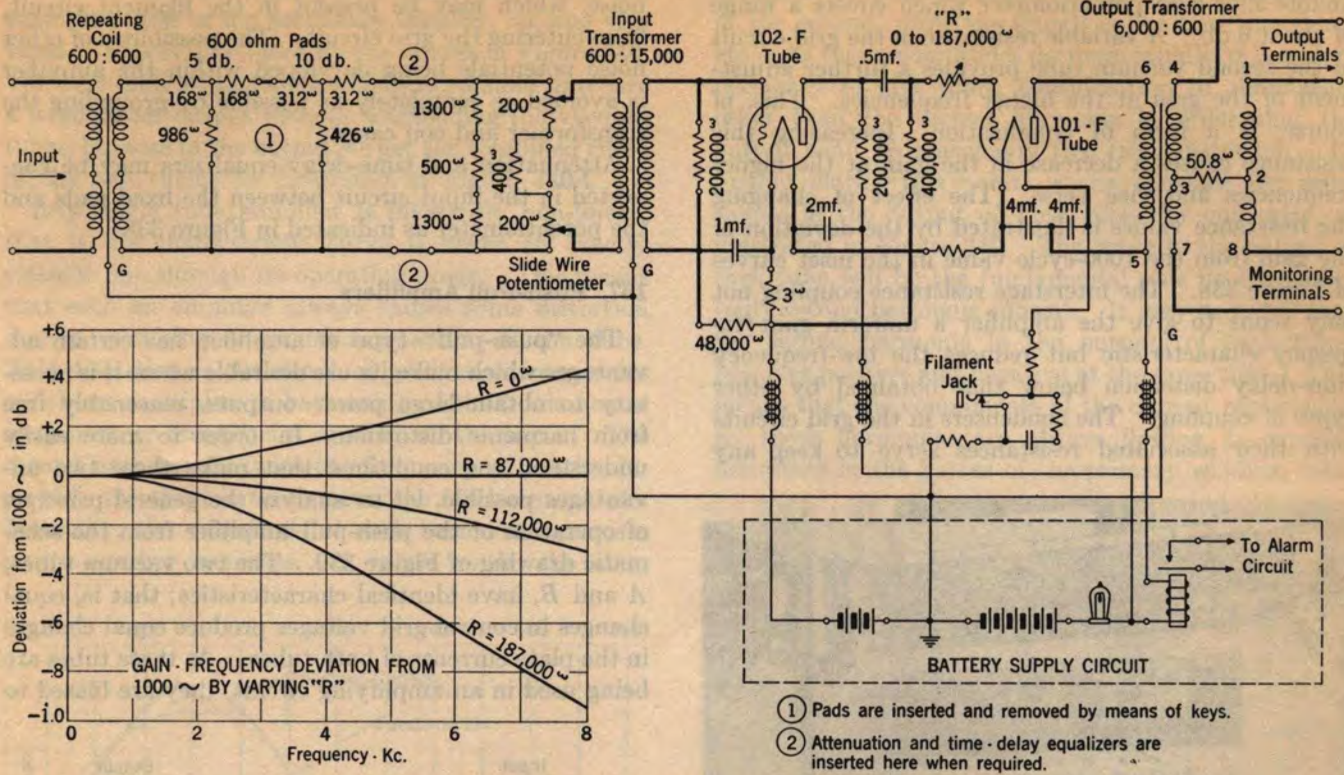


FIG. 338. AMPLIFIER USED IN PROGRAM TRANSMISSION CIRCUITS

matched to the line impedance by the use of suitable repeating coils. When regulating networks (see Article 180) are used, they are connected in the input circuit at the points designated.

The overall gain of the amplifier may be adjusted at three places. Steps of 4.7 db may be obtained from taps on the secondary of the input transformer. (Once the repeater has been initially adjusted, however, it is not ordinarily necessary to change these connections.) Finer gain adjustments are obtained from the taps on the secondary of the interstage transformer, which are operated by a dial switch. Each tap on this switch corresponds to .95 db. Still smaller steps, of either $\frac{1}{3}$ or $\frac{2}{3}$ db, may be obtained by inserting dummy plugs in either of the two jacks in the output circuit. In one case, a 50-ohm resistance is put in series with the output coil and in the other, a 9000-ohm resistance is bridged across the output coil in addition. The shape of the gain-frequency characteristic is controlled by an equalizing network of inductance and capacitance connected in series at the mid-point of the primary of the input transformer.

The first stage of amplification in this repeater is a voltage amplifier using a 102-F type vacuum tube. The output of this tube operates into the interstage transformer which has a voltage ratio of 1.25:1. The second stage, which employs a 101-F type tube, is a power amplifier. The output impedance of this tube, 6000 ohms, matches the impedance of the output transformer in order to secure the maximum transfer of power.

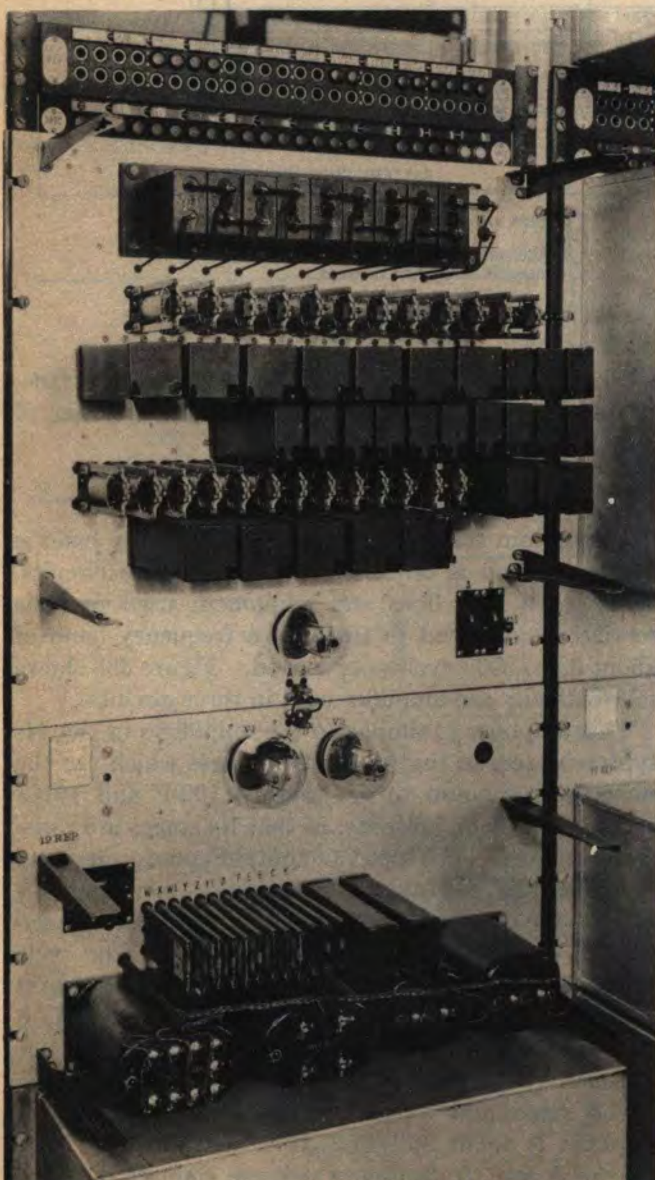
166. Program Amplifiers

In program transmission, a wider frequency band is necessary than is required for regular voice circuits. As a result, the lines and equipment used for this service are designed to transmit a frequency band of about 35 to 8000 cycles per second. Figure 338 shows schematically the amplifier used in these circuits.

This amplifier is similar to the amplifiers in the 44-type repeaters in that it has two stages which use the same type vacuum tubes, that is, 102-F and 101-F types. It differs, however, in that its stages are resistance coupled. The input and output transformers and the repeating coil at the input have very high inductance so as to give the amplifier very uniform transmission performance at all frequencies in the wide transmitted band. The use of permalloy for the cores of the transformers and coils makes it possible to obtain the necessary high inductance without going to unreasonable coil dimensions.

The maximum 1000-cycle gain of the program amplifier is about 36.5 db. It is controlled by fixed pads of 5 and 10 db loss, which are connected in the input circuit by means of keys, as noted in the drawing of Figure 338. Finer adjustments are made by a double slide wire potentiometer which covers a range of about 6 db. A variable resistance in the grid circuit of the second vacuum tube provides a further adjustment of the gain at the higher frequencies. This, of course, is a form of equalization. Increasing this resistance causes a decrease in the gain at the higher frequencies and vice versa. The effect of changing the resistance values is illustrated by the deviation of the gain from the 1000-cycle value in the inset curves of Figure 338. The interstage resistance coupling not only tends to give the amplifier a uniform gain-frequency characteristic but reduces the low-frequency time-delay distortion below that obtained by other types of coupling. The condensers in the grid circuits with their associated resistances serve to keep any noise, which may be present in the filament circuit, from entering the grid circuit. The possibility of other noise potentials being developed within the amplifier is avoided as completely as possible by grounding the transformer and coil cases.

Attenuation and time-delay equalizers may be connected in the input circuit between the fixed pads and the potentiometer as indicated in Figure 338.



Cable Program Repeater and Associated Regulating Network

167. Push-Pull Amplifiers

The "push-pull" type of amplifier has certain advantages which make its use desirable where it is necessary to obtain large power outputs, reasonably free from harmonic distortion. In order to more easily understand the conditions that make these two advantages possible, let us analyze the general principle of operation of the push-pull amplifier from the schematic drawing of Figure 339. The two vacuum tubes, *A* and *B*, have identical characteristics; that is, equal changes in control grid voltages produce equal changes in the plate currents of both tubes. As these tubes are being used in an amplifying circuit, they are biased to

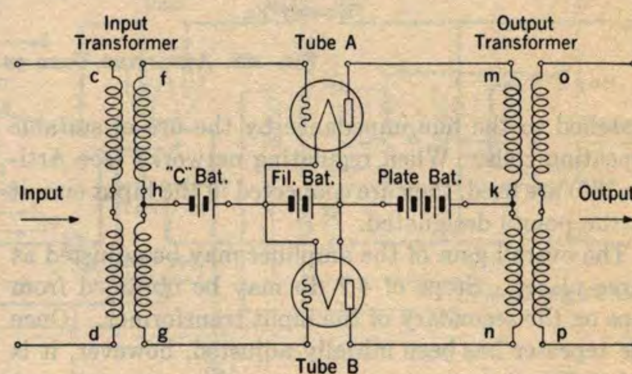


Fig. 339. Push-pull Amplifier

operate on the approximate straight line portion of their characteristic curve (control grid voltage vs. plate current curve). An alternating voltage applied at the input, *cd*, impresses voltages of equal magnitude but opposite polarity upon the control grids of tubes *A* and *B*. As the control grid of one tube becomes less negative (more positive), its plate current increases; at the same time, the control grid of the other tube becomes equally more negative, which decreases its plate current; and vice versa. Since the plate battery is connected to the mid-point, *k*, of the primary winding, *mn*, of the output transformer, the plate currents flow in opposite directions in each half of the primary winding. When the two plate currents are equal, therefore, there is no current in the secondary winding, *op*. On the other hand, **a decreasing plate current in one half of the primary winding, and an increasing plate** current in the other half, induce equal currents in the same direction in the secondary winding, *op*, and vice versa. This is illustrated by the curves of Figure 340. The push-pull arrangement makes it possible to secure a large power output without overloading the vacuum tubes, because in the output we get the combined effect of the two tubes, as illustrated in Figure 340-C.

Any single tube amplifier is inherently non-linear; that is, its characteristic curve is never an exactly straight line through its operating range. This means that such an amplifier always causes some distortion due to rectifying or modulating action in the tube. This distortion is the result of additional frequencies which are multiples, or harmonics, of the fundamental frequency or frequencies applied at the input. The second harmonic is twice the frequency of the fundamental, the third is three times the fundamental, etc. Normally, the magnitude of each harmonic decreases as its number increases. In other words, the magnitude of the third harmonic is less than the magnitude of the second, while the fourth is less than the third, etc. As these magnitudes decrease rapidly, it is seldom necessary to give much consideration to eliminating other than the second harmonic in order that the output shall be reasonably free from distortion. To further clarify this point, let us consider the wave shapes in the output of both tubes for one cycle, as illustrated in Figure 340. The wave in the output of each tube will be the fundamental plus its harmonics (only second harmonic shown). It may be noted that the second harmonic in the outputs of both tubes becomes positive and negative at the same time. This means that the components of the current represented by these harmonics are always flowing in opposite directions in the halves of the primary winding, *mkn*, and accordingly produce no effect in the secondary winding, *op*. In other words, the second harmonic is suppressed or balanced out. It can be shown that all even numbered harmonics are in phase and, therefore, balanced out; while on the other hand, the odd numbered harmonics are 180° out of phase in each half of the primary winding, *mkn*, which results in their being added to the fundamental. (If one of the windings *mk* or *nk* was reversed, the reverse condition would be true; that is, all odd numbered harmonics, including the fundamental, would be suppressed while the even numbered harmonics would appear in the output.)

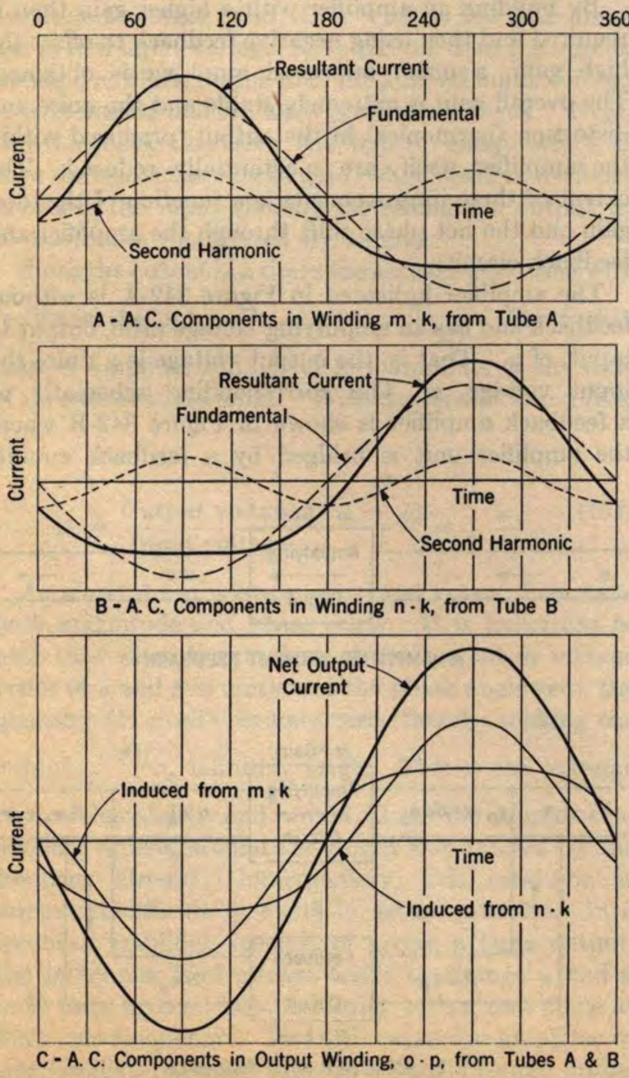
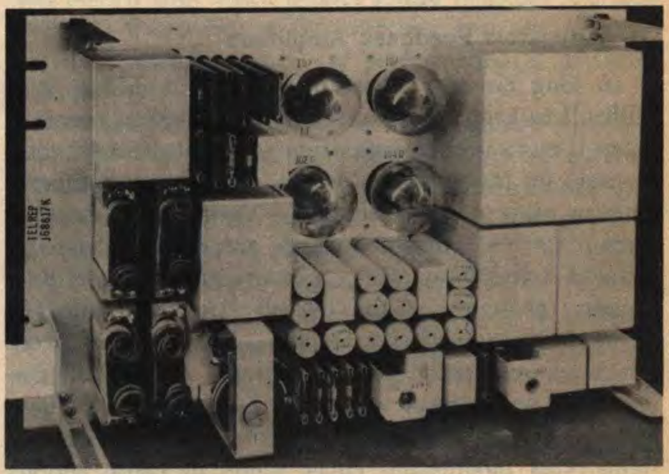


Fig. 340. Principle of Push-pull Amplifier



Push-pull Amplifier Used on Open Wire Program Circuits

When connected as shown in Figure 339, the push-pull arrangement provides a large power output reasonably free from distortion. Still a larger output may be obtained by operating additional tubes in parallel as illustrated by the schematic drawing in Figure 341. It is important to note that in order to suppress the even numbered harmonics the tubes must be evenly balanced, that is, have identical characteristics. This

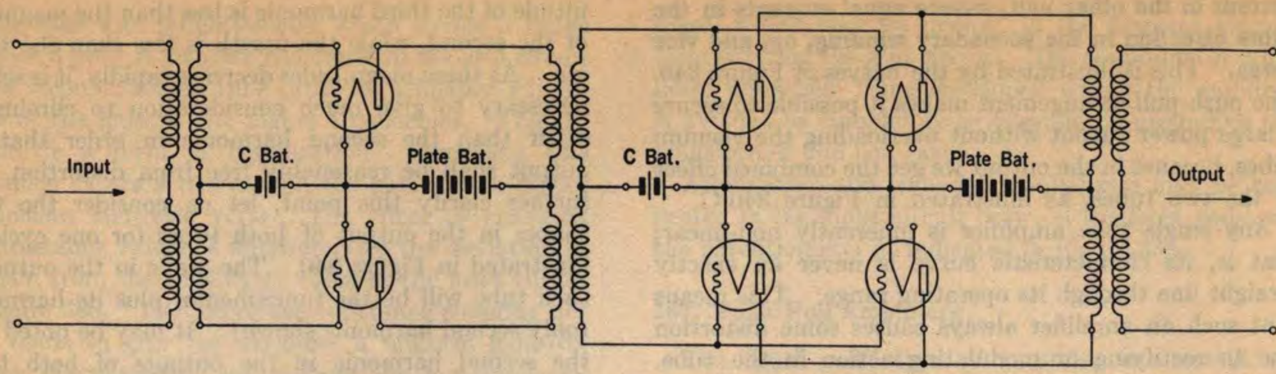


FIG. 341. PUSH-PULL AMPLIFIER WITH SECOND STAGE TUBES IN PARALLEL

requirement, of course, means increased maintenance where the push-pull amplifier is used.

168. Negative Feedback Amplifiers

In long circuits containing many amplifiers, it is difficult to keep the overall circuit equivalent constant even though proper regulation is maintained to compensate for the variation in line losses due to temperature or other natural changes. Variations in battery potentials and currents, which are small when considered individually, add up to produce serious transmission changes for the overall circuit. While the ordinary vacuum tube amplifier has an approximately linear characteristic, some additional modulation frequencies are always generated in the tube. Where such amplifiers are used in multi-channel carrier systems, this modulation effect tends to cause serious interference between channels. To keep this interference down to the point where it is not objectionable, involves sacrificing effective amplifier capacity and adds to the maintenance difficulties.

All of the above problems may be overcome by using amplifiers of the "negative feedback type". The principle of the "feedback" amplifier is to return a portion of the output back to the input. The gain of the amplifier is then increased or decreased, depending upon the magnitude and phase relation of the impulses fed back. When the impulses fed back are **in phase** with the input, the loss in the feedback path has to be greater than the gain of the amplifier; otherwise, each time an impulse is fed back to the input its magnitude will be increased, thereby causing the amplifier to build up a sustained oscillation or "singing" around the closed loop. This, of course, makes it inoperative. When the impulses fed back increase the initial input (i.e., are fed back in phase) the feedback is called **positive feedback.** If the feedback circuit is so designed, in conjunction with the amplifier itself, as to feed back the impulse **out of phase** with the initial input, the actual input to the amplifier is reduced, which reduces the overall gain. This holds true even though the loss in the feedback circuit is less than the gain of the amplifier and there is no tendency to sing. When the impulse fed back decreases the initial input in this manner, it is called **negative feedback.**

By building an amplifier with a higher gain than is required and then using negative feedback to offset the high gain, a much improved amplifier is obtained. The overall gain is extremely stable and the noise and distortion (harmonics) in the output, produced within the amplifier itself, are substantially reduced. The extent of these improvements is a function of the total gain and the net phase shift through the amplifier and feedback circuit.

The amplifier indicated in Figure 342-A is without feedback and has an amplifying voltage ratio, output to input, of μ. That is, the output voltage is μ times the input voltage, e. The corresponding schematic of a feedback amplifier is shown in Figure 342-B where the amplifier unit is bridged by a feedback circuit.

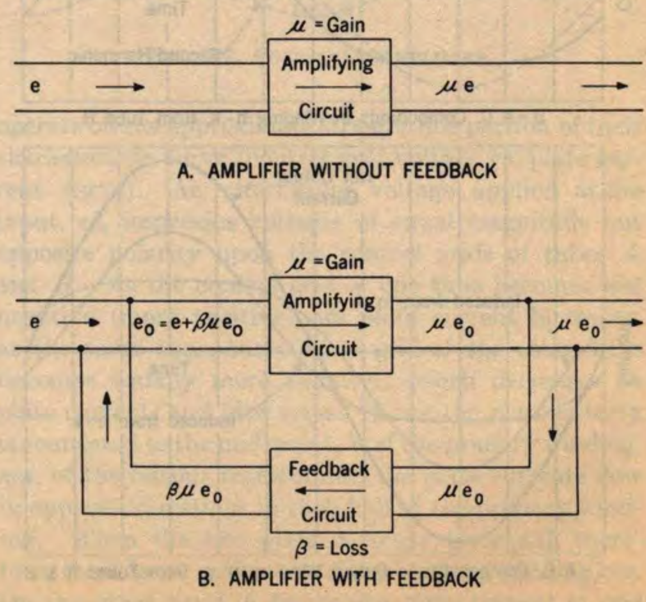


FIGURE 342

This feedback, or β circuit, provides both a loss and phase shift to the impulses fed back to the input. The shift in phase is to bring the impulses back out of phase with the applied input voltage, e. As feedback takes place in the closed loop thus formed, the actual input voltage, e_0, of the amplifier unit is no longer equal to e, but now becomes the sum of the applied voltage, e, and the feedback voltage, $\mu\beta e_0$. That is

$$e_0 = e + \mu\beta e_0 \qquad (152)$$

Solving this for e_0, we get—

$$e_0 = \frac{e}{1 - \mu\beta} \qquad (153)$$

In other words, by inserting the feedback circuit, the applied voltage, e, is divided by the quantity $(1 - \mu\beta)$ to give the actual input voltage, e_0, to the amplifier unit. If the magnitude of this quantity $(1 - \mu\beta)$ is greater than unity, the actual input voltage, e_0, is less than the applied voltage, e, and we have negative feedback; on the other hand, if $(1 - \mu\beta)$ is less than unity, the actual input voltage, e_0, is greater than the applied voltage, e, which means that the applied input voltage, e, is increased and we have positive feedback. Negative feedback decreases the input voltage which in turn decreases the overall gain, while positive feedback increases the input thereby increasing the output up to the limits established by singing.

Since the output is μ times the actual input, then with feedback the output is μe_0, or $\frac{\mu e}{1 - \mu\beta}$ The overall gain or amplification, A, of any amplifier is the ratio of its output to its input. With the feedback amplifier it is—

$$A = \frac{\text{Output voltage}}{\text{Input voltage}} = \frac{\frac{\mu e}{1 - \mu\beta}}{e} = \frac{\mu}{1 - \mu\beta} \qquad (154)$$

As μ and β are voltage or current ratios, they have both magnitude and phase angle. It is important to note that when the product of the current or voltage ratios of μ and β is unity and the phase angle zero, the quantity $(1 - \mu\beta)$ becomes zero, thereby making the output $\frac{\mu e}{1 - \mu\beta}$ infinitely large. This is the extreme case of instability and would cause the amplifier to oscillate or sing around the closed loop formed by the feedback circuit. Unfortunately, this condition is somewhat difficult to avoid in actual practice. In a feedback amplifier, capable of giving a large output, the increasing load always tends to change μ and β until their product becomes unity with a zero angle at some one frequency. This will cause the amplifier to sing (oscillate) at this frequency which, of course, makes it wholly inoperative. Great care in controlling the

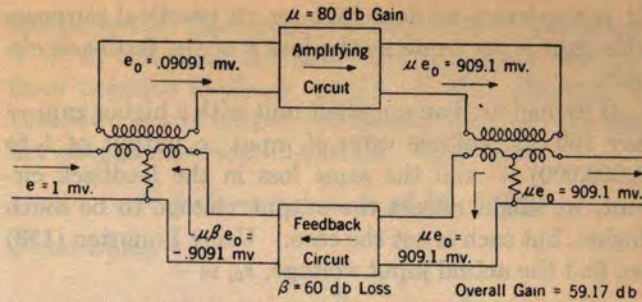


FIG. 343. EXAMPLE OF NEGATIVE FEEDBACK PRINCIPLE

phase shifts in the amplifier and feedback circuit are, therefore, required. This applies to a wide range of frequencies above and below the useful band, as well as to the useful band itself. Singing will occur unless these relations are maintained. However, experience has shown that when proper phase relations are provided in the design of the amplifier unit and its associated feedback circuit, its performance is perfectly reliable.

The practical operation of the negative feedback amplifier may be more easily understood by considering a numerical example. Let us consider the amplifier in Figure 343 where the gain, μ, of the amplifier unit is 80 db (voltage ratio of input to output of 1 to 10,000) and the loss in the feedback circuit is 60 db (voltage ratio of 1000 to 1). If the applied input voltage, e, is 1 millivolt, the actual input voltage to the amplifying unit, e_0, is — from Equation (153)—

$$e_0 = \frac{1}{1 - \frac{10{,}000}{-1{,}000}} = \frac{1}{1 - (-10)} = \frac{1}{11} = .09091 \text{ millivolt.}$$

The output voltage is—

$$\mu e_0 = 10{,}000 \times .09091 = 909.1 \text{ millivolts.}$$

This output of 909.1 millivolts is also impressed on the feedback circuit which allows 1/1000 of it to be fed back to the amplifier input. In passing through the feedback circuit its phase is shifted until it is out of phase with the applied input of 1 millivolt, which gives it a minus sign. We then have $-.9091$ millivolt combining with the initial 1 millivolt to form the actual input voltage to the amplifier, which, therefore, is—

$$1.000 - .9091 = .0909 \text{ millivolt.}$$

This checks the value of e_0 obtained above, which means that the amplifier is stable and as long as the applied input of 1 millivolt is maintained, there will be 909.1 millivolts in the output. The overall gain of the amplifier under these conditions is—

$$20 \log_{10} \frac{\text{Output voltage}}{\text{Input voltage}} = 20 \log_{10} \frac{909.1}{1}$$
$$= 20 \times 2.9586 = 59.17 \text{ db.}$$

It is important to note that for all practical purposes this gain is the same as the loss β of the feedback circuit.

If we had used an amplifier unit with a higher gain — say 100 db (voltage ratio of input to output of 1 to 1,000,000) — and the same loss in the feedback circuit, we might expect the output voltage to be much higher, but such is not the case. Using Equation (153) we find the actual input voltage, e_0, is—

$$e_0 = \frac{1}{1 - \frac{100,000}{-1,000}} = \frac{1}{101} = .009901 \text{ millivolt.}$$

The output voltage, μe_0, would now be—

$$100,000 \times .009901 = 990.1 \text{ millivolts.}$$

The overall gain of the amplifier is now—

$$20 \log_{10} \frac{990.1}{1} = 20 \times 2.9957 = 59.91 \text{ db.}$$

This is still practically equal to the loss in the feedback circuit. What is happening here is that **as the amplification is increased, the feedback circuit feeds back a larger impulse out of phase, which combines with the applied input voltage to form a lower actual input to the amplifier unit, thereby reducing the output voltage to the point where the overall gain (output to applied input) is practically the same as the loss in the feedback circuit.** On the other hand, **if the amplification is decreased, the feedback circuit feeds back a smaller impulse out of phase, which combines with the applied input voltage to form a higher actual input to the amplifier unit, thereby increasing the output voltage to the point where the overall gain (output to applied input) is practically the same as the loss in the feedback circuit.**

The above examples show that as the amplifier gain, μ, is increased, using the same loss in the feedback circuit, the overall amplification, for all practical purposes, remains the same as β, the loss in the feedback circuit. Table XVII further illustrates this fact where the gain, μ, of the amplifier is changed but the feedback circuit remains unchanged at 60 db loss. While the applied input is 1 millivolt, the actual input voltage, e_0, to the amplifier unit is much less and decreases rapidly as μ is increased, thereby keeping the overall amplification, $\frac{\mu e_0}{e}$, practically equal to 60 db. Of course, actual amplifiers could not be built to the extreme degree of accuracy indicated by this table. The essential point here is that **even if the gain, μ, of the amplifier unit changes due to variations in the battery supply, changing vacuum tube characteristics, etc., the overall gain of the amplifier remains the same for all practical purposes and is equal to the fixed loss in the feedback circuit.**

This same point can be shown mathematically from Equation (154) where the amplification, A, is equal to $\frac{\mu}{1 - \mu\beta}$. If $\mu\beta$ is large as compared to 1 — as in the above example, where $\mu = 80$ db, $\beta = 60$ db and the product of their voltage ratio is $\frac{100,000}{1,000}$, or 100—we can disregard the term 1 in the denominator, and the amplification becomes approximately equal to $\frac{1}{-\beta}$. In other words, this shows that when the product of μ and β is substantial, the overall gain of the negative feedback amplifier is for all practical purposes equal to the value of the loss in the feedback circuit.

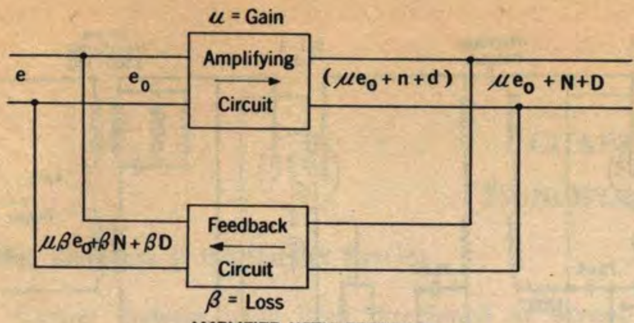
TABLE XVII

Overall Amplification of Feedback Amplifier Using Initial Input of 1 Millivolt and β Circuit of 60 db.

μ		e_0	Amplification	
db	Voltage Ratio	$\frac{e}{1-\mu\beta}$	$\frac{\mu e_0}{e}$	db
70	$\frac{3163}{1}$	.2402	759.8	57.61
80	$\frac{10,000}{1}$	.09091	909.1	59.17
90	$\frac{31,628}{1}$	.030649	969.4	59.73
100	$\frac{100,000}{1}$	.009901	990.1	59.91
110	$\frac{316,280}{1}$	.003152	996.9	59.97
120	$\frac{1,000,000}{1}$	.000999	999.0	59.99
130	$\frac{3,162,800}{1}$	.000316	999.4	59.995
140	$\frac{10,000,000}{1}$	.0000999	999.9	59.999

Throughout this discussion, we have considered the impulses fed back through the feedback circuit as being 180 degrees out of phase with the applied input voltage. This condition is not fully attainable in practice for all frequencies. The amplifiers and feedback circuits are designed, however, to approach this ideal condition as nearly as practicable. The phase shift produced by the amplifier unit must, of course, be taken into consideration in designing the associated feedback circuit.

In the output of an ideal or perfect amplifier, we would find a true reproduction of all the frequencies in the input but at a higher voltage or volume level. Practical amplifiers, unfortunately, are not perfect.



AMPLIFIER WITH FEEDBACK

FIGURE 344

In their output we usually find noise and distortion voltages which are developed within the amplifier itself. For any given amplifier gain, this noise is independent of the input and appears in the output even when the input voltage is zero (amplifier input open). Distortion, on the other hand, depends on both the input voltage and output volume. As we have already noted, this distortion represents additional frequencies which are developed within the amplifier unit and are the modulation components of the frequency or frequencies in the input.

An important feature of the negative feedback amplifier is its ability automatically to reduce to a negligible magnitude any noise or harmonic distortion developed within the amplifier itself. A small part of this noise and distortion appearing in the output is led back to the input through the feedback circuit where it reenters the amplifier in such a phase relation that when it is reamplified and again appears in the output it is out of phase with the original noise and distortion, thereby reducing its effect. This may be more easily understood if we consider the general case schematically illustrated in Figure 344 where n and d represent the noise and distortion voltages in the output before feedback takes place, while N and D represent the final noise and distortion voltages in the output. In the final output then, we have the noise and distortion that appear before feedback takes place, $n + d$, added to the noise and distortion resulting from negative feedback, $\mu\beta\ (N + D)$. The sum of these two sets of output voltages is, of course, the final net output noise and distortion. That is—

$$N + D = n + d + \mu\beta\,(N + D)$$

From which—

$$N + D = \frac{n}{1 - \mu\beta} + \frac{d}{1 - \mu\beta}$$

or—

$$N = \frac{n}{1 - \mu\beta} \qquad (155)$$

and—

$$D = \frac{d}{1 - \mu\beta} \qquad (156)$$

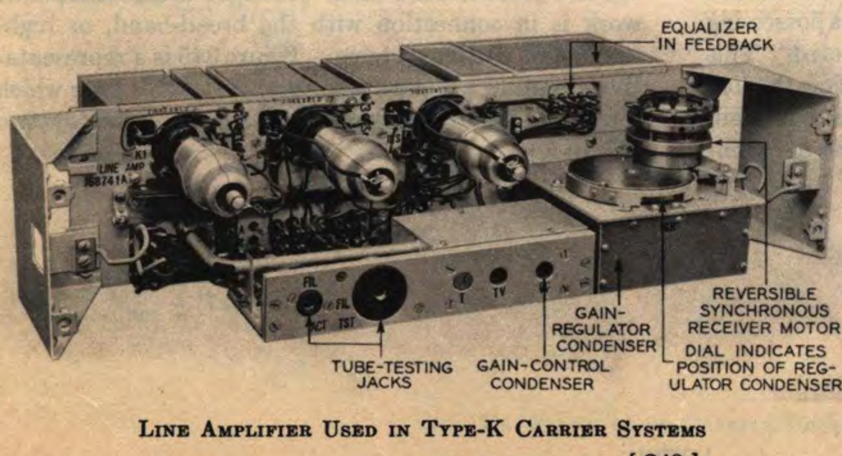
In other words, negative feedback reduces the noise and distortion by the factor $\frac{1}{1 - \mu\beta}$. As an example, let μ = 100 db (voltage ratio of 100,000 to 1) and β = 60 db (voltage ratio of 1 to 1,000). Then with negative feedback, the noise and distortion becomes—

$$N = \frac{n}{1 - \dfrac{100{,}000}{-1{,}000}} = \frac{n}{101}$$

and

$$D = \frac{d}{1 - \dfrac{100{,}000}{-1{,}000}} = \frac{d}{101}$$

Under these conditions, negative feedback reduces the noise and distortion ($n + d$) developed within the amplifier itself to approximately one one-hundredth of its original value. The extent of the reduction in the noise and distortion in any particular case is determined by the values of the amplifier gain, μ, and the loss, β, in the feedback circuit. Since a common source of noise in amplifiers is in the power supply and power tubes for the last stages of amplification, it may be seen that by using the negative feedback amplifier it is possible to meet noise requirements heretofore considered impracticable. However, **this applies only to noise or distortion developed within the amplifier**; any noise or distortion present in the input of the amplifier circuit will naturally be amplified along with and to the same extent as the regular transmission.



LINE AMPLIFIER USED IN TYPE-K CARRIER SYSTEMS

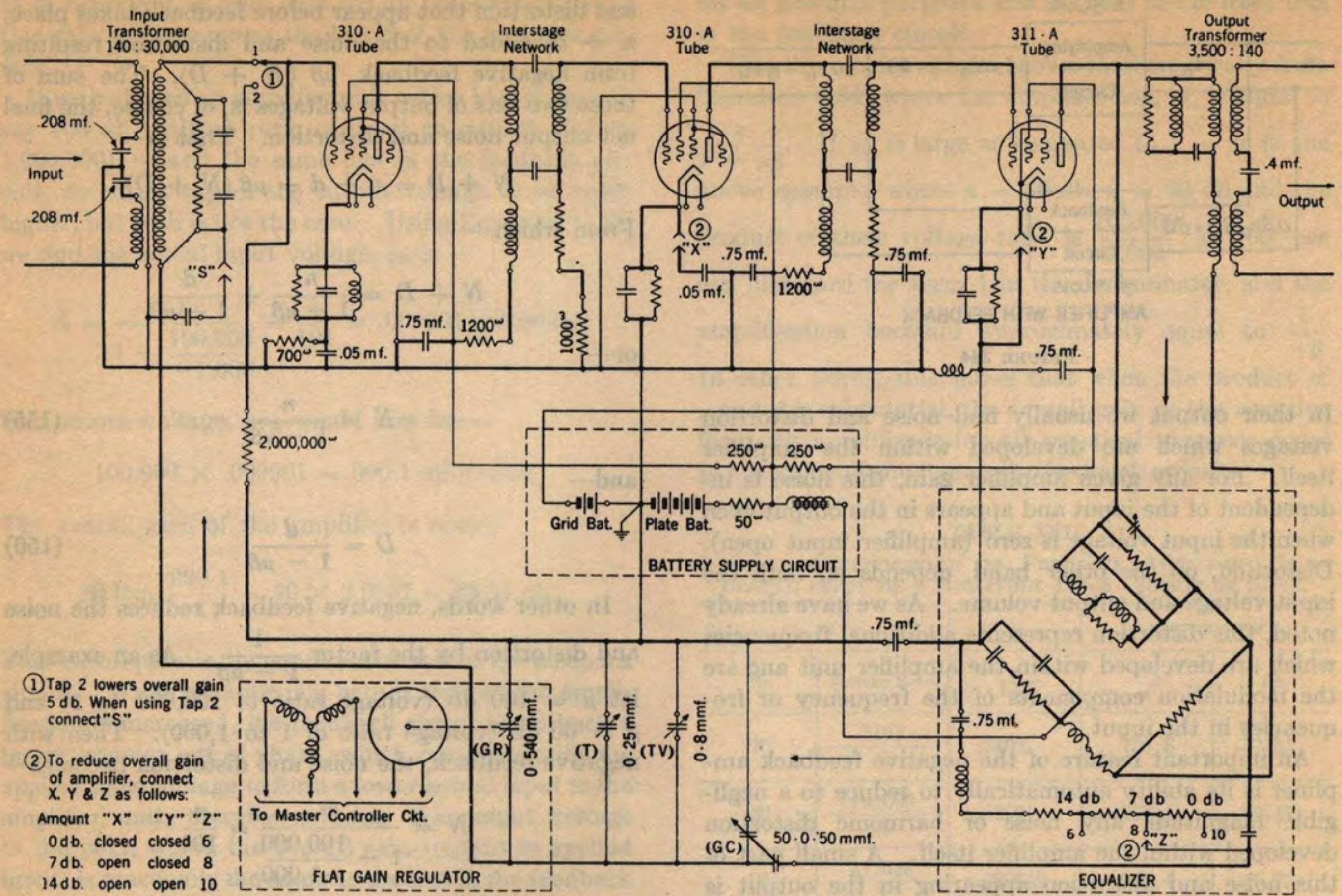


FIG. 345. THREE-STAGE NEGATIVE FEEDBACK LINE AMPLIFIER USED WITH TYPE K CARRIER SYSTEMS

Another interesting feature of the negative feedback amplifier circuit is the manner in which line equalization can be effected by the insertion of an appropriate circuit in the feedback circuit. In circuits equipped with ordinary amplifiers (no feedback), it is common practice to obtain equalization by connecting in the direct path of the signal a network having a frequency characteristic which is the inverse of that to be corrected. On the other hand, when the negative feedback amplifier is used, equalization may be obtained by inserting in the feedback circuit apparatus possessing the **same** characteristics as that to be corrected. This can be seen if we keep in mind that increasing the loss in the feedback circuit one db raises the overall gain of the amplifier one db and vice versa. The net gain-frequency characteristic of the amplifier is therefore the same as the loss-frequency characteristic of the feedback circuit with its included equalizer. In some cases such equalizing networks are easier to build than the ordinary types having characteristics inverse to those of the line.

Because of their excellent operating characteristics, negative feedback amplifiers have a very broad field of application in all kinds of communication circuits. At the present time their principal use in telephone work is in connection with the broad-band, or high-frequency, carrier systems. Figure 345 is a representative circuit of a three-stage amplifier of this type which forms one-half of the repeater in a cable carrier system (Type-K).

CHAPTER XXVII

PRINCIPLES OF CARRIER

169. Elements of the Carrier System

Carrier systems have been mentioned at several points in the preceding chapters. The carrier principle is used in both telephone and telegraph transmission and in either case the object is the simultaneous, independent transmission of several messages over a single circuit, usually without affecting the circuit's ordinary message carrying capacity.

The term "carrier" derives from the fact that alternating currents of certain selected frequencies are employed "to carry" the messages. More specifically, the variations of current making up the normal telephone or telegraph message are impressed on the carrier current, and are transmitted over the line by currents whose frequencies are of the order of the carrier frequency rather than of the initial message current. In other words, the carrier system acts to shift the frequencies of the message currents to a different range, the position of which is usually above the maximum normal voice-frequency band and dependent on the frequency of the carrier itself. It is well to note, however, that we cannot reduce the total number of frequencies (that is, the total width of the frequency band) included in the original message—we can only change its position in the frequency "spectrum". We might, for instance, shift the 2500-cycle band of voice frequencies between 200 and 2700 cycles to a band of the same width between, say, 16,000 and 18,500; or the band of telegraph frequencies between zero and 25 cycles to a band between, say, 475 and 500 cycles, but the message must always occupy at least its initial amount of space in the frequency spectrum, no matter how it is transmitted over the line.

If now we select several carrier frequencies far enough apart so that the message currents which we next impress upon them will not interfere with each other, we may simultaneously transmit the several carriers, with their impressed messages, over a single circuit just as independently for practical purposes as if a separate circuit were provided for each. Then, provided we can find a way to select the message bearing carrier currents at the receiving end of the circuit and take from them the message currents in their original form, we have a system that will handle simultaneously as many messages as we have carriers. The first problem is solved by the use of filters, and the second by a process similar to that necessary for impressing the messages on the carriers. The steps required for accomplishing the total result may be summarized as follows:

1. Providing by means of vacuum tube oscillators, or otherwise, the currents of different selected frequencies to be used as **carriers.**
2. Impressing upon each carrier the message current from the terminal telephone or telegraph station. This process is called **modulation.**
3. Separating or selecting the several modulated carrier currents at the receiving end by means of selecting circuits known as **filters.**

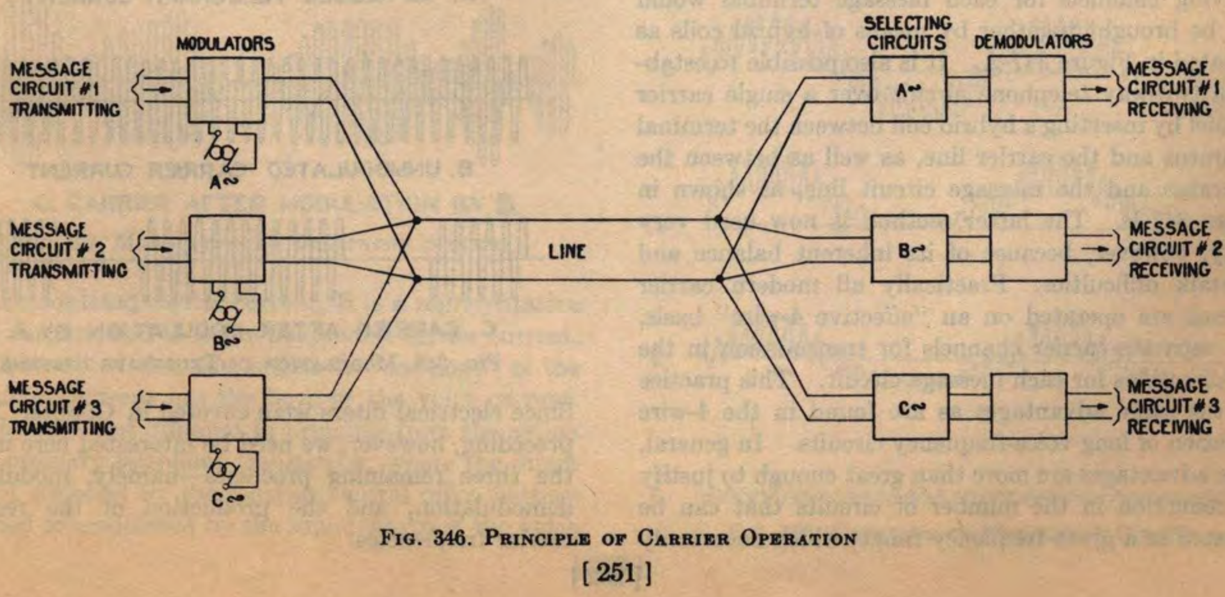


FIG. 346. PRINCIPLE OF CARRIER OPERATION

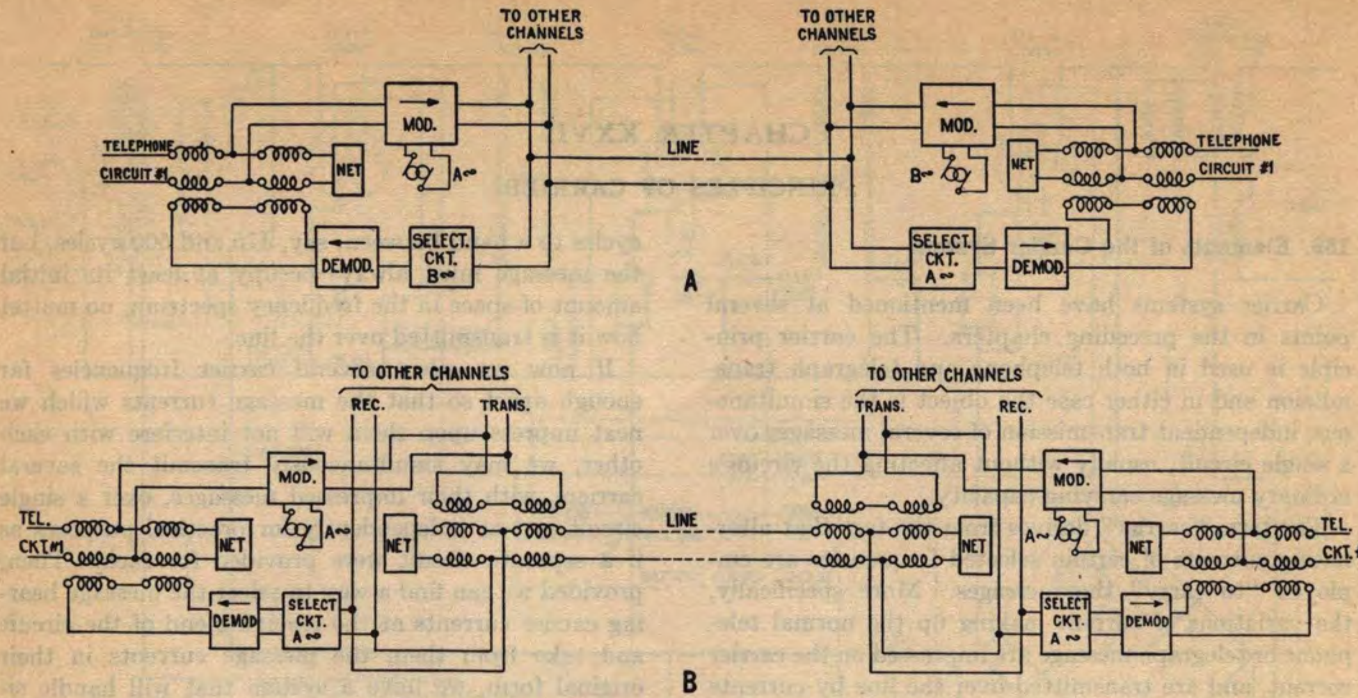


FIG. 347. METHODS OF OBTAINING TWO-WAY OPERATION

4. Separating or restoring from the selected carrier current the original message current for transmission to the receiving terminal telegraph or telephone station. This process is called **demodulation.**

Figure 346 illustrates the general arrangement required graphically. It will be noted that this schematic provides for transmission in one direction only over each carrier channel. If, as would ordinarily be the case, it is desired to transmit in both directions, three additional channels transmitting in the opposite direction could be used. In telephone systems, the sending and receiving channels for each message terminal would then be brought together by means of hybrid coils as indicated in Figure 347-A. It is also possible to establish a two-way telephone circuit over a single carrier channel by inserting a hybrid coil between the terminal apparatus and the carrier line, as well as between the apparatus and the message circuit line, as shown in Figure 347-B. The latter method is now used very rarely, however, because of its inherent balance and crosstalk difficulties. Practically all modern carrier systems are operated on an "effective 4-wire" basis, with separate carrier channels for transmission in the two directions for each message circuit. This practice has the same advantages as are found in the 4-wire operation of long voice-frequency circuits. In general, these advantages are more than great enough to justify the reduction in the number of circuits that can be operated in a given frequency range, which necessarily results from the use of two carrier channels for each message circuit.

The various types of carrier systems used in current practice are discussed in later chapters along with other kinds of transmission systems. In this chapter we shall be concerned with the principles of the circuits and apparatus employed to effect the several processes necessary to carrier operation enumerated above.

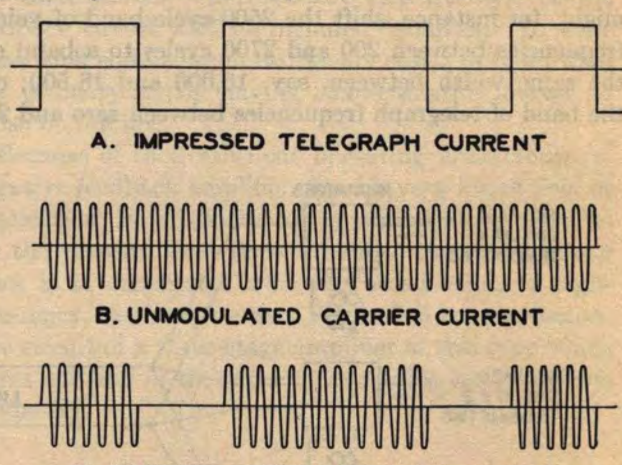


FIG. 348. MODULATION IN TELEGRAPH SYSTEMS

Since electrical filters were covered in Chapter XXIV preceding, however, we need be interested here in only the three remaining processes—namely, modulation, demodulation, and the production of the required carrier frequencies.

170. Modulation

Modulation has been defined as the process of impressing upon a carrier current, usually of a relatively high frequency, message currents of lower frequencies. The degree of difficulty involved in such a process depends upon the nature of the message current. For a telegraph current such as that shown in Figure 348-A, the method is very simple and consists merely in interrupting the supply of carrier frequency to the line during negative impulses of the telegraph signal and permitting it to flow during positive impulses. The result is to apply to the carrier line a series of "spurts" of current of the frequency of the particular carrier channel, as indicated in Figure 348-C.

In telephony, since the variations in voice current are much more complex than those of telegraph current, the process is somewhat more involved. Within certain limits, it may be thought of as a process whereby the amplitude of the carrier current is varied to correspond to the variations of the voice currents. This is illustrated in Figure 349 where A is a representation of

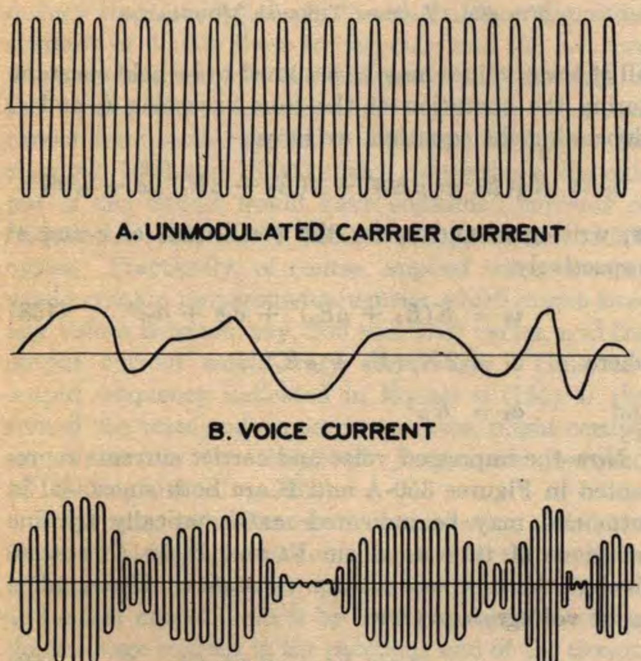


Fig. 349. Modulation in Telephone Systems

the unmodulated carrier current, B is a representative voice current, and C is the modulated carrier current. It will be noted that the outline or "envelope" of the modulated current has the form of the voice current. This effect is not different in principle from the action of an ordinary telephone transmitter, where the direct current supplied by the local or central office battery is varied or modulated by the sound waves of the voice impinging on the transmitter button. The output current from the transmitter is then a varying direct current consisting of the initial unvarying battery current, with the changing voice current superimposed upon it.

In the same way, the current of Figure 349-C could be obtained by connecting a transmitter in series with the carrier current generator, just as the battery is in series with the transmitter in the ordinary subset. The disadvantages of such a scheme will be apparent, however, and in practice vacuum tubes or other kinds of rectifying devices are used entirely for this purpose. In our study of the vacuum tube in Chapter XXV, we found that by using suitable circuit arrangements and working on a straight line portion of the control grid voltage vs. plate current characteristic of the tube, a small voltage impressed on the grid of the tube was capable of controlling a substantial current in the plate circuit, which varied in exactly the same way as the voltage impressed on the grid. In other words, the tube acted as a powerful amplifier. If now we bias the tube so that we are no longer working on a

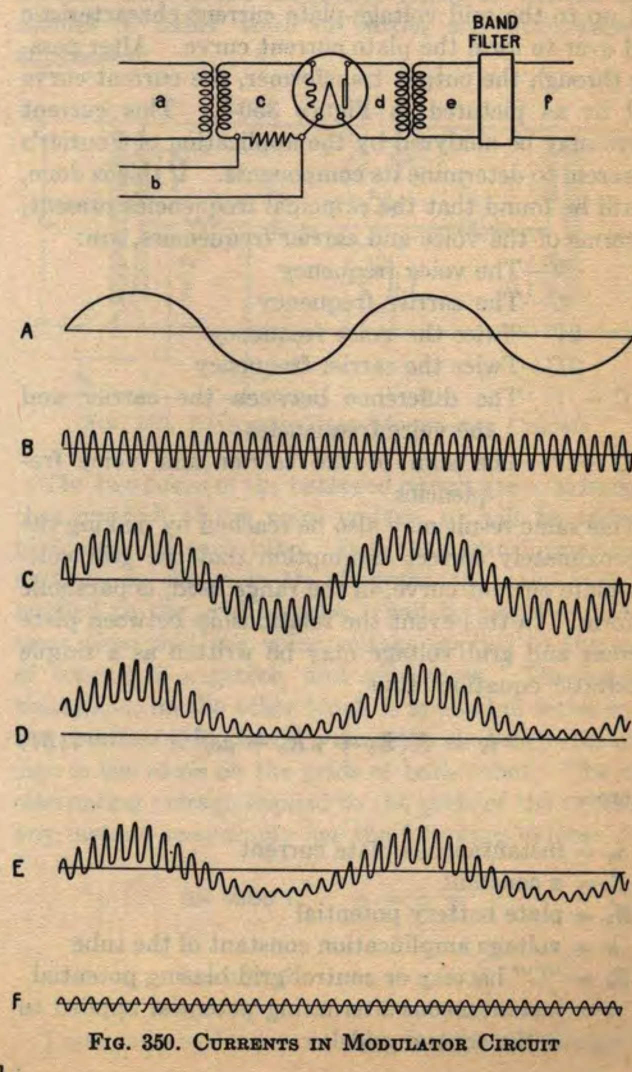


Fig. 350. Currents in Modulator Circuit

straight line portion of the characteristic curve but on a definitely curved portion, the output current will no longer vary directly with the input voltage. While there will still be some amplification, it will no longer be constant but will depend on the value of the instantaneous input voltage. This distorting or rectifying action of the tube is made use of in modulation.

In the simple circuit of Figure 350, let us assume that a voice voltage such as is represented by A is connected to the circuit through a transformer, together with the carrier voltage represented by B. For simplicity we have here assumed the voice voltage to be sinusoidal in form although this, of course, would not generally be the case. These two voltages, being in series, may be added together to give the voltage represented by C impressed on the grid of the tube. Now if the C battery or bias of the tube is given the value indicated by Figure 351, and the characteristic curve of the tube is as there shown, the impressed control grid voltage will cause a plate current of the form shown in Figure 350-D. This may be seen by projecting each instantaneous value of the grid voltage curve of Figure 351 up to the grid voltage-plate current characteristic and over to form the plate current curve. After passing through the output transformer, the current curve will be as pictured in Figure 350-E. This current curve may be analyzed by the application of Fourier's Theorem to determine its components. If this is done, it will be found that the principal frequencies present, in terms of the voice and carrier frequencies, are:

- V—The voice frequency
- C—The carrier frequency
- $2V$—Twice the voice frequency
- $2C$—Twice the carrier frequency
- $C - V$—The difference between the carrier and the voice frequencies
- $C + V$—The sum of the carrier and voice frequencies.

This same result may also be reached by making the approximately correct assumption that the grid voltage-plate current curve, in the range used, is parabolic in form. In this event the relationship between plate current and grid voltage may be written as a simple quadratic equation, thus:

$$i_b = K(E_b + \mu E_c + \mu e)^2 \qquad (157)$$

where—

i_b = instantaneous plate current
K = a constant
E_b = plate battery potential
μ = voltage amplification constant of the tube
E_c = "C" battery or control grid biasing potential
e = instantaneous alternating potential applied to the control grid.

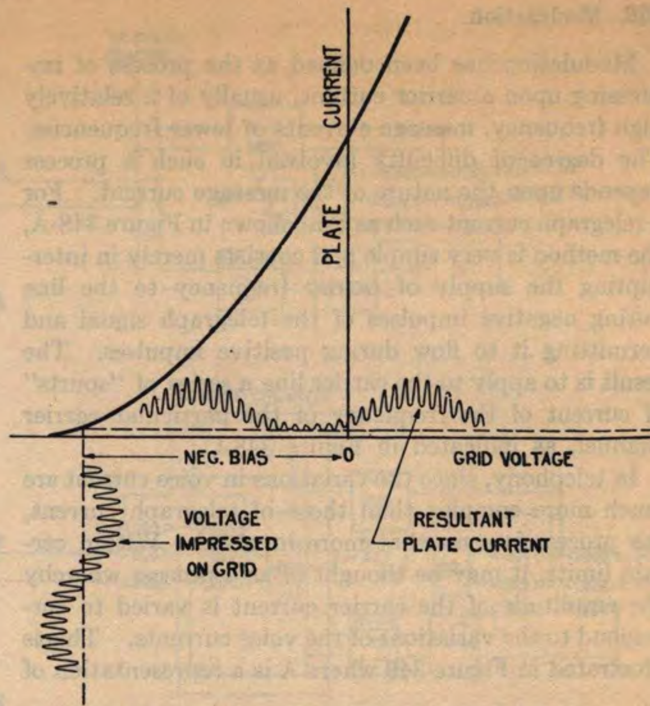


Fig. 351. Vacuum Tube as Modulator

All of these values may be assumed to be held constant during the operation of the tube excepting i_b and e. Expanding the equation, we have—

$$i_b = K[(E_b + \mu E_c)^2 + 2(E_b + \mu E_c)\,\mu e + \mu^2 e^2]$$

or, writing a_1 and a_2 for the coefficients of e and e^2 respectively,

$$i_b = K(E_b + \mu E_c)^2 + a_1 e + a_2 e^2 \qquad (158)$$

where $\quad a_1 = 2K\mu(E_b + \mu E_c)$

and $\quad a_2 = K\mu^2$

Now the impressed voice and carrier currents represented in Figures 350-A and B are both sinusoidal in form and may be indicated mathematically by sine functions of time as $A \sin Vt$ and $B \sin Ct$ respectively, where A and B are constants. The applied input voltage, e, is then—

$$e = A \sin Vt + B \sin Ct. \qquad (159)$$

Substituting (159) in Equation (158), we have for the output current—

$$i_b = K(E_b + \mu E_c)^2 + a_1(A \sin Vt + B \sin Ct) + a_2(A \sin Vt + B \sin Ct)^2$$

and, expanding—

$$i_b = K(E_b + \mu E_c)^2 + a_1 A \sin Vt + a_1 B \sin Ct + a_2 A^2 \sin^2 Vt + 2a_2 AB \sin Ct \sin Vt + a_2 B^2 \sin^2 Ct. \qquad (160)$$

Making use of the trigonometric relationships—

$$\sin^2\theta = \tfrac{1}{2} - \tfrac{1}{2}\cos 2\theta$$

and

$$\sin\theta\sin\phi = \tfrac{1}{2}\cos(\theta - \phi) - \tfrac{1}{2}\cos(\theta + \phi)$$

we may expand further to obtain—

$$\begin{aligned} i_b = {} & K(E_b + \mu E_c)^2 + a_1A\sin Vt + a_1B\sin Ct \\ & + \tfrac{1}{2}a_2A^2 - \tfrac{1}{2}a_2A^2\cos 2Vt + a_2AB\cos(C - V)t \\ & - a_2AB\cos(C + V)t + \tfrac{1}{2}a_2B^2 - \tfrac{1}{2}a_2B^2\cos 2Ct \\ = {} & K(E_b + \mu E_c)^2 + \tfrac{1}{2}a_2(A^2 + B^2) + a_1A\sin Vt \\ & + a_1B\sin Ct - \tfrac{1}{2}a_2A^2\cos 2Vt - \tfrac{1}{2}a_2B^2\cos 2Ct \\ & + a_2AB\cos(C - V)t - a_2AB\cos(C + V)t. \quad (161) \end{aligned}$$

An analysis of this equation shows the first and second terms to be constants representing direct current which, of course, will not appear on the line side of the output transformer. The third and fourth terms are merely amplified currents of voice and carrier frequency respectively; the fifth and sixth are sinusoidal currents of double these frequencies; and the last two represent respectively the difference and the sum of the carrier and voice frequencies. If the voice and carrier frequencies applied to the grid had been, for example, 1000 and 10,000 cycles respectively, the output of the circuit would have contained currents of frequencies 1000, 10,000, 2000, 20,000, 9000 and 11,000 cycles. Practically, of course, applied voice currents would contain numerous frequencies which might have any values between, say, 200 and 2700 cycles, and the output current would vary accordingly. Thus, the output frequency indicated in Equation (161) as the sum of the voice and carrier frequencies, might occupy any value in the band of frequencies between $(C + 200)$ and $(C + 2700)$.

These sum and difference frequencies are called the upper and lower modulation components, respectively, or, more commonly, the **upper and lower side-bands,** and **either one of them is by itself capable of carrying the message current to the receiving end of the circuit.** In practice, accordingly, it is customary to suppress by means of filters or otherwise, all of the frequencies in the output of the modulators except one side-band for transmission over the line, although our theoretical diagrams of Figures 346 and 347 show the output of the modulators connected directly to the line. Thus in Figure 350-F the band filter has blocked all frequencies except the upper side-band, $(C + V)$.

It is obviously desirable also to so arrange the modulator circuit that the current to be transmitted over the line has the largest possible value, and the currents that are not needed have relatively small values, thus making feasible the utilization of the greatest possible part of the modulator tube's output energy. This result can be to a degree achieved by properly adjusting the values of the constants a_1, a_2, A and B in Equation (161). Referring to this, it will be noted that if a_1 is made very small, the voice and carrier frequencies may be practically eliminated from the output. This may be accomplished within limits by giving E_c a large negative value, in which case the factor $(E_b + \mu E_c)$, in the expression $(a_1 = 2K\mu[E_b + \mu E_c])$ may be made to approach zero, reducing a_1 correspondingly.

In most of the vacuum tube modulator circuits now in service, however, the method generally employed to control both the absolute and relative magnitudes of the output components depends on the use of a balanced tube arrangement similar to that described in Article 167 under push-pull amplifier circuits. Such a modulator circuit is shown schematically in Figure 352. Under ideal conditions, the output of this circuit includes only the voice frequency and the two side-bands. The carrier frequency itself, as well as harmonics of either voice or carrier are automatically suppressed.

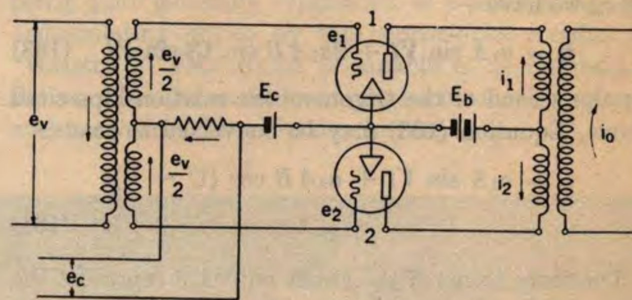


FIG. 352. BALANCED TUBE MODULATOR CIRCUIT

The two tubes of the balanced circuit are so arranged that one-half of the voice voltage, e_v, will be applied to the grid of each tube. But, with transformer connections as shown in the diagram, the voice voltage applied to the grid of tube 1 will be positive at the same time that the voice voltage applied to the grid of tube 2 is negative, and vice versa. The carrier voltage, e_c, on the other hand, is applied in series with the common grid biasing voltage so that its value and sign is the same on the grids of both tubes. The net alternating voltage applied to the grids of the tubes at any instant accordingly has the following values:

$$\text{for tube 1} \qquad e_1 = e_c + \frac{e_v}{2}$$

$$\text{for tube 2} \qquad e_2 = e_c - \frac{e_v}{2}$$

These values of input voltage may be substituted in

Equation (158) to determine the plate current of each tube. This gives, for tube 1—

$$i_1 = K(E_b + \mu E_c)^2 + a_1\left(e_c + \frac{e_v}{2}\right) + a_2\left(e_c + \frac{e_v}{2}\right)^2$$

and, for tube 2—

$$i_2 = K(E_b + \mu E_c)^2 + a_1\left(e_c - \frac{e_v}{2}\right) + a_2\left(e_c - \frac{e_v}{2}\right)^2$$

These currents, it will be noted, flow in opposite directions in the primary winding of the output transformer. Therefore, to obtain the value of the current in the secondary, we must subtract them. Thus, neglecting the direct-current components since these will not pass through the transformer, we have for the output current—

$$i_0 = i_1 - i_2 = a_1e_c + \frac{a_1}{2}e_v + a_2e_c^2 + a_2e_ce_v + \frac{a_2}{4}e_v^2$$

$$- a_1e_c + \frac{a_1}{2}e_v - a_2e_c^2 + a_2e_ce_v - \frac{a_2}{4}e_v^2$$

$$= a_1e_v + 2a_2e_ce_v \qquad (162)$$

Then, by substituting $A \sin Vt$ for e_v and $B \sin Ct$ for e_c, we have—

$$i_0 = a_1A \sin Vt + 2a_2AB \sin Ct \sin Vt \qquad (163)$$

By the second of the trigonometric relationships cited above, Equation (163) may be converted to read—

$$i_0 = a_1A \sin Vt + a_2AB \cos (C - V)t$$

$$- a_2AB \cos (C + V)t \qquad (164)$$

The three terms of the above equation represent the voice frequency and the two side-bands of the carrier frequency. As in the single tube modulator circuit discussed earlier, the voice-frequency term may be kept small in comparison to the side-band terms by making a_1 as low in value as practicable. Since there is no carrier-frequency term in this equation, an even more effective method of insuring that the greater part of the output energy shall be represented by the side-band terms is to make B much larger than A—that is, to make the amplitude of the applied carrier voltage much larger than that of the applied voice voltage.

Before leaving this subject, it should be noted that in all of the above it has been assumed that the characteristic curve of the modulator tubes had the ideal parabolic form. This is only approximately true in practice and in so far as the curve departs from this ideal, frequency components additional to those indicated in the above mathematical expressions will appear in the output. Further, we have assumed a single frequency for the applied signal voltage in all cases. Actually, a voice signal will usually include several different frequencies which will be applied simultaneously to the modulator input. There will be a certain amount of inter-modulation between these signal frequencies and some of the resultant harmonics or sum and difference components may have frequencies within the range of the useful side-band, thus tending to cause distortion. However, it may be seen from both Equations (161) and (164) that the magnitude of these disturbing frequencies will be proportional in all cases to A, the amplitude of the applied signal voltage. By making the carrier voltage much larger than the signal voltage, accordingly, these frequencies may be kept low enough in value so that their distorting effect is practically negligible.

171. Demodulation

The action of the demodulator is identical in principle with that of the modulator, as may be seen from an examination of Figure 353. The carrier-frequency B, identical in frequency to the carrier frequency employed at the sending end of the line, adds to the in-

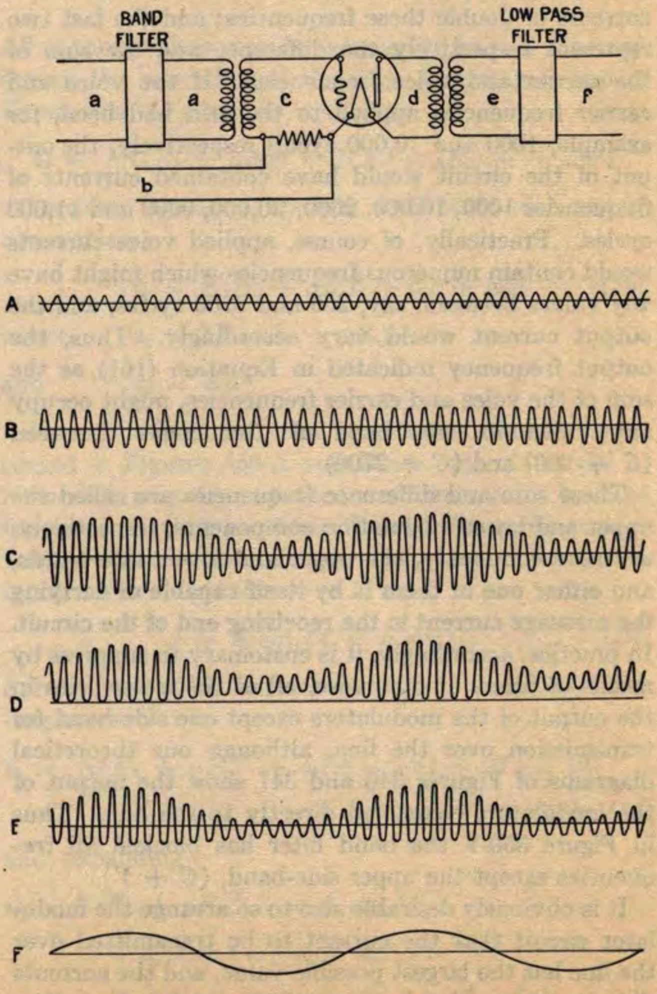


Fig. 353. Currents in Demodulator Circuit

coming side-band A, which in this case is assumed to be carrying a voice current of a single frequency, to give the net voltage C impressed on the grid. Assuming the upper side-band is transmitted over the line, the impressed voltage then is equal to $(C + V)$ plus C. If we substitute these values for e in Equation (158) and expand, we will find that the resultant output currents are—

C—The carrier frequency.
$(C + V)$—The impressed side-band frequency.
$2C$—Twice the carrier frequency.
$2(C + V)$—Twice the impressed side-band frequency.
$(C + V) + C = 2C + V$—The sum of carrier and side-band.
$(C + V) - C = V$—The difference of carrier and side-band, which is the **voice frequency.**

All of these currents are present in Figure 353-D; and Figures 353-E and F represent respectively the complex current on the drop side of the output transformer and the voice current itself, after the higher frequencies have been eliminated by means of a **low-pass** filter.

Similarly, in the case of the balanced tube circuit employed as a demodulator, if the upper side-band $(C + V)$ is applied, the output frequencies will be V, $C + V$, and $2C + V$. This will be apparent from Equation (164) where $(C + V)$ may be substituted for V to determine the demodulation products.

In telegraph systems the process of demodulation is relatively simple, as in the corresponding modulation. As noted in Article 170, the modulated current transmitted over the line consists of a series of "spurts" of alternating current of the frequency of the carrier. This incoming current, represented by Figure 354-A, after being selected by the proper filter, is led to a vacuum tube, the grid of which is so strongly biased that it acts as a rectifier. The resultant output is a unidirectional varying current, as shown in Figure 354-B. This current obviously consists of two components, a direct current and a superimposed alternating current of the carrier frequency. The alternating current is filtered out by a simple condenser arrangement, leaving only a series of pulses of direct current corresponding in duration to those applied at the sending end of the circuit, as illustrated by Figure 354-C. These direct-current impulses are then used to operate a relay, the contacts of which control the battery connections to the usual telegraph repeating apparatus and establish the polar operating current of Figure 354-D.

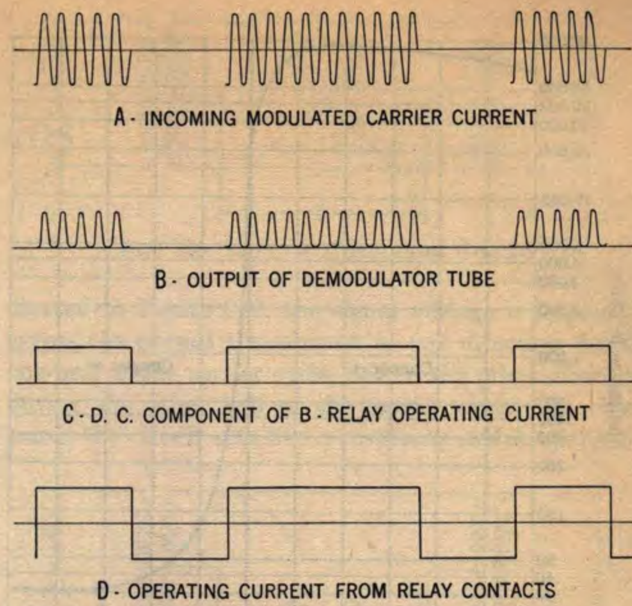


Fig. 354. Demodulation in Telegraph Systems

172. Copper-Oxide Varistors as Modulators and Demodulators

In more recent carrier practice, the vacuum tube is being quite generally superseded as a modulating and demodulating device by the copper-oxide rectifier or "varistor". This device is capable of accomplishing the same results as we have just been considering, and with a considerable reduction in cost.

Copper-Oxide Varistor Unit Used as Modulator and Demodulator

The principle of the copper-oxide unit as a rectifier for converting alternating to direct currents in power supply circuits was discussed briefly in Article 53. Its essential characteristic for the present purpose is that, as shown in Figure 355, its resistance varies with the magnitude and polarity of the applied voltage. This is a typical curve for a single disc-shaped copper-oxide unit having a diameter of $\frac{3}{16}$ inch. It will be noted that the resistance of the unit varies from a relatively low value when the copper is negative with respect to the copper oxide, to a very high value when the voltage polarity is reversed.

For use as modulators and demodulators in carrier

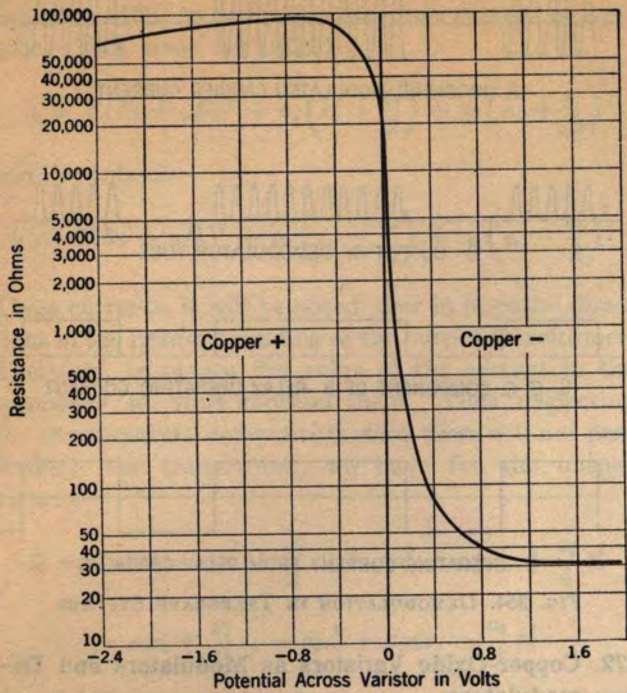


FIG. 355. RESISTANCE CHARACTERISTIC OF COPPER-OXIDE VARISTOR

systems, four of these tiny copper-oxide units are mounted in a sealed container having a maximum dimension of less than one inch, as shown in the accompanying photograph. The characteristics of such units are very stable and their useful life is apparently indefinite.

In the channel modulator and demodulator circuits of carrier systems, the varistor units are connected in the simple Wheatstone bridge arrangement illustrated in Figure 356. (In the symbols used here for the varistor units, the copper oxide is represented by the arrow, and the copper by the crossbar. The conducting direction of the unit is thus in the direction of the arrow point.) The carrier voltage, *C*, is made very large as

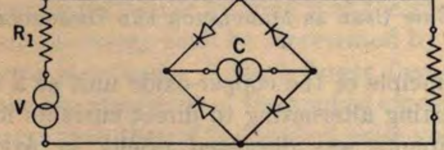


FIG. 356. BALANCED BRIDGE MODULATOR CIRCUIT

compared with the signal voltage, *V*, so that the resistance presented by the varistor units is effectively under the control of the carrier voltage alone. In other words, the resistance of the varistors varies from a low value to a high value at the frequency of the applied carrier voltage.

Under these circumstances, and assuming perfect rectification, the network of varistors will act so as virtually to short-circuit the line during the positive halves of the carrier voltage cycle; and to present an open circuit across the line during the negative halves of the carrier voltage cycle. This is illustrated by the two diagrams of Figure 357 where the varistors are

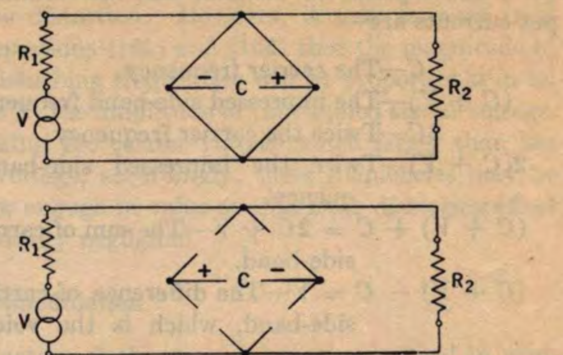


FIG. 357. OPERATING PRINCIPLE OF CIRCUIT OF FIG. 356

indicated as perfect conductors during the positive pulse and as opens during the negative pulse. The effect on the applied signal voltage, *V*, is therefore to block it completely during the positive half of the carrier cycle and to permit its free transmission during the negative half of the carrier cycle. In this ideal case, therefore, the varistors act effectively like a short-circuiting switch, opening and closing at the frequency of the carrier voltage. The resultant output current is shown in Figure 358.

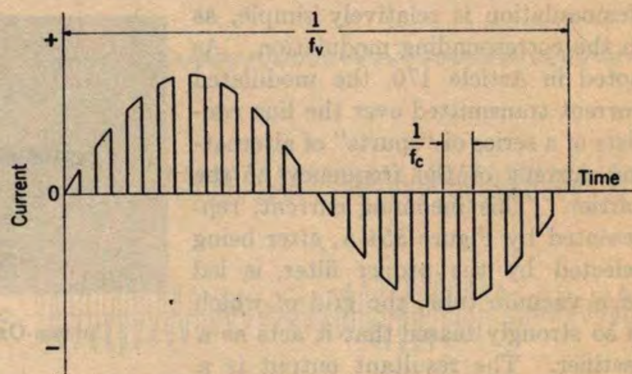


FIG. 358. OUTPUT CURRENT OF BALANCED BRIDGE MODULATOR

An analysis of this current curve would show that its principal components are the signal frequency and the upper and lower side-bands of the carrier frequency. If we assume for the signal voltage a sine wave of the form—

$$e = A \sin Vt$$

where *A* represents the amplitude of the signal and *V* is 2π times the signal frequency, an approximate equation for the output current represented by Figure 358 may be written as follows:

$$I = \frac{A \sin Vt}{2(R_1 + R_2)} + \frac{2A}{\pi(R_1 + R_2)} [\sin Vt \sin Ct$$
$$+ \tfrac{1}{3} \sin Vt \sin 3Ct + \tfrac{1}{5} \sin Vt \sin 5Ct + \cdots] \quad (165)$$

Here R_1 and R_2 are respectively the input and output resistances as indicated in Figure 356 and C is 2π times the carrier frequency.

Making use of the trigonometric relationship—

$$\sin \theta \sin \phi = \tfrac{1}{2} \cos (\theta - \phi) - \tfrac{1}{2} \cos (\theta + \phi),$$

the above equation may be rewritten as—

$$I = \frac{A \sin Vt}{2(R_1 + R_2)} + \frac{A}{\pi(R_1 + R_2)} [\cos (C - V)t$$
$$- \cos (C + V)t + \tfrac{1}{3} \cos (3C - V)t - \tfrac{1}{3} \cos (3C + V)t$$
$$+ \tfrac{1}{5} \cos (5C - V)t - \tfrac{1}{5} \cos (5C + V)t + \cdots] \quad (166)$$

The first term of this equation represents the original signal voltage with a reduced amplitude. The first two terms inside the brackets are the lower and upper side-bands of the modulated carrier wave, and the remaining terms in the brackets represent similar upper and lower side-bands of odd multiples of the carrier frequency. The equation does not include any term for the carrier frequency itself, showing that the carrier is suppressed by the balanced arrangement of the varistors.

In practice, as we know, only one of the side-bands of the carrier frequency is made use of and this is selected from the several frequency terms appearing in the output by means of a suitable band-pass filter. A demodulator arrangement, identical to that shown in Figure 356, is used at the receiving end of the carrier line to restore the original signal frequency. In this case, the frequencies applied to the varistor circuit (demodulator) are the received side-band and a locally generated carrier identical in frequency to that supplied to the modulator at the sending end. Thus, if we assume that the lower side-band is transmitted, the signal frequency applied to the demodulator may be indicated in the form, $K \cos(C-V)t$. When this term is substituted in Equation (165) in place of $A \sin Vt$, the first term inside the brackets in Equation (166) will become:

$$\cos [C - (C - V)]\, t = \cos Vt$$

This is the desired original signal and it can be selected from the other components of demodulation by the use of a simple low-pass filter.

For the group modulators and demodulators of the Types-J and K carrier systems, a somewhat different arrangement of the varistor units is employed. This is illustrated in Figure 359. It is also a balanced bridge arrangement but the circuit connections and the configuration of the varistors are such that, as in-

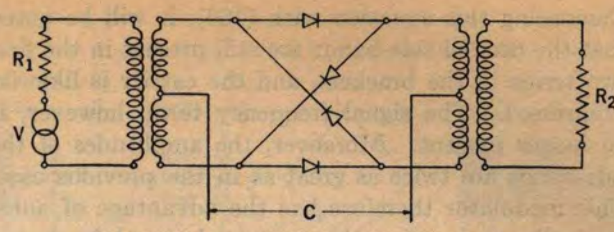


FIG. 359. LATTICE MODULATOR CIRCUIT

dicated in Figure 360, the signal voltage is impressed across the output transformer in one direction during one-half of the carrier cycle, and in the other direction during the other half of the carrier cycle. In other words the circuit acts like a reversing switch operating

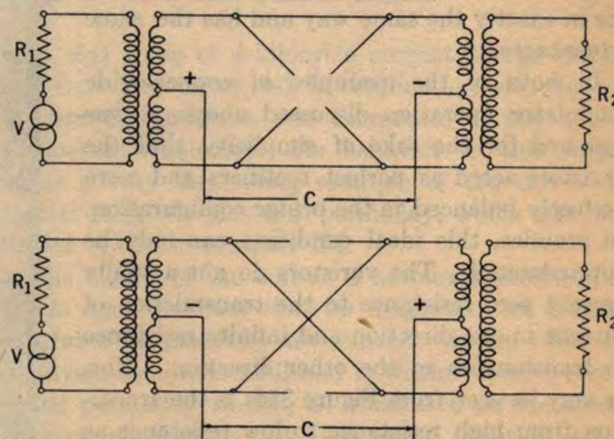


FIG. 360. OPERATING PRINCIPLE OF CIRCUIT OF FIG. 359

at the carrier frequency and results, in the ideal case, in the output current wave shown in Figure 361.

Using the same terminology as in the preceding discussion, the approximate equation for the curve of Figure 361 is—

$$I = \frac{2A}{\pi(R_1 + R_2)} [\cos (C - V)t - \cos (C + V)t$$
$$+ \tfrac{1}{3} \cos (3C - V)t - \tfrac{1}{3} \cos (3C + V)t$$
$$+ \tfrac{1}{5} \cos (5C - V)t - \tfrac{1}{5} \cos (5C + V)t + \cdots] \quad (167)$$

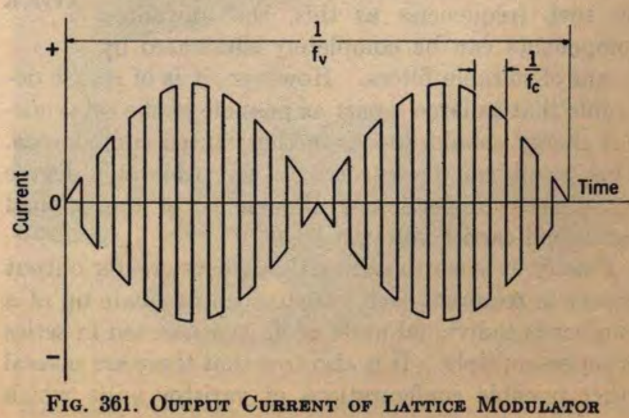


FIG. 361. OUTPUT CURRENT OF LATTICE MODULATOR

Comparing this equation with (166), it will be noted that the desired side-bands are still present in the first two terms in the brackets, and the carrier is likewise suppressed. The signal frequency term, however, is no longer present. Moreover, the amplitudes of the side-bands are twice as great as in the previous case. This modulator therefore has the advantage of automatically suppressing the unwanted signal frequency components and of providing a larger output of the desired side-bands. These characteristics are particularly desirable in group modulators where the wide band transmitted makes maximum side-band output, and the reduction of the number of unwanted products, very important. This arrangement of course operates as a demodulator in exactly the same way and has the same advantages.

In both of the examples of copper-oxide modulator operation discussed above, it was assumed for the sake of simplicity that the varistors acted as perfect rectifiers and were perfectly balanced in the bridge configuration. In practice, this ideal condition can only be approximated. The varistors do not actually present zero resistance to the transmission of current in one direction and infinite resistance to transmission in the other direction. Nor, as may be seen from Figure 355, is the transition from high resistance to low resistance as sharp as might be desired. Exact balance between the four varistors in the bridge connections is also a condition which can only be approached in practice.

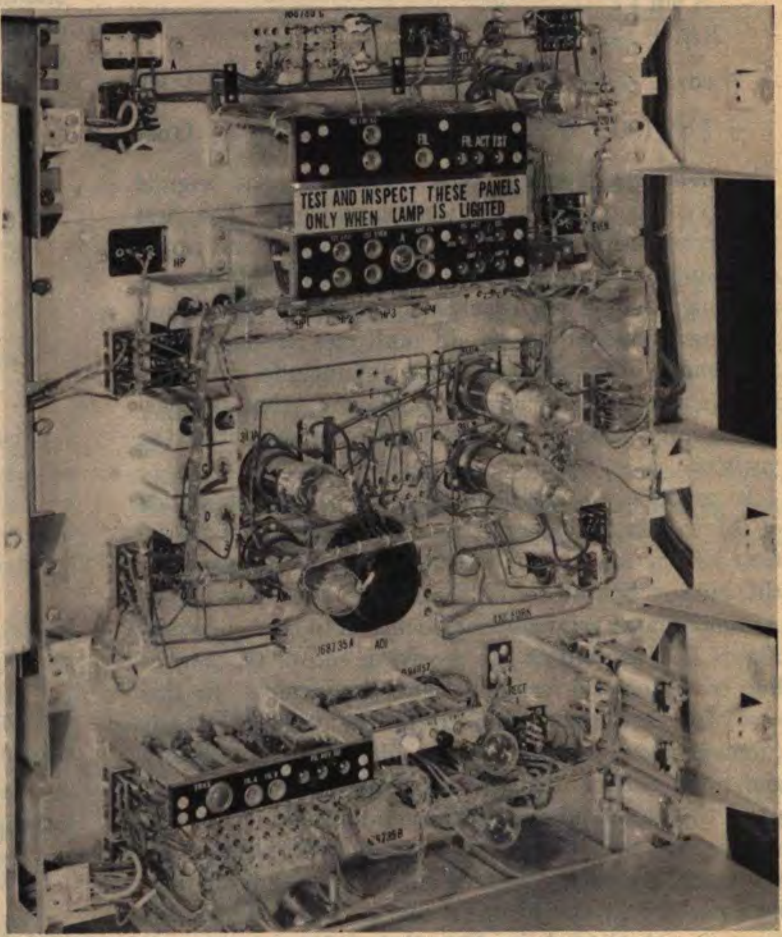
As a result of the above practical facts, the modulator and demodulator outputs always contain numerous components additional to those indicated by Equations (166) and (167), including the carrier frequency itself. Most troublesome of these unwanted components, probably, are harmonics of the signal frequency which may fall within the range of the useful side-band and thus cause distortion. Except for such frequencies as this, the unwanted components can be completely eliminated by means of suitable filters. However, it is of course desirable that as large a part as possible of the total output energy should appear in the wanted components. This result can be effected to a considerable degree by properly proportioning the values of the applied signal and carrier voltages.

Finally, it is worth noting that where greater output energy is required, each varistor can be made up of a number of individual units or discs connected in series or series-multiple. It is also true that there are several other possible configurations of varistor units which will give results as modulators similar to those discussed above. The particular arrangement of varistors to be used, as well as the number of discs required, is a matter of fundamental design which must be determined in relation to the circuit design as a whole.

173. Sources of Carrier Frequencies

In practically all of the carrier telephone systems now in use, the signal transmitted over the line consists of one of the side-bands alone. As we have just seen, the voice current can be obtained from this side-band only by the use of a demodulator circuit to which is applied the side-band and a carrier voltage exactly equal in frequency to that which was applied at the transmitting end of the channel.

Since the carrier frequency is not transmitted over the line, extremely reliable methods for generating carrier frequencies must be employed in order that the respective channel frequencies at the two ends of the line shall be exactly equal at all times. Vacuum tube oscillator circuits with the required degree of stability



Type-K Carrier Supply Panels—Carrier Generator in Center—120 kc. Amplifier and Filter Above—Transfer Circuit Below

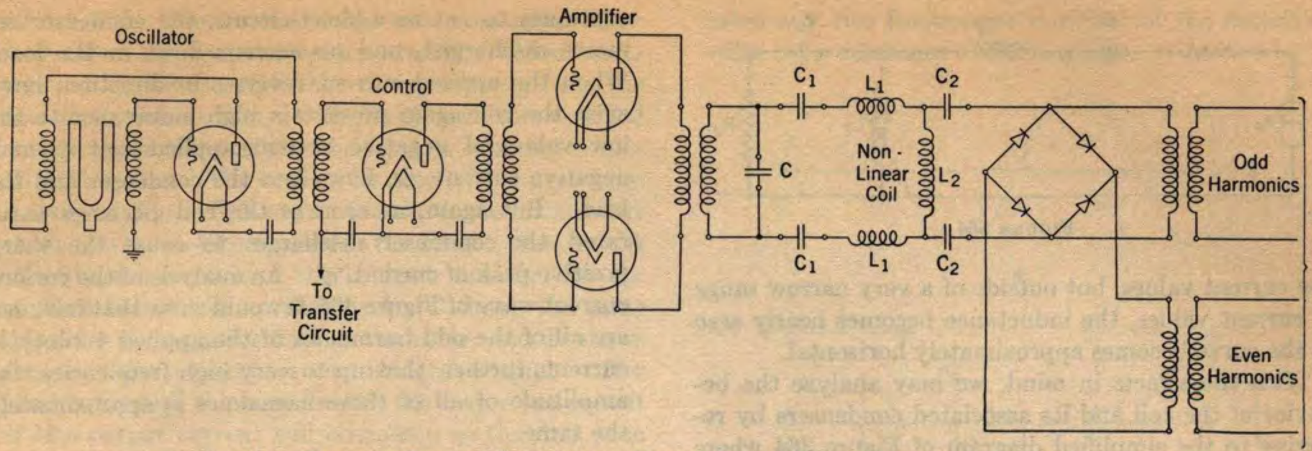


FIG. 362. CARRIER SUPPLY CIRCUIT FOR TYPES J AND K BROAD-BAND CARRIER SYSTEMS

in operation can be built, and such oscillators are in fact used in the lower frequency open wire carrier systems, such as Type-C. For the broad-band carrier systems, both open wire and cable, it is the practice to use a single base frequency generated by a tuning-fork controlled vacuum tube oscillator, and to obtain the various needed carrier frequencies from this base frequency by means of harmonic producing devices.

The essential circuits and apparatus units employed in the broad-band systems are shown schematically in Figure 362. The generated frequency of the oscillator tube is controlled in this case by a 4-kilocycle tuning-fork. This tuning-fork is made of an alloy having a low temperature coefficient, and its stability is such as to hold the frequency of oscillation accurate to within plus or minus one part in one million. The oscillator output is amplified in two stages to a value of about 4 watts by the control tube and two power tubes operating in a push-pull arrangement. The control tube also acts in conjunction with an auxiliary transfer circuit, not shown in the drawing, to automatically put into service an emergency oscillator circuit in case of failure of the regular circuit.

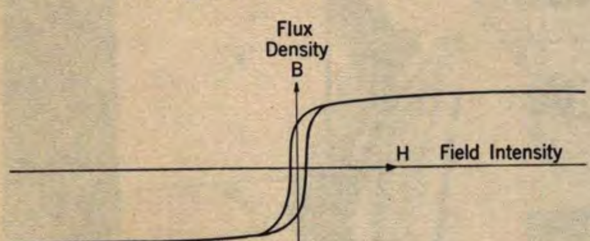
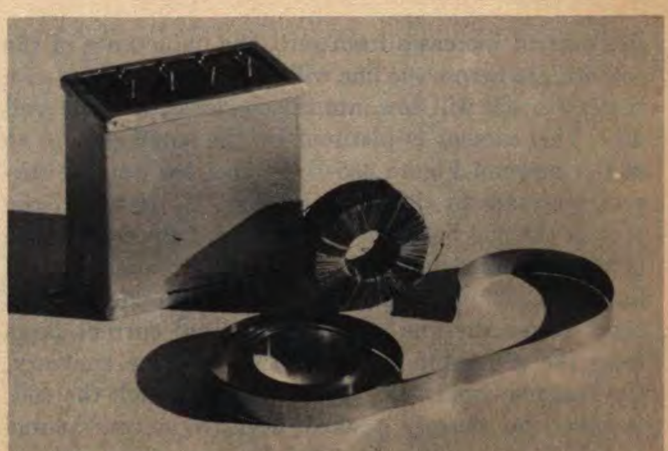


FIG. 363. B–H CURVE OF CORE OF HARMONIC PRODUCING COIL

The secondary of the amplifier output transformer and the parallel condenser C are designed to be anti-resonant at 8 kilocycles, and thus to short out any second harmonics developed in the amplifier tubes. The series condensers and inductances, C_1 and L_1, are resonant at 4 kilocycles and thus favor transmission of a pure sine wave of 4-kilocycle current to the bridged coil L_2. This latter coil, in conjunction with the condensers C_2, produces odd harmonics of the applied 4-kilocycle frequency. Its behavior in this respect offers a very interesting example of the use of the special magnetic material, permalloy, which was mentioned briefly in Chapter III.

The action of the coil as a harmonic producer depends upon the fact that its magnetic core becomes saturated at relatively low current values. The coil is physically quite small and is wound upon a core of coiled permalloy ribbon, as illustrated in the accompanying photograph. A B–H curve (refer to Article 30) for the core is given in Figure 363 from which it will be noted that the magnetic field passes from negative to positive saturation for a comparatively small change in the field intensity. (The small hysteresis loop indicated is of no importance in this application.) In other words, the curve shows that the coil becomes saturated very quickly and with a comparatively low value of current in its winding. Since the inductance of the coil is proportional to the slope of this curve, this means that the inductance has a high value for



HARMONIC PRODUCER COIL SHOWING PERMALLOY TAPE CORE

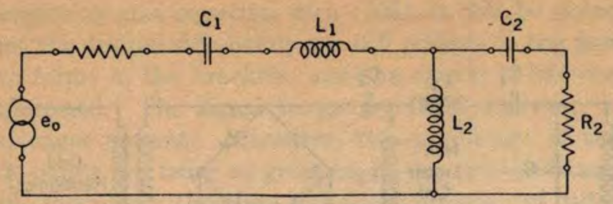


Figure 364

low current values, but outside of a very narrow range of current values, the inductance becomes nearly zero as the curve becomes approximately horizontal.

With these facts in mind, we may analyze the behavior of the coil and its associated condensers by referring to the simplified diagram of Figure 364 where all of the circuit to the right is indicated by the single load resistance R_2. To do this let us follow through what happens during a single cycle of the applied 4-kilocycle voltage e_0. One cycle of this applied voltage is shown in the usual manner in Figure 365–A. As

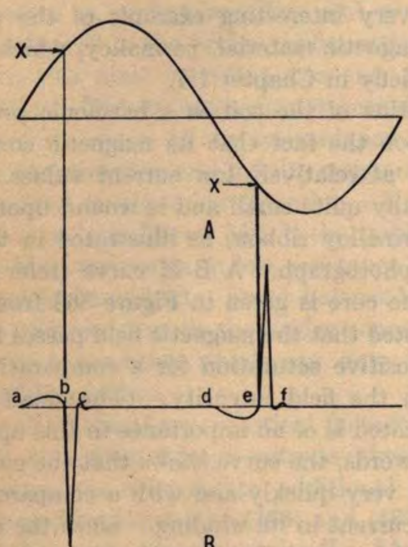
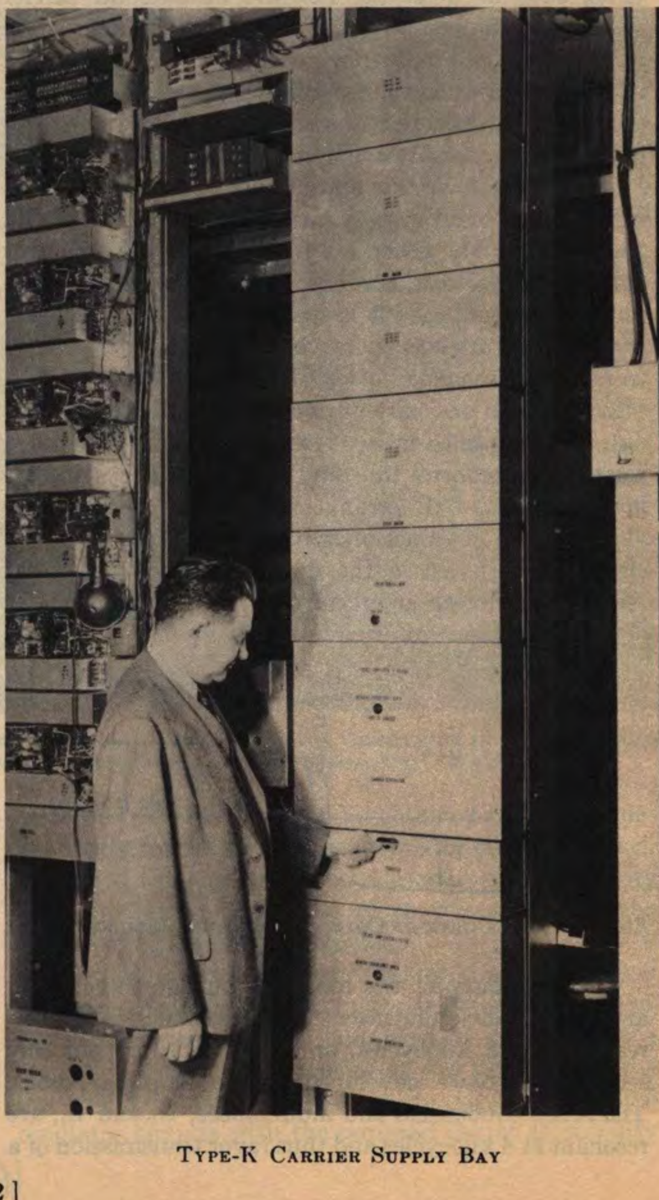


Fig. 365. Output Current of Harmonic Producing Coil Circuit

this current increases from zero, the inductance of the coil bridged across the line will at first be high and as a result, current will flow into the condenser and the load R_2. This current is pictured by the small section *ab* of the curve of Figure 365-B. When the applied current increases to the critical value, *X*, however, the core of the coil becomes saturated and the inductance of the coil immediately decreases to zero. As the coil has quite low resistance, it then becomes effectively a short across the line and no additional current flows from the generator into the load. On the contrary, the charged condenser C_2 discharges through the coil, causing the sharply peaked negative current surge shown in the section *bc*. For the remaining part of the positive pulse of the applied voltage, the coil continues to act as a short-circuit, the condenser remains discharged, and no current flows in the load. When the applied current reverses in direction, however, the coil again presents a high inductance to the low values of negative current applied and a small negative current, *de*, flows into the condenser and the load. But again, as soon as the coil becomes saturated, the condenser discharges to cause the sharp positive peak of current, *ef*. An analysis of the curious current wave of Figure 365-B would show that included are all of the odd harmonics of the applied 4-kilocycle current; further, that up to very high frequencies, the amplitude of all of these harmonics is approximately the same.

As indicated by Figure 362, these odd harmonics are separated for use in the various carrier channels by means of filters. Even harmonics are obtained by



Type-K Carrier Supply Bay

means of the Wheatstone bridge arrangement of varistors shown bridged across the circuit. These varistors rectify about half of the energy coming from the non-linear coil L_2 and, in the ideal case, this rectified energy appears entirely in the form of even harmonics of 4 kilocycles. The reason for this will appear from a study of Figure 366 where A is the curve for any one of the odd harmonics applied to the bridge, B shows the effective connections of the bridged varistors for the positive half of the currnte cycle, C shows the same thing for the negative half of the cycle, and D shows the resultant current in the bridged output. After passing through the transformer, the major component of this output current will obviously be the first even harmonic of the applied current and its other components will be higher even harmonics. The even harmonics are of course likewise selected for application to their particular carrier channels by means of filters. The complete separation of odd and even harmonics by the method described above tends to simplify the design of the selecting filters since it automatically separates any two frequencies in either of the output circuits by a minimum of 8 kilocycles.

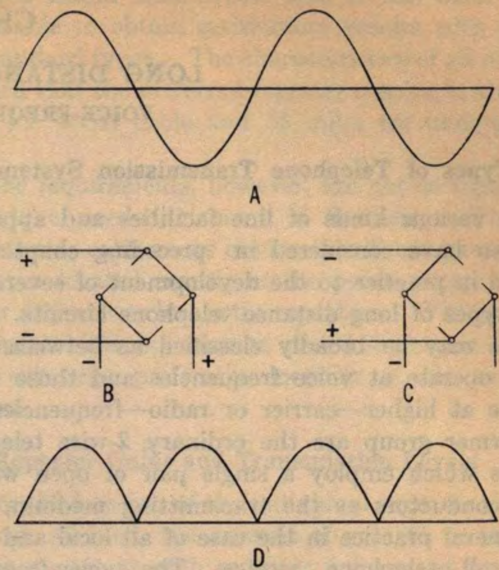


FIGURE 366

CHAPTER XXVIII

LONG DISTANCE TRANSMISSION SYSTEMS

VOICE-FREQUENCY TELEPHONE CIRCUITS

174. Types of Telephone Transmission Systems

The various kinds of line facilities and apparatus that we have considered in preceding chapters are applied in practice to the development of several distinct types of long distance telephone circuits. Such circuits may be broadly classified as between those which operate at voice frequencies and those which operate at higher—carrier or radio—frequencies. In the former group are the ordinary 2-wire telephone circuits which employ a single pair of open wire or cable conductors as the transmitting medium, as is the general practice in the case of all local and short haul toll telephone service. The voice-frequency group also includes the 4-wire cable circuits in which a separate pair of cable conductors is used for transmission in each direction, as was discussed in Article 165. If they are of any considerable length, both the 2-wire and 4-wire circuits require the insertion of telephone repeaters at regular intervals in order to maintain transmission at satisfactory levels.

Except in the case of the special coaxial type of conductor, carrier circuits employ the same, or the same kinds of, wire facilities for transmission as do the voice-frequency circuits. They also require the use of amplifiers or repeaters at regular intervals along the line. The use of radio circuits is limited in general to those situations where it is impossible or impracticable to build suitable wire facilities between the points in question.

It is not possible to make an unqualified statement as to the particular kinds of situations in which each of the above types of circuits may be best applied in practice. In general, however, the 2-wire circuits are commonly used for relatively short distances—in the order of 1000 miles maximum for open wire circuits, and 150 miles maximum for cable circuits. Four-wire cable circuits are used for somewhat longer distances—up to something in the order of 1200 miles when equipped with "echo suppressors"—and carrier circuits are used for the longest distances. The length of radio circuits of course depends entirely upon the distance which the geographical requirements make it necessary to span.

In this and the following two chapters, we shall consider some of the more essential problems involved in the design and operation of the long voice-frequency and carrier telephone circuits. Because of the special types of apparatus used and their relatively limited application in long distance telephone work, radio circuits will not be discussed in this book. It may be pointed out, however, that a survey of the applications of electrical principles to carrier work may be distinctly helpful in connection with the study of radio transmission.

175. Repeater Spacing

The different types of circuit facilities, whose principal characteristics are discussed in Chapter XXII, show a wide diversity in their relative transmission efficiencies. It may be noted, for example, that at voice frequencies a loss greater than one db is caused by one mile of 19-gage non-loaded cable side circuit while a 165 open wire phantom circuit causes a loss of only .025 db per mile. In other words, one mile of the former gives rise to as great a loss as nearly forty miles of the latter.

Prior to the advent of the telephone repeater in 1915, large gage open wire facilities were used for all very long circuits; furthermore, such facilities were usually loaded. But even with the use of loaded 165 facilities, the maximum practicable range for long distance telephony was limited to about two-thousand miles. The application of the telephone repeater had two fundamental and far reaching effects; first, it made possible an indefinite extension of the maximum range of telephonic communication; and second, it permitted smaller wire gages for long distance service and so made economically and physically feasible the great expansion in the number of long distance circuits that has occurred since its introduction.

The first transcontinental telephone service was furnished by loaded 165 open wire facilities with repeaters inserted at 500 to 600-mile intervals. This same service was later improved by removing the loading from the open wire facilities and reducing the repeater spacing. Here the repeater served another purpose; it improved the quality of the circuit by making possible the elimination of the inherently troublesome open wire loading. Repeaters are now used in practically all long distance cable and open wire circuits. Since open wire facilities must for mechanical reasons be of relatively large gage and suspended with considerable separation between conductors, their resistance and capacity values are relatively low. As a result repeaters need only be spaced at intervals of the order of 150 to 350 miles to compensate for the energy attenuation caused by the conductors. This means that even in the longest circuits the number of repeaters in tandem is not very great. On the other hand, in cable facilities the conductors are usually of 16 or 19-gage and even though loading is used, repeaters must be inserted at 50 or 100-mile intervals, depending on the gage of the conductors. It follows that a very long cable circuit must include a considerable number of repeaters in tandem. In either case, it is the usual practice to employ repeaters at the terminals as well as at intermediate points along the circuit.

In an open wire circuit, the 165 wire will generally have the best electrical and mechanical characteristics, with 128 wire circuits next, and 104 wire circuits last. This is due largely to the differences in attenuation. By increasing the repeater spacing when larger wires were used, these three types of circuits could be made practically identical from a transmission standpoint. However, most open wire lines include different sizes of wire and the spacing of the repeaters is determined by the losses of the smallest wires. Nevertheless, even though short repeater sections are used, climatic conditions may be such as to put occasional severe strains on the wires and thus necessitate the use of 128 wire or even 165 wire to obtain greater mechanical strengths than is possible with 104 wire. The final decision as to the size of the open wire, and hence the repeater spacing, must be based on the proper consideration of both the economic and electrical factors. The latter include repeater balance, transmission variation due to temperature and other weather changes, and echo effects, all of which are discussed in following articles. In practice it has been found that in most cases the repeater spacing on open wire facilities should not exceed 350 miles for 165 wire, 225 miles for 128 wire, and 150 miles for 104 wire.

The latest types of long toll cables for voice-frequency use employ 19 and 16-gage conductors. The former gage is used for both 2 and 4-wire circuits, while the latter is used for program services (radio broadcasting networks) and to some extent for 2-wire message circuits. These conductors are loaded to reduce their attenuation and thereby permit longer repeater spacing. The type of loading used depends somewhat upon the lengths of the circuits and the uses that are made of them. Although this might imply that it is desirable to have a different type of facility for each length and circuit use, it has been found practicable to obtain satisfactory results with only a few standard types. The characteristics of all of these are such that the preferred repeater spacing is about 50 miles for aerial cable and 55 miles for underground cable.

These requirements, however, are not so rigid as to preclude a needed element of flexibility. Thus an open wire or cable route obviously will not have towns and cities located exactly at the points where it may appear desirable to locate the repeaters. Within limits, the repeater spacing may be varied somewhat to conform with the preferable location of the repeater stations.

176. Repeater Gains and Transmission Levels

After the location of the individual repeater stations has been selected, the amount of gain to be inserted in each circuit at each repeater point must be determined. It is generally desirable to keep the energy of the message currents at the highest possible level in order to reduce the possibility of noise interference. If the transmission level of the message currents is too low, any small noise currents that may be induced into the circuit from external sources may be relatively great enough to cause excessive interference when they are amplified by the repeaters along with the message currents. On the other hand, by keeping the transmission level of the message currents high with respect to the level of the induced noise currents, the effect of the latter is minimized. However, it should not be forgotten that there is a limit to the amount of energy

Open Wire Repeater Station at Kingman, Arizona that any particular amplifying circuit can handle, and even before reaching this limit, distortion is introduced.

In adjusting the gains of 22-type repeaters, the ordinary limits are as follows: With the volume of transmission at the switchboard at the sending terminal of a circuit defined as "zero transmission level", a 22-type repeater having B battery of constant voltage may be operated to deliver a volume of transmission not exceeding the zero level by more than 6 db for 22-A-1 repeaters or 10 db for 22-B-1 types, but no repeater, regardless of spacing, should be expected to give more than an 18 db net gain. However, these limiting factors apply only under ideal conditions. If, for instance, there is a fluctuating B battery voltage, as in the case of the ordinary telegraph battery, the zero level should not be exceeded by more than one db.

In the 44-type repeaters used in 4-wire cable circuits, much higher gains are possible. It may be recalled that these repeaters have two stages of amplification and it is important that the gain be so adjusted between the first and second stages as to not overload the first stage. When so adjusted, the two stages will give a combined gain of about 50 db without appreciable distortion. However, crosstalk considerations usually prevent the attainment in practice of this maximum. It is permissible to operate 44-type repeaters between an input volume not lower than about 25 db below zero level and a delivered output not greater than about 10 db above zero level. This means a possible gain of 35 db, under which condition the energy delivered is nearly two-thousand times as great as the energy received. This extreme energy ratio is the reason for the crosstalk limitation. If we imagine a case where an incoming cable pair is adjacent to an outgoing cable pair of any other 4-wire circuit and there exists a small crosstalk unbalance from one pair to the other, the highly energized circuit may transfer a quantity of energy which although an almost negligible fraction of its own energy, may nevertheless be quite appreciable as compared with the energy in the other circuit, which is only about 1/2000th as great in value. This crosstalk energy is applied to the repeater with the incoming transmission and is amplified along with and to the same degree as the incoming transmission, thereby tending to become audible.

In laying out long circuits containing a number of repeaters in tandem, use is made of an "energy level diagram" which shows in a single chart not only the losses in each line section and the gain of each repeater, but also the level of the voice energy at each point along the circuit as compared with the energy originally applied to the circuit terminal (zero level). Figure 367 gives such a diagram for a typical long 2-wire circuit on open wire facilities. The ordinates represent energy levels in decibels above and below zero level, losses being measured downwards and gains upwards. The gains of the repeaters are naturally represented by straight vertical lines, while the line attenuation losses are indicated by lines between repeater stations sloping downward in the direction of transmission. A separate set of zigzag lines is required to show transmission in each direction, even when the net equivalent of the

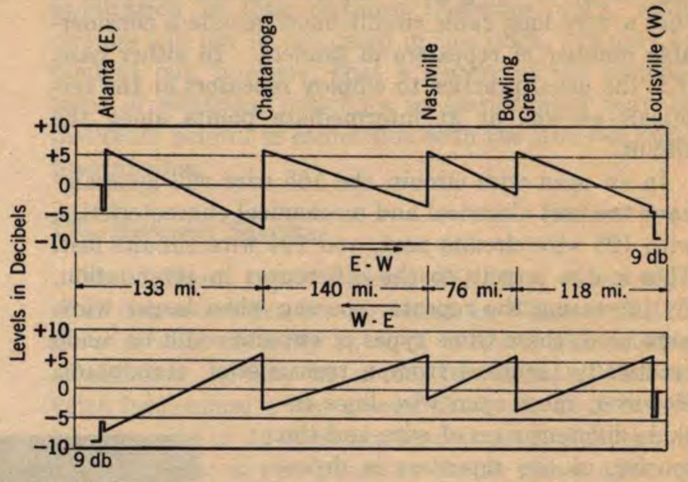


Fig. 367. Energy Level Diagram for Open Wire Circuit

circuit and the gains of each repeater are the same in each direction. Such a chart is valuable not only in facilitating the original engineering design of the circuit, but also as maintenance information to enable the repeater stations to know both the gain at which each repeater is to be operated and the proper output energy level for transmission in each direction.

While it is customary to describe the gain of a telephone repeater by a single value in decibels, this is to be understood to mean the gain of the repeater at only the single frequency of 1000 cycles. At any other frequency the gain of the repeater may or may not be the same value. In order to completely describe the characteristics of a repeater and to analyze its entire effect when inserted in a circuit, it is therefore necessary to know the gain of the repeater not only at 1000 cycles but at all frequencies through the normal voice range. This information is usually given in the form of a curve between frequency and gain, known as the "gain-frequency characteristic" of the repeater.

Figure 368 shows the gain-frequency characteristic of a 2-wire (22-type) repeater, set for maximum gain at 1000 cycles, with four different filters in the output circuit. Curve A is the characteristic of a type of repeater that has been generally employed on open wire circuits. This shows a nearly constant gain for frequencies between 300 and 1700 cycles, a gradually decreasing gain for frequencies between 1700 and 2300 cycles, and a sharply decreasing gain for frequencies

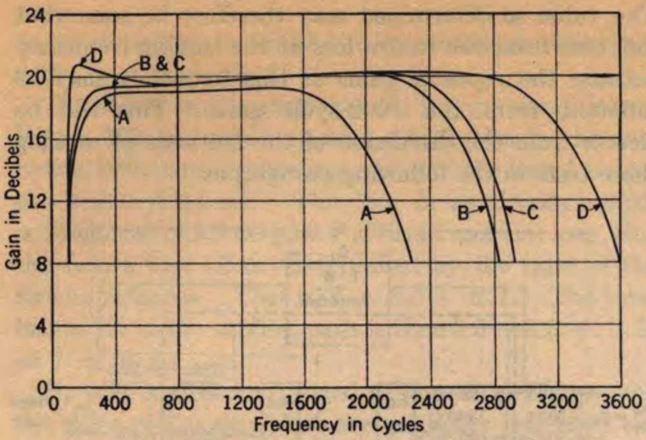


FIG. 368. GAIN-FREQUENCY CHARACTERISTICS OF 22-TYPE REPEATERS

below 300 cycles. Contrasted with this, Curves B, C and D are the gain-frequency characteristics of later types of repeater circuits which employ filters having a much wider pass band. These are now generally used on both cable and open wire circuits. The repeater with the characteristic illustrated by Curve A, since its range of uniform amplification includes the most important of the voice frequencies, may be used on relatively short circuits for ordinary service without causing serious impairment in quality. The repeaters of Curves B, C and D, with their fairly constant amplification over a much wider band of frequencies, naturally cause less distortion and are more satisfactory for use on longer circuits. The repeater of Curve D is, of course, the most desirable when conditions permit its use.

As we shall see in the next article, however, the use of wider frequency bands complicates the problem of balance somewhat because the balancing networks associated with the repeaters must balance the lines over a correspondingly greater range of frequencies. Cases may occur where impairment of quality on account of unsatisfactory balance conditions is liable to be more severe than that caused by a lack of uniformity in the gains over the wider frequency range. In such cases the use of repeaters having gain-frequency characteristics such as Curve B, or even Curve A, may be preferable to those of Curves C or D.

Moreover, it is important to keep in mind that the gain at the various frequencies passed by the filters in the output circuit, may be changed within certain limits by changing the equalization at the input transformer (see Article 143). All the curves in Figure 368 use the same equalization arrangement at the input, and they represent only the effect of using different filters in the output circuit.

In the case of 4-wire circuits, the gain-frequency characteristics of the 44-type repeaters are made approximately equivalent to the loss-frequency characteristics of the line and equipment in the preceding repeater section by the use of equalizing networks in the input circuit (see Article 165). That is, if the line loss is higher for certain frequency ranges, the gain is also higher, and vice versa.

177. Return Loss

In 2-wire circuits, repeater gains are usually limited by the degree of balance which it is possible to secure between each line and its balancing network, rather than by the maximum energy output of the amplifying tubes. In other words, the allowable amplification of a 22-type telephone repeater depends upon the gains that make the repeater circuit oscillate or "sing", or appreciably impair quality because of unbalance between the line and the associated network. Definite impairment of quality is quite noticeable just before the "singing" point is reached.

As already seen from our study of the hybrid coil in Article 117, if identical impedances are connected to the line and network terminals of the hybrid coil, no power can pass from the series winding to the bridge taps; in other words, there is infinite loss across this path. If, however, there is an inequality between the line and network impedances, power can pass and a finite loss may be measured between these points. This loss which we may designate as L, is made up as follows: the power first divides about equally between the line and network giving a loss of 3.25 db (3 db for

CABLE REPEATER STATION AT WAUKEGAN, ILLINOIS the power division loss and 0.25 db for the coil loss). Assuming the line impedance to differ from the network impedance, a portion of the power entering the line will be reflected back towards the hybrid coil. The part so reflected back is less than the power sent out on the line by the amount of the so-called "return loss" ($R.L.$), the magnitude of which is determined by the relation between the line impedance and the network impedance. The greater the departure of the line impedance from its normal value (which the network simulates), the more the power reflected back, and the smaller the return loss. This reflected power enters the hybrid coil in the same manner as normal incoming transmission and in the same way divides equally between the bridge taps and the series winding, thus incurring another 3.25 db loss. The total loss L between the series winding and bridge taps is then $3.25 + 3.25 + R.L.$, or $R.L. + 6.5$. Thus, **the return loss at a given frequency is the measured transmission loss across the hybrid coil at that frequency, less 6.5 db.**

The value of the return loss is thus a measure of the similarity between the line and network impedances, and is the kind of quantity "singing point tests" are designed to measure to a certain approximation. Its value in db may be determined by the formula—

$$R.L. = 20 \log_{10} \frac{Z_N + Z_L}{Z_N - Z_L} \qquad (168)$$

where Z_N is the impedance of the network and Z_L is the impedance of the line circuit. If the network perfectly balances the circuit, that is, if $Z_N = Z_L$, then Equation (168) shows that the return loss is infinite. When an unbalance exists, the loss takes a finite value. As an example let us assume Z_N to be 600 ohms, and $Z_L = 400$ ohms, then from Equation (168)—

$$R.L. = 20 \log \frac{600 + 400}{600 - 400}$$

$$= 20 \log 5 = 20 \times .7 = 14 \text{ db}$$

Since both Z_N and Z_L may vary with frequency, a return loss measurement or computation must be made in terms of a single frequency, and the gains and losses in the measuring circuit expressed for the particular frequency used. Such measurements or computations must be made for a number of frequencies in the voice range if it is desired to determine at what point in the range balance conditions are worst.

In singing point tests, however, the repeater automatically selects the frequency most favorable for singing, and thus in a single measurement gives the approximate balance condition at the worst frequency. In this case, the gains of the repeater used in making the tests are ordinarily measured at 1000 cycles and for convenience these 1000-cycle values are used in determining the numerical value of the singing points. The value so determined may therefore be somewhat different from the return loss at the singing frequency because the repeater gains at this frequency may be different from the 1000-cycle gains. This will be clearer from the discussion of the methods of making these tests in the following paragraphs.

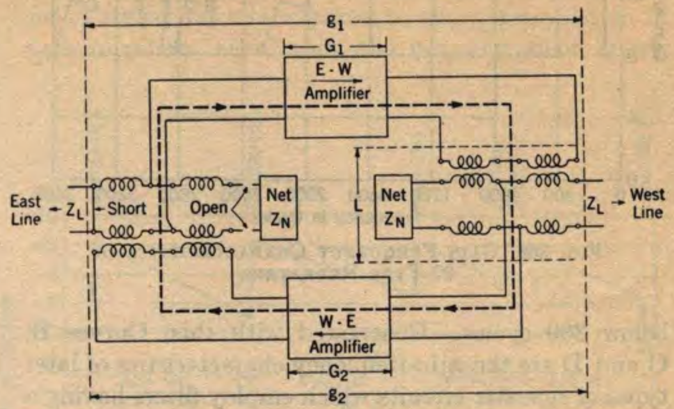


Fig. 369. Principle of Repeater Balance Test

From the foregoing discussion, the singing path of the 22-type repeater circuit as indicated by the heavy dashed line in Figure 369 may be readily understood. Let us assume that the degree of electrical balance between the impedances Z_L and Z_N connected to the line and network terminals of the west hybrid coil is under test. The line and network terminals of the east hybrid are shorted and opened respectively, thus converting the hybrid coil into a simple repeating coil. If the impedances Z_L and Z_N are unequal, the loss across the west hybrid coil will be finite and power can be transferred across this coil, through the W-E amplifier, through the east hybrid (now a repeating coil) to the E-W amplifier, through this amplifier, and back to the west hybrid coil. A round trip path is thus established and as soon as the sum of the gains of the two amplifiers becomes equal to or slightly larger than the total losses in the path, a sustained circulating current will be set up.

The only gains in the singing path are those of the two amplifier units. These gains may be designated G_1 and G_2. In making singing point tests, it is impracticable to measure the gains of the individual amplifiers. Instead, measurements are made from one line to the other, that is, from the line terminals of one hybrid coil to the line terminals of the other hybrid coil. These are the "calibrated gains" and may be designated g_1 and g_2. In going from one line to the other, the power suffers a 6.5 db loss, due to the two hybrid coils. The gain as measured between line terminals in one direction is therefore 6.5 db less than the gain of the amplifier unit itself. That is, $G_1 = g_1 + 6.5$. The total gain in the singing path is merely the sum of the gains of the two amplifying elements, $G_1 + G_2$, or, expressed in terms of the measured or calibrated gains, $g_1 + g_2 + 13$.

The losses in the singing path for the condition of Figure 369, are made up of a 0.5 db loss in the east hybrid coil (now a repeating coil) and the loss L across the west hybrid coil. This loss L, as already noted, is composed of 6.5 db power division and coil loss, plus the return loss ($R.L.$) determined by the ratio of the two impedances. That is, $L = 6.5 + R.L.$ The total loss in db in the singing path is then $0.5 + 6.5 + R.L.$ or $7 + R.L.$

If, with a circuit arranged as shown in Figure 369, the gains of the amplifiers are gradually increased the circuit will start to sing at the point where the total gains in the singing path become equal to the total losses. That is when

$$g_1 + g_2 + 13 = 7 + R.L. \qquad (169)$$

Furthermore, the circuit will automatically select the frequency at which singing will occur. This is obviously the frequency at which the return loss has the lowest value, assuming that the gains of the amplifiers are constant through the transmitted frequency range. Equation (169) may be solved for the return loss to give

$$R.L. = g_1 + g_2 + 6 \qquad (170)$$

178. Singing Points

In Equation (170) the calibrated gains, g_1 and g_2, are assumed to be measured at the singing frequency, so that the substitution of these values in the equation would give the return loss at that frequency. In making practical singing point tests, however, the repeater gains are measured at 1000 cycles as a matter of convenience. When these 1000-cycle gains are substituted in Equation (170), the result will be something other than the true return loss. The difference will depend on the difference between the 1000-cycle and the singing frequency gains, and hence on the gain-frequency characteristic of the measuring repeater. The singing point, $S.P.$, so measured, may be represented by an equation similar to (170)—

$$S.P. = g_1 + g_2 + 6 \qquad (171)$$

in which, however, g_1 and g_2 now represent the 1000-cycle calibrated gains of the amplifiers when the circuit is adjusted to the point where singing begins.

As noted, singing points may differ from return losses due to the amplifier characteristics. In addition, phase relationships may sometimes be such as to prevent singing from occurring at the frequency where the balance conditions are poorest. However, singing point tests give results sufficiently accurate for practical maintenance purposes. Commonly known as "21 Circuit Balance Tests", they provide a ready means of ascertaining what is the maximum safe working gain of a 22-type repeater when connected to a given circuit. The measurement also gives a direct check on the effectiveness of the network balance, since a high singing point means that at no single frequency within the voice range is there an appreciable dissimilarity between the impedance of the network circuit and the impedance of the line and its associated equipment.

Such a satisfactory balance between a line and its network depends, among other things, upon the termination of the line at the next adjacent repeater point. When making tests this termination may consist of a network or of a "passive repeater"—that is, a repeater so arranged as to present its nominal impedance to the circuit, but with its gains set at zero so as to prevent irregularities in succeeding repeater sections from causing reflected currents to return to the test repeater. A repeater may also be made passive by replacing with balanced resistances the line and network connected to the hybrid coil on the side of the repeater away from that to which the circuit under test is connected. The balance measured under this condition is called the **passive singing point,** which means fundamentally that the test repeater is the only repeater in the circuit that amplifies the reflected power, or that only one amplification path is involved.

Now suppose that instead of being terminated at the adjacent office in a network or in a passive repeater, the circuit at that office goes through an active repeater (one in operating condition) and on to another repeater section beyond. The reflected power in the first section will still return to the test repeater, but in addition, part of the sent power will enter the second repeater, be amplified and sent into the second repeater section; if the second section contains irregularities, part of the power entering this section will be reflected back to the second repeater, through this repeater and into the first section, and then back to the hybrid coil of the test repeater, thus adding to the power returned from the first section. There are now two points in the circuit where reflected power is amplified. In other words, there are two amplification paths and with more repeaters in the circuit, there may be a third and fourth path, etc. All of these returned powers combine at the hybrid coil of the test repeater to enter the circulating path of this repeater. The greater this total power, the less the gain required to sustain singing and, accordingly, the lower the singing point. The balance for this condition is termed the **active singing point,** meaning, as already indicated, that one or more repeaters, in addition to the test repeater, amplify reflected powers; that is, two or more amplification paths are involved. Since these are the prevailing conditions when the circuit is in operation, the value of this active singing point is of fundamental practical importance.

179. Balancing Networks

From the preceding discussion it will be noted that in 2-wire repeatered circuits we are concerned as much with the impedance of the line as with its attenuation. The extent to which the repeater may improve transmission depends directly upon the degree to which the network balances the line. In turn the degree of balance depends first upon the "smoothness" of the telephone line's impedance throughout the working range of voice frequencies, and second, upon the adjustments that it is practicable to make for the effect that terminating conditions have upon this impedance.

The basic requirements as to balance may be understood by referring to Figure 370. Here we have the R and X components of the characteristic impedance,

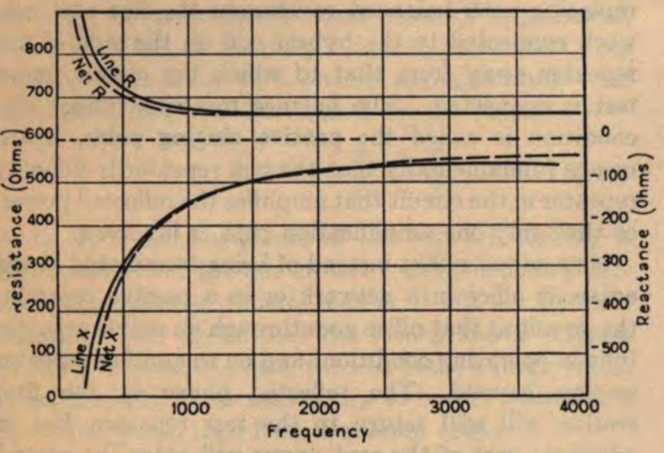


Fig. 370. Impedance Characteristics of Open Wire Circuit and Its Balancing Network

Z_0, of a 104 open wire side circuit plotted (solid line) with respect to the voice-frequency band. It will be seen that the resistance component of the characteristic impedance becomes appreciably lower at the higher frequencies and that there is likewise a marked change in the value of the negative reactance.

Now to balance such a circuit, a network must be designed with impedance components that not only equal those of the line at some one frequency, but vary similarly with the impedance of the line at all frequencies within the voice band. The dashed curves in this same figure compare the R and X components of the impedance of the standard network used to balance this type of line.

The schematic design of the basic network for an open wire circuit is illustrated by Figure 372-A. This simple arrangement, with proper values of resistance and capacity, will very closely approximate the impedance components of the line itself. It will not, however, take care of near-end terminating conditions such as toll entrance cable, etc. Furthermore, it balances only the characteristic impedance of the circuit, i.e., the circuit must be in effect infinite in length; or in other words, terminated at the distant end in an impedance equal to the characteristic impedance. Consequently, balance, even in the open wire circuit case, involves considerations other than the mere design of a basic network that has an impedance approximating that of the characteristic impedance of the line. These balance requirements, however, are general and will be discussed after considering the basic network for the loaded cable circuit.

A basic network for a loaded circuit usually has a more complex design than a basic network for a non-loaded circuit. In this design some assumption must be made regarding the loaded circuit's near-end termination, i.e., the basic network must be chosen to balance a loaded circuit terminating at a mid-section point, or at some fraction of the loading section other than mid-section. Figure 371-A shows the resistance components of the impedance of an ideal loaded line having no resistance, for various forms of termination, the frequency band being that up to and including the critical frequency. (The scale for frequency is shown as fractions of the critical frequency rather than as cycles in order that the curves may apply to any case.) Figure 371-B shows the corresponding reactance components.

An inspection of Figure 371-A shows that for a .2 or .8 section termination, a plain non-inductive resistance will approximate the resistance component of the circuit, as this resistance component remains nearly constant through the band of frequencies that the loaded circuit would be expected to transmit. This is true only for these two terminating conditions. Accordingly, if we choose the .2 section sending-end termination as that for which the basic balancing network is to be designed, we only need to connect in series with a resistance some combination of inductance and capacity that will approximate the corresponding reactance component shown in Figure 371-B in order to obtain a network which will simulate almost exactly the ideal loaded line; and, except at very low frequencies where the resistance of the actual line causes the impedance to depart appreciably from that of the ideal line, will closely approximate an actual loaded line. This combination is found to be a capacity value in parallel with an inductance value. A schematic diagram of a simple basic network for a loaded circuit at .2 section termination is shown in Figure 372-B.

For balancing loaded cable circuits in practice, more exact simulation at the low frequencies is usually required than can be obtained with this simple network,

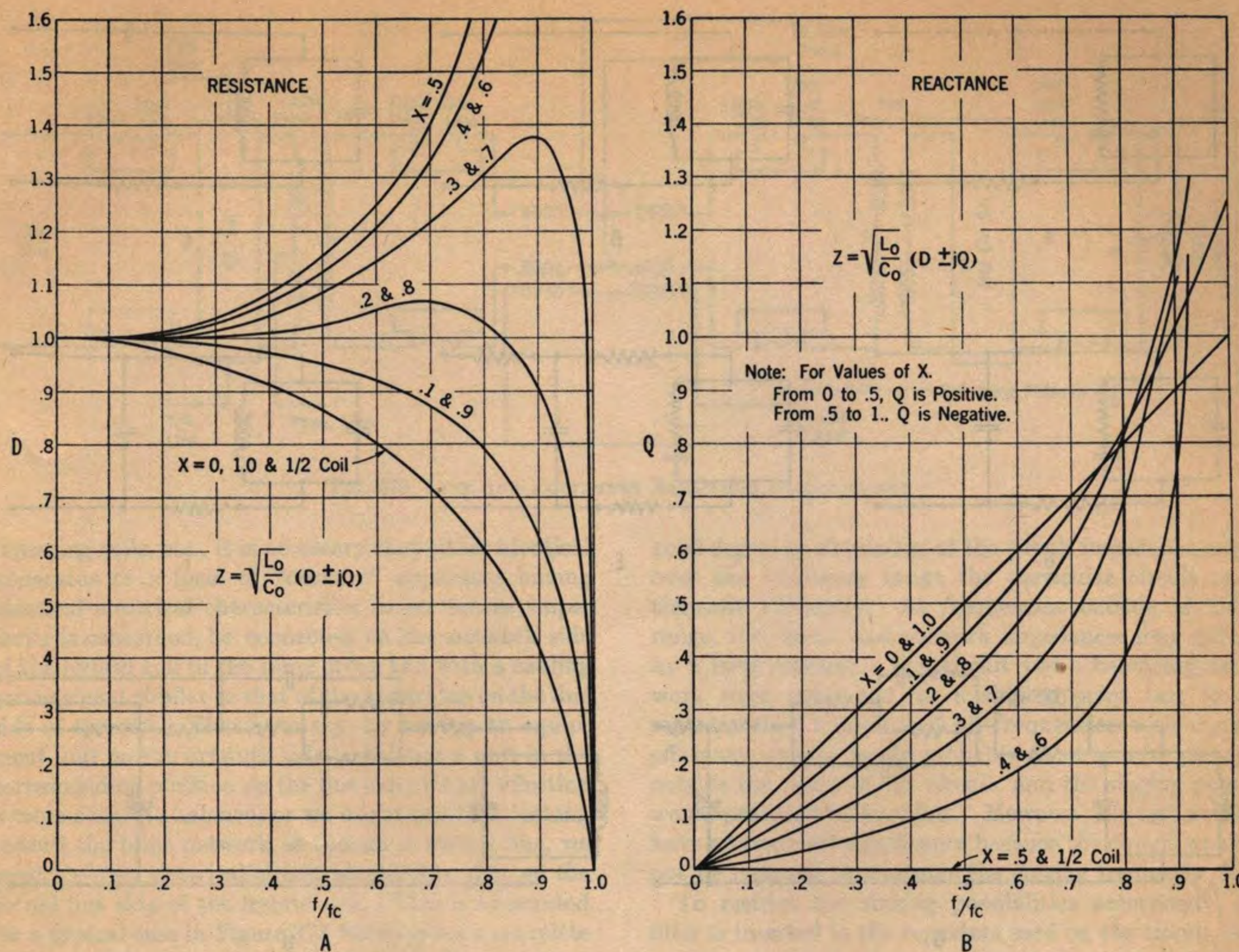


Fig. 371. Impedance Components of Ideal Loaded Circuit at Various End Sections

and networks of somewhat more complex design are employed. Figure 372-C illustrates the design of the standard type of network used for balancing certain heavier loaded types such as 16 and 19-gage H-174-S and H-63-P cable circuits. The design of the standard type of network used for balancing 19-gage H-44-S and H-25-P cable circuits at .2 section is the same as that shown by Figure 372-B. A somewhat more elaborate arrangement is used for balancing H-88-50 and B-88-50 facilities. Figure 372-G gives the schematic circuit of the former network. The inductance, capacity and resistance values are of course different for each different type of facility.

Balancing networks are usually so encased as to be mounted on coil racks or relay racks and are similar in their external appearance to repeating coils when so mounted. They are designated in a manner that permits them to be easily identified as balancing networks and not mistaken for repeating coils.

As stated in the foregoing, the basic network is only intended to balance the characteristic impedance of a smooth line of infinite length in the case of open wire circuits, or the .2 section termination sending end impedance for a smooth line of infinite length in the case of cable circuits. But the actual sending end impedance of the circuit may vary widely from the particular impedance which the basic network is designed to balance due to various reasons:

a. At the repeater station open wire circuits may be brought in through toll entrance cable, etc.
b. In the case of loaded cable circuits, the termination may not be at the .2 section point.
c. Circuits may be equipped with terminating apparatus such as composite sets, repeating coils, composite ringers, etc.
d. Circuits may have irregularities due to intermediate submarine cables, etc.
e. The terminations of the circuits at the distant end may introduce irregularities which do not permit them to act as circuits of infinite length.

It is the practice to make adjustments on the network sides of the telephone repeater's hybrid coils to take

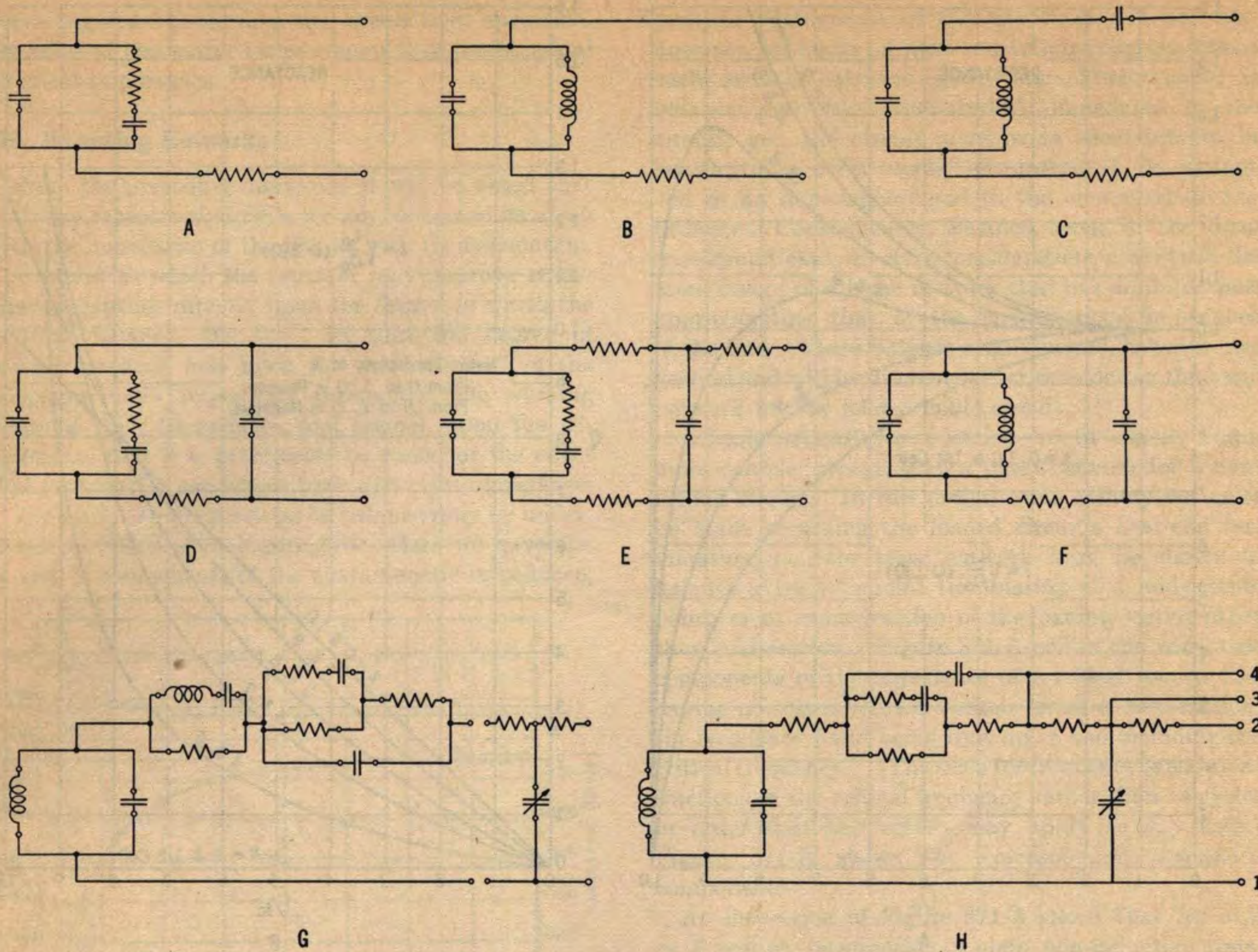


Fig. 372. Standard Balancing Network Circuits

care of (a) and (b) above by means of building-out sections. That is, if an open wire circuit has a short section of non-loaded toll entrance cable, there will be a capacity value equal to that of the capacity of this section, bridged directly across the network as illustrated by Figure 372-D. If it has a long section of toll entrance cable, it may be necessary to compensate for the resistance as well as the capacity, and accordingly, a building-out section with both a resistance and condenser, as shown in Figure 372-E, may be used.

Similarly, in the case of the loaded cable circuit, if the capacity on the office side of the last loading point is greater than that corresponding to .2 loading section, it is necessary to build out the basic network to adjust for this capacity, as shown in Figure 372-F. If, on the other hand, the circuit should be so terminated that the capacity from the office side of the last loading coil was less than that of .2 loading section, it would be necessary to add bridged capacity to the line of such value as to make the termination equivalent to .2 of a section.

In determining the proper value of building-out section to be used in any particular case, not only must we consider the capacity of the last section of the cable itself but all office cabling from the protectors to the first apparatus unit installed on the circuit. That is, to the capacity of the cable terminated at the distributing frame, must be added the capacity of the office cabling from the frame to the testboard and from the testboard to some equipment unit such as a composite set.

In one of the more recent designs, the basic network and the associated building-out section are assembled as a single piece of apparatus and provision is made for varying certain of the resistance values of the composite unit. Such a network, suitable for balancing circuits of the H-88-50 type, is illustrated by Figure 372-H. The network terminals are 1 and 2, and the resistance values may be varied by suitable strapping between terminals 2, 3 and 4.

In order to balance any apparatus that may be associated with the circuit such as composite sets,

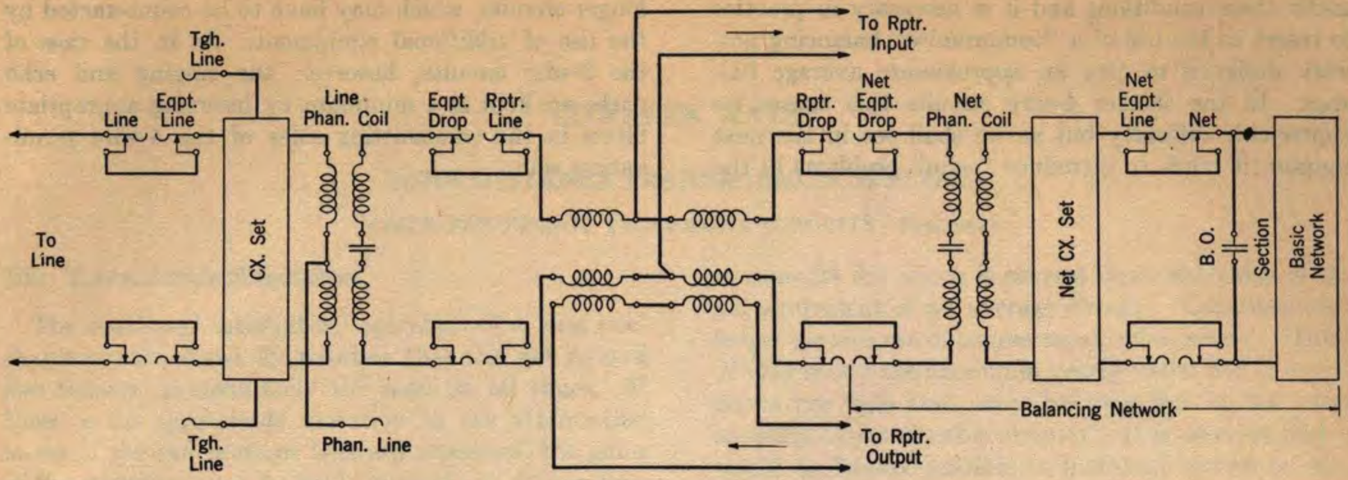


FIG. 373. LINE AND EQUIPMENT BALANCING ARRANGEMENT

repeating coils, etc., it is necessary that either identical apparatus or a form of "dummy" apparatus having identical electrical characteristics in so far as impedance is concerned, be connected on the network side of the hybrid coil in the same order and with a cabling arrangement similar to that of the apparatus on the line side of the coil. That is to say, by having an equipment unit on the network side to balance a unit in the corresponding position on the line side, we are effecting a more accurate balance; or we might say that looking toward the basic network as though it were a line, we must have an electrical circuit identical to that on the actual line side of the hybrid coil. This is illustrated for a typical case in Figure 373 which gives a complete diagram of the network and the line circuits.

Referring now to (d) above, it is not usually feasible to make any network adjustments to compensate for irregularities in the line other than near-end terminating ones. Such irregularities must be dealt with by actually clearing them if they are due to some line trouble, by building-out the capacity of the line as described in Article 137, or by working the telephone repeaters at correspondingly lower gains. The location and seriousness of line irregularities can be determined by a series of tests that are described in Chapter XXXIII. The remaining reason given in the foregoing for the unbalance between network and line is the effect of the termination at the distant end of the circuit, or in other words, the condition where the line does not act as though it were infinite. In some cases this is a situation that must be tolerated, with a resultant decrease in the maximum permitted repeater gain. It may be minimized, however, by employing repeaters whose passive impedance closely approximates the characteristic impedance of the circuit.

Some circuits transmit a wider band of frequencies than others, and in general a network is designed for a good degree of simulation of the circuit impedance only over the frequency range the particular circuit can transmit efficiently. At frequencies outside of this range, the circuit and network impedances may differ by a large amount. If a circuit and a balancing network were connected for a singing point test to a repeater that transmitted all frequencies with equal efficiency, singing would probably occur at a frequency outside the range of the circuit, and the singing point would probably be very low. However, the test would have no practical significance because the circuit would not be required to transmit the singing frequency.

To restrict the singing possibilities accordingly, a filter is inserted in the repeaters used on the circuit, so as to suppress the frequencies outside the range the circuit is designed to transmit. Singing in the suppressed range is then impossible and a singing point test made with a repeater equipped with the proper filter would accordingly give a satisfactory indication of the balance condition. When repeaters of the same type are equipped with different types of filters, depending on the type of circuit involved, it is important that the repeater used for singing point tests should have the proper filter for the particular kind of circuit under test.

In the case of 4-wire circuits, the problem of balance is no greater in magnitude than that involved in a 2-wire circuit equipped with a single repeater. This is true because the 4-wire circuit, however long it may be, employs only two hybrid coils (4-wire terminating sets). The only balance required, accordingly, is that between the 2-wire terminal of the circuit and its associated network. Unfortunately, the impedance of this 2-wire terminal is usually rather a variable quantity because at various times it may be connected to different types of local trunks or subscribers' loops. Any great precision of balance is of course impossible under these conditions and it is necessary in practice to resort to the use of a "compromise" balancing network designed to give an approximate average balance. In the shorter 4-wire circuits this causes no appreciable difficulty but as we shall see in the next chapter, it tends to introduce "echo" problems in the longer circuits, which may have to be counteracted by the use of additional equipment. As in the case of the 2-wire circuits, however, the singing and echo paths are kept at a minimum by inserting appropriate filters in the transmitting sides of the 4-wire terminating sets.

CHAPTER XXIX

LONG DISTANCE TRANSMISSION SYSTEMS

VOICE-FREQUENCY TELEPHONE CIRCUITS—*Continued*

180. Transmission Regulation

The continued satisfactory operation of a long telephone circuit obviously requires that the net overall loss remain approximately the same at all times. If there is no appreciable variation in the attenuation losses of the line sections between repeaters, the gains of the repeaters must be held constant; or, failing this, any variations in attenuation must be promptly compensated for by equal variations in repeater gains. Telephone repeaters and routine maintenance methods have been developed to a point where it is not difficult in practice to hold the gains of repeaters constant at any desired value. However, certain variations in the attenuation of line conductors, due to temperature changes, are inevitable.

The magnitude of net variation in total equivalent of a circuit, caused by temperature changes, is of course proportional to the total gross attenuation of the line circuit, since it depends entirely on the variation in the resistance of the copper line wires. In open wire and aerial cable circuits, a daily change in resistance value of some 5 per cent, which corresponds to a temperature change of about 22°F., may be expected. On a 1000-mile 165 open wire circuit, the total line attenuation of which is about 30 db, this would mean a variation in net overall equivalent of only about 1.5 db, assuming repeater gains to be held constant; on the other hand, 5 per cent of the total wire attenuation of a 1000-mile 19-gage H-44-S cable circuit, about 480 db, amounts to some 24 db, which is several times the value of the net equivalent of an average circuit. Variations over longer periods are of course much more severe. Table XVIII shows the maximum yearly variations in equivalents per mile that may be expected in the more common types of cable circuits. It is obvious that it would be hardly possible to maintain service on long cable circuits without the aid of some automatic means of changing the gains of repeaters to compensate for changes in line attenuation due to temperature variation.

Long cable circuits are broken up into sections averaging about 150 miles in length, known as "circuit units" One of the repeaters in each circuit unit is a "regulating repeater", the gain of which is automatically changed by a "master regulator" in accordance with changes in temperature. The master regulator employs the principle of the balanced bridge. One arm of the bridge consists of a cable pair known as a "pilot wire", which extends through the same length of cable as the circuit units to be regulated. When the resistance of the pilot wire changes due to a temperature change along the cable line, the galvanometer of the master regulator tends to deflect. By means of an auxiliary circuit, this causes a shaft driven by a small motor to turn until the bridge is again balanced. The rotation of this shaft also causes the operation of certain master relays which in turn control the operation of relays in all the regulating repeaters, causing the gains

TABLE XVIII

TRANSMISSION EQUIVALENTS IN DECIBELS PER MILE OF 16 AWG & 19 AWG CABLE CIRCUITS AT 55° F. SHOWING YEARLY VARIATIONS IN EQUIVALENTS ON ACCOUNT OF TEMPERATURE CHANGES

LOADING	TYPE	16 AWG				19 AWG			
		Side		Phantom		Side		Phantom	
		At 55°F.	Yearly Variation	At 55°F.	Yearly Variation	At 55°F.	Yearly Variation	At 55°F.	Yearly Variation
H-172-63	AE.	.16	± .018	.16	± .018	.28	± .031	.28	± .032
	U.G.		± .006		± .006		± .010		± .011
H-88-50	AE.	.19	± .022	.16	± .019	.35	± .041	.30	± .035
	U.G.		± .007		± .006		± .014		± .012
H-44-25	AE.	.25	± .029	.21	± .024	.47	± .055	.39	± .046
	U.G.		± .010		± .008		± .018		± .015

Note: The loss at any other temperature, T, is approximately the 55°F. loss shown above plus the quantity $\frac{T-55}{55}$ times the Yearly Variation.

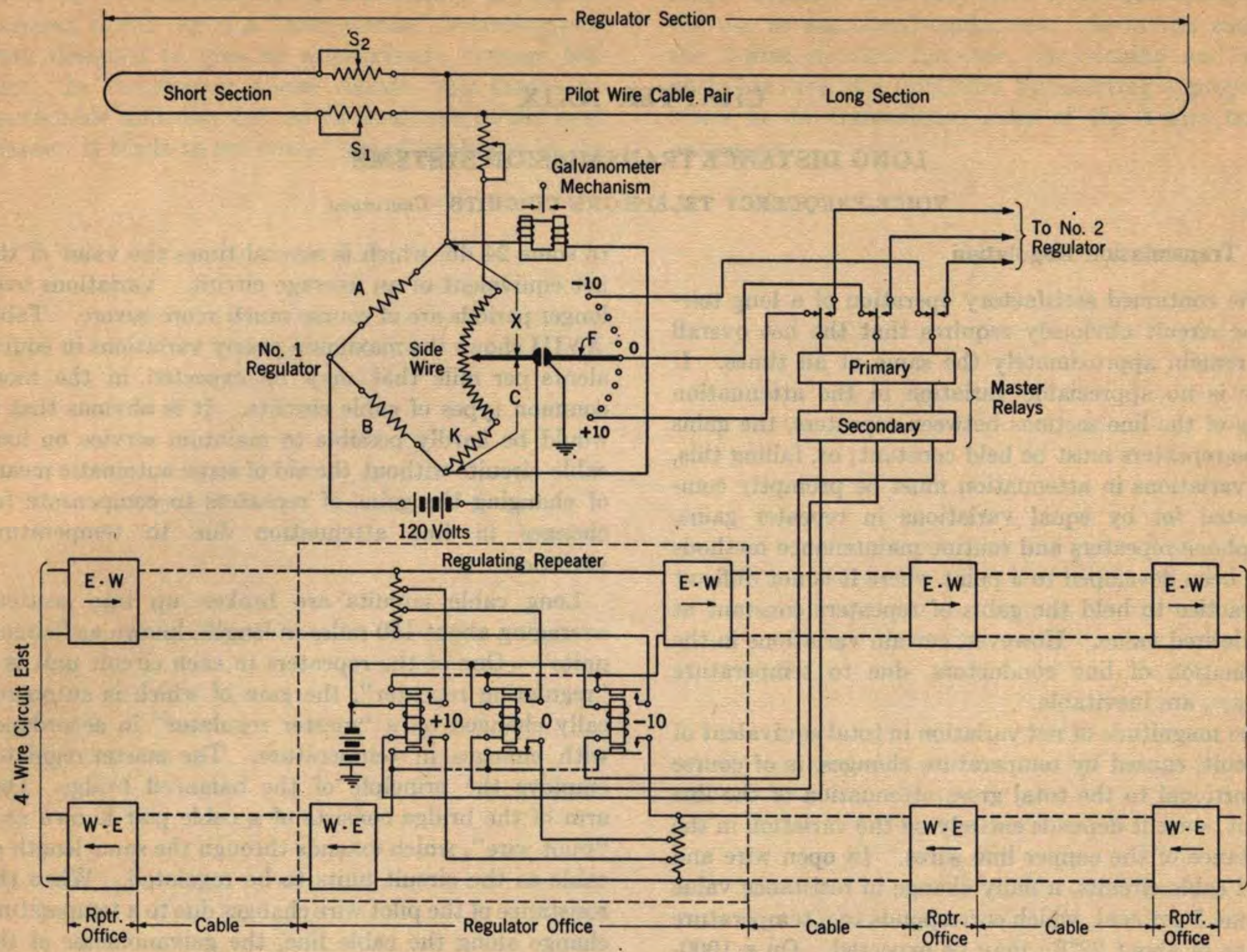


FIG. 374. PILOT WIRE TRANSMISSION REGULATOR CIRCUIT

of these repeaters to be changed in proportion to the change in temperature of the pilot wire. This change in gain is accomplished by means of regulating networks, or potentiometers, associated with the repeaters. In the case of 4-wire circuits, the regulating networks consist of potentiometers connected across the inputs of the 44-type repeaters (see Figure 337). In 2-wire repeaters, the regulating networks consist of artificial lines, or H type pads, placed in the repeater circuit between the bridge points and the manual potentiometers (see Figure 336).

Regulating repeaters are ordinarily arranged to vary their gain in 1 db steps a total of plus or minus 10 db from their nominal designated values. Thus, for example, a 4-wire regulating repeater may be adjusted for a gain of 20 db at an average temperature of 55°F. and this gain may be automatically lowered to 10 db in cold weather or increased to as much as 30 db in hot weather.

In order to follow the operation of the regulating system in more detail, let us refer to Figure 374. Here it will be noted that the bridge proper has two equal ratio arms, *A* and *B*, while the third arm, *X*, consists of the combination of the two sections of the pilot-wire circuit in parallel (which are made equal at 55°F. by the adjustable resistances, S_1 and S_2) together with a part of the slide-wire resistance. In the fourth arm, *C*, is a fixed resistance, *K*, and the remainder of the slide-wire resistance. Since arms *A* and *B* are equal, balance of the bridge is secured when the total resistance of the third arm *X* and the fourth arm *C* are equal.

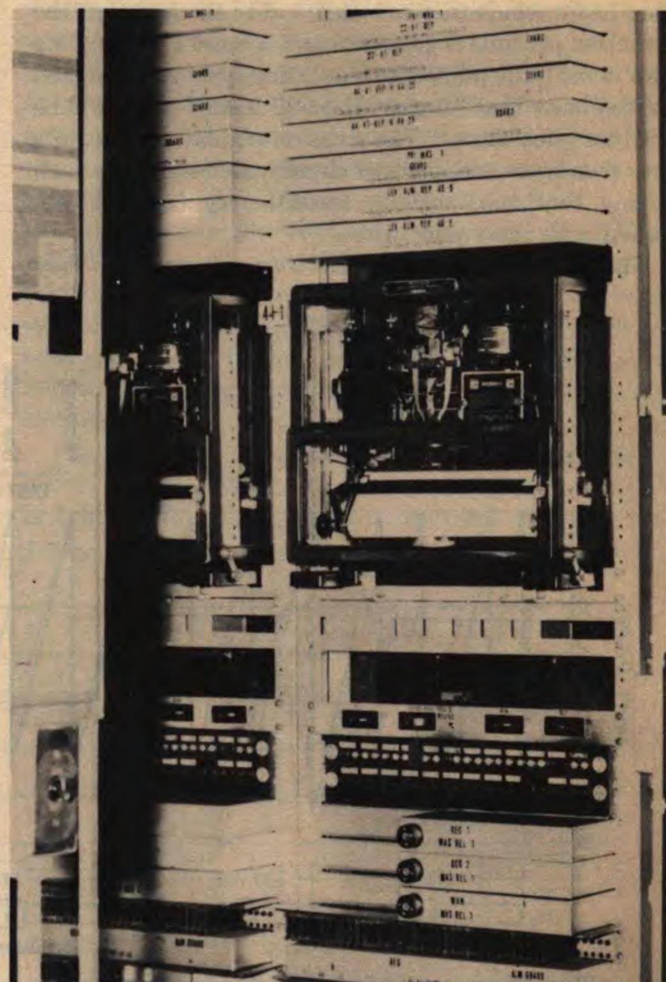
The regulating networks of the repeaters and the equalizing arrangements are designed on the basis of 55°F. being the average cable temperature. Consequently, the regulator is normally adjusted so that the slide wire will be at its mid-position (step 0) when the cable is at this temperature.

In observing the operation of the regulator, we may assume that the temperature of the pilot wire is initially 55°F. and that at this temperature, balance at the middle (zero) position of the slide wire is secured. Now assume that the temperature increases and thus increases the resistance of the pilot wire. Current then flows through the galvanometer and causes it to deflect. This will have no effect until the increase in temperature is great enough to cause an increase in the overall loop resistance of the pilot wire of as much as 180 ohms. At this point, the deflection of the galvanometer will become sufficiently large to cause movement of a mechanism to take place in such a way that the shaft on which the slide wire is mounted is caused to turn in the proper direction to restore balance. When the slide wire has moved sufficiently to restore balance, current no longer flows through the galvanometer and the movement of the mechanism stops. Since the increase of 180 ohms in the total loop resistance of the pilot wire produces an increase of 45 ohms in the joint parallel resistance of the two sections, to restore balance it is evident that the slide-wire contact is required to transfer 22.5 ohms from arm X of the bridge in which the pilot wire is included, to arm C.

As the slide-wire contact transfers 22.5 ohms from one arm of the bridge to the other, the brush arm, F, moves from one stud on the dial switch to the next adjacent stud. This movement is spoken of as a movement of one step and, as is evident from the above description, it corresponds to a change of 180 ohms in the overall loop resistance of the pilot-wire circuit. There are 21 studs, consisting of a zero step and 10 steps each side of the zero step (only 10 steps shown in Figure 374). The movement of the brush arm causes the master relays to operate and so adjust the regulating networks of all the regulating repeaters.

In Figure 374, the brush arm of the slide-wire mechanism is shown on the 0 stud of the dial switch, and ground is then connected to the master relays. As a result, these relays are so operated that the center relay in each regulating repeater is likewise operated. The latter relays are then connecting the proper resistances in the input potentiometers of the repeaters to provide the prescribed gain for the 55°F. temperature. When the brush arm moves a step, other master relays are operated and, in turn, other relays in the regulating repeaters. These change the potentiometer resistances in such a way as to effect the appropriate change in gain.

One master regulator is capable of controlling a very large number of regulating repeaters. However, since the change in the loss of a circuit with varying temperature depends upon the gage of conductors and type of loading, the proper gain variations for a given temperature change may not be the same for all the circuits under the control of a single master regulator. In order to use the same master regulator under these conditions, the systems are designed so that a given movement of the master regulator produces different changes in the gains of the regulating repeaters on the different types of facilities. This is the purpose of the two sets of master relays, primary and secondary. A separate chain of secondary relays is provided for each of the different types of regulating networks. The primary relays are operated directly from the dial switch on the master regulator. These relays are numbered from +10 to zero to −10, corresponding to the stud from which they are operated. The contacts on the primary relays are used only to operate the secondary relays, and do not control directly any of the regulating repeaters. Each of the different groups of secondary relays then controls directly the relays of all of the like regulating repeaters with which it is associated.



Transmission Regulator Equipment for Voice-Frequency Cable Circuits

At each regulator point a second complete master regulator mechanism is provided, with its own bridge battery supply and pilot wire, to guard against failure. There are also duplicate sets of master relays, either of which can be controlled by either regulator. Furthermore, in the somewhat unlikely event of the failure of both master regulators, the relay circuits are so arranged that manual control is possible.

Perfect compensation for temperature changes requires that the pilot wire be loaded with the same kind of coils as the transmission circuits which are to be regulated. This would, in general, require separate regulators and pilot wires for sides and phantoms, as well as for the different types of loading and for the different gages. Since 19-gage H-44-25 4-wire circuits require the most accurate regulation (due both to the greater lengths for which they are used and to the fact that the variation per mile is greater than for most other loading now in use), the pilot-wire regulating system is designed on the basis that 19-gage H-44-25 loaded pairs will be used for pilot wires. The steps on regulating repeaters for other types and gages of circuits are laid out in such a way that substantially accurate regulation is obtained with these pilot wires.

The pilot-wire circuits may be spare pairs or they may be obtained by compositing side circuits. In the latter case, the method of deriving the pilot-wire circuit is similar to that employed in deriving a D.C. metallic telegraph circuit, where the metallic telegraph circuit is by-passed at the telephone repeater points from repeater inputs to repeater inputs or from repeater outputs to repeater outputs without going through by-pass coils.

Where intermediate composite sets are used, the use of a by-pass on only one side of a quad will introduce considerable unbalance to ground, which may tend to introduce noise in the phantom circuit if the by-pass is from repeater input to repeater input. Accordingly, it is desirable to arrange the circuits so as to maintain approximately the same impedance to ground on both sides of the pilot-wire quad in such cases. Where the lengths of the long and short sections of pilot wire differ considerably, better balance may be secured at the regulator office with intermediate composite sets by

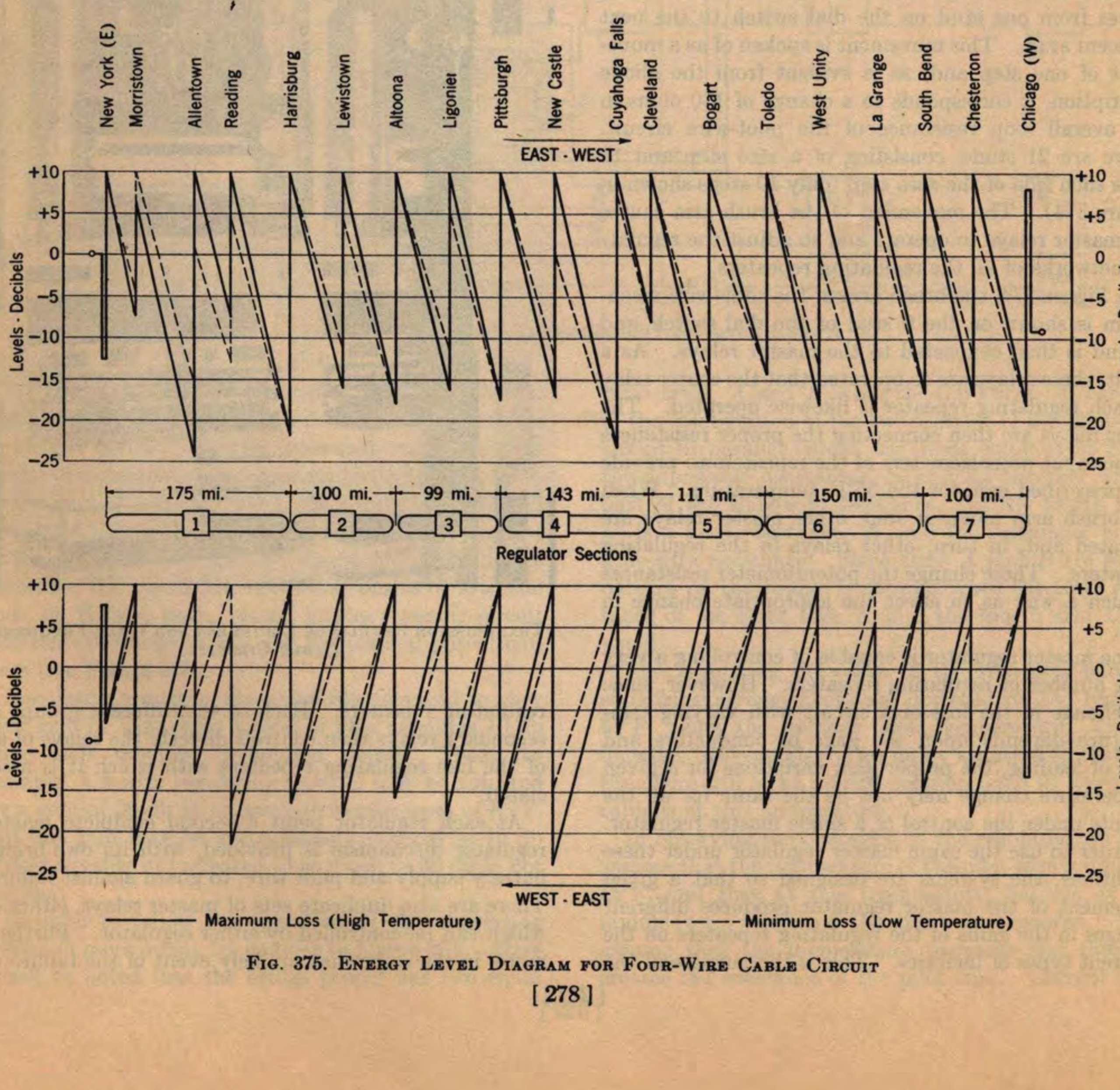


FIG. 375. ENERGY LEVEL DIAGRAM FOR FOUR-WIRE CABLE CIRCUIT

leaving the telegraph branches on the opposite side of the quad open. Where the by-pass is from repeater output to repeater output, or where terminal composite sets are used, exact balance is of less importance and in such cases, the opposite side of the pilot-wire quad may be connected to a telegraph repeater set, or left open, or by-passed in the same manner as the pilot-wire pair.

Each regulator is equipped with a recording device which gives a continuous record of the regulator setting. This consists of the following:

(1) A pen which is controlled by the movement of the shaft to which the dial switch is connected.

(2) A roller, connected to the motor by a worm gear, which turns a roll of recording paper under the pen.

(3) A reroll device for rolling up the record paper after it has received the record.

(4) A scale which extends the full length of the roller and travel of the pen.

Figure 375 is a transmission level diagram of a long 4-wire circuit made up of seven regulator sections or circuit units. It will be noted that at all repeater stations except those at the terminals of the circuit units, the input and output levels vary with temperature. It follows that it is necessary to know the setting of the regulating repeater in order to know the proper levels at any of these repeaters at any particular ime. Levels at the ends of the units are constant, however, under the normal condition where each circuit unit is completely regulated by the regulating repeater that it includes. Regulating repeaters are usually located at the repeater station nearest the mid-point of the circuit units, although this is not strictly necessary in every case.

181. Echo Control

Another series of problems, largely peculiar to the longer cable circuits arises from the fact that the velocity of propagation over such circuits, as now loaded, is relatively low. The time required for transmission over circuits of this type may thus become quite appreciable. If when a conversation is being carried on, some portion of the speaker's voice energy is returned toward him from a sufficiently distant point, the effect will be like that of an ordinary echo. This will obviously be rather disconcerting to the speaker if the time factor is great enough so that he hears entire syllables repeated back to him. In any case, such an echo effect tends to degrade the quality of transmission and it must be guarded against in long, low-velocity circuits.

Figure 376-A shows schematically a long 4-wire circuit layout. When the person at the east terminal is talking, the voice currents are sent through the 4-wire terminating set to both sides of the circuit.

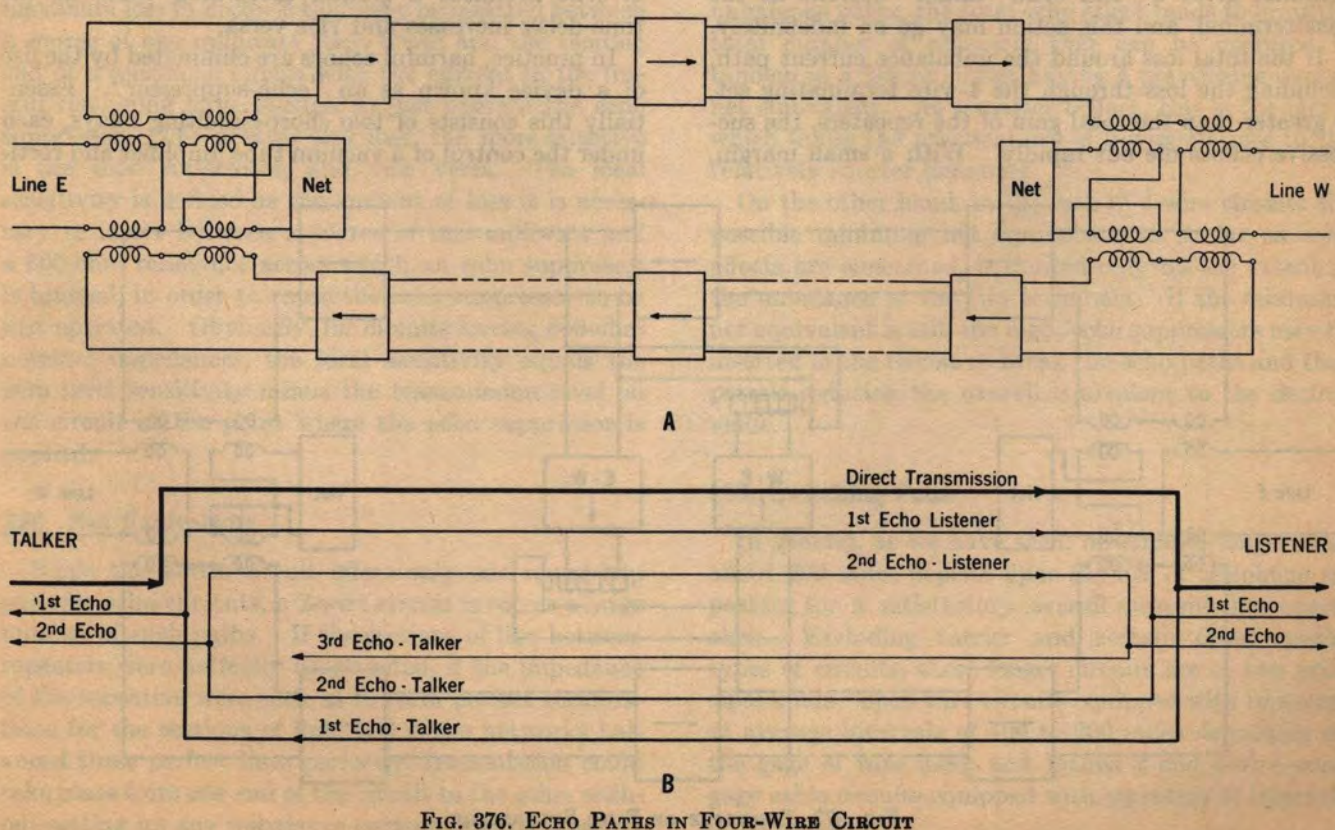


FIG. 376. ECHO PATHS IN FOUR-WIRE CIRCUIT

Transmission over the lower (receiving) side stops at the output side of the terminal repeater, but the voice currents are transmitted over the upper side until the 4-wire terminating set at the west terminal is reached, where the energy is divided between the network and the 2-wire line connected at that terminal. This transmission is indicated by the heavy line in Figure 376-B marked "direct transmission". With perfect balance between the network and the 2-wire circuit at the west terminal, no further effects will be produced.

As we have already noted, however, there is always some unbalance at the 4-wire terminating sets and, consequently, a small current passes into the lower branch of the circuit at the west terminal and is propagated back to the talking station at the east terminal, as is indicated by the line marked "1st echo—talker" This is heard at the east terminal either as side-tone, or as a distinct echo if the time of transmission around the circuit is great enough. Such currents are called unbalance or echo currents affecting the talker.

Due to unbalance at the east terminal, another current, derived from the first echo affecting the talker, is propagated from the east to the west, forming another echo which is heard by the listener. This is indicated by the line marked "1st echo—listener" in Figure 376-B. Such currents are called unbalance or echo currents affecting the listener. The first echo current affecting the listener through the unbalance at the west terminal gives a "2nd echo—talker" current at the east terminal, and this action may go on indefinitely.

If the total loss around the unbalance current path, including the loss through the 4-wire terminating set, is greater than the total gain of the repeaters, the successive echoes die out rapidly. With a small margin, there may be several echoes of sufficient magnitude to affect the persons at each end of the circuit, and if the loss becomes equal to or less than the gain, the circuit will sing.

If the transmission to and fro in a circuit could be accomplished instantaneously and the unbalance did not vary with frequency, the waves constituting the echoes would occur simultaneously with the waves of the direct transmission and would tend either to reenforce or weaken the direct transmission, depending upon the poling around the unbalance path. The net results of the echo currents would then be merely to increase or decrease the transmission equivalent between the east and west terminals and to introduce more or less side-tone at both terminals.

However, as shown in Table XIII, the velocity of propagation of the various types of facilities most commonly employed for cable circuits is actually less than 20,000 miles per second, and an appreciable time is therefore required for propagation over the longer circuits. Each successive echo accordingly arrives after a definite time interval, depending upon the length of the circuit and the velocity of propagation. Tests have been made to determine the maximum echo currents that may be permitted, without undue interference to either the talker or listener, for different times of delay. The interfering effect depends on both the volume of the echo and the time-delay. A given volume of echo produces a greater disturbing effect as the time-delay increases and vice versa.

In practice, harmful echoes are eliminated by the use of a device known as an "echo-suppressor". Essentially this consists of two short-circuiting relays, each under the control of a vacuum tube amplifier and recti-

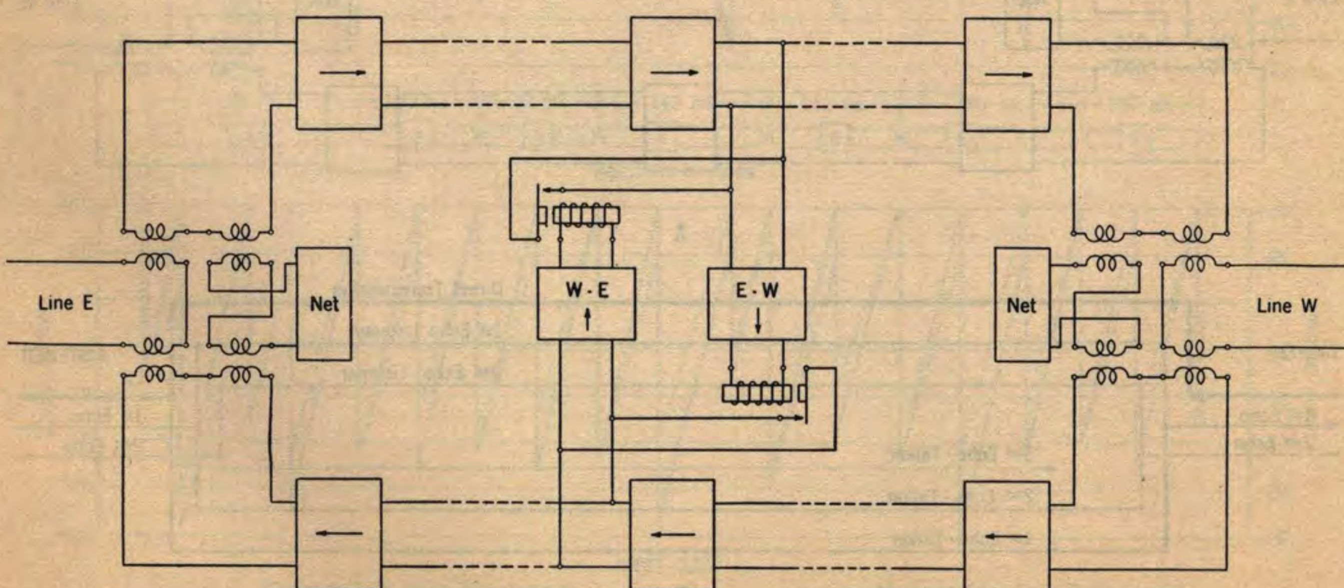


FIG. 377. PRINCIPLE OF ECHO SUPPRESSOR

fier bridged across the other side of the 4-wire circuit, as schematically illustrated in Figure 377. When conversation is being transmitted in one direction, the other side of the circuit is automatically shorted out so that any currents crossing the 4-wire terminating set cannot be returned to the transmitting end of the circuit.

The two halves of the echo-suppressor are, of course, bridged on opposite sides of the 4-wire circuit. Each half includes two vacuum tubes, one of which amplifies the small amount of energy taken from the telephone line while the other acts as a combined amplifier and detector to convert the voice currents to values sufficient for operating the relays. The input circuit of each amplifier contains series resistance which provides a high impedance and permits bridging the suppressor on the circuit without appreciably affecting direct transmission. The input circuit also includes a tuned circuit to provide maximum efficiency at frequencies in the neighborhood of 1000 cycles and to block frequencies below about 500 cycles. This provides protection against operation by low-frequency noise currents that may be present on the telephone circuit.

The effectiveness of an echo suppressor depends upon its ability to be operated by the weak voice currents. This is called its "sensitivity" and may be specified as either the "zero level sensitivity" or the "local sensitivity". The zero level sensitivity is defined as the maximum loss in decibels that may be inserted between a source of one milliwatt (zero level) and the sending end of a telephone circuit with the current in the line still remaining large enough to just operate the echo suppressor. The greater this loss, the more sensitive is the echo suppressor, and vice versa. The local sensitivity is defined as the amount of loss it is necessary to insert between a source of one milliwatt and a 600-ohm resistance across which an echo suppressor is bridged, in order to cause the echo suppressor to be just operated. Obviously, for circuits having 600-ohm nominal impedances, the local sensitivity equals the zero level sensitivity minus the transmission level on the circuit at the point where the echo suppressor is applied.

182. Net Equivalents

While the 4-wire circuit offers only one round-trip path for echo currents, a 2-wire circuit involves a large number of such paths. If the sections of line between repeaters were perfectly constructed, if the impedance of the repeaters were such as to form perfect terminations for the sections of line, and if the networks balanced these perfect lines perfectly, transmission could take place from one end of the circuit to the other without setting up any unbalance currents except those reflected from the terminals. Such an ideal circuit would give the same performance as a 4-wire circuit with equal terminal unbalances.

In practice, however, there is more or less unbalance on each side of each repeater so that, as the direct transmission passes through each successive repeater, an unbalance current is set up which travels back toward the talker, giving an echo current for each repeater in the circuit. Each of these first echo currents in turn sets up an echo current traveling toward the listener at every repeater that it encounters, and each of these echo currents sets up another echo current at each repeater it encounters. This process continues indefinitely but the successive echo currents are attenuated rapidly to inappreciable magnitudes. The action, however, does give rise to a very large number of echo currents. The final effect upon the talker or listener naturally depends upon the resultant of all these currents.

These unbalance currents increase in volume as the overall net equivalent of a circuit is decreased, due to raising the repeater gains. For each circuit, therefore, there will be a certain minimum permissible net overall equivalent because of the unbalances which are present in the circuit. Any further increase in the repeater gains will cause the unbalance currents to become too large to be tolerated. Furthermore, since each additional repeater in the circuit increases the number of unbalance paths, the total echo effect tends to limit the total number of repeaters that can be operated in tandem in a 2-wire circuit having a practicable overall net equivalent. As a matter of fact, this is the factor which limits the practical use of 2-wire circuits to the relatively shorter distances.

On the other hand, in the case of 4-wire circuits the possible minimum net equivalent, in so far as echo effects are concerned, is limited only by the extent of the unbalance at the two terminals. If the minimum net equivalent is still too high, echo suppressors may be inserted in the circuit to break the echo paths and thus permit reducing the overall equivalent to the desired value.

183. Switching Pads

In general, as we have seen, all circuits longer than about 200 miles depend upon the use of telephone repeaters for a satisfactory overall transmission equivalent. Excluding carrier and certain other special types of circuits, these longer circuits are of two principal kinds—open wire circuits equipped with repeaters at average intervals of 100 to 200 miles depending on the gage of wire used, and loaded 2 and 4-wire small gage cable circuits equipped with repeaters at intervals of 50 miles.

Both of these two main types of circuits are commonly arranged with repeaters at the ends or terminals of the circuit, as well as at intermediate points. This plan permits fairly equal spacing between all repeater stations, whether located at terminals or intermediate points, and so makes possible a flexible layout of circuit facilities. The use of such terminal repeaters on these voice-frequency circuits, as well as on carrier circuits, is also taken advantage of to improve transmission on switched connections. Here they are used to produce the same effect as would result from the insertion of a "cord-circuit" repeater at the switching point.

The terminal repeaters are operated at a higher gain than necessary and "switching pads" are inserted between the terminal repeaters and the circuit drops. These cause sufficient loss to keep the equivalents of the circuits to their prescribed values when they are being used for terminating traffic. When two such circuits are connected together at the switchboard for a through connection, the pads are automatically cut out and the total gain at the switching point is increased correspondingly. Thus, for example, assume that two circuits terminating at a certain point are designed to have equivalents of 9 db each. This means that if the two circuits are connected together with no special provision, the total equivalent of the connection will be 18 db. Now let the terminal repeaters on the two circuits be arranged to give gains of, say, 3 db more than necessary for the 9 db equivalent. Then, if 3 db pads are at the same time inserted in each circuit, the net equivalent of each circuit will remain 9 db but when the two circuits are connected together, the pads will be eliminated and the total equivalent of the overall circuit will be 12 db rather than 18 db. The arrangement of these switching pads in the circuit is shown in the Switching Pad Circuit of Figure 140.

184. Signaling Systems

In order to provide some means for the operators at the two ends of a circuit to signal each other, signaling equipment, or "ringers", are required for each telephone circuit. The arrangements generally used may be divided into three classes, depending upon the type of current which is used to transmit the signal over the line; namely, 20-cycle, 135-cycle, and 1000-cycle signaling systems.

Signaling with 20 cycles is used only on 2-wire circuits which are not composited for telegraph. When 20-cycle signaling is used on repeatered circuits, an intermediate ringer is required at each repeater point to relay the signal around the repeater because the repeaters do not amplify at this frequency (see Figure 368). The distance over which satisfactory signals may be transmitted is obviously limited.

Ringers using 135 cycles are employed on certain lines composited for telegraph, but in the case of cable circuits equipped with metallic telegraph, the signaling current must be limited in order to avoid interference with the telegraph signals. Intermediate ringers are generally employed at repeaters on account of the inefficiency of the repeaters at this frequency as well as at 20 cycles. However, this system of signaling has a greater range than the 20-cycle system.

In the 1000-cycle signaling system, the signal is transmitted by means of a 1000-cycle current interrupted 20 times per second. Since the frequency of this signaling current is within the voice range, no intermediate equipment is required at any repeater point and the system is of universal applicability wherever it is economical. It is generally used on the longer telephone circuits of all types. False operation from voice currents is prevented by making the receiving ringers selective to both the frequency of 1000 cycles and the frequency of interruption, 20 cycles, and by the introduction of a time element requiring that the signal be sustained for several tenths of a second.

It is sometimes desirable to make use of one type of signaling on one part of a circuit and another type on another part. To take care of this situation, standard arrangements are provided for converting the signal from any one of the three types to any other at intermediate points on the circuit.

CHAPTER XXX

LONG DISTANCE TRANSMISSION SYSTEMS

CARRIER TELEPHONE AND TELEGRAPH CIRCUITS

185. Types of Carrier Systems

For a number of years most of the longer telephone circuits, and many of the telegraph circuits, operated on open wire lines have been obtained by carrier methods. As noted in Article 174, this tends to become true in current practice in the case of cable facilities also. It therefore may reasonably be anticipated that a great many of the long circuits—both telephone and telegraph—will be operated over carrier channels in the future.

Economy is of course the principal reason for the use of this carrier technique. Interestingly enough, however, it is usually practicable to obtain higher quality circuits by employing carrier-frequency transmission than by the use of the ordinary voice-frequency methods. This is in part merely a result of the way in which the art has developed. But it is also due in part to the fact that the elimination of loading, which is generally necessary for carrier transmission, permits higher velocities of propagation and also tends to increase the stability of the circuit characteristics.

The general principles of carrier operation are outlined in Chapter XXVII and these principles apply alike to all of the several types of carrier systems used in the long distance plant. With respect to the frequency allocations employed, the systems in use range all the way from the voice-frequency carrier telegraph system, which operates entirely within the voice range, up to the coaxial systems where frequency bands millions of cycles in width may be involved. Indeed, by including radio systems, which by a slight expansion of the definition may properly be classified among carrier systems, the range of frequencies used in carrier operation is extended almost indefinitely.

Figure 378 shows the frequency allocations of the principal current types of telephone and telegraph carrier systems, for the frequency range extending from just above the ordinary voice band up to 140 kilocycles. The voice-frequency telegraph system mentioned above would occupy a position at the extreme left of this figure, while carrier systems operating on coaxial facilities would extend far to the right—to perhaps 2000 or more kc. Of the systems indicated in Figure 378, we shall be interested chiefly in the telegraph systems coded B, the "low-frequency" telephone systems coded C, and the so-called "broad-band" telephone systems coded J and K.

It may be noted that in both the Type-B telegraph systems and the Type-C telephone systems, several different sets of frequency allocations are used. The purpose of this is to reduce the possibilities for intersystem crosstalk where a number of systems are operated on the same pole line. In the case of the Type-J telephone system, which, like both of the above systems, is applied only to open wire lines, it is expected that additional "staggered" frequency allocations will eventually be employed for the same reason.

As indicated in the figure, all of the open wire carrier systems, including the short-haul Types-D, -G, and -H, use separate channels for transmission in opposite directions, thus operating effectively on a 4-wire basis. The Type-K system, which is designed for application to cable facilities, employs the same channel frequencies for transmission in both directions but actually uses two pairs of conductors, one transmitting in each direction. This is also true of the voice-frequency telegraph system, which either operates on a 4-wire voice-frequency cable circuit or on the two one-way channels of an open wire or cable carrier telephone circuit.

AUXILIARY CABLE CARRIER REPEATER STATION

The detailed design features of the several carrier systems are covered in standard instructions. All systems make use of one or another of the various types of pads, filters, equalizers, modulators, demodulators, carrier supply circuits, and amplifiers that have been discussed in preceding chapters. We shall be concerned here, therefore, only with a brief survey of the general design of the principal types of systems, and

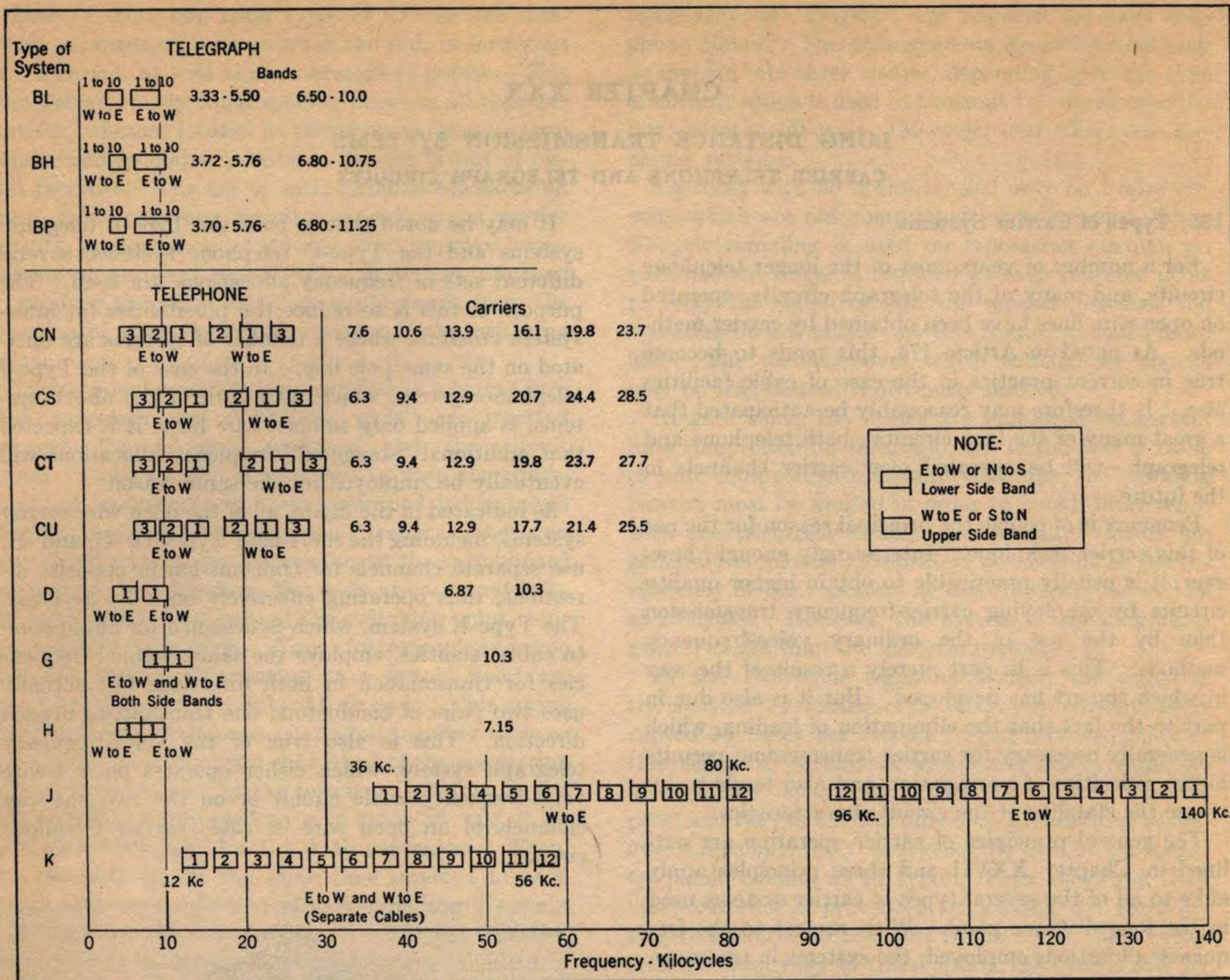


FIG. 378. CHANNEL FREQUENCY ALLOCATIONS FOR TELEPHONE AND TELEGRAPH CARRIER SYSTEMS

with certain features which will serve to bring out some interesting applications of the electrical principles with which this text is primarily concerned.

186. "Low-Frequency" Carrier Systems

On the basis of their historical order of development and the general techniques employed, it is convenient to divide carrier systems into two general groups. The first of these groups we shall designate as low-frequency systems because the systems included make use of frequencies up to a maximum of only about 30,000 cycles. The second group includes the broad-band systems which, in general, provide more channels per system and employ frequencies ranging from 12 kc. up to some two megacycles (2000 kc.). Included in the first or low-frequency group, are the voice-frequency and Type-B telegraph systems and the Types-C, -D, -G, and -H telephone systems.

The first of the telegraph systems in the order of development is the Type-B system. This provides for superimposing ten two-way telegraph channels on an open wire telephone circuit, which may also be composited for ordinary grounded telegraph operation. As indicated in Figure 378, three frequency allocation groups are used, coded BL, BH, and BP. Each group employs twenty carrier frequencies, the lower ten frequencies being used for transmission in one direction (West to East) and the higher ten for transmission in the opposite direction (East to West). The separation between carrier frequencies ranges from slightly less than 200 cycles at the lower end of the frequency band, to nearly 1000 cycles at the upper end. With the carrier frequencies grouped for transmission in the two directions, separation at terminal and repeater points is secured in each case by a pair of directional filters, one of which passes only frequencies below 6000 cycles and the other only frequencies above that value. The

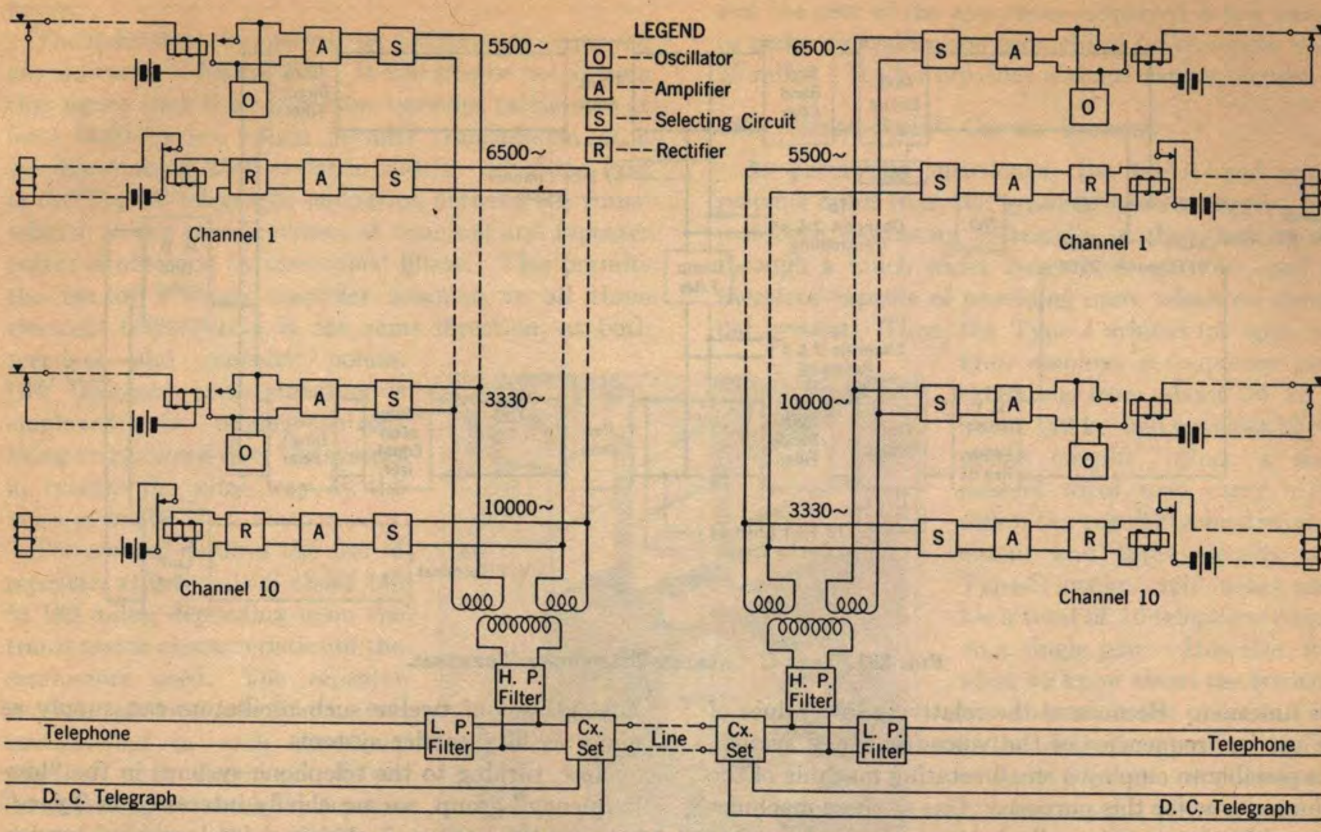


FIG. 379. HIGH-FREQUENCY CARRIER TELEGRAPH SYSTEM

general arrangement of the Type-B system terminal is indicated in the block diagram of Figure 379. Since only the single frequency of the carrier itself is transmitted over the line, the channel filters need be merely simple tuned circuits. The modulator and demodulator circuits are also relatively simple.

The second type of telegraph system is the voice-frequency system, so called because the carrier frequencies used are within the ordinary voice-frequency band. The principle of this system is not essentially different from that of the Type-B system but due to the lower frequencies employed, there is considerable variation in the details of the apparatus. The system is operated on a 4-wire cable circuit or on a channel of a carrier telephone system, and since its operation would naturally interfere with ordinary telephone transmission, it is not superimposed on a telephone circuit. As the transmitting medium is either actually or effectively a 4-wire circuit in all cases, there is no problem of separating the transmitting and receiving channels. The same carrier frequencies are used for transmission in both directions. The system provides 12 two-way telegraph circuits. The carrier frequencies employed are multiples of 85 cycles beginning with 425 cycles and extending to 2295 cycles, the separation between adjacent channels being 170 cycles.

The carrier sending and receiving apparatus is similar to that of the Type-B system. Spurts of the carrier current are sent over the line as the transmitting amplifier (modulator) is shorted out by the telegraph impulses, and a rectifying device (demodulator) at the receiving end converts the spurts of carrier back again to ordinary direct-current telegraph signals.

One of the chief differences between the two types of telegraph systems has been the method of generating the carrier frequencies. In the Type-B system, as in all telephone systems, vacuum tube oscillators perform

VOICE-FREQUENCY CARRIER TELEGRAPH PANELS

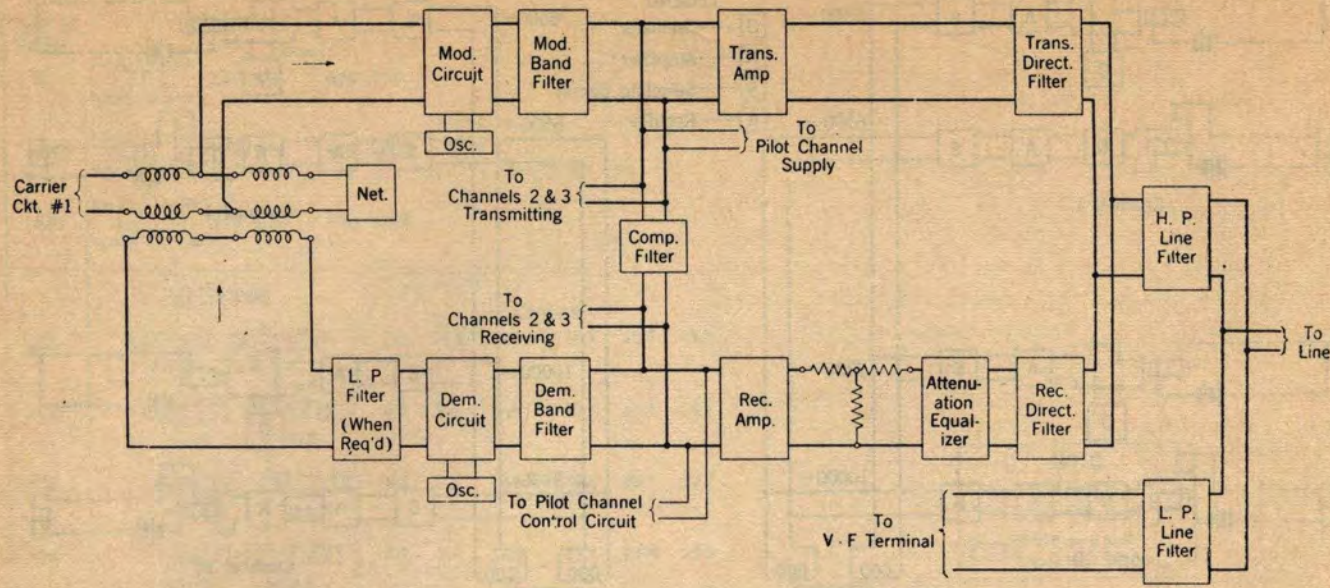


Fig. 380. Type-C Carrier Telephone Terminal

this function. Because of the relatively low values of the carrier frequencies of the voice-frequency system, it is possible to employ a small rotating machine of the inductor type for this purpose. One of these machines is capable of generating all of the required carrier frequencies for twenty systems. The speed of the machine which, of course, determines the frequencies of the output currents, is controlled by a mechanical governor, and an associated electronic indicating device. However, vacuum tube oscillators may also be used and these have some advantages with respect to stability. Furthermore, the failure of an oscillator affects only one carrier channel whereas a failure of the machine generator affects all channels together. A single set of twelve such oscillators can supply as many as fifty carrier systems.

Now, turning to the telephone systems in the "low-frequency" group, we are chiefly interested in Type-C because this system is designed for long haul service. It operates on open wire facilities and provides three telephone circuits additional to the normal voice-frequency circuit. The general layout of the Type-C carrier terminal is shown schematically in Figure 380. The transmission over each channel consists of a single side-band, the carrier frequency being suppressed in the modulator. The required carrier frequency at each end of each channel is supplied by individual vacuum tube oscillators whose stability is such as to main-

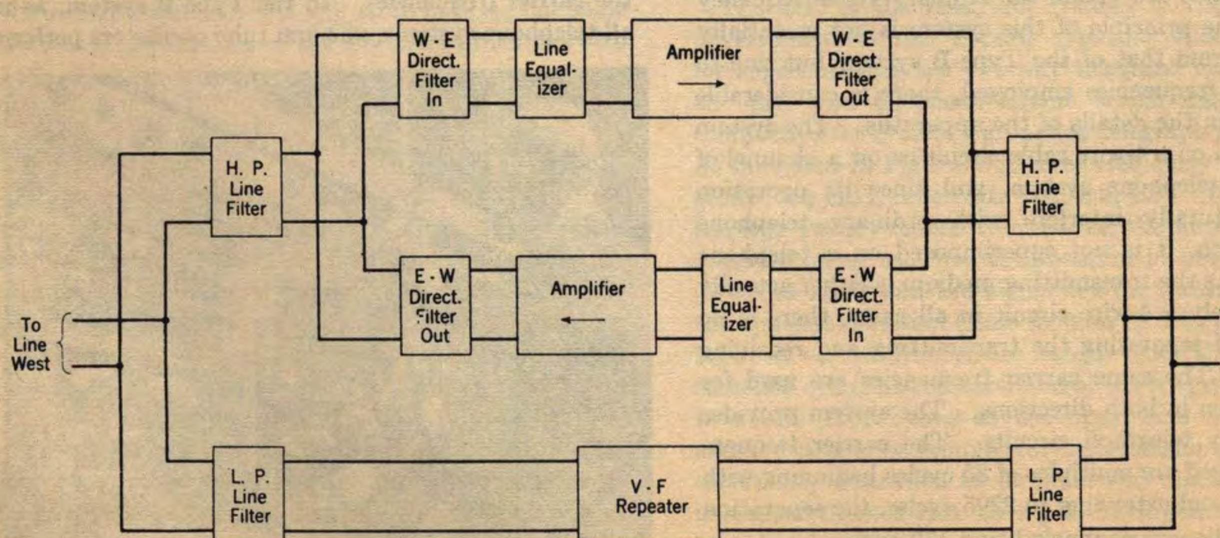


Fig. 381. Type-C Carrier Repeater Arrangement

tain satisfactory frequency synchronization at all times.

The individual channel carrier frequencies employed are indicated in Figure 378. It will also be noted from this figure that the separation between carriers is at least 3000 cycles, which permits transmission of a side-band about 2500 cycles in width. As in the case of the Type-B telegraph, separation between the transmission in the two directions at terminal and repeater points is obtained by directional filters. This permits the use of a single amplifier common to all three channels transmitting in the same direction, at both terminal and repeater points. One thousand cycle signaling is employed, the ringing current being transmitted over the system in exactly the same way as the voice currents.

The system requires the use of repeaters at intervals of about 140 to 180 miles, depending upon the transmission characteristics of the conductors used. The repeater consists of two amplifiers, one transmitting in each direction, together with directional filters for obtaining the necessary separation, as shown schematically in Figure 381. The amplifiers are of a high-gain type arranged to operate with output levels up to as high as 15 to 20 db above zero. Equalizers having characteristics generally similar to that illustrated in Figure 296 of Chapter XXIII are associated with each amplifier.

In the older systems the repeater amplifiers, as well as the transmitting and receiving amplifiers at the terminals, are of the push-pull type illustrated in Figure 339 of Chapter XXVI. In more recent systems, negative feedback amplifiers generally similar to that illustrated in Figure 384 are employed. Similarly, balanced vacuum tube modulator and demodulator circuits are used in the older systems, while more recent systems employ copper-oxide modulator units. Both arrangements are discussed in Chapter XXVII.

TERMINAL EQUIPMENT OF TYPE-C CARRIER SYSTEM

The three remaining low-frequency carrier systems indicated in Figure 378—namely, Types-D, -G, and -H—have a relatively limited field of application in long distance work. The D and H systems are designed to provide a single carrier circuit and can be economically applied over short distances of the order of 50 miles. The G system is even simpler in design and the cost of the apparatus employed is low enough to make its application economical for distances under 25 miles. It also provides a single carrier circuit.

187. "Broad-Band" Carrier Systems

As previously mentioned, the broad-band carrier systems differ from the types of systems that we have just been discussing principally in that they operate through a much wider range of frequencies, and are therefore capable of providing more telephone circuits per system. Thus, the Type-J system for open wire lines employs a frequency range extending from about 36 kc. to about 140 kc. and provides 12 telephone circuits. Since a single pair of wires may carry in addition the regular voice-frequency circuit and three circuits of a Type-C system, this makes possible a total of 16 telephone circuits on a single pair. However, from what we know about the transmission characteristics of line facilities at the higher frequencies, it will be evident that the application in practice of such a broad-band system necessarily introduces new problems in controlling losses and avoiding noise and crosstalk.

From Figure 382 which is a block schematic of the essential elements of a Type-J system terminal, it may be noted that the principle of operation does not differ from that employed in Type-C and other carrier telephone systems. Perhaps the most striking new feature is the use of more than one stage of modulation and demodulation. The basic purpose of this is to permit the use of the most desirable group of frequencies for the initial channel modulation, while at the same time applying to the line the bands of frequencies most suitable for transmission. The twelve carrier channel frequencies employed are 64, 68, 72, 76 and so on, up to 108 kc. The separation between carrier frequencies is thus 4000 cycles.

There are a number of reasons for the selection of this group of carrier frequencies. In the first place, it happens that high-grade crystal filters can be most economically built for operation in this general range. Also important is the fact that the range is high enough so that the lowest harmonic of the lowest frequency is

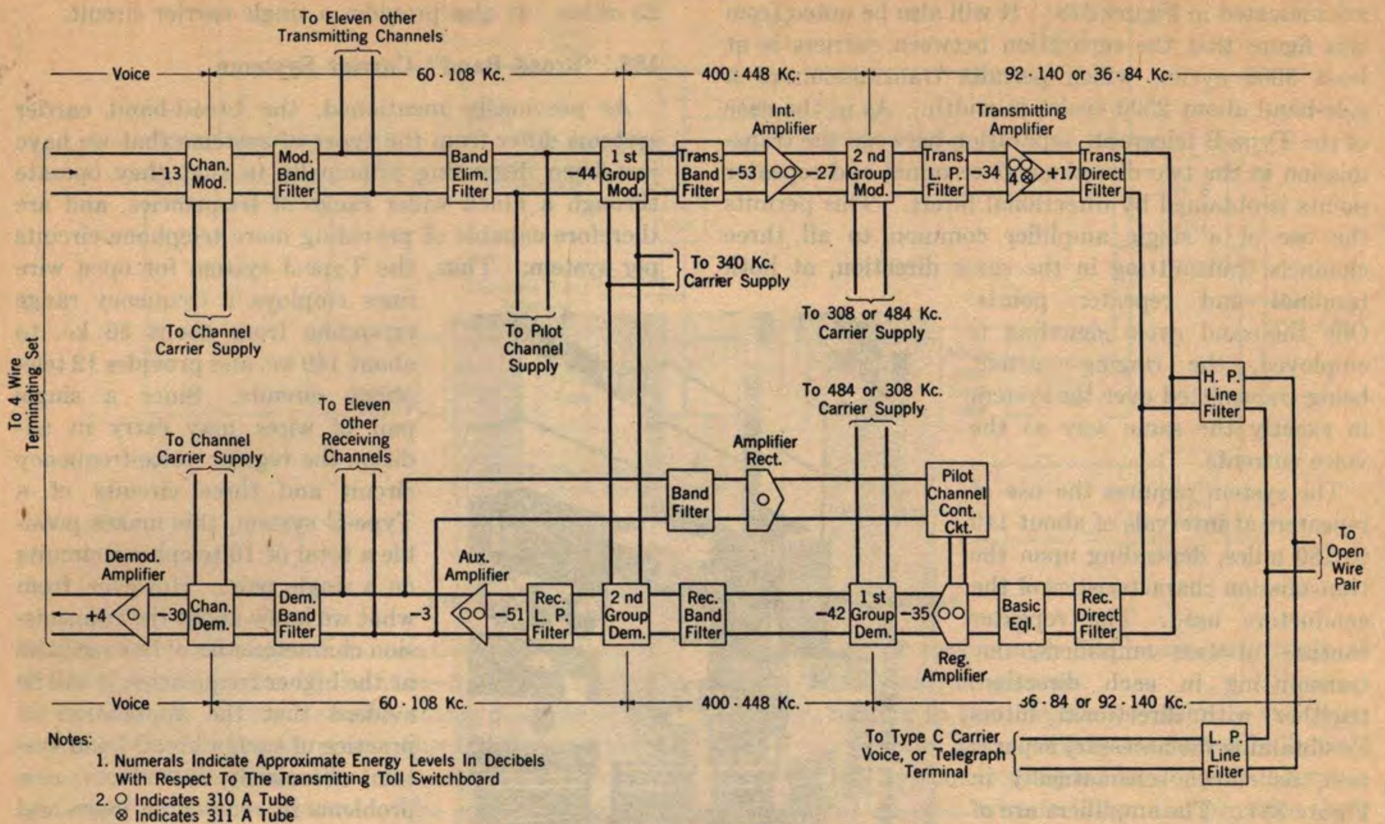


FIG. 382. TYPE-J CARRIER TELEPHONE TERMINAL

above the highest frequency of the band. Thus, the second harmonic of 60 kc., which is the lowest frequency in the lower side-band of the 64 kc. carrier, is 120 kc., which is well above the top frequency of 108 kc. This obviates the possibility of any harmonics that may be generated in the channel modulators interfering with other channels. Finally, a general design and manufacturing economy is obtained by using this same group of channel carrier frequencies for all of the broad-band systems—Types-J, -K, and Coaxial.

Having modulated the twelve channel carriers with voice frequencies and eliminated everything but the lower side-bands by means of appropriate band filters, the entire group of frequencies—48 kc. in width—is translated by an additional modulation process to the band of frequencies that it is desired to transmit over the line. In the case of the Type-J system, the range of this final band depends upon the direction of transmission since transmission in both directions is over the same pair of wires. Thus, for the frequency allocation shown in Figure 378, transmission West to East occupies the band between 36 and 84 kc. and transmission East to West is in the range from 92 to 140 kc.

The crystal band filters associated with the channel modulators are sharp enough so that a band of voice frequencies more than 3000 cycles wide is transmitted. This, it may be noted, is an appreciable improvement over the 2500-cycle band transmitted by the older types of C systems and the usual loaded cable circuit.

Due to the fact that the two frequency bands transmitted over the line in the Type-J system both overlap the initial 60 to 108 kc. band, a direct translation is not practicable. Instead, it is necessary to make the translation in two modulation stages as indicated in Figure 383. From a study of this figure, along with Figure 382, it may be observed that for transmission East to West the initial 60–108 kc. band is delivered to the first group modulator along with a carrier frequency of 340 kc. The output of the modulator includes the upper 400–448 kc. side-band which is selected by the transmitting band filter and passed on, through the two-stage intermediate amplifier, to the second group modulator. The carrier frequency applied to this modulator is 308 kc. and its output therefore includes a lower side-band of 92–140 kc. The upper side-band and other extraneous frequencies above the 92–140 kc. band are eliminated by the transmitting low-pass filter, so that the input currents applied to the transmitting amplifier, and thence to the line, are in this desired frequency band. The frequency trans-

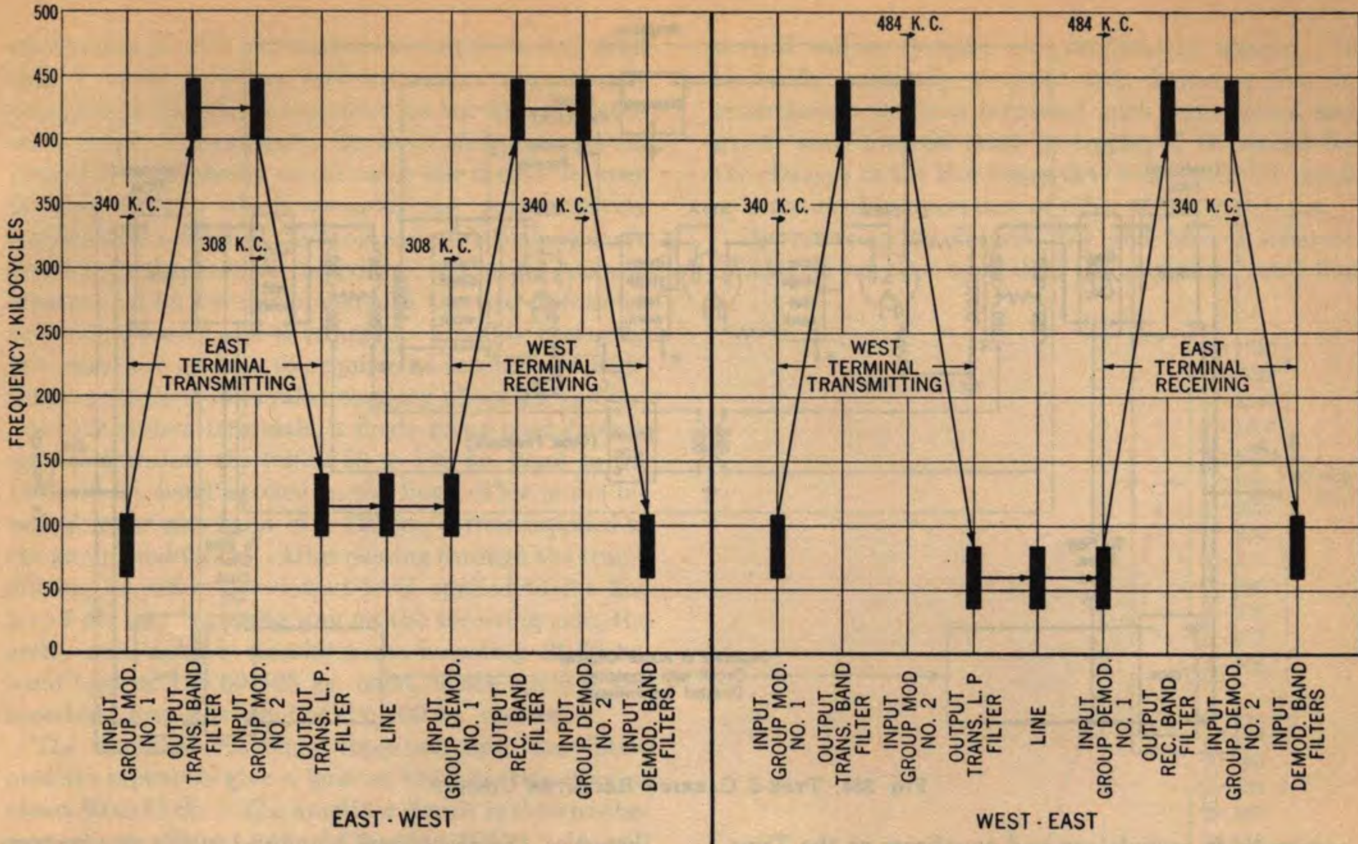


Fig. 383. Frequency Translations in Type-J Carrier Systems

lations occurring on the receiving side of the circuit are, of course, in the opposite direction, so to speak, and may be followed through in a like manner. For transmission West to East, the processes are exactly similar except that here the carrier frequency applied to the second group modulator in the transmitting circuit (and the first group demodulator in the receiving circuit) is 484 kc. instead of 308 kc. The resultant lower side-band of this is 84–36 kc., or, when turned over, the desired 36–84 kc. band which is applied to the line for transmission in this direction.

The copper-oxide modulator and demodulator units employed are capable of handling only small amounts of energy, so that the transmission levels at both their inputs and outputs are necessarily rather low. Thus, as may be seen by referring to Figure 382, the input level to the channel modulator is −13 db and the input levels at the first and second group modulators are considerably lower than this. As the transmitting level applied to the line is about +17 and the receiving level applied to the voice terminal is +4, this means that several amplifiers are required in both the transmitting and receiving legs of the circuit. All of these amplifiers are of the stabilized negative feedback type, capable of giving substantial gains. The demodulator amplifier (single-stage) is adjustable through a plus or minus 5 db range by means of a potentiometer mounted in the voice-frequency jack panel. This permits convenient manual adjustment of the receiving levels when the circuits are lined up.

Since the line losses at the high frequencies employed are relatively very high, the Type-J system requires the use of repeaters at considerably closer spacings than does the Type-C. Just what this spacing must be depends upon the weather conditions prevailing in the territory through which the line extends. Sleet, frost, or ice forming on the line wires will greatly increase their attenuation, and the repeater spacing should be close enough so that there will be sufficient margin to take care of the most adverse conditions that may reasonably be anticipated. For most of the lines where these systems are likely to be applied, the average spacing is expected to be about 70 miles. At the repeater points, the transmission in the two directions is separated by directional filters and each repeater includes two amplifiers, one "pointed" in each direction. The general arrangement is practically the same as is shown in Figure 381 for the Type-C repeater, except that in this case the equalizers form part of the amplifier circuit. The amplifiers are three-stage negative feedback devices of the type illustrated schematically in Figure 384. The circuit, although somewhat different in detail, is in principle essentially the same as that used in the Type-K systems.

The Type-K systems for application to cable facilities also make use of the same types of crystal filters,

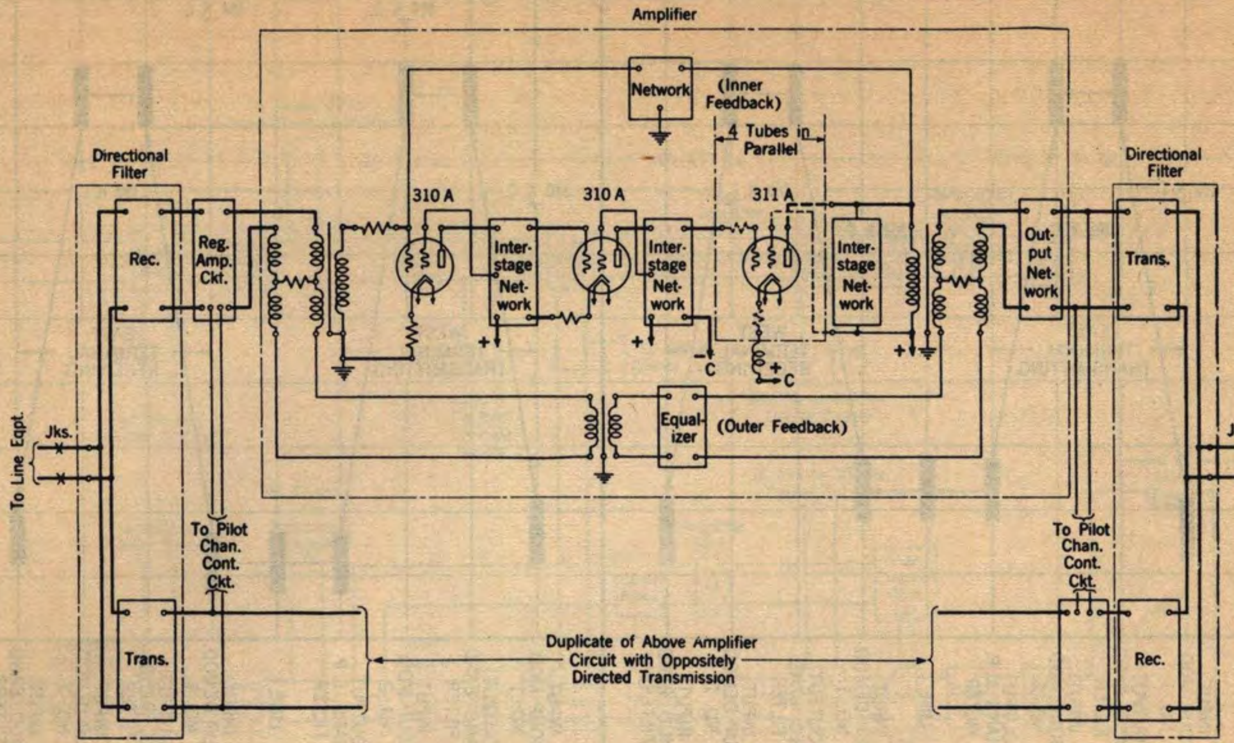


FIG. 384. TYPE-J CARRIER REPEATER CIRCUIT

copper-oxide modulators and equalizers as the Type-J systems. Indeed, as illustrated in Figure 385, the channel modulating and demodulating circuits are identical.

So far as transmission over the line is concerned, however, the techniques employed in the two systems are necessarily quite different. Since any practicable type of cable circuit loading cuts off at a comparatively low frequency, it is necessary to use non-loaded cable conductors for broad-band transmission. Because the

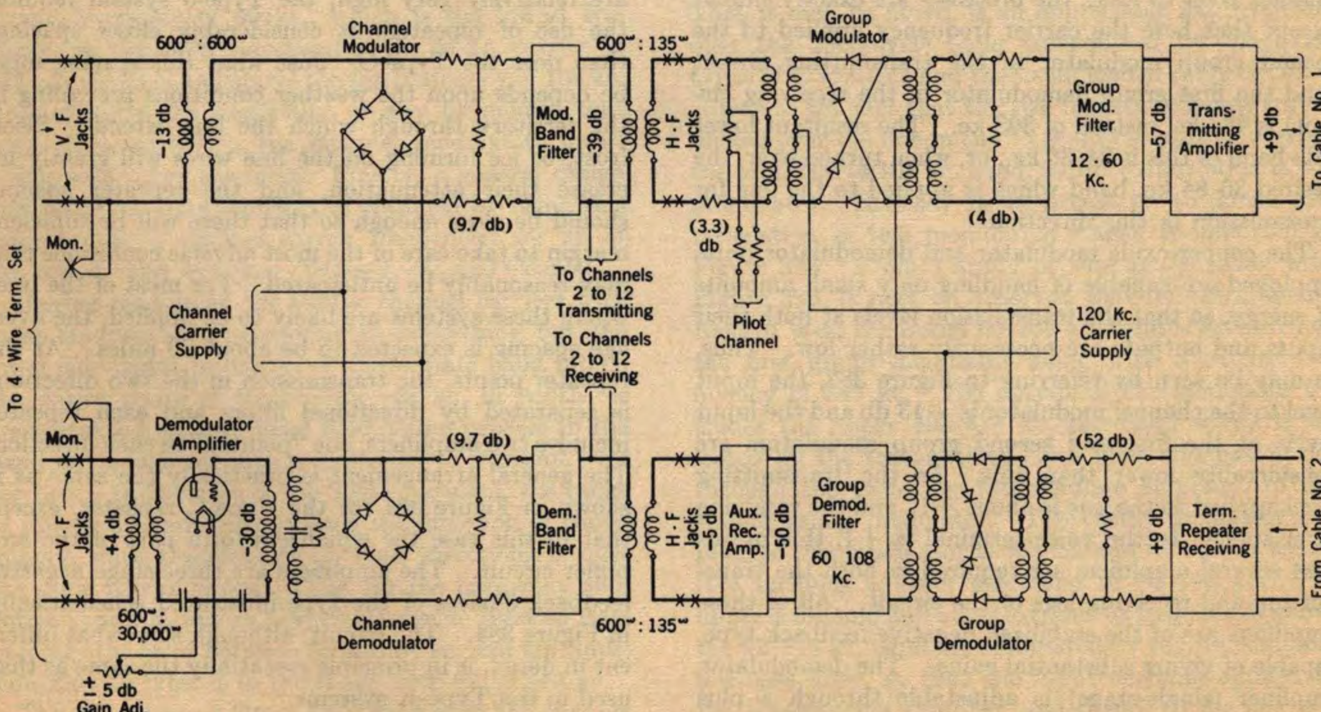


FIG. 385. TYPE-K CARRIER TELEPHONE TERMINAL

attenuation of such non-loaded conductors is very high and, of course, increases with frequency, it is desirable to keep the maximum frequency to the lowest practicable value. Accordingly, the band of frequencies selected for transmission on the cable line is that between 12 and 60 kc., which occupies the comparatively straight-line portion of the attenuation-frequency curve just above the knee of the curve. By using pairs in separate cables for transmission in the two directions, only one 48 kc. band is required. On this basis, the line losses are such as to require the insertion of high-gain repeaters at intervals averaging about 16 miles.

At the system terminals, a single group modulator is used to translate the initial 60 to 108 kc. band to the 12 to 60 kc. band applied to the line. This is the inverted lower side-band of a 120 kc. carrier supplied to the group modulator. After passing through the transmitting amplifier, the output level applied to the line is +9 db. In the same way on the receiving side, the group demodulator translates the incoming 12–60 kc. band back to the 60–108 kc. band, which again is the inverted lower side-band of the 120 kc. carrier.

The terminal receiving repeaters and the intermediate repeaters give a gain at the top frequency of about 50 to 75 db. The amplifier circuit is shown schematically in Figure 345 of Chapter XXVI. It will be noted that both the line equalizers and the gain control apparatus are inserted in the feedback path, the latter being under the control of a receiver motor, as is discussed in a later article.

At the time this is written, broad-band carrier systems for application to the coaxial type of conductor are still in the experimental stage. It is anticipated, however, that such systems when used will employ the same initial 12 carrier channels as are used in the J and K systems, and that groups of 12 channels will be raised by a second modulation process to successively higher positions on the frequency scale in the general manner indicated in Figure 386. In most cases it may be presumed that the number of individual telephone circuits which will be so carried on a single coaxial conductor, or on a pair of such conductors, will depend on circuit requirements rather than on technical limitations. By spacing amplifiers in the conductors at intervals as close as five miles, present indications are that it is theoretically possible to obtain as many as 500 telephone circuits from a pair of coaxial conductors.

188. Transmission Regulation in Open Wire Carrier Systems

In discussing the requirements for intermediate and terminal repeaters in the two preceding articles, the question of the overall transmission stability of the several carrier systems was deliberately ignored. It is hardly necessary to point out, however, that in transmission systems involving such large losses and gains, some method must be employed to correct for the changes in the line losses that will inevitably occur as a result of temperature or other weather changes.

By referring to Figures 279 and 280 of Chapter XXII, it may be seen that the minimum total line

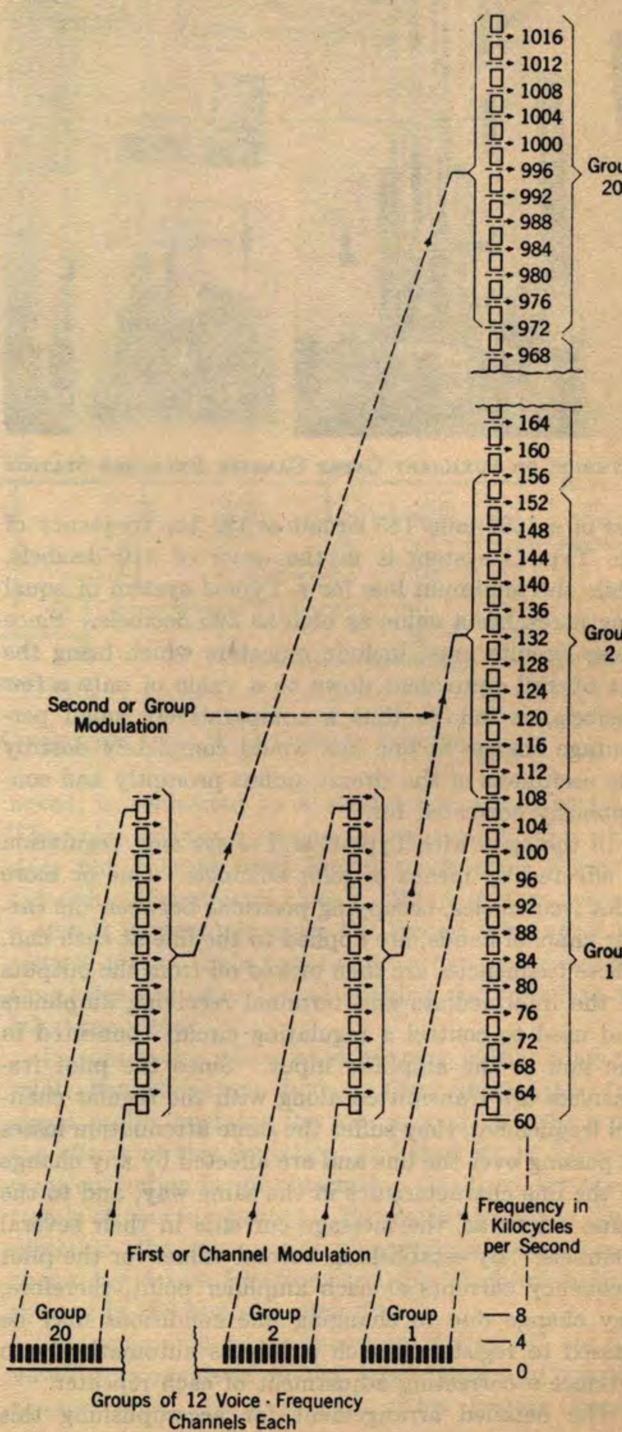
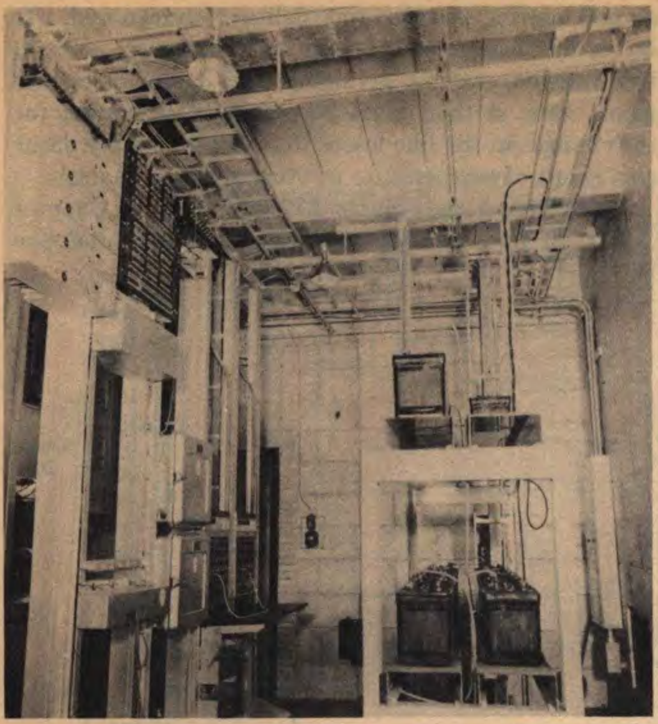


Fig. 386. Frequency Translations in Coaxial Systems



INTERIOR OF AUXILIARY CABLE CARRIER REPEATER STATION

loss of a 1000-mile 165 circuit at the top frequency of the Type-C system is in the order of 110 decibels, while the minimum loss for a Type-J system of equal length reaches a value as high as 205 decibels. Since these circuits must include repeaters which bring the net overall equivalent down to a value of only a few decibels, it follows that a comparatively small percentage change in line loss would completely destroy the usefulness of the circuit, unless promptly and continuously corrected for.

In the open wire Type-C and -J systems, regulation is effected by means of pilot channels. One or more pilot frequencies, occupying positions between the carrier channel bands, are applied to the line at each end. These frequencies are then picked off from the outputs of the intermediate and terminal receiving amplifiers and used to control a regulating circuit connected in the line at the amplifier input. Since the pilot frequencies are transmitted along with the regular channel frequencies, they suffer the same attenuation losses in passing over the line and are affected by any change in the line characteristics in the same way, and to the same extent as, the message currents in their several channels. By establishing normal values for the pilot frequency currents at each amplifier point, therefore, any change due to changing line conditions may be caused to register in such a way as automatically to produce a correcting adjustment of each repeater.

The detailed arrangement for accomplishing this result varies somewhat as between different carrier system designs, but Figure 387 illustrates schematically the regulating circuit used in the J systems. It is also practically identical with the circuit employed in the most recent design of C systems. As was shown schematically in Figure 382 for the Type-J terminal circuit, the pilot channel current is applied to the line at the input of the first group modulator. At repeater points, this single-frequency current is tapped off at the output of the line amplifier and led back through an amplifier and rectifier circuit to a pilot channel control circuit which controls the net loss or gain of a regulating amplifier connected into the main transmission path in front of the line amplifier. As shown in Figure 382, the same general plan applies at the receiving terminal, except that here the pilot channel current is taken off at the output of the auxiliary amplifier following the second group demodulator.

The regulating amplifier circuit consists essentially of a variable attenuator in series with a two-stage amplifier. The attenuator, known as the regulating network, is designed to have loss-frequency characteristics similar to those of the line and is divided into three units of equal loss. Its net loss to through transmission is varied by means of a condenser whose movable plate is rotated under the control of the pilot channel current. The rotor of the condenser is connected to the control grid of the first vacuum tube of the amplifier so that the voltage applied to the amplifier depends upon the position of the condenser plate. When the condenser rotor is at its extreme left position, the regulating network is effectively out of the transmission path. At its extreme right position, on the other hand, the entire network is in the transmission path; and at any other position, the loss inserted is some definite fraction of the total loss of the regulating network. This arrangement provides a very smooth control of the net loss or gain of the regulating circuit and avoids the use of sliding contacts or relays in the transmission path.

It should be noted that the regulating amplifier circuit is not ordinarily intended to introduce any net gain. This function is taken care of by the fixed-gain line amplifier. As a matter of fact, the regulating amplifier circuit usually introduces a net loss, but it varies this loss in such a way as to counteract any changes in the loss of the preceding line section.

The upper part of Figure 387 shows how the pilot channel current controls the position of the regulating condenser in the regulating amplifier circuit. As may be seen, the incoming pilot channel current is first passed through a highly selective band filter and a tuned circuit. It is then amplified by a single-stage feedback amplifier, rectified, and led through the winding of the control relay. This latter is a highly sensitive type of relay designed to act very positively by

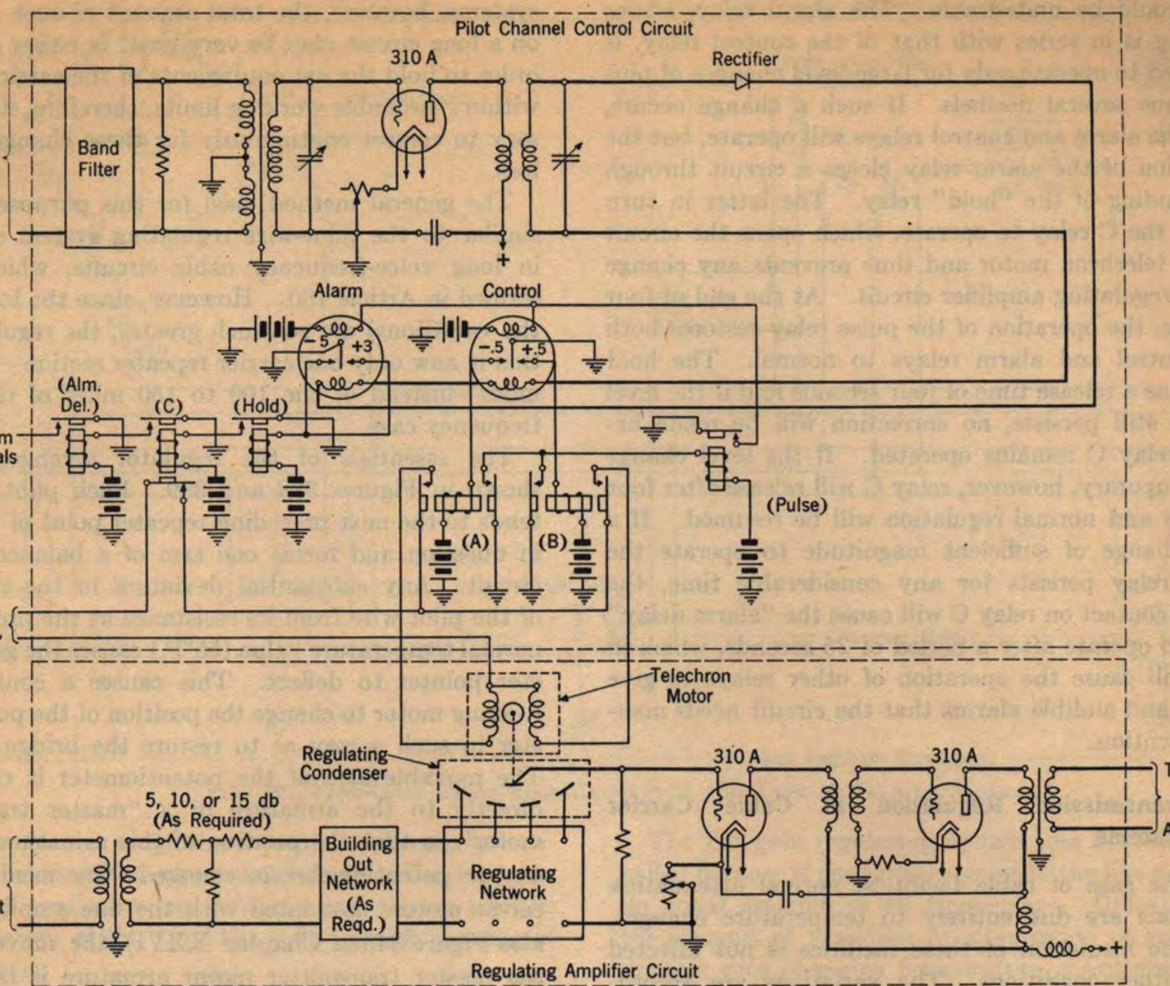


FIG. 387. REGULATING AMPLIFIER AND CONTROL CIRCUIT FOR OPEN WIRE CARRIER SYSTEMS

means of the attraction of magnetic material on its armature to a small magnet on each of the fixed contacts. The relay is given a mechanical bias so that its armature is centered between the contacts when the rectified pilot current is at the normal level. Deviations from this normal level cause the armature to move sharply to one or the other of the contacts, where it will stay until released by the action of the second winding of the relay.

The direction of movement of the armature depends, of course, upon whether the pilot level is increased or decreased. Thus, if the level should increase by .5 db or more, the armature would be moved to the right contact. This would cause the operation of relay B which in closing connects 60-cycle current to the right winding of the "telechron" motor in the regulating amplifier circuit, causing it to rotate in such a direction as to increase the loss of the regulating network. The telechron motor will continue to operate until the control relay is released. This release is effected by means of the "pulse" relay, the winding of which, it may be noted, is connected to a second contact of relay B. The pulse relay is a special slow-operating type which does not operate until four seconds after the path through its winding is closed. When it does operate, a circuit is closed through the second winding of the control relay. This restores its armature to normal, thus releasing relay B and opening the circuit to the telechron motor.

To summarize, then, what happens is that when the pilot channel current deviates from normal, the telechron motor operates for four seconds to counteract the effects of this deviation, and then stops. If sufficient correction is not obtained in this time, the operation is of course repeated. For deviation in the minus direction, the same series of operations occur except that relay A now functions and the telechron motor is rotated in the opposite direction.

The remaining relays shown in Figure 387 are provided to take care of sudden large changes in the pilot channel level. Such changes either require manual attention or are of such short duration that a correction would be undesirable. The alarm relay, whose winding is in series with that of the control relay, is adjusted to operate only for large level changes of plus or minus several decibels. If such a change occurs, both the alarm and control relays will operate, but the operation of the alarm relay closes a circuit through the winding of the "hold" relay. The latter in turn causes the C relay to operate, which opens the circuit to the telechron motor and thus prevents any change in the regulating amplifier circuit. At the end of four seconds, the operation of the pulse relay restores both the control and alarm relays to normal. The hold relay has a release time of four seconds and if the level change still persists, no correction will be made because relay C remains operated. If the level change was temporary, however, relay C will release after four seconds and normal regulation will be resumed. If a level change of sufficient magnitude to operate the alarm relay persists for any considerable time, the second contact on relay C will cause the "alarm delay" relay to operate after a period of 25 seconds, which in turn will cause the operation of other relays to give visible and audible alarms that the circuit needs manual attention.

189. Transmission Regulation in Cable Carrier Systems

In the case of cable facilities, normal attenuation variations are due entirely to temperature changes, since the insulation of these facilities is not affected by weather conditions. The variations are accordingly larger in aerial cable than in underground cable. At the relatively high frequencies of the Type-K carrier systems, however, the total amount of such variation on a long circuit may be very great in either case. In order to hold the net equivalents of the carrier circuits within reasonable working limits, therefore, it is necessary to correct continuously for these changes in line loss.

The general method used for this purpose is quite similar to the pilot-wire regulating system employed in long voice-frequency cable circuits, which is described in Article 180. However, since the losses (and the variations) are so much greater, the regulator section is now only one carrier repeater section—about 16 miles—instead of the 100 to 150 miles of the voice-frequency case.

The essentials of the regulator arrangement are shown in Figures 388 and 389. Each pilot wire extends to the next preceding repeater point of the cable in question and forms one arm of a balanced bridge circuit. Any substantial deviation in the resistance of the pilot wire from its resistance at the pre-selected normal temperature value (55°F.) causes the galvanometer pointer to deflect. This causes a continuously running motor to change the position of the potentiometer in such a way as to restore the bridge balance. The movable arm of the potentiometer is connected directly to the armature of a "master transmitter motor" so that the position of this armature changes as the potentiometer is changed. By means of receiver motors associated with the line amplifiers (see also Figure 345 of Chapter XXVI), the movement of the master transmitter motor armature is translated to the movable plates of the flat gain regulator condensers in the feedback circuits of the amplifiers. The

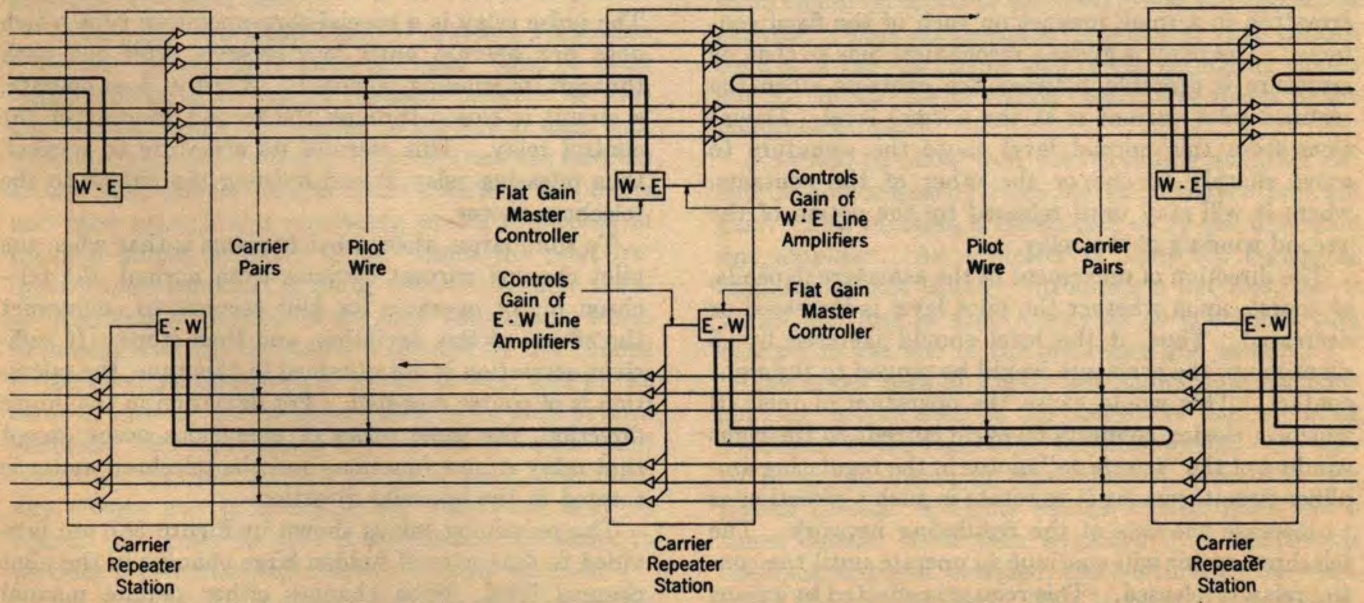


FIG. 388. ARRANGEMENT OF PILOT WIRES IN TYPE-K CARRIER SYSTEM FOR FLAT-GAIN REGULATION

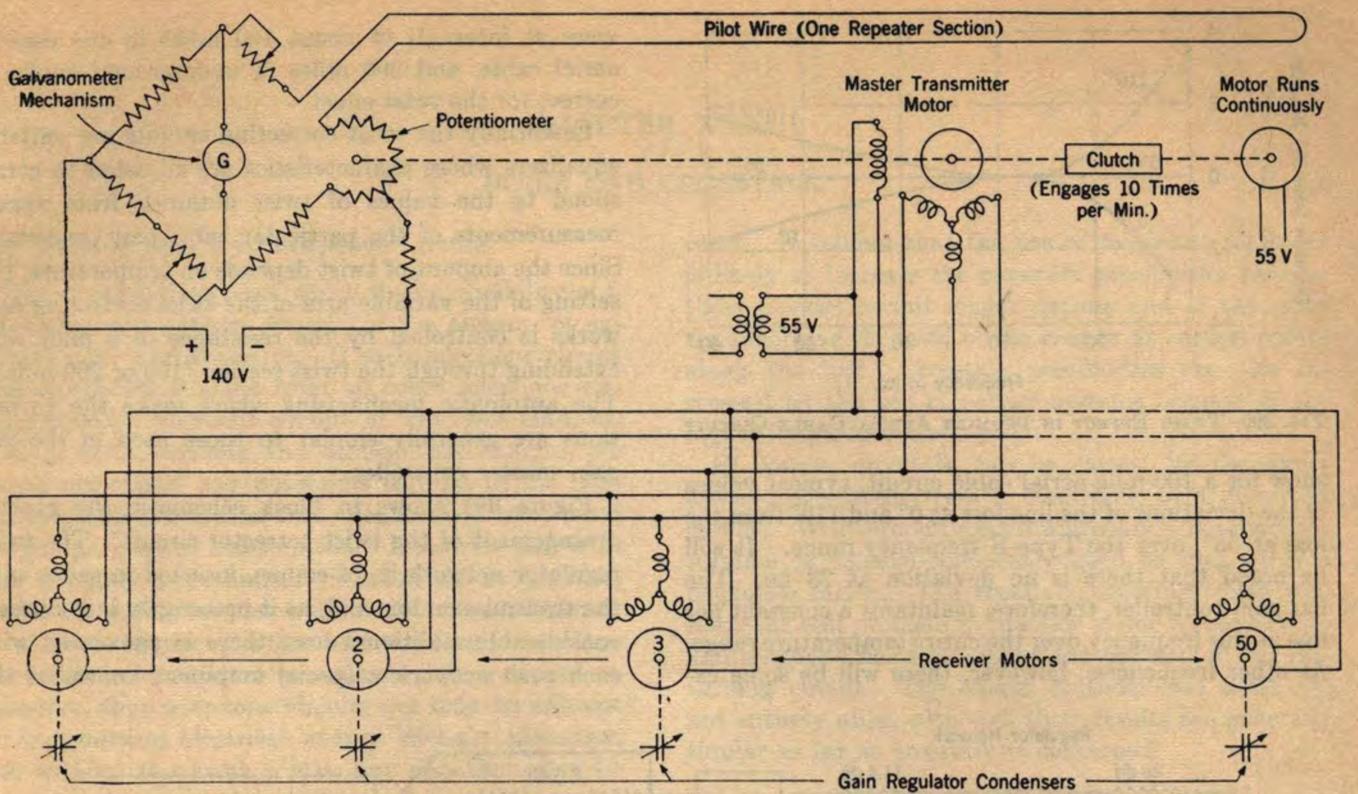


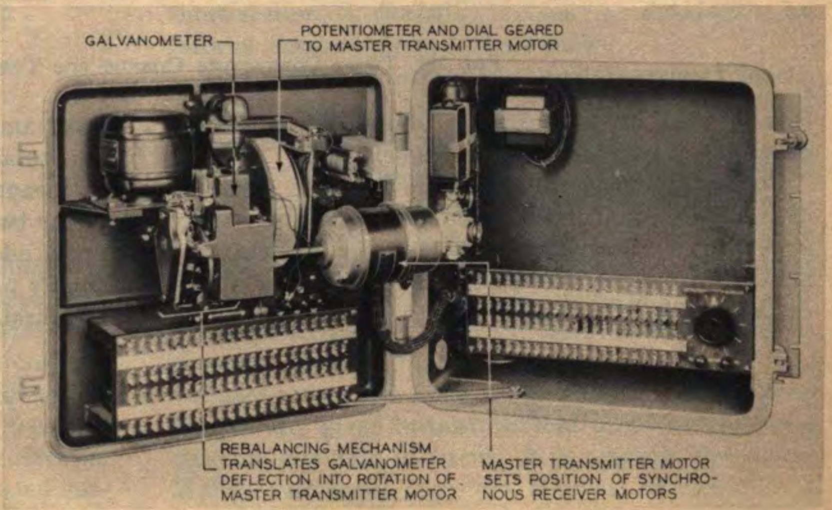
FIG. 389. FLAT-GAIN PILOT-WIRE REGULATOR FOR TYPE-K CARRIER SYSTEMS

net gain of the amplifiers is thus continuously controlled by the pilot-wire temperature. The flat gain regulator condenser has sufficient capacity to provide for variation of the amplifier gain through a range of 14 decibels.

The master transmitter motor and the receiver motors which it controls are merely an electrical means of gearing the bridge potentiometer shaft to the shafts of the flat gain regulator condensers in the line amplifiers. They are small synchronous motors whose armatures are normally stationary and are all held in exactly the same position when a common source of 60-cycle alternating current is applied to their field windings. However, if the armature of one motor (in this case the master transmitter motor) is moved by an external source, the armatures of all the other motors are moved at the same time by exactly the same amount, just as if all the armatures were connected directly together by a perfect mechanical gearing system.

A single master transmitter motor is capable of controlling 50 receiver motors and their associated regulator condensers. This means, of course, that one flat gain master controller is sufficient to take care of 50 carrier systems (600 circuits) in the same cable.

The flat gain regulating system just described is so called because it changes the gains of the line amplifiers an equal amount at all frequencies. Unfortunately, the amount of variation in attenuation of cable conductors with changing temperature is not exactly the same at all frequencies. In other words, the shape of the attenuation-frequency curve for cable circuits is usually slightly different at different temperatures. This effect is known as "twist". An idea of its magnitude may be had from the curves of Figure 390, which



TYPE-K CARRIER SYSTEM FLAT GAIN MASTER CONTROLLER

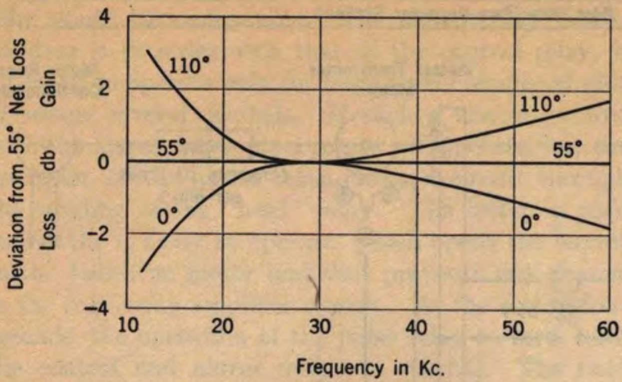


FIG. 390. TWIST EFFECT IN 100-MILE AERIAL CABLE CIRCUIT

show for a 100-mile aerial cable circuit, typical values of the deviations of the line loss at 0° and 110° from the loss at 55°, over the Type-K frequency range. It will be noted that there is no deviation at 28 kc. The flat gain controller, therefore, maintains a constant net loss at this frequency over the entire temperature range. At other frequencies, however, there will be some excess of gain at temperatures above 55° and some deficiency of gain at temperatures below 55°.

Compared in magnitude with the flat 28 kc. variation of loss with temperature, the twist variations may appear to be practically negligible. Nevertheless, they are too large to be allowed to accumulate over a long carrier system. It is necessary to employ special devices at intervals of about 100 miles in the case of aerial cable, and 200 miles in underground cable, to correct for the twist effect.

Essentially the twist correcting circuits are variable equalizers whose characteristics are adjusted to correspond to the values of twist obtained from actual measurements of the particular cable pair concerned. Since the amount of twist depends on temperature, the setting of the variable arm of the twist correcting networks is controlled by the resistance of a pilot wire extending through the twist section (100 or 200 miles). The automatic mechanisms which make the corrections are generally similar to those used in the flat gain master controller.

Figure 391 shows in block schematic the general arrangement of the twist corrector circuit. The twist regulator network is, of course, inserted in series with the transmission line and, as it necessarily introduces a considerable additional loss, there is associated with each such network a special amplifier, known as the twist amplifier, which has a fixed gain of 60 decibels. The total loss of the twist regulator circuit is adjusted by means of pads to be approximately equal to the gain of the twist amplifier so that in net effect the arrangement adds neither loss nor gain to the circuit, but simply provides the relatively small amount of special equalization required to overcome the twist effect.

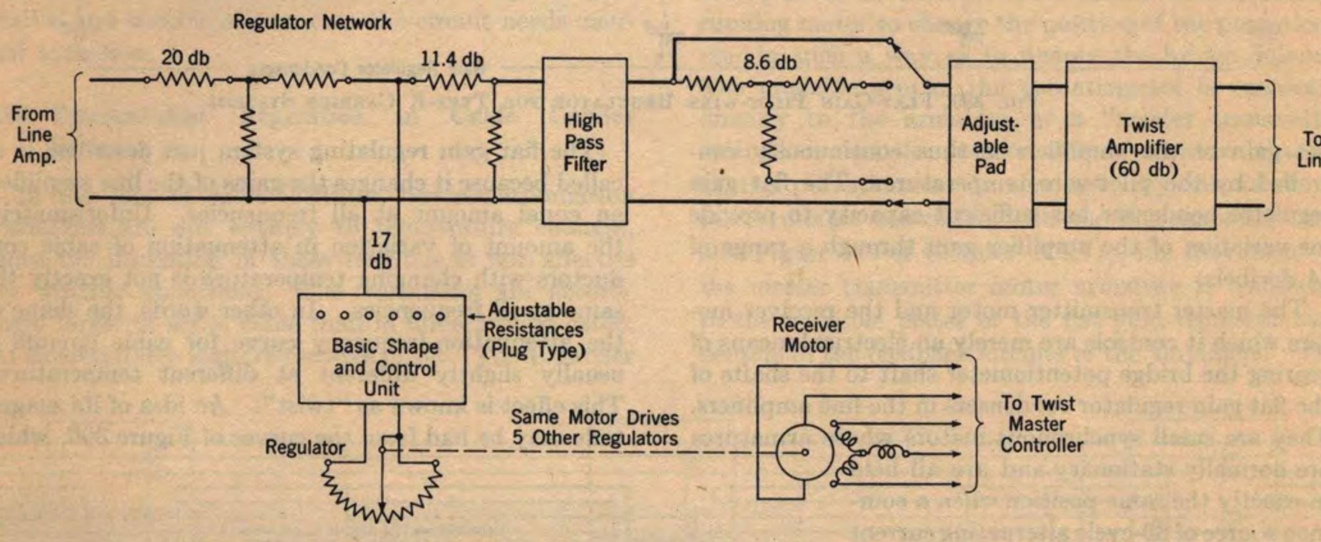


FIG. 391. TWIST CORRECTING CIRCUIT FOR TYPE-K CARRIER SYSTEM

CHAPTER XXXI

NOISE AND CROSSTALK

190. Induced Effects in Telephone Circuits

One of the factors upon which the intelligibility of a telephone conversation depends is the absence of excessive noise and crosstalk. If each telephone circuit was completely isolated from all other telephone circuits or other electrical circuits of whatever kind, including earth currents and atmospheric charges, we would not expect any potentials to exist in the telephone circuit other than those deliberately introduced for the purpose of transmission. However, this is in fact a purely hypothetical situation as, in practice, nearly every long telephone circuit is in close proximity to other telephone circuits, and sometimes to other electrical circuits such as power lines. It is necessary, therefore, that telephone circuits not only be efficient in transmitting electrical energy without distortion and without too great a loss, but also that they be protected against induced electrical currents coming from adjacent telephone circuits or from other electrical circuits.

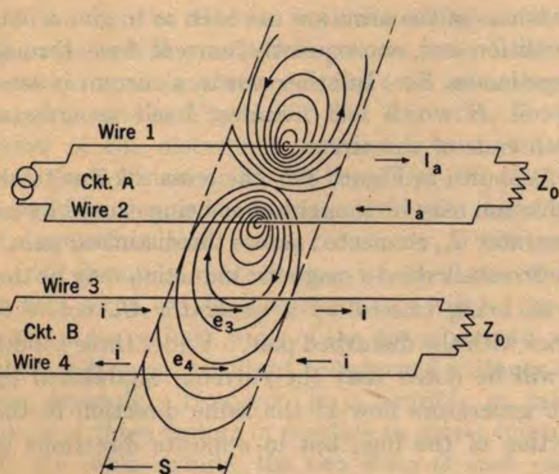


FIGURE 392

As a matter of fact, any two long paralleling telephone circuits that are not "balanced" against each other by means of transpositions, or otherwise, will crosstalk to such a degree as seriously to interfere with their practical use. Furthermore, because crosstalk is largely an inductive effect, its magnitude tends to increase with (1) the length of the paralleling circuits, (2) the strength (energy level) of the transmitted currents, and (3) the frequency of the transmitted currents. It follows that the use of telephone repeaters is likely to increase the crosstalk possibilities because these devices permit longer circuits and at the same time increase the level of the energy at certain points along the line. Crosstalk possibilities are also increased by the use of carrier systems because of the higher frequencies employed.

Of course, crosstalk can be caused by the direct leakage of current from a disturbing to a disturbed circuit. With properly maintained lines, however, insulation is usually sufficiently good to make this a negligible factor. The crosstalk coupling which presents the real problem in practice is due to the electric and magnetic fields set up by the currents in the disturbing circuit. The effects of these two fields are not entirely alike, although their results are generally similar as far as crosstalk is concerned.

191. Causes of Crosstalk

The effect of the magnetic field of one circuit on a second paralleling circuit is called "magnetic induction". Similarly, the effect on the second circuit of the electric field of the first circuit is called "electric" or "electrostatic induction". How magnetic induction causes crosstalk can be seen by referring to Figure 392. This shows the four wires of two telephone circuits running parallel to each other in the usual flat configuration employed on standard open wire lines. If we assume that an alternating voltage is applied to the left end of the disturbing circuit, A, which is made up of wires 1 and 2, the magnetic field existing about a short section, S, of this circuit may be represented by lines of magnetic induction as shown.

At a particular instant in the alternating cycle, the current in wires 1 and 2 may be represented by the equal and opposite vectors designated I_a. As I_a increases or decreases in value, the associated lines of magnetic induction will cut wires 3 and 4 of the paralleling telephone circuit, B. But with the relative spacing of the wires shown in the diagram, more lines will cut wire 3 than cut wire 4. Accordingly, the voltage induced by the magnetic field in wire 3 will be somewhat greater than that induced in wire 4. The voltages induced in both wires are in the same direction at any given instant, so that they tend to make currents circulate in circuit B in opposite directions. If they were equal, therefore, their net effect would be zero. But in so far as the induced voltage e_3 exceeds the induced voltage e_4, there is an unbalanced voltage, e_3-e_4, tending to make a current circulate in circuit B. If the circuit is terminated at both ends in its characteristic impedance, Z_0, the current resulting from this unbalanced voltage induced in a short section of the circuit may be written as—

$$i = \frac{e_3 - e_4}{2Z_0} \qquad (172)$$

The summation of the circulating currents caused by the voltages induced in each small section of the line may result in an appreciable current through both terminals of the circuit, which is due entirely to magnetic induction.

It should be noted that although the current in the disturbing circuit was considered in the above as being transmitted from the left end to the right end of the circuit, the crosstalk current induced in the disturbed circuit appears at both ends of the circuit. The crosstalk appearing at the left end of the disturbed circuit is known as **near-end crosstalk** and that appearing at the right end is known as **far-end crosstalk.**

Now turning our attention to electric induction, Figure 392 may also be used to show the equipotential lines of the electric field established about circuit A under the same conditions as in the previous example. This electric field will set up potentials on the surfaces of wires 3 and 4 and, with the configuration and spacing of the wires shown in the drawing, these potentials will not be equal. The resultant difference in potential between wires 3 and 4 will tend to cause crosstalk currents to flow to both ends of circuit B.

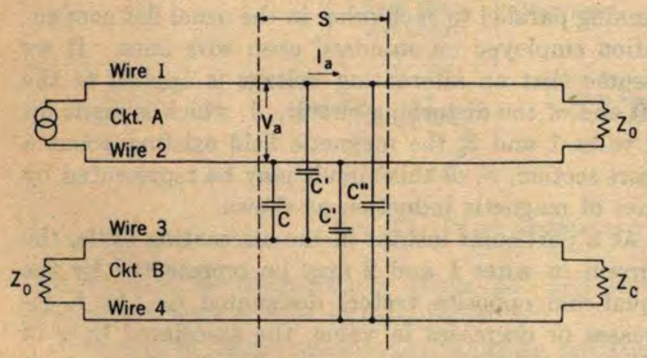


FIGURE 393

The crosstalk effect of electric induction may also be analyzed by consideration of the capacity relationships between the wires of the disturbing and disturbed circuits. Thus, referring to Figure 393, we know that in any unit length of the two circuits there is a definite capacity between wire 1 and wire 2 and between wire 3 and wire 4. Moreover, if the wires are equally spaced as shown, the separation between wires 2–3 is the same as that between wires 1–2 or 3–4,

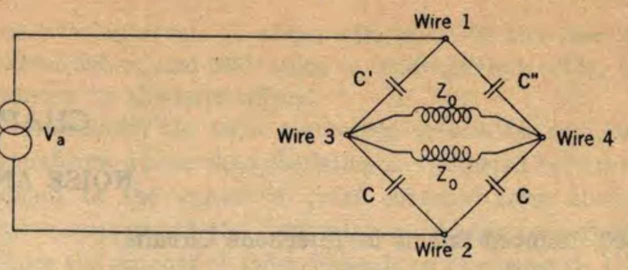


FIGURE 394

and there is therefore the same capacity between wires 2–3 as between the wires of either pair. This capacity is represented in the figure by the small condenser designated C. Similarly, the capacities between wire 1 and wire 3 and between wire 2 and wire 4 are designated by condensers whose capacity, C', is less than C because the separation between these wires is greater. There remains also the still smaller capacity between wire 1 and wire 4, which is indicated by the condenser, C''. Now if we assume an alternating current flowing in circuit A, there will be a difference of potential between wires 1 and 2, which will tend to cause small currents to flow through the various condenser connections to the wires of circuit B. The net effect can best be analyzed by redrawing the diagram of Figure 393 in the form of a Wheatstone bridge network as shown in Figure 394. A study of the capacity values of the arms of this bridge shows that the impedances of the arms are not such as to give a balanced condition and, consequently, current flows through the impedances Z_0. In other words, a current is set up in circuit B which will manifest itself as crosstalk at both ends of the circuit.

As shown in Figure 395, the crosstalk due to electric induction may be thought of as being caused by a small generator G_e, connected across the disturbed pair, while the crosstalk due to magnetic induction may be thought of as being caused by a generator G_m connected in series with the disturbed pair. Under these conditions, it will be noted that the currents established by the two generators flow in the same direction in the left portion of the line, but in opposite directions in the

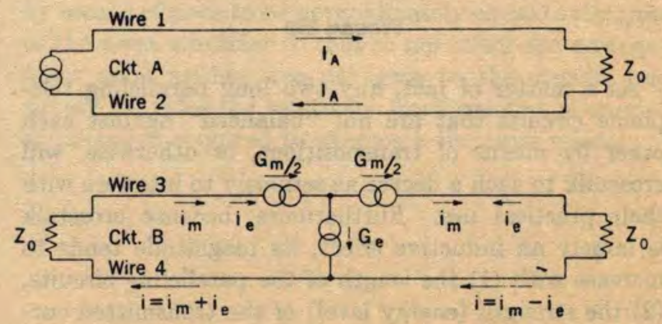


FIGURE 395

right portion. In other words, the crosstalk effects of magnetic and electric induction are additive in the case of near-end crosstalk, but opposed to each other in the case of far-end crosstalk.

192. Principles of Crosstalk Reduction

There are a number of possible ways of eliminating, or at least substantially reducing, the crosstalk induction discussed in the preceding article. One possibility is to arrange the paralleling wires in such a configuration that the effect of the field of one pair will be the same at both wires of the other pair, thus leaving no residual difference to cause currents in the second pair. Two possible ways to effect such a non-inductive configuration are shown in Figure 396. For a number of reasons, however, such wire configurations are not usually practicable.

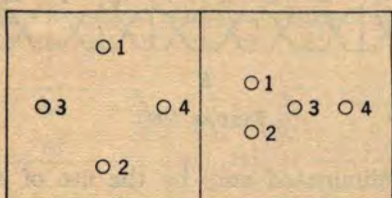


FIGURE 396

Another partial solution is to reduce the separation between the wires of either or both disturbing and disturbed pairs and, if practicable, at the same time to increase the separation between the two pairs themselves. A glance at Figure 392 will show that if the two wires of the disturbing pair are spaced closely together, the fields set up by the two wires will occupy approximately the same position and will therefore tend to neutralize each other. Similarly, if the two wires of the disturbed pair are close together, the effect of any field setup by the disturbing pair will be practically the same on both wires of the disturbed pair, so that there will be no resultant unbalanced voltages to produce crosstalk. However, it is possible to take advantage of these factors in practice to only a limited extent. In cable circuits, the two wires of each pair are close together, but so are the pairs themselves; in open wire circuits, there is considerable separation between pairs, but the two wires of a pair cannot be placed close together.

A third alternative is the use of **transpositions.** The principle involved here can be understood by referring to Figure 397, which shows the same four wires as were indicated in Figure 392. In this case, however, the relative pin positions of wires 1 and 2 are interchanged by means of a transposition in the middle of the short section, S. Under these conditions, it will be evident that when equal and opposite currents are flowing in the two wires of circuit A, the voltages induced in wires 3 and 4 will be in opposite directions on the two sides of the point where circuit A is transposed. Thus, as indicated in the figure, while e_3 is larger than e_4 and e_3' is larger than e_4', e_3 is exactly equal and opposite to e_3' and e_4 is exactly equal and opposite to e_4'. There is therefore no net voltage induced in either wire 3 or wire 4 and, consequently, no crosstalk from circuit A.

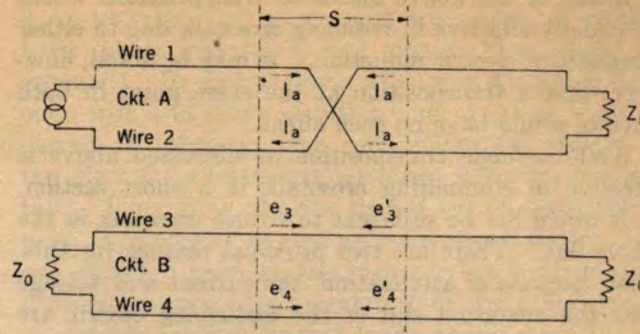


FIGURE 397

The same net effect would be obtained by inserting the transposition in the disturbed circuit B and leaving the wires of the disturbing circuit running straight through, as shown in Figure 398. In this case, the voltage induced in the wire nearer wire number 2 is broken into two equal parts represented by the vectors e_3 and e_3'. Similarly, the voltage induced in the wire farther away from wire number 2 is broken into two parts, e_4 and e_4'. But with the transposition as shown, voltage e_3 combines with voltage e_4' and voltage e_4 combines with voltage e_3'. The induced or crosstalk current in the section, therefore, is—

$$i = \frac{(e_3 + e_4') - (e_4 + e_3')}{2Z_0} \tag{173}$$

But with the transposition in the center of the section as shown, it is obvious that—

$$e_3 + e_4' = e_4 + e_3'$$

The numerator of Equation (173) is therefore equal to zero and there is no resultant crosstalk.

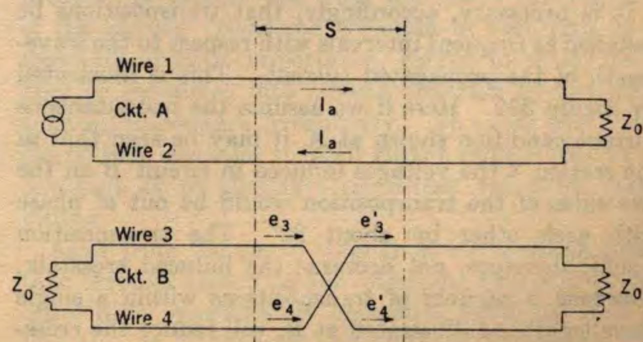


FIGURE 398

Either of the above discussed transpositions would be equally effective in reducing crosstalk due to either magnetic or electric induction. It may be noted, however, that a transposition at the same point in both circuits would have no such effect.

While a single transposition as discussed above is effective in eliminating crosstalk in a short section, *S*, it would not be sufficient to reduce crosstalk in the whole line. There are two principal reasons for this. First, because of attenuation, the current and voltage near the energized end of the disturbing circuit are many times as great as the current and voltage near the other end of the circuit. It could not be expected, therefore, that the induced crosstalk on the energized side of the transposition would be neutralized by the weaker crosstalk induced on the other side of the transposition. As a matter of fact, even in a short section, the transposition will not completely eliminate **near-end** crosstalk because the currents coming back from the far side of the transposition are necessarily attenuated somewhat more than the currents coming back from the near side of the transposition. On the other hand, the transposition is completely effective in the case of **far-end** crosstalk because the slightly higher currents induced on the energized side of the transposition are attenuated more in reaching the far end of the circuit than are the currents induced on the far side of the transposition.

The second reason why a single transposition is not effective in reducing crosstalk to the desired minimum is the phase change of the transmitted currents. In a long circuit, several wave-lengths may be included in the propagation of a voice current from one end to the other. Since crosstalk is an induced effect, its instantaneous value in any small section *S* depends upon the position of *S* with respect to the cycle of current in the disturbing circuit. If *S* is so located that the current or voltage in it has a maximum value, either positive or negative, we cannot expect the crosstalk induced here to be neutralized by the crosstalk in some other similar section, which is located at a point in the line where the voltage or current has a value nearly zero at the same instant.

It is necessary, accordingly, that transpositions be installed at frequent intervals with respect to the wave-length of the propagated current. This is illustrated by Figure 399. Here if we assume the instantaneous current condition shown at A, it may be seen that in the section *S* the voltages induced in circuit *B* on the two sides of the transposition would be out of phase with each other by about 90° The transposition would, therefore, not decrease the induced crosstalk. **However,** a number of transpositions within a single wave-length, as illustrated at B, will reduce the crosstalk to practical limits, although the crosstalk could be

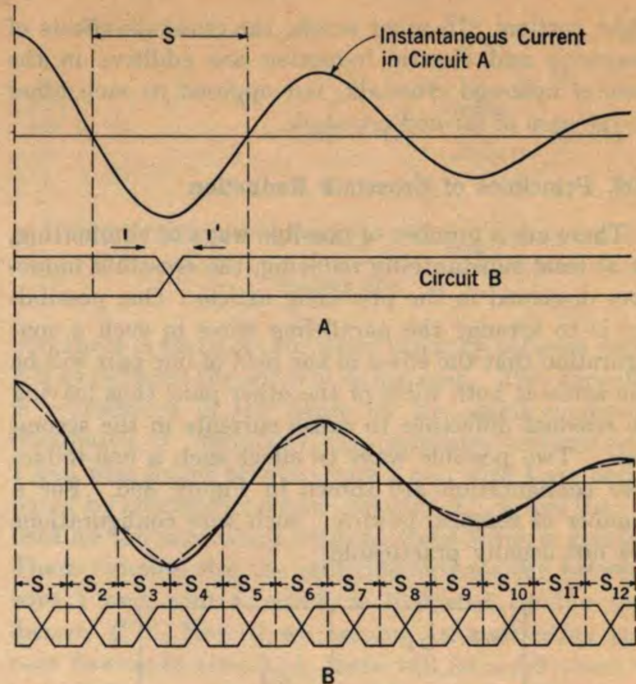


FIGURE 399

completely eliminated only by the use of an infinite number of transpositions. The dotted curve would then become identical with the solid curve.

For voice-frequency transmission, where the frequencies are relatively low and the wave-lengths correspondingly long, it is not difficult to obtain a sufficiently good approximation to this condition. Where high-frequency carrier systems are used, on the other hand, the wave-lengths are so short as to require very closely spaced transpositions. In open wire lines, spacings as close as every second pole are used in practice where the wires are carrying frequencies up to 140 kc. The degree of effectiveness of such closely spaced transpositions is of course dependent upon accurate spacing of the poles themselves, and is finally limited in practice by economic rather than theoretical factors.

There is another fundamental consideration in connection with crosstalk that is of the first order of importance. In what has been said above regarding transpositions, it has been tacitly assumed that the four wires which we were considering were of the same gage and material—and particularly, that the two wires of each pair were electrically identical. The latter is of course ordinarily the case in practice. However, there is always a possibility in open wire lines of the series resistance, or the insulation resistance, of one wire of a pair being slightly different from that of its mate due to imperfect joints, defective insulators, etc. When this occurs, the assumption made in connection with Figures 397 and 398 that the currents flowing in the two wires of the disturbing pair, or the induced currents in the two wires of the disturbed pair, were exactly equal and opposite is no longer true.

Thus, an unbalanced series resistance in wire 1 of Figure 397 would have the effect of reducing the current in wire 1 as compared to the current in wire 2 in an amount depending upon the value of the resistance. Under these circumstances, it is clear that the opposing currents induced in wires 3 and 4 would no longer be exactly equal, and there would therefore be some crosstalk despite the transposition. A resistance or leakage unbalanced in the 3–4 pair would cause a similar result. In this case, any resistance unbalance would cause one of the induced currents to be smaller than the other, with the result that some crosstalk current would flow through the terminal. It is extremely important, accordingly, that the two wires of every talking pair be so constructed and maintained as always to have identical electrical characteristics.

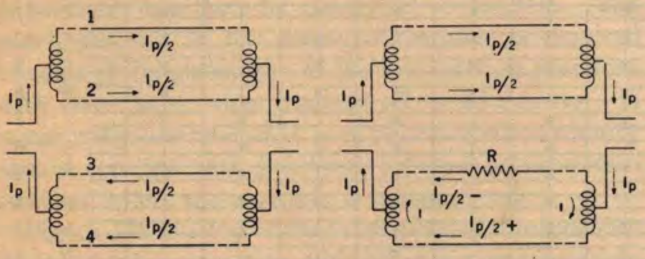


Figure 400

Resistance unbalance is particularly important when two pairs are used to create a phantom circuit. Here, under ideal conditions, exactly half of the phantom circuit current flows in each of the four wires, as shown in Figure 400. The phantom currents in wires 1 and 2 are then equal and in the same direction, and they therefore cause no current to flow through the terminal stations connected to these wires. The same is true of the side circuit made up of wires 3 and 4. A bad joint, or resistance unbalance of any other kind, in any one of the four wires will reduce the current in that wire somewhat. As a result, the phantom currents in the two wires of the pair concerned will no longer be equal and an unbalance current will flow through the side circuit terminal. In other words, the phantom circuit will crosstalk into the side circuit, or vice versa. The effect of resistance unbalance in this situation is ordinarily much more serious than its effect on crosstalk between two side circuits or two non-phantomed circuits.

193. Crosstalk Reduction Practices

In considering practical methods for keeping the crosstalk in long toll circuits at a reasonable minimum, it is desirable first to consider the effects of certain basic design features of long circuits with respect to crosstalk. In general, these will apply equally to both open wire and cable facilities, and at either voice or carrier frequencies. One such important feature is the effect of the location of telephone repeaters on crosstalk. Thus, it is obvious that if two circuits are in close proximity at a point near a repeater station, and one circuit is carrying the high current levels coming from the output of a repeater while the other circuit is carrying the low current levels approaching the input of a repeater, the tendency of the first circuit to interfere with the second circuit is very great. The very small percentage of the current in the first circuit which may be induced into the second circuit will be amplified by the repeater on that circuit along with, and to the same degree as, the normal transmission. The best practical remedy for this condition, of course, is to avoid such situations by keeping circuits carrying high level energy away from low level circuits as much as possible. Where such physical separation between circuits is not feasible, differences in energy level between adjacent circuits can frequently be minimized by proper adjustment of repeater gains when the circuit is designed.

Another basic element of circuit design is that in most of the longer voice-frequency cable circuits and in all carrier circuits, the effect of near-end crosstalk is minimized by the use of separate paths for transmission in the two directions. In cable circuits, the wires carrying the transmission in the two directions are physically separated as much as possible by placing them in different layers or segments of the cable; or, in the special case of cable carrier circuits, in different cables. An equally effective separation is obtained in open wire carrier circuits by using entirely different bands of frequencies for transmission in the two directions.

Furthermore, any near-end crosstalk occurring in spite of these physical separations is returned on the disturbed circuit to the output of an amplifier. Since the amplifier is a one-way device, the crosstalk can proceed no farther and does not reach the terminal of the circuit. Near-end crosstalk in such circuits is therefore of little importance, except in so far as it may be converted into far-end crosstalk by reflection from an impedance irregularity. To avoid this latter effect, it is essential that all circuit impedances be so matched as to eliminate important reflection possibilities.

Aside from the above techniques for avoiding crosstalk through circuit design methods, practical procedures differ considerably depending upon the type of facility. It is desirable, accordingly, to analyze the problem separately for open wire and cable facilities.

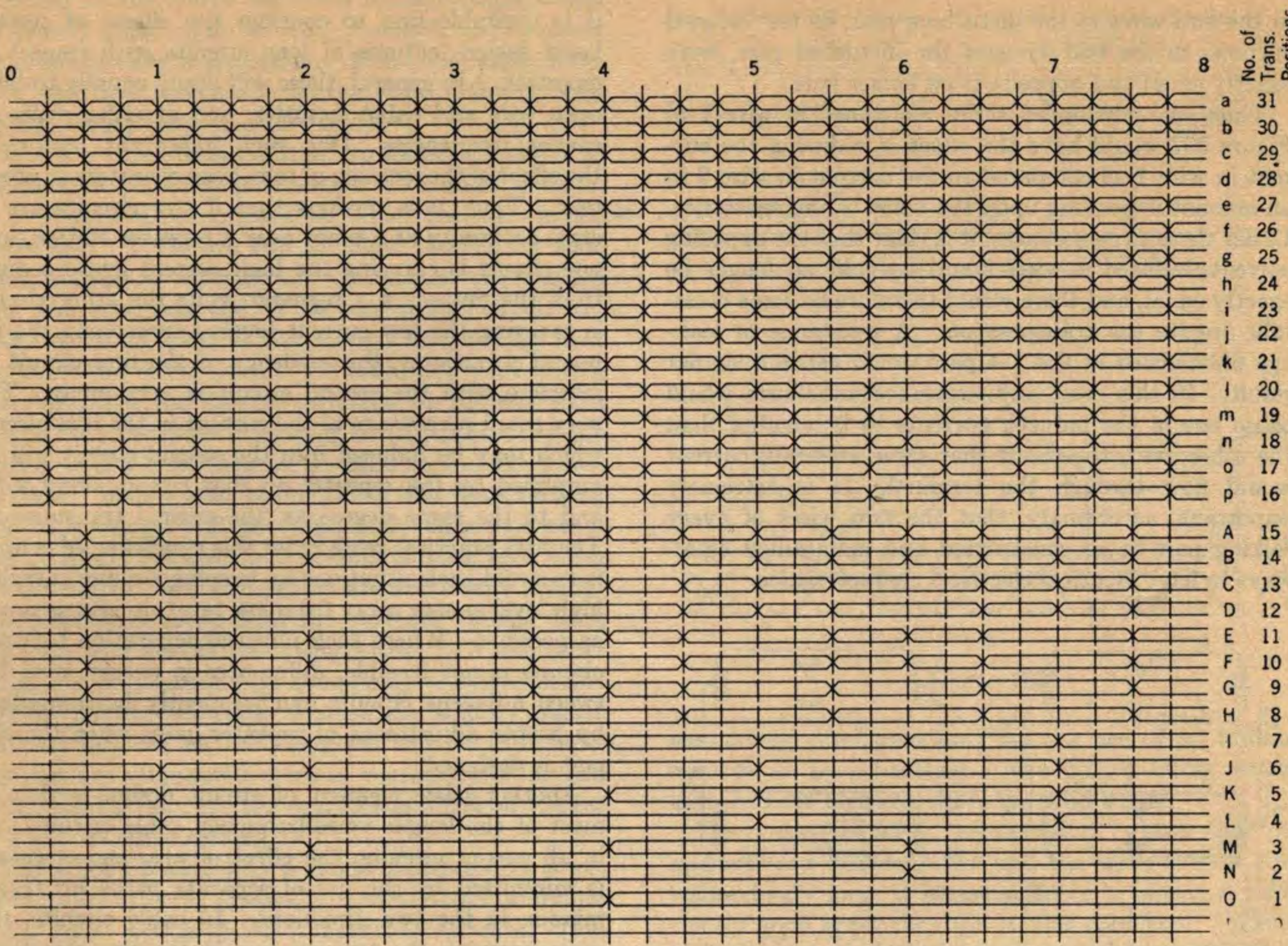


Fig. 401. Standard Transposition Code

In the case of open wire lines, crosstalk reduction depends upon three principal factors—namely, wire configuration on the poles, transpositions, and resistance balance. Resistance balance is primarily a question of maintenance and ordinarily presents no great difficulty. The use of high-frequency carrier systems, with their much greater crosstalk possibilities, has led to the development of new configurations of open wire lines in which the wires of individual pairs are spaced closer together and the pairs are spaced farther apart. One standard pole-head configuration of this kind is illustrated in Figure 276 of Chapter XXII, where it may be noted that the separation of the wires of each pair is 8 inches and the horizontal separation on the crossarm between any two wires of different pairs is at least 26 inches. The vertical separation between crossarms is 36 inches. Six inch spacing between the wires of a pair has also been used in a limited number of cases, but the danger of such closely spaced wires swinging together in the spans tends to restrict this practice to situations where weather conditions are particularly favorable.

The basic principle of transpositions was outlined in the preceding article. It was noted there that a large number of transpositions was needed in any long section of line to reduce crosstalk to the desired extent. In the entire discussion, moreover, only two pairs were considered. In practice an open wire line usually carries many more wires than this, and obviously there are crosstalk possibilities between any two pairs on such a line. These possibilities are greater between the pairs that are adjacent to each other, but all of the other possibilities are sufficiently large that they must be taken into consideration in designing a transposition system for the line. A practical system must also guard against crosstalk between side and phantom circuits and between the phantoms themselves, when such circuits are used.

There is still another extremely important factor which has not been considered up to this time. This is the possibility of crosstalk from one circuit to another via a third circuit. In a line carrying many circuits, there are a large number of these tertiary circuits via which crosstalk might be carried from any one pair to any other pair. Even the hypothetical line that we considered in the first place, carrying only four wires, has two such tertiary circuits. These are the phantom circuit, made up of the two wires of one pair transmitting in one direction and the two wires of the other pair transmitting in the opposite direction; and the "ghost" circuit, made up of the four wires acting as one side of a circuit, with a ground return. (Note that these circuits exist as tertiary crosstalk paths regardless of whether a working phantom circuit is actually applied to the four wires.) Needless to say, the presence of these tertiary circuits in a line complicates the problem of designing effective transposition systems. So much so, indeed, that no attempt can be made here to analyze this problem in detail.

Transposition systems for open wire lines are designed for unit lengths ranging from a few hundred feet to some six or eight miles. The purpose of the design is to approach as closely as possible to a complete crosstalk balance in each such unit section. Any number of sections can then be connected in tandem. The non-uniformity in the length of sections is required because of discontinuities in the line, such as junctions with other lines, wires dropped off or added, etc. It is naturally desirable that such points of discontinuity should coincide with junctions between transposition sections, where the crosstalk is balanced out.

Figure 401 illustrates the fundamental transposition designs used in a section with 32 or less transposition poles. These fundamental types are frequently extended to include 64, 128, or even as many as 256 transposition poles per section, on lines to which high-frequency carrier systems are applied.

Physically, there are two standard methods for effecting transpositions between wires on pole lines. These, known as "point type" and "drop bracket" transpositions, are shown respectively in Figures 402 and 403. The former (point type) is widely used on lines carrying carrier systems because it does not change the configuration of the wires in the adjacent spans, as does the drop bracket type. Where very high frequencies are used, as in the Type-J carrier system, this becomes extremely important. In fact, the sensitivity of these carrier systems to crosstalk is so great that every possible effort has to be made to avoid even slight deviations in the amount of sag of the wires in the spans between poles.

Turning now to cable, the most striking feature of this type of facility with respect to crosstalk is that the conductors are crowded closely together. This is particularly true of the two wires of each circuit pair, which are separated by only thin coatings of paper insulation. As we have already seen, this close spacing of the two wires of a pair in which equal and opposite currents are flowing tends to minimize the external

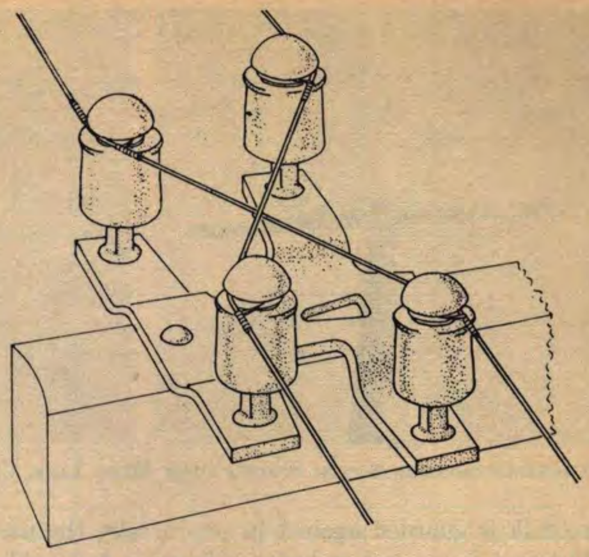


Fig. 402. Point-Type Transposition

effect of the electromagnetic field of the pair. Moreover, in the process of manufacture, the cable conductors are very thoroughly transposed by twisting the two wires of each pair together, by twisting the two pairs of each group of four wires together to form quads, and by spiralling the quads in opposite directions about the cable core. Cables are also so manufactured and installed that their conductors are practically free from series resistance unbalances or insulation leakages. On the other hand, the close spacing of many circuits within the cable sheath, as well as their proximity to the sheath itself, offsets the above advantages to a considerable extent.

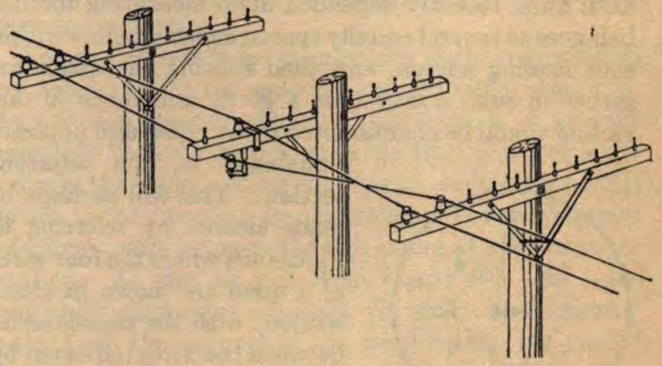
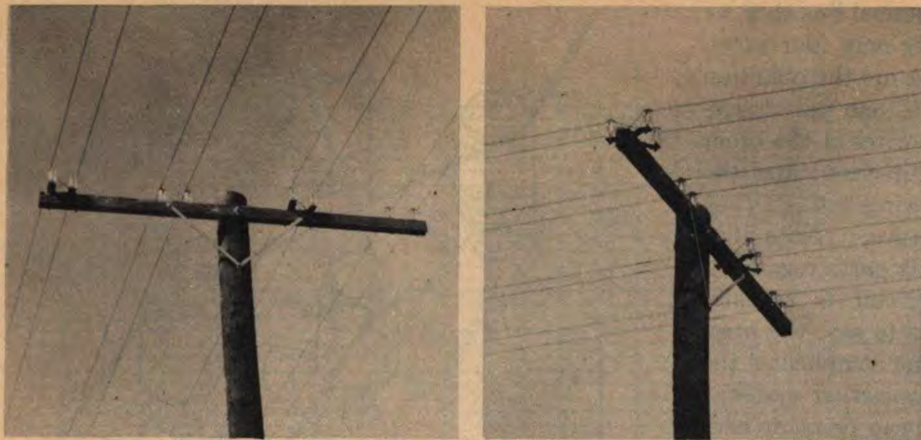


Fig. 403. Drop-Bracket Transposition

At voice frequencies, magnetic induction (inductive coupling) between circuits in a cable is normally so small as to be of relatively little importance in creating crosstalk. The same cannot be said of electric induction (capacity coupling). Despite the most careful manufacturing methods, the capacity unbalances between cable conductors usually remain large enough to cause objectionable crosstalk in long circuits. This



Transpositions in 8-inch Spaced Open Wire Line Using Point-Type Fixtures

crosstalk is guarded against in practice by the use of additional balancing techniques when a toll cable is installed.

Voice-frequency crosstalk between circuits in different quads of a cable can be reduced to a satisfactory minimum at that time by splicing the successive lengths of cable in a more or less random manner such that no two quads are adjacent for more than a small part of their total length. This technique of course has no effect upon the crosstalk between circuits in the same quad. To reduce this crosstalk, it is necessary to measure the capacity unbalances of each quad at the time of installation and then to correct such unbalances as are found large enough to be likely to cause serious crosstalk.

There are two principal methods of effecting this latter correction. The method most extensively used until quite recently depended upon measuring the unbalances at several equally spaced splicing points within each loading section, and then splicing the quads together in such a way that a given unbalance in one section would be counteracted by an equal and opposite unbalance in the adjacent section. This will perhaps be made clearer by referring to Figure 404 where the four wires of a quad are shown in cross-section, with the capacitances between the wires indicated by small condensers. The wires marked 1 and 2 form one pair of the quad, and the wires marked 3 and 4 the other. (The capacities between the pairs themselves are not shown because they have no effect on crosstalk.) The ideal condition in such a quad is that the values of all four capacities A, B, C, and D shall be equal, and that capacity E shall equal capacity F and capacity G shall equal capacity H. In this case there is no unbalance within the quad and no crosstalk. However, if it is found, for example, that capacity A in a certain quad of one section of the cable is too low, this quad can be spliced to a quad in an adjacent section of equal length in which capacity A is too high by an approximately equal amount. The net unbalance of the connected quad over the two sections will thus be made to approach zero.

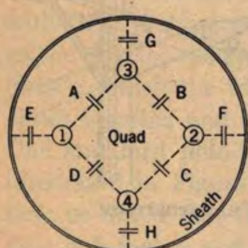
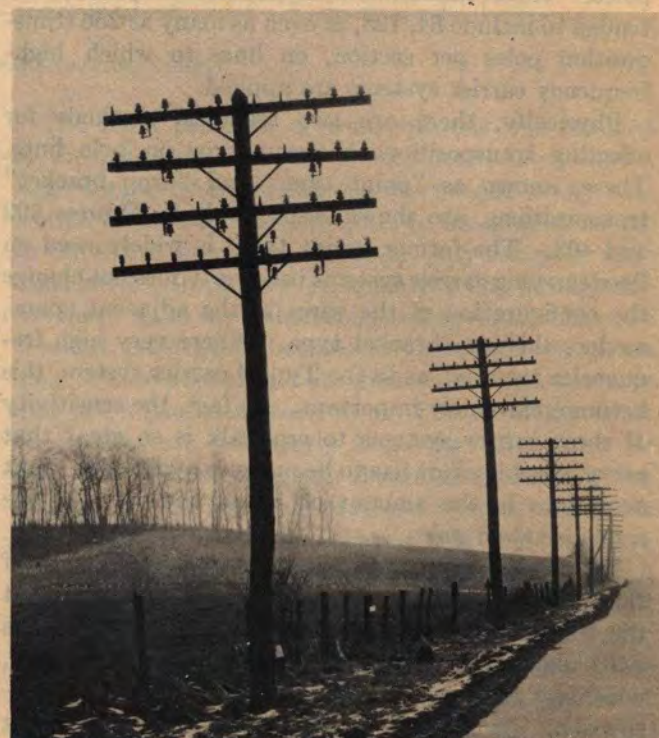


Figure 404

In current practice, it is usually found more economical to counteract the unbalances in part by connecting small balancing condensers into the circuits at one or two points in each loading section. This, combined with a limited number of "test splices" as above, effects the net result desired with greater accuracy, and reduces the number of capacity unbalance tests that have to be made when a cable is installed. These balancing condensers consist of short lengths of two parallel insulated fine-gage wires wound helically around a non-conducting core. Two terminals of this tiny condenser are connected across the two line conductors whose capacity it is desired to increase, and the other ends of the wires can be cut off at whatever point is necessary to give the condenser the precise value of capacity required. A large num-



Transpositions in 12-inch Spaced Open Wire Line Using Drop-Bracket Fixtures

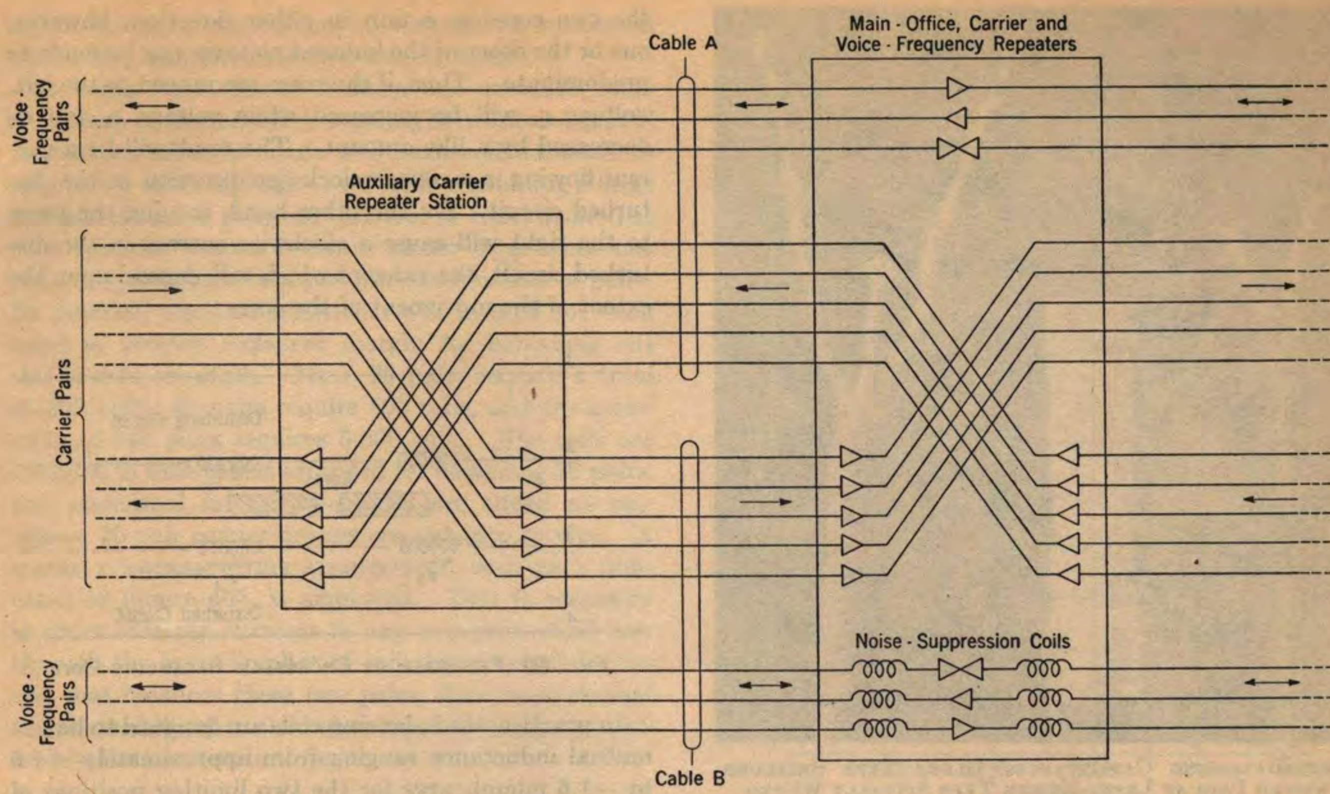


FIG. 405. ARRANGEMENTS FOR REDUCING NOISE AND CROSSTALK IN TYPE-K CARRIER SYSTEMS

ber of these condensers can be included within the lead sleeve at a splicing point. In certain cases, where the cable conductors are to be used for 4-wire circuits, it is practicable to balance the capacities for a whole repeater section by adding condensers of this type at one end.

The capacity balancing methods outlined above have been found adequate in practice for keeping crosstalk to a tolerable minimum in **voice-frequency** cable circuits. When **carrier** systems are applied to cable circuits, the crosstalk problem becomes much more severe. In this case, while capacity coupling is still of consequence, inductive coupling becomes much more important as a cause of crosstalk. In fact, at the highest frequencies of the Type-K carrier systems, it predominates over capacity coupling as a cause of crosstalk in the ratio of about 3 to 1. Accordingly, additional crosstalk reduction measures must be applied to cable conductors used for carrier systems.

The crosstalk possibilities at these high frequencies are so great, in fact, that a number of basic changes in circuit design are required. In the first place, the carrier pairs are used for carrier transmission only. Next, the transmitting paths in the two directions are kept entirely separated by using separate cables for transmission East to West and West to East (or a special cable, with a shield between pairs transmitting in opposite directions). The circuits in the two directions are likewise kept separated within the terminal offices and repeater stations, and shielded office wiring is used in all cases. This means that the energy levels of the carrier currents are approximately the same in all physically adjacent conductors, and that near-end crosstalk possibilities are completely eliminated (assuming that reflection effects have been properly guarded against).

Far-end crosstalk between carrier pairs is minimized by balancing out the capacity and inductive couplings. In addition, special precautions are taken to prevent crosstalk between carrier pairs via the voice-frequency pairs in the cable. Most effective in accomplishing this latter, is the complete transposition of the entire group of carrier pairs between the two cables at each repeater station. As may be seen from Figure 405, this automatically eliminates crosstalk via the voice-frequency pairs from the outputs of the amplifiers in the carrier pairs to the inputs of amplifiers in other carrier pairs. Carrier filters or noise suppression coils are also inserted in the voice-frequency pairs at voice-frequency repeater stations, and certain other points, to discourage the transmission of induced currents of carrier frequencies over the voice-frequency conductors.

The methods of balancing out capacity coupling between the carrier pairs themselves are essentially the same as were discussed above in connection with voice-frequency transmission. In balancing out crosstalk

Cable Balancing Capacitances—Older Type Shielded Twisted Pair at Left—Newer Type Spirally Wound Pair at Right—Unit Used at End of Repeater Section in Center

due to inductive coupling, entirely new methods have had to be devised. The fundamental problem is to balance every carrier pair against every other carrier pair in the same cable, in each repeater section. The method used depends upon counteracting the crosstalk currents with equal currents flowing in the opposite direction. Thus, if in a given disturbed circuit a crosstalk current is flowing in a clockwise direction, it is desired to set up an equal current in the circuit flowing in a counterclockwise direction.

This result can be effected by means of tiny transformers connected between each carrier pair and every other carrier pair. However, since it is necessary to control the magnitude of the artificially induced currents and also to cause them to flow in either direction, depending upon the direction of the crosstalk current, the transformers must be designed so that the coupling between circuits can be adjusted and so that they can be poled in either direction. The method used to obtain this result is indicated schematically in Figure 406. Here, it may be noted that there are really two separate transformers, one having a reversed winding in the disturbing circuit so that a current, I, flowing in the disturbing circuit will induce oppositely poled voltages in the disturbed circuit. If the cores of the two transformers are centered as shown in the drawing, the induced voltages will be exactly equal and the net effect on the disturbed circuit will be nil. By moving the two cores as a unit in either direction, however, one or the other of the induced voltages can be made to predominate. Thus, if the cores are moved to the left, voltage e_1 will be increased while voltage e_2 will be decreased by a like amount. The result will be a current flowing in a counterclockwise direction in the disturbed circuit. On the other hand, moving the cores to the right will cause a clockwise current in the disturbed circuit, the value of which will depend upon the extent of the movement of the cores.

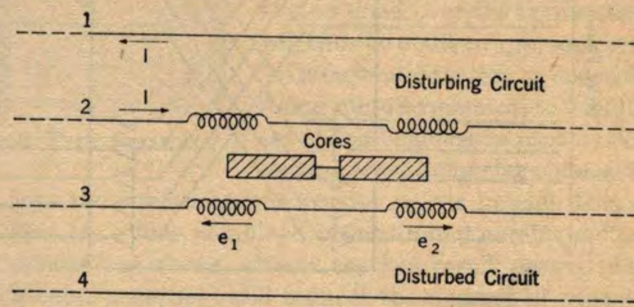
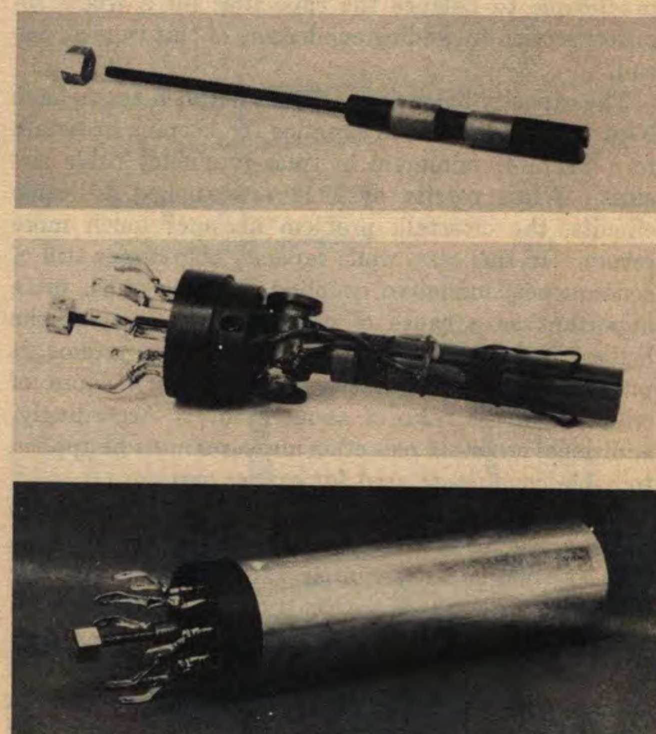


Fig. 406. Principle of Crosstalk Balancing Coil

In practice, the balancing coils are designed to have a mutual inductance ranging from approximately +1.6 to −1.6 microhenrys for the two limiting positions of the cores. The coils are mounted in small cylindrical containers about $4\frac{1}{2}$ inches long and $1\frac{3}{8}$ inches in diameter and arranged for rack mounting. The position of the coil cores is controlled by a screw by means of



Construction of Crosstalk Balancing Coil which the core can be moved through its maximum travel of ½ inch in about 16 complete turns.

In using these coils to balance out crosstalk, measurements of the inductive coupling between each pair of conductors must be made and each coil adjusted to counteract this coupling. In a cable containing a large number of carrier pairs, the number of coils required at each repeater station becomes rather large since one coil is required for every possible combination of pairs. In practice, also, an additional coil is used for each quad to provide sufficient margin for balancing out side-to-side crosstalk. Thus, 20 pairs require a total of 200 coils, 40 pairs require 800 coils, and the maximum of 100 pairs requires 5,000 coils. The coils are installed in unit panels arranged for balancing 20 pairs, and additional intergroup panels are added as successive 20 pair carrier groups are put into service. A special crisscross wiring arrangement, such as is indicated in Figure 407, is employed. This is necessary in order that the currents in any two pairs shall flow through the same number of coils before reaching the coil that balances these two pairs, thus insuring that the phase shift up to the balancing coil will be approximately the same on both pairs.

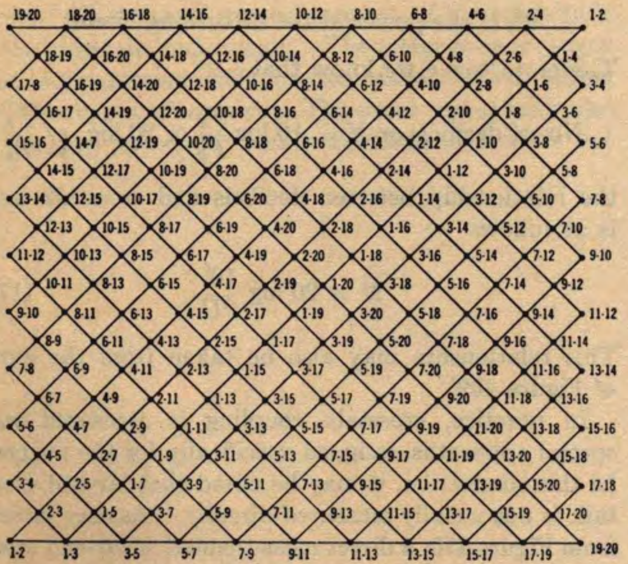


FIG. 407. METHOD OF CONNECTING CROSSTALK BALANCING COILS

194. Noise

Currents within the voice-frequency range, induced into a telephone circuit from electrical power circuits, are manifested to a listener on a disturbed telephone circuit as noise. In many cases crosstalk currents may also appear merely as noise. This is particularly true in the case of cable circuits where any crosstalk heard is likely to come simultaneously from a considerable number of other circuits, and appears to the listener on the disturbed circuit as a special form of noise, called "babble". In other words, it is just an unintelligible conglomeration of speech sounds coming from a large number of sources.

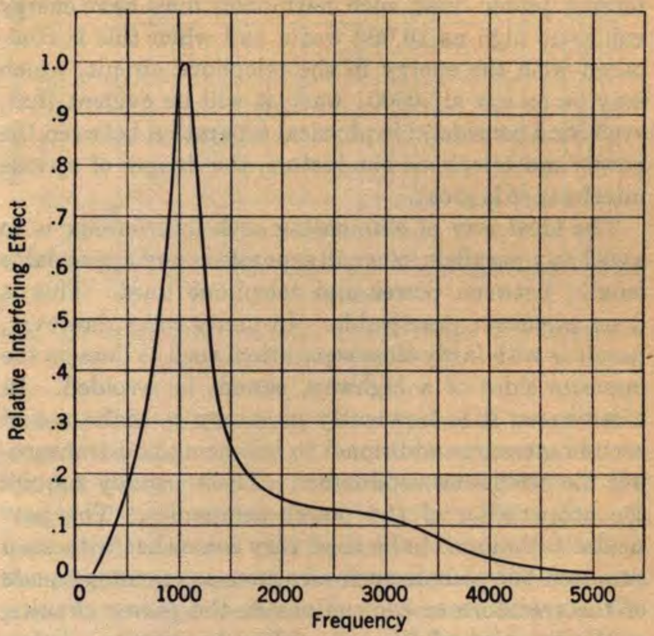


FIG. 408. RELATIVE INTERFERING EFFECT OF NOISE AT DIFFERENT FREQUENCIES

The disturbing effect of noise to a listener depends first, of course, upon its volume. It also depends upon the frequency of the noise currents. Figure 408 shows the results of tests that have been made to determine the relative disturbing effects of various noise frequencies. It will be noted that the disturbing effect peaks up rather sharply in the neighborhood of 1100 cycles. Where noise is of appreciable volume—particularly in the more sensitive frequency range—it is naturally annoying to the telephone user and may seriously reduce the intelligibility of conversation. It is accordingly necessary to keep the noise in working telephone circuits below those limits where its interfering effect on conversation will be important.

Since noise is essentially an induced effect like crosstalk, similar measures are used to counteract it. Careful resistance balancing of the telephone conductors, the use of transposition systems, and other measures taken to avoid crosstalk, are likewise effective in reducing noise. However, such measures alone may be inadequate to keep noise within the desired limits. This is a result of the fact that paralleling power lines are the principal source of noise, and the power carried over such lines is greater by tremendous percentages than that carried over any telephone circuit.

Of course, the usual fundamental frequency of power transmission is 60 cycles and this frequency is too low to interfere appreciably with telephone transmission. Unfortunately, however, the currents transmitted over the power line usually include several harmonics of the fundamental frequency, and these may lie well within the range of telephone frequencies. In high tension power lines, such harmonics may have energy values as high as 10,000 watts and when this is compared with the energy in the telephone circuit, which may be as low as .00001 watt, it will be evident that, even for a considerable physical separation between the power and telephone conductors, the danger of serious interference is great.

The ideal way of eliminating such interference is to avoid any parallels, of small separation and appreciable length, between power and telephone lines. This is done whenever practicable. In many cases, however, parallels with fairly close separation, such as lines on the opposite sides of a highway, cannot be avoided. In these cases, it is frequently necessary to make use of certain measures additional to balancing and transposing the telephone conductors. These usually require the cooperation of the power companies. The particular techniques to be used vary somewhat with each situation but include such measures as rearrangements of the transformer connections in the power circuits, or the insertion of filters to reduce harmonics. Other methods frequently employed include changes in the configuration of the power wires on their poles to make for better electrical balance, and transpositions of the power wires. Such power line transpositions have essentially the same effect in balancing out the magnetic fields as do transpositions in a disturbing telephone circuit.

195. Crosstalk and Noise Measurements

The ideal objective of the various methods discussed above for counteracting crosstalk and noise induction in telephone circuits is, of course, to eliminate their effects altogether. In practice this ideal is rarely attained. But certain practical limits are established, and every reasonable effort is made to keep the crosstalk and noise below these limits. In designing and maintaining circuits, therefore, it is necessary to be able to make definite quantitative measurements of both crosstalk and noise. As in any other kind of measurement, this requires the establishment of definite units.

Since crosstalk represents the transmission of telephone currents from one circuit to another, it can be measured directly in terms of the transmission loss between the two circuits in the same way as direct telephone transmission. This loss may be expressed either in terms of decibels or as "crosstalk units". The number of crosstalk units (CU) is equal, by definition, to 10^6 times the ratio of the current in the disturbed circuit to that in the disturbing circuit at the

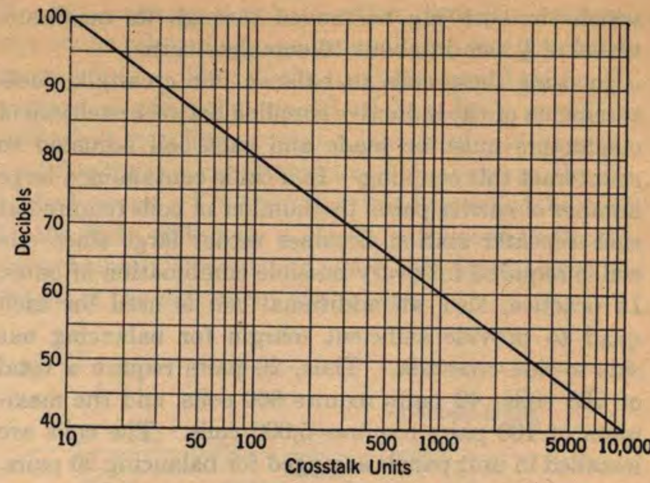


FIG. 409. RELATION BETWEEN DECIBELS AND CROSSTALK UNITS

two points under consideration; or, if the circuit impedances are not identical, 10^6 times the square root of the power ratios. Expressed as an equation—

$$\text{No. of crosstalk units or } CU = 10^6 \times \sqrt{\frac{P_R}{P_S}} \qquad (174)$$

where P_R is the power in the disturbed circuit and

P_S is the power in the disturbing circuit.

Therefore, since we know that—

$$\text{No. of decibels or } N = 10 \log \frac{P_S}{P_R} \text{ or } 20 \log \sqrt{\frac{P_S}{P_R}}$$

the relationship between decibels and crosstalk units is as follows—

$$N = 20 \log \frac{10^6}{CU} \qquad (175)$$

This relationship may also be taken from the curve of Figure 409.

In practice, crosstalk coupling is measured with special apparatus designed specifically for the purpose. In this connection, it may be noted that far-end crosstalk is not usually measured directly. As may be seen from Figure 410, a direct measurement of far-end crosstalk would involve determining the ratio of the current I_a at one end of the disturbing circuit A to the current

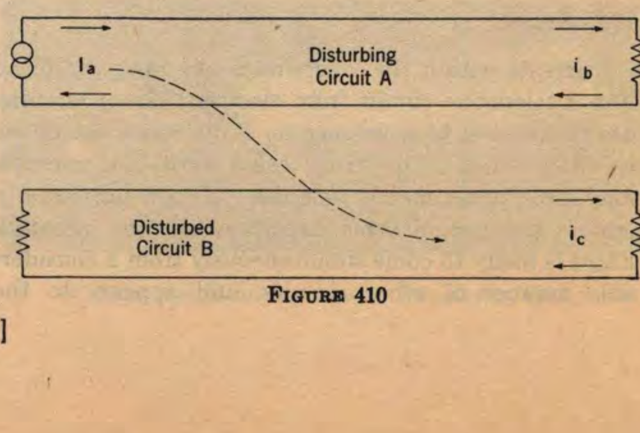


FIGURE 410

i_c at the other end of the disturbed circuit B. It is usually more convenient to measure the ratio between i_c and the current i_b at the far end of the disturbing circuit. This ratio is called the "measured far-end" crosstalk. If desired, it can be readily converted into true far-end crosstalk by multiplying by the current ratio $\frac{i_b}{I_a}$ of the current in the disturbing circuit—or, if the measured crosstalk is expressed in decibels, by simply adding the transmission loss of the disturbing circuit in decibels.

For measuring noise, a basic reference point has been selected, which is equal to 10^{-12} watts of 1000-cycle power. Noise may then be measured as the **number of decibels above this reference noise.** A unit known as the Noise Unit is also employed. Its value is such that the reference noise defined above is equal approximately to seven Noise Units. The relationship between Noise Units and decibels above reference noise is given approximately by the curve of Figure 411.

Noise measuring meters are now in general use by means of which the noise on a circuit can be read directly from a scale calibrated in decibels above reference noise (see Article 207). These same meters are also used to measure crosstalk volume in the same terms. This gives a general measure of the magnitude of the crosstalk on a circuit. The meter is so arranged that either measurement can be made without taking the circuit out of service. The noise meter may also be used to measure crosstalk coupling in decibels but

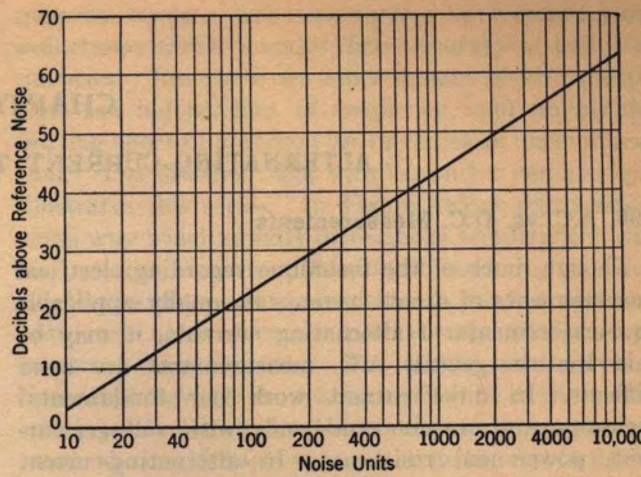


Fig. 411. Relation Between Noise Units and Decibels Above Reference Noise

such a measurement requires removing the two circuits in question from service. The power source applied to the disturbing circuit is preferably an oscillator of the "warbler" type, as described in Article 208. Then if the output volume of this oscillator (which is applied to the disturbing circuit) is first measured with the noise meter in terms of decibels above reference noise and the crosstalk coming from this same source is measured on the disturbed circuit, the difference between the two measurements gives the crosstalk coupling in decibels directly. This, of course, can be readily converted to crosstalk units if desired.

CHAPTER XXXII

ALTERNATING-CURRENT TESTS AND MEASUREMENTS

196. A.C. vs. D.C. Measurements

Though much of the technique regarding electrical measurements of direct currents is equally applicable to measurements of alternating currents, it may be said that in general A.C. measurements are more difficult. In direct-current work our fundamental measurements are concerned only with voltage, current, power and resistance. In alternating-current work, while voltage, current, and power are still the fundamental quantities, their interrelationship is no longer simple but involves considerations of phase, frequency, etc. Again, the measurement of alternating voltages and currents must presuppose some standard wave shape and make some supposition as to the basis of measuring a quantity which is ever varying, i.e., we may measure an instantaneous value, an effective value or a maximum value. Furthermore, in dealing with wave shapes other than sine waves, we must effect some analysis into a fundamental sine wave and harmonics of this fundamental, in order to analyze the conditions correctly (see Appendix IV). These new conditions are responsible for complications incidental to the measurement of the quantities which correspond to those we encounter in direct-current work, and introduce the necessity for more elaborate and painstaking methods for the complete analysis of A.C. phenomena. Moreover, it may be said that a degree of instability is inherent in certain of the properties met with in A.C. work, whereas in D.C. work this difficulty is not encountered. To illustrate, the direct-current resistance of a coil winding remains practically fixed, with the exception of minor changes in values due to temperature, while the alternating-current resistance of the coil may be less stable due to certain additional factors upon which it depends. These include the magnetic properties of the iron core and the physical relationship of the winding to the iron core, both having to do with certain power losses which in turn affect the resistance to alternating current.

It is not always the practice in alternating-current work to make measurements with the basic units, i.e., ampere, volt, watt, etc. We may employ as standards other units based either directly or indirectly upon the fundamental units. For instance, in telephone transmission work it is quite possible to determine the attenuation from the relationship—

$$\alpha = 2.303 \log \frac{I_s}{I_r}$$

by making current measurements, but this method is seldom used in practice. Instead, as we know, the decibel is used as a comparison standard and the measurement is expressed in decibels rather than as a numerical value of α.

On the other hand, the instruments designed to measure fundamental quantities are none the less important because apart from their field use, which may be limited in some cases, the same principles of operation employed in these devices are frequently employed in connection with other measuring apparatus. We shall, therefore, consider first of all the measurement of the fundamental alternating-current quantities.

197. A.C. Ammeters, Voltmeters, and Wattmeters

Indicating instruments, such as ammeters, voltmeters, and wattmeters for alternating-current measurements, are similar in appearance and in manipulation to direct-current instruments but have certain differences in design. A direct-current ammeter, such as was described in Chapter IV, if connected in series with an alternating E.M.F., would tend to indicate the instantaneous value of the current in the circuit. Now this value is constantly changing, and in the case of a 60-cycle power circuit, for example, the change is from zero to a maximum to zero 120 times a second, 60 such changes occurring while the current flow is in one direction, and the remaining 60 while the current flow is in the opposite direction. It is not possible for the needle to fluctuate so rapidly and consequently it would stand at zero, not responding to any value of current through the instrument. However, by substituting a coil winding for the permanent magnet of the D.C. instrument, we can obtain a definite deflection on the ammeter scale and this indication will depend upon the **effective** value of the current.

This, briefly, is the fundamental difference between D.C. and A.C. ammeters and voltmeters of this type. The coil winding and the moving element are connected in series so that whenever the current reverses in one, there is a similar reversal in the other. Consequently, the reaction between the magnetic field of the coil and the current in the moving element is always such as to turn the moving element in the same direction. This is illustrated by Figure 412. Here a movable coil is suspended within, and by means of a spiral spring is held perpendicular to, a stationary coil. The magnetic

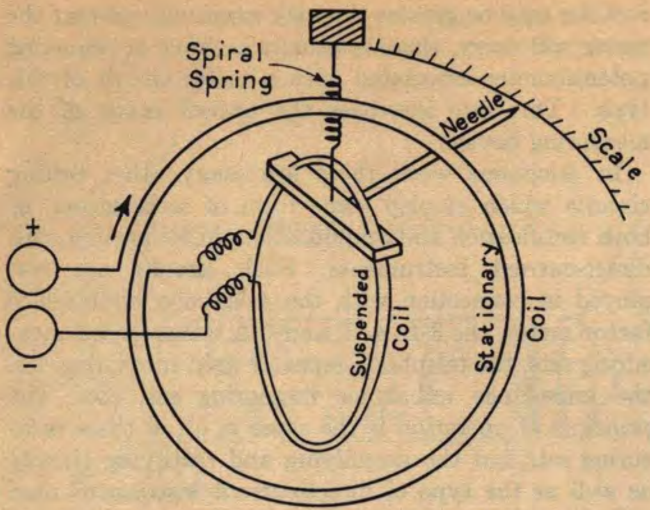


FIG. 412. DYNAMOMETER TYPE A.C. METER

field of the movable coil tends to align itself with that of the stationary coil, and the direction of rotation of the movable coil is the same regardless of the direction of the current through the two coils in series. If the instrument is a voltmeter it must, of course, have a high resistance; if an ammeter, it must have a low resistance, which can be secured by the use of a shunt.

In alternating-current work, unless the angle of lead or lag between the voltage and current is known, it is not possible to use an ammeter and a voltmeter for the measurement of power, inasmuch as the power equation is—

$$P = EI \cos \theta$$

instead of

$$P = EI$$

The power, therefore, cannot be determined by simply multiplying together the measured voltage and current values; a wattmeter must be used for accuracy. The A.C. wattmeter likewise employs two coils, but in connecting such an instrument in the circuit, one coil is connected in series so that the current in it varies as the line current, while the other coil is connected across the circuit so the current in it is proportional to the voltage. Such an arrangement automatically takes care of any phase difference between voltage and current, and the indication of the wattmeter, therefore, depends upon the power in the circuit.

The ammeter, voltmeter, and wattmeter described in the foregoing are said to employ the "dynamometer" principle. The commercial types of these instruments for alternating-current work are usually designed for a single frequency, or at best a narrow band of frequencies. It is not possible, for example, to use the rotating coil mechanism designed for 60-cycle power circuits in connection with telephone current frequencies because such instruments have considerable inductance which impairs their accuracy at high frequencies. But there are other designs of instruments that are independent of frequency, and employ the heating effect of a current as the basis of their operation. The so-called "hot wire" ammeter perhaps best illustrates this series. In Figure 413, h represents a small wire which rapidly increases in temperature with an increase in current flowing through it. W_1 and W_2 are the instrument connections to this hot wire, and both ends of the wire are permanently fixed, though insulated from the case. The middle of the wire is connected through the insulating link L to the needle which is pivoted at P. As in other indicating instruments the needle has a spring attached to it, but unlike other indicating instruments, when the temperature of the wire is above normal this spring tends to make it stand at full scale reading. However, when the hot wire is at normal temperature, it is so constructed as to pull the needle to the zero position. As the current flows through the wire and increases its temperature, the wire expands and the needle is allowed to give a scale reading. The scale is so calibrated as to indicate

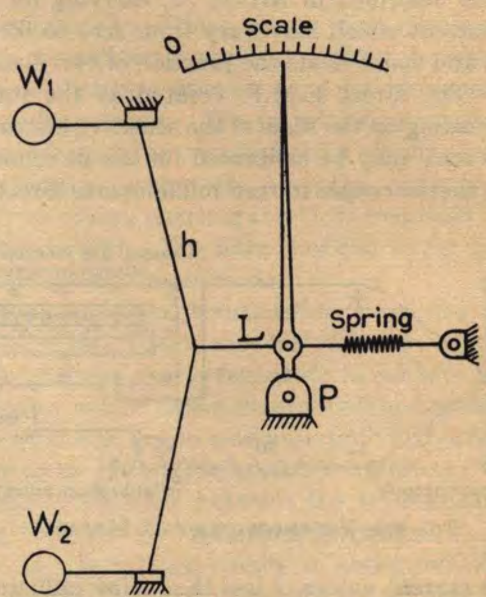


FIG. 413. HOT WIRE TYPE A.C. METER

the effective value of the current flowing through the wire. To take care of changes in atmospheric temperatures, the wire is so mounted on the case that the expansion between the mountings compensates for the effect of changes in temperature on the wire itself.

Although the hot wire type of instrument is independent of frequency, it has other practical limitations. It is not only sluggish in action, since time is required for the heating of the wire, but there is danger of burning out the instrument because any appreciable overload will produce a temperature great enough to melt the wire.

Neither the dynamometer nor the hot wire type of instrument is suitable for measuring extremely small alternating-current quantities such as are often encountered in communication work. The actual voice current, when transmitted over a telephone circuit, may vary from less than 10 milliamperes at the talking station to $\frac{1}{10}$th of one milliampere at the receiving station. For the high degree of sensitivity that is required for such measurements, amplifying and rectifying devices are often used in connection with direct-current meters in preference to the types of instruments we have discussed in the foregoing.

198. The Use of Rectifying and Amplifying Devices in Connection With Measuring Instruments

The simplest method of measuring high-frequency alternating currents, employing a direct-current meter, is to use a thermocouple connected to a sensitive millivoltmeter or microammeter. Such an arrangement is used in various transmission measuring sets and is illustrated in Figure 414. Here we have a thermocouple, as described in Article 52, carrying an alternating current which may vary from zero to 60 milliamperes and which heats the junction of two dissimilar metals. The direct E.M.F. created at the junction gives a reading on the scale of the sensitive instrument, and this scale may be calibrated for use in connection with the thermocouple to read milliamperes direct.

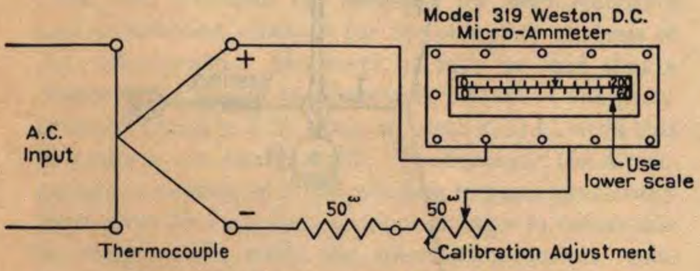


Fig. 414. Thermocouple A.C. Meter

Where current values of less than a few milliamperes are to be recorded, the same model or a similar but more sensitive model of instrument may be used in connection with an amplifying and rectifying circuit. Such an arrangement is commonly employed in transmission measuring circuits. In this case only the sensitive D.C. meter is used in the output circuit of the rectifying device, no thermocouple being necessary. The amplifying circuit, of course, amplifies the weak alternating voltages impressed on its input while the rectifying circuit converts these amplified voltages into direct current to be measured on the D.C. meter. To take care of the possibility that the output current of the rectifier may be greater than the maximum current the meter will carry, there is usually a shunt or adjusting potentiometer associated with a meter circuit of this type. This also increases the overall range of the measuring device.

In telephone work there are many other testing circuits which employ some form of rectification, or both rectification and amplification, in connection with direct-current instruments. Such circuits are employed in connection with the telephone interference factor meter, the 3-B, 4-B, and 6-A transmission measuring sets, the telephone repeater gain measuring set, the impedance unbalance measuring set, etc. The principle of operation is the same in all of these measuring sets, but the amplifying and rectifying circuits as well as the type of direct-current instrument used may vary in detail.

199. Transmission Measurements

Although the apparatus used in making transmission measurements is somewhat complex, this is due entirely to the difficulties inherent in the measuring of high-frequency currents of low value, and not in any way to the theory involved. The most obvious way of determining the attenuation of a circuit is to measure the currents sent and received at the two terminals, and thus learn the value of the current ratio, from which the attenuation may be determined. However, this method does not lend itself to rapid work and for routine measurements, transmission measuring sets reading directly in decibels are much more suitable. These sets are arranged either to make measurements directly, or to compare the loss of the circuit under test with a known calibrated loss. The former types are considered in the next article. The general arrangement of the comparison types for making transmission measurements is illustrated in Figure 415. The set is first calibrated by connecting a voltage to a fixed artificial line which causes a definite known loss. The entering current, after passing through this line, is amplified and rectified and passes through a potentiometer to a D.C. meter. The value of the applied voltage is then adjusted to such a value as to give any desired deflection of the D.C. meter, usually mid-scale. After calibrating, connections are changed so that the same voltage is applied to a variable artificial line in series with the circuit whose equivalent is to be determined. By cutting out sections of the artificial line the total loss in the circuit is made the same as that in the calibrating circuit, so that the D.C. meter gives the same deflection in both cases. The dials are arranged to read the loss in the "unknown" circuit directly. This is accomplished by constructing the variable artificial line to cause a maximum loss exactly equal to

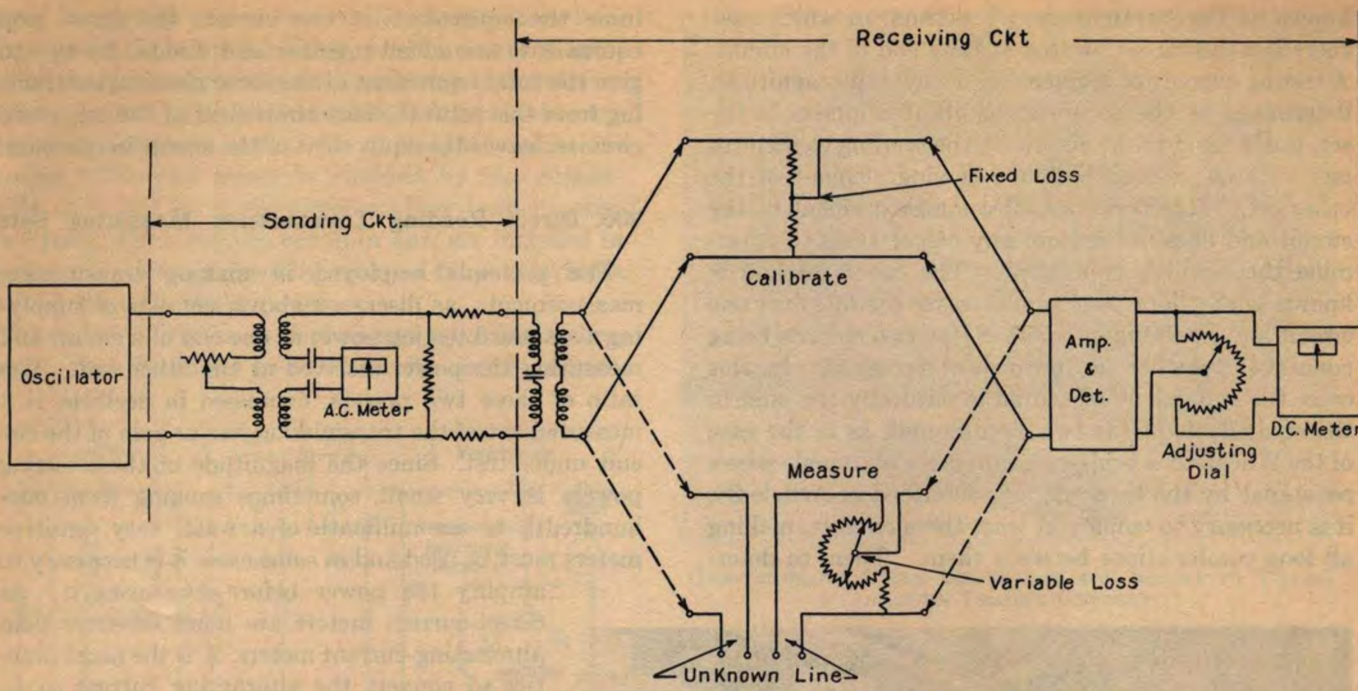


FIG. 415. PRINCIPLE OF TRANSMISSION MEASURING SET

the loss caused by the fixed artificial line. When the measuring dials are set at zero, the variable line will cause the maximum loss; with the dials set at 10 db, the total loss caused by the variable line will be the maximum minus 10, and so on.

Considering the component parts of the complete circuit, we have a generator or sending element, a load or receiving element, and a line connecting the two, which may be the calibrating standard or the circuit under test, depending upon the position of the switching key. In order that any reflection losses on the connection may be reduced, there are associated with the sending and receiving elements variable networks which may be adjusted to have impedance values closely approximating those of the circuits to which the elements are connected. Also, in order that line noises caused by inductive effects may not appreciably impair the accuracy of the readings, a filter which stops all currents except those between 850 and 1150 cycles is connected in the receiving element. It is obvious that noise will cause an error inasmuch as it increases the received energy and the ratio of energy received to energy sent will, therefore, be incorrect.

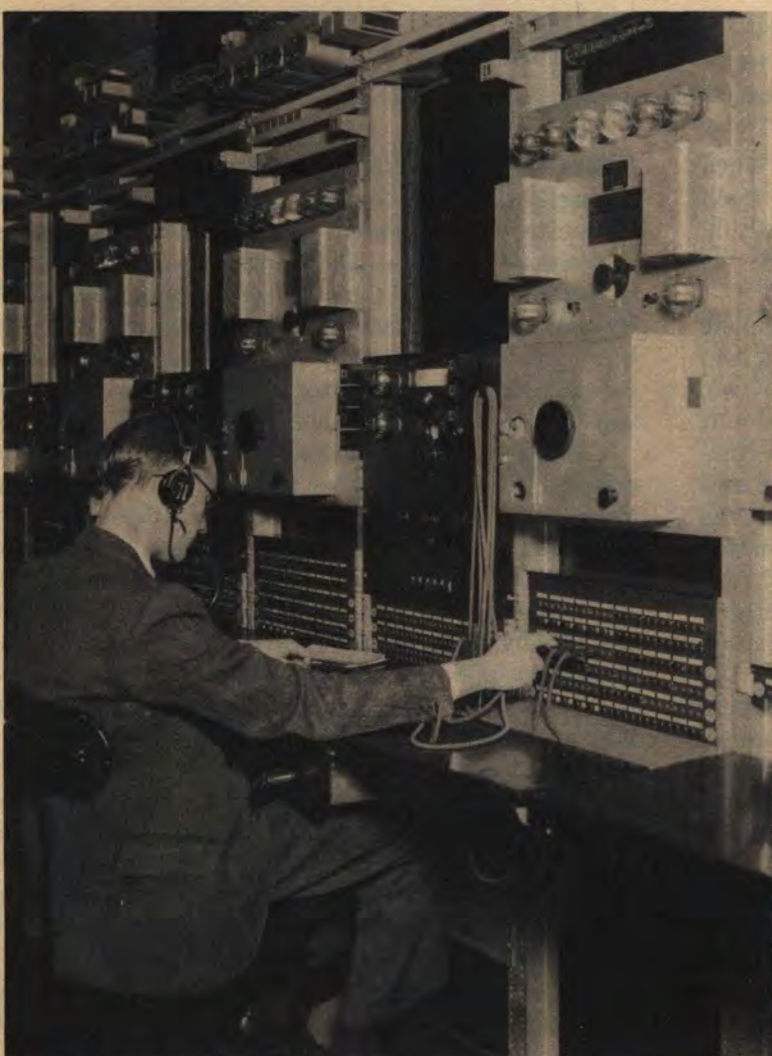
Since all artificial lines, networks, etc. in this apparatus are made up of resistances, it is possible to make measurements at any desired frequency in the voice band. A study of the equivalents of the majority of circuits measured at a single frequency and measured with a band of voice frequencies for comparison, has led to the standardization of 1000 cycles for tests of this nature. This particular single frequency gives for the various conditions, results most closely approximating the talking equivalents. This, however, does not mean that measurements are not sometimes made at other frequencies. The transmission measuring set is often employed in the study of a circuit's "quality" by making measurements over the entire voice-frequency band and charting the "loss-frequency" curves (see Article 139). The filter is of course cut out when such measurements are made.

Certain models of transmission measuring sets are adapted to permanent office installations. The testing circuits of these models terminate in cords or jacks for the sending and receiving connections, and are equipped with a reversing key to send and receive in either direction without interchanging the connections. Trunks are usually installed between the transmission measuring set and the testboard, switchboard, repeater bays, etc. A talking circuit is also provided at the measuring set, which may be used to talk over a circuit connection to the trunks or over a call circuit to a test operator in the toll operating room. Portable transmission measuring sets are also available which can be conveniently used for measuring the transmission equivalents of cord circuits, composite sets, switching trunks, and miscellaneous equipment units that affect the transmission features of a talking connection. These do not have the range of the permanently mounted sets but their principle of operation is not widely different.

In making measurements with the types of sets discussed above two methods are employed. One is known as the "straight-away" method, in which case there is a testing set located at each end of the circuit. A testing current of a given frequency and magnitude, determined by the accurately calibrated meters in the set, is delivered to the circuit by the sending element of one set and received by the receiving element of the other set. This permits a direct measurement of the circuit and does not require any calculations to determine the circuit's equivalent. The other method is known as the "loop test" and involves sending over one circuit and receiving on another, the two circuits being connected together at the distant terminal. In this case, the equivalent measured is naturally the sum of the equivalents of the two circuits and, as in the case of the Wheatstone bridge measurement of a single wire's resistance by the loop method, described in Article 38, it is necessary to employ at least three circuits, making all loop combinations between them. Then, to determine the equivalent of one circuit, the three loop equivalents are added together and divided by two to give the total equivalent of the three circuits; subtracting from this total the loop equivalent of the other two circuits, leaves the equivalent of the circuit in question.

200. Direct Reading Transmission Measuring Sets

The principle employed in making transmission measurements, as discussed above, consists of supplying a standard testing power at one end of a circuit and measuring the power received at the other end. The ratio of these two powers, expressed in decibels, is a measurement of the transmission loss or gain of the circuit under test. Since the magnitude of these testing powers is very small, sometimes ranging from one-hundredth to one-millionth of a watt, very sensitive meters must be used and in some cases it is necessary to amplify the power before measuring it. As direct-current meters are more sensitive than alternating-current meters, it is the usual practice to convert the alternating current to direct current by means of vacuum tube or copper-oxide rectifiers and then measure with direct-current meters.



Transmission Testboard Showing Variable Oscillators and Transmission Measuring Sets

In accordance with these principles, the simplest transmission measuring arrangement would consist in connecting a known source of power to one end of the unknown circuit and a direct reading meter, calibrated in decibels, to the other end of the circuit. Until quite recently, such a simple arrangement has not been feasible, either because sufficiently sensitive meters were not available or because the amplifiers used for increasing the currents to the required values were not sufficiently stable. It has been necessary accordingly to employ the comparison method, described in the preceding article. This method obviates errors due to variations in the amplifier circuit because it consists in making two successive measurements, which merely compare the unknown circuit with a known loss, and does not measure absolute values.

The development of highly stable negative feedback amplifiers, copper-oxide rectifiers, and more sensitive direct-current meters has made possible direct reading measuring systems. Meters employing some of the newer alloy steels are now available, which will measure transmission losses of around 20 db below one milliwatt without an amplifier. Figure 416 is a simplified circuit of such a measuring set that accurately measures losses up to 20 db. In using this set, one milliwatt of power at 1000 cycles is applied at one end of the circuit under test. This power is supplied by a small magneto generator designed especially for furnishing transmission testing power. The transmission testing set is connected at the other end of the circuit, where the incoming 1000-cycle power is rectified by the copper-oxide units and led to the direct reading decibel meter. Two pads, which may be cut in or out, are included in the input circuit to extend the range of the set to values beyond the scale of the meter.

Where transmission measurements involving values greater than 20 db are to be made, the simple copper-

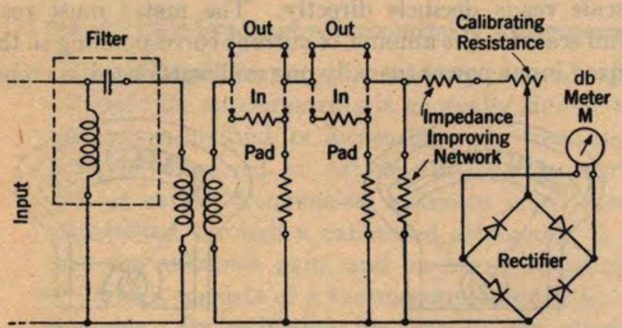
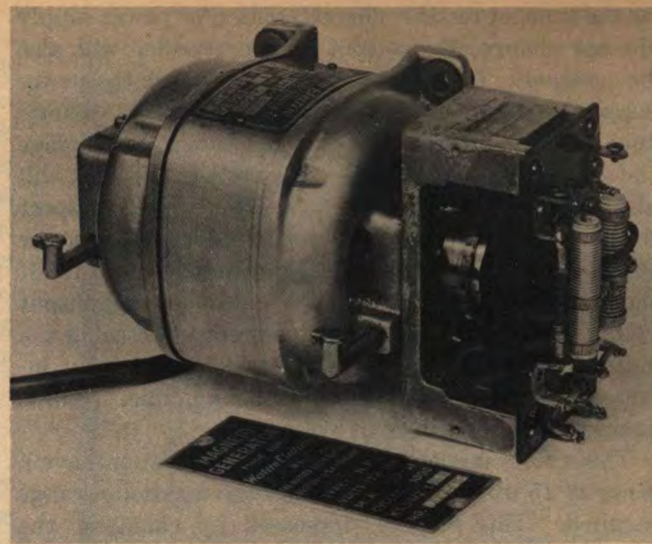


Fig. 416. Direct Reading Transmission Measuring Set

oxide rectifier and meter circuit illustrated in Figure 416 is inadequate. To meet this condition, a negative feedback amplifier is included in the measuring set. As we know, such an amplifier is very stable and may be operated over long periods without adjustments. The amplifier is shown in Figure 417. It consists of a high impedance input transformer, *T*, bridged across a 600-ohm terminating resistance (which may be removed when level measurements are made), two pentode tubes, a copper-oxide rectifier, *R*, and a meter, *M*, calibrated directly in decibels. The negative feed-



One-thousand Cycle Generator for Supplying Transmission Testing Current

back voltage is introduced into the control grid circuit of the first tube through resistances *A*, *B*, and *C* by connecting one of the rectifier terminals to contact *P* of the potentiometer. These resistances, together with resistance *D*, form the cathode drop resistance of the control grid circuit, and any potential applied across them affects the potential on the control grid of the first tube. With the potentiometer, *P*, contact as shown in Figure 417, the negative feedback is a maximum and the net overall amplification of the amplifier is therefore a minimum. Changing the potentiometer contact to a lower step gives less feedback and results in a greater net gain of the amplifier.

When a constant potential is applied to the input terminals of the amplifier, a constant voltage is also applied to the control grid of the first tube. As long

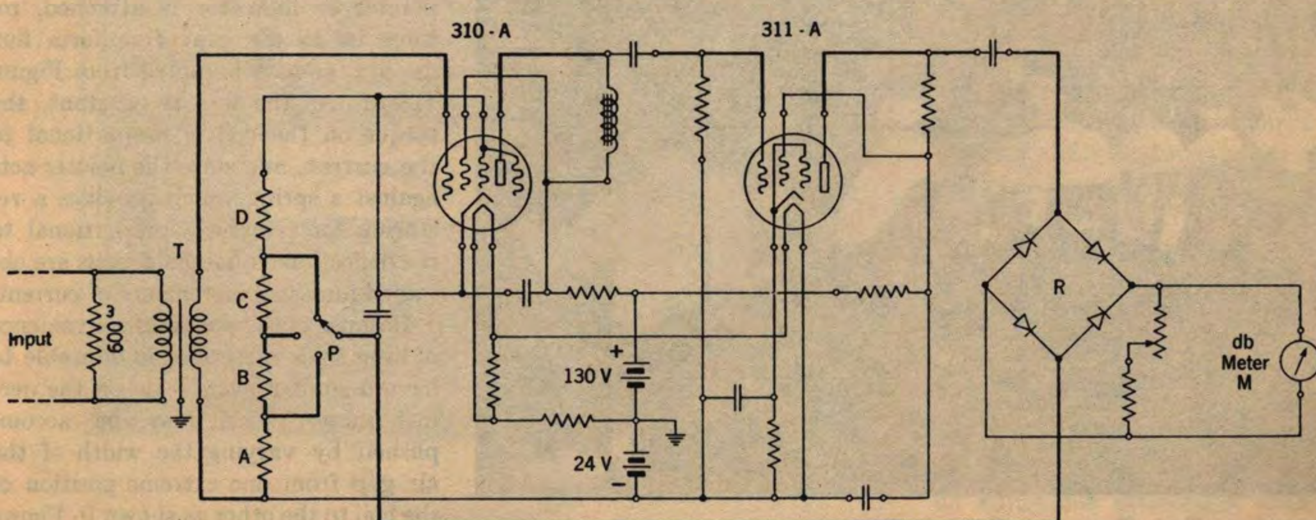


Fig. 417. Direct Reading Transmission Measuring Set with Amplifier

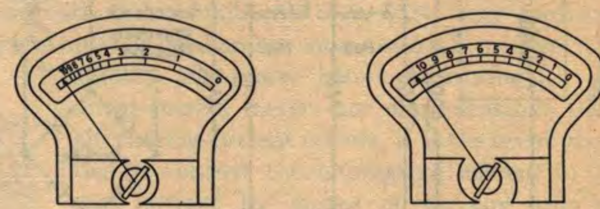
as the tube, or rectifier characteristics, or power supply do not change, the output of the rectifier will also be constant. If, however, due to any cause the power supply changes, or the characteristics of a vacuum tube or the rectifier changes in such a way as to change the output to a higher value, the voltage fed back will also be increased. Because it is negative with respect to the input, this reduces the input sufficiently to lower the net output voltage to its proper value again. If the change were in the reverse direction, the output voltage would first be lowered thereby decreasing the voltage fed back to the input, which effectively increases the input voltage and the amplifier returns again to its proper output voltage.

The meters used with this measuring system have a range of 15 db, which is less than the maximum range required. This range is increased by changing the amplifier gain in steps of 10 db. The resistances *A*, *B* and *C* in the feedback path are used for this purpose. In practice the contact arm of the potentiometer, *P*, may be controlled by relays which in turn are remotely controlled by keys, jacks, or dials at various points in the office. The db meters may be located where desired, without reference to the amplifier-rectifier. They may also be placed in lantern slide projectors to throw a greatly enlarged meter scale on a screen, which can be read from a distance of 50 feet or more. These arrangements are extremely flexible since a single amplifier-rectifier and one or more associated decibel meters may be used to make measurements from any one of several points in an office.

The amplifier-rectifier circuit is designed to give a constant gain through a wide range of frequencies. It may, therefore, be used, with an appropriate variable oscillator at the sending end, to measure the loss-frequency characteristics of any circuit through the voice-frequency range. It may also be used in connection with a recording device to measure the variations in the loss of a circuit over a period of time.

201. Decibel Meters

The decibel meter, associated with the amplifier-rectifier circuit discussed in the preceding article, is merely a special D.C. ammeter designed so that its scale reads decibels directly. The meter must read full scale for the amount of current corresponding to the fixed input power (usually one milliwatt) used in trans-



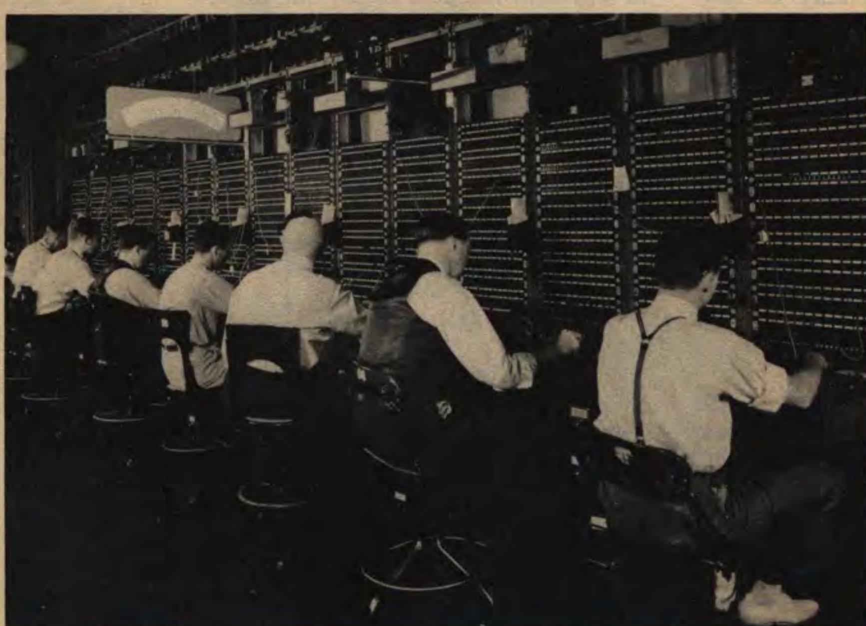
A B

Fig. 418. Decibel Meter

mission measuring, and this full scale deflection of the meter is marked "zero" loss. If the meter is of the usual ammeter design, the points on the scale for subsequent units of db loss are not evenly spaced, but crowded together as illustrated in Figure 418-A, because the decibel is a logarithmic unit.

Of course, the D.C. ammeter is ordinarily designed to have equal spacing on its scale for equal increments of current. This is a result of the fact that the moving coil, to which the pointer or indicator is attached, rotates in an air gap of uniform flux density, as may be noted from Figure 418-A. As the flux is constant, the torque on the coil is proportional to the current, and since the pointer acts against a spring which provides a restoring force directly proportional to the deflection, equal deflections are obtained for equal increments of current.

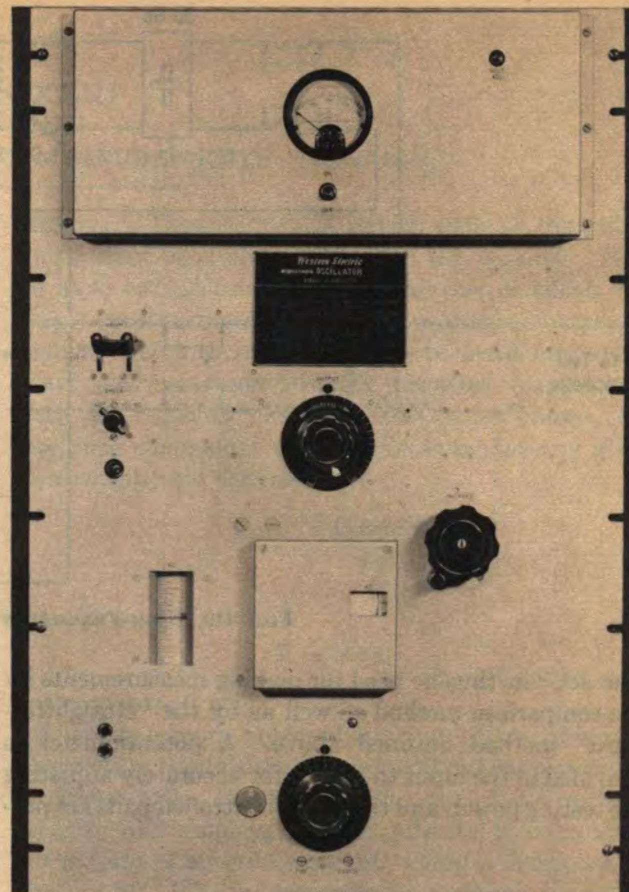
Because of the practical convenience of even scale meters, it is desirable to have a similar even scale on the decibel meter. This may be accomplished by varying the width of the air gap from one extreme position of the coil to the other as shown in Figure 418-B, so that the flux instead of being



Secondary Toll Testboard Showing Projection Meter for Transmission Measurements the same at all positions of the moving coil, increases logarithmically toward the position of greater loss. With this arrangement, the position for maximum current on the right-hand end of the scale is still marked zero db loss. But in order that the movement of the pointer along the scale shall be the same for each decibel, the air gap is so proportioned as to make the flux correspondingly larger toward the larger loss end, thus offsetting the decreasing increments of current. However, a meter of this type requires a higher operating current than the ordinary ammeter because the flux is less at maximum current.

202. Carrier-Frequency Transmission Measurements

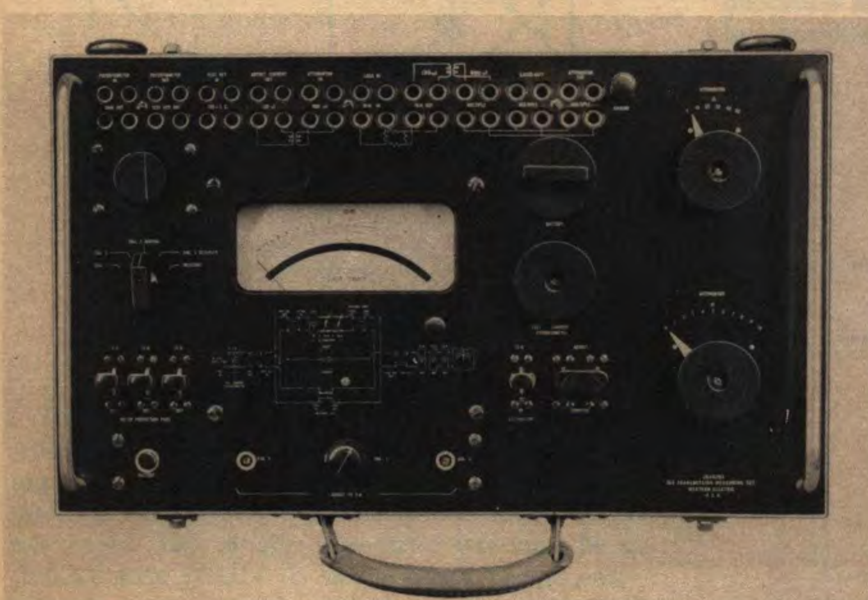
The principle employed in making carrier-frequency transmission measurements is generally similar to the comparison method for measuring voice-frequency circuits, as discussed in Article 199. In measuring a gain at carrier frequencies, a known input power is transmitted through a calibrated attenuator in series with the unknown gain, and on into a metering circuit which consists of a thermocouple and D.C. milliammeter. By adjusting the loss in the attenuator until the output of the unknown gain device is equal to the input power at the attenuator, the loss in the attenuator is made equal to the unknown gain. In the case of measuring a loss, the same basic principle is employed. The known input power is now transmitted through a calibrated attenuator, the unknown loss, and a fixed known gain. Then when the attenuator is adjusted so that the output of the known gain device is equal to the input power, the unknown loss is equal to the difference between the known gain and the calibrated loss.



HIGH-FREQUENCY OSCILLATOR (50 TO 150,000 CYCLES) FOR BROAD-BAND CARRIER TESTING WORK

A transmission measuring set using this principle was developed primarily for making carrier-frequency measurements on the Types-C, -J, and -K carrier telephone systems. It consists essentially of a thermocouple, milliammeter, attenuators, keys, jacks and repeating coils, arranged in such a manner as to facilitate the various types of tests which are required. Because of the wide range of frequencies involved heterodyne oscillators, capable of furnishing frequencies in the carrier range, are required for making measurements.

A simplified schematic of this measuring set is shown in Figure 419. Two paths, which are controlled by a switching key, are provided between the input to the set and the thermocouple and meter. One path contains an adjustable attenuator while both paths contain suitable jacks for inserting losses or gains to be measured.



TRANSMISSION MEASURING SET FOR TESTING BROAD-BAND CARRIER CIRCUITS

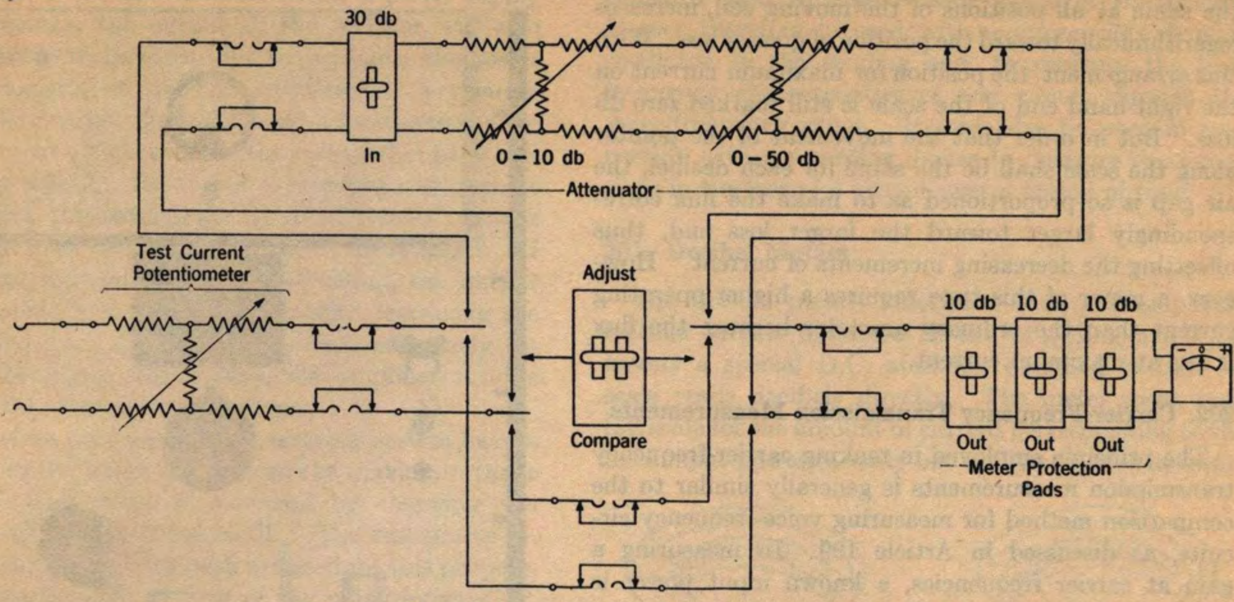


FIG. 419. HIGH-FREQUENCY TRANSMISSION MEASURING SET

The set can thus be used for making measurements by the comparison method as well as by the "straightforward" method outlined above. A potentiometer is provided in the input to the set for accurately adjusting the testing power, and three key-controlled pads are provided ahead of the meter for protection against burning out the sensitive thermocouple. The thermocouple and milliammeter are arranged to read with reference to a test power of one milliwatt. The set is calibrated on direct current from a dry cell battery contained in the set.

CHAPTER XXXIII

ALTERNATING-CURRENT TESTS AND MEASUREMENTS—(*Continued*)

203. The Impedance Bridge

In Chapter V we first encountered the Wheatstone bridge and learned the theory of "balance" which permits us to measure the D.C. resistance of any circuit. Later we saw that the principle of balance can also be applied to alternating currents and in Article 117 we discussed briefly an A.C. Wheatstone bridge circuit. In telephone work it is frequently necessary to make A.C. bridge measurements, and to this end we utilize the "impedance bridge" illustrated in Figure 420. It can be seen that this is merely a modification of the Wheatstone bridge, with an alternating source of E.M.F. substituted for the battery, a telephone receiver (where the frequencies are within the voice-current range) substituted for the galvanometer, and a variable impedance consisting of resistance and inductive or capacitive reactance, substituted for the adjustable resistance arm of the Wheatstone bridge.

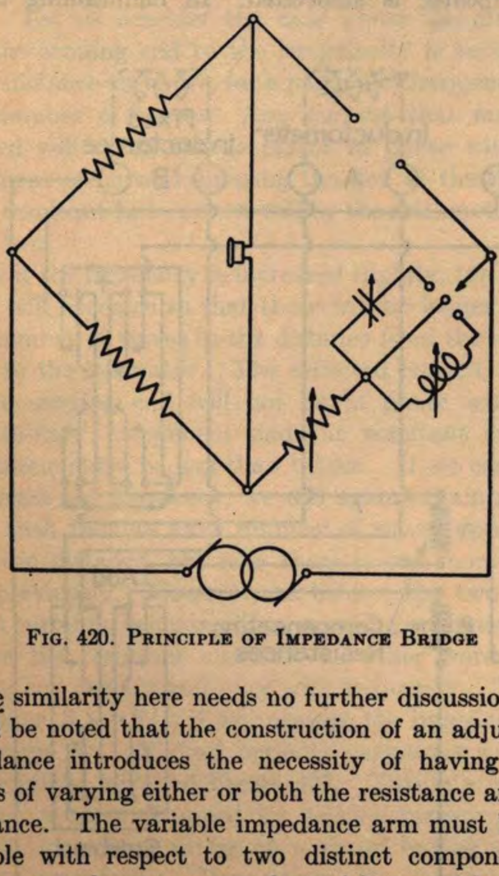


FIG. 420. PRINCIPLE OF IMPEDANCE BRIDGE

The similarity here needs no further discussion, but it will be noted that the construction of an adjustable impedance introduces the necessity of having some means of varying either or both the resistance and the reactance. The variable impedance arm must be adjustable with respect to two distinct components—resistance and reactance, and provision must be made to make the balancing reactance either positive or negative. Theoretically, a device such as that illustrated may be used in obtaining the balance. Here we have two distinct adjustments, one of which is a simple resistance and the other of which is an arrangement by which the reactance may be made any value, either positive (inductive) or negative (capacitive). With the capacity or inductance values known, the reactance component can be calculated for any given adjustment from the formula—

$$X = -\frac{1{,}000{,}000}{2\pi fC}$$

or

$$X = 2\pi fL$$

as the case may be.

There is a more practicable arrangement, however, which will accomplish the same result as that illustrated by Figure 420. It eliminates the use of the variable capacity and employs only an adjustable inductance in the form of an inductometer. This is illustrated in Figure 421. The inductometer is a device consisting of two similar coils, one of which may be rotated with respect to the other. Such an arrangement permits of a variation of inductance from almost zero, when the magnetic fields of the two coils oppose each other, to a maximum when the fields aid each other and the inductances add directly. The condenser is eliminated by introducing a transfer switch by which the in-

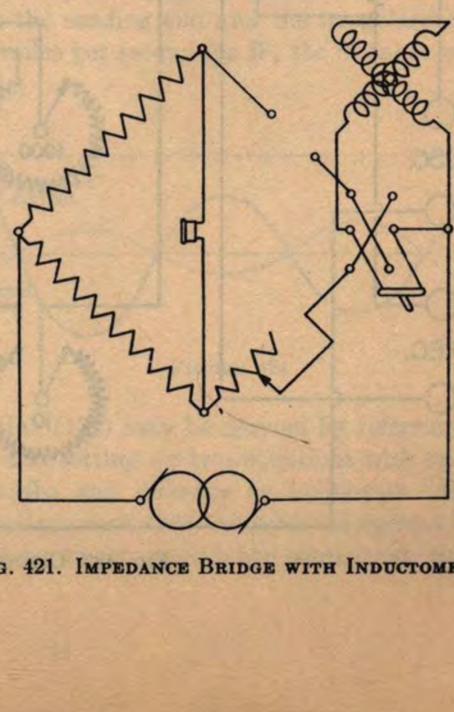


FIG. 421. IMPEDANCE BRIDGE WITH INDUCTOMETER

ductometer may be connected in series with the unknown impedance so as to neutralize a negative reactance component rather than balance it. Inasmuch as the inductometer is calibrated in terms of millihenrys or henrys, as the case may be, the reactance component must be calculated by the use of the formula—

$$X = 2\pi fL$$

This component will be negative when the switch is so thrown that the inductometer is in series with the unknown impedance, and positive when the switch is so thrown that the inductometer is in series with the adjustable resistance arm.

In the actual construction of an impedance bridge it is of course not possible to employ an inductometer having zero resistance. It is necessary, therefore, to have a compensating resistance which is similarly connected to switch contacts so as to always be in the opposite arm to the one in which the inductometer is connected. Figure 422 is a diagram of connections of the 1-B Impedance Bridge used in the telephone plant for measuring line impedances. Due to the fact that it is not possible to secure absolute zero inductance with the inductometer, a second fixed inductance is used, which may be switched from one arm to the other. Zero inductance can then be secured by throwing the fixed inductance on one side of the bridge and the inductometer on the opposite side, so that the fixed one neutralizes its value on the scale of the other. Since either inductance unit may be switched to either side of the bridge, it can be seen that the reactance values that can be measured range from zero to a value of $\pm 2\pi f (L_a + L_b)$, where L_a and L_b denote the values of the two inductance units. Like the inductometer, the fixed inductance also has a compensating resistance which is so wired to the transfer switch that it will always be in the proper arm of the bridge to balance the resistance of the coil.

A vacuum tube oscillator is used in connection with the bridge as the A.C. supply, and to permit measurements over the entire range of voice frequencies the oscillator has an adjustable resonant circuit which allows any desired frequency value within this range to be obtained. It is also equipped with a filter for eliminating the harmonics of the particular frequency used, thereby affording a sine wave testing current.

204. Line Impedance Measurements

In Chapter XXVIII we learned that the successful operation of 22-type telephone repeaters necessitates the use of a balancing network for each line with which the repeater is associated. In maintaining the re-

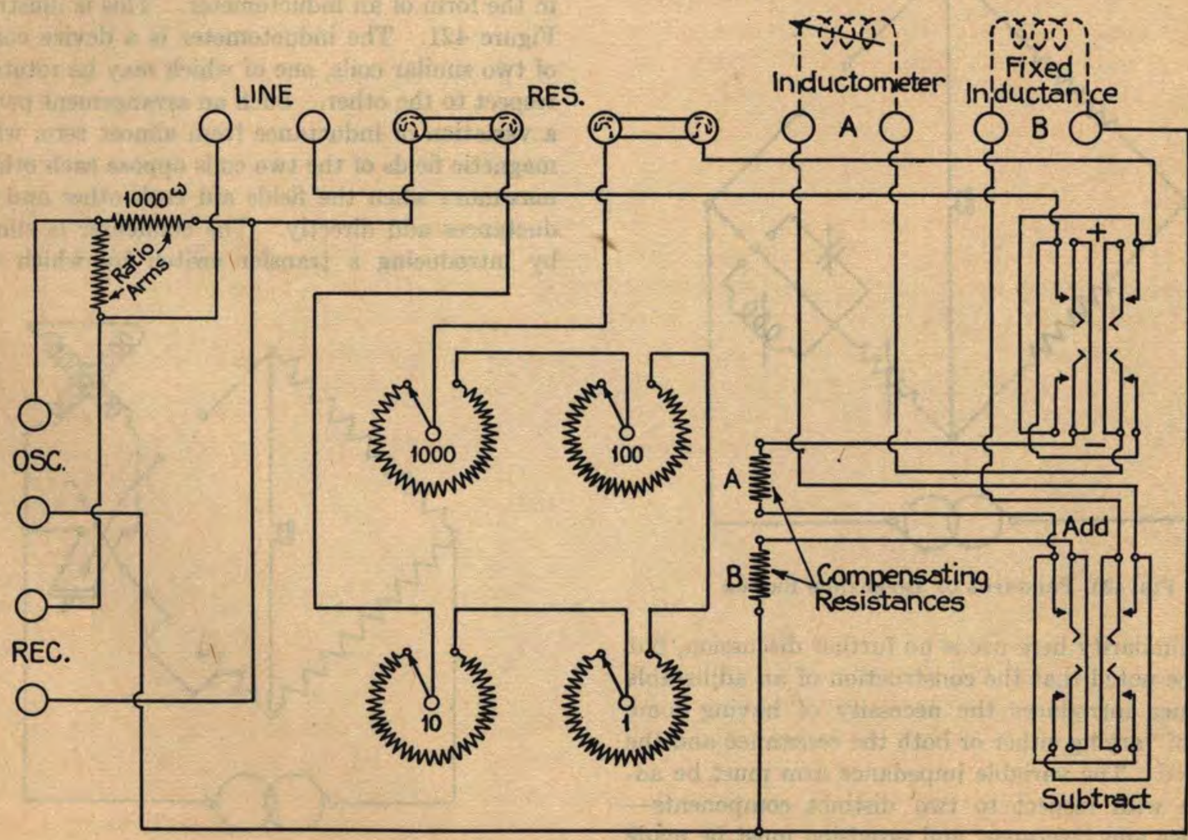


FIG. 422. CIRCUIT OF NO. 1-B LINE IMPEDANCE BRIDGE

quired degree of balance between the network and the line, measurements throughout the voice range of frequencies are sometimes required in order to check the impedance values and to locate any irregularities in the line circuit that may seriously affect the balance. Here an impedance irregularity means any condition throughout the length of the line that may cause a partial reflection of the electrical wave. From our earlier considerations of wave propagation we know that any wave will suffer both attenuation and phase displacement as it travels along a line, and this is, of course, also true of the reflected wave. Consequently, the magnitude of this wave and its phase position relative to the wave just leaving the generator, will change as it moves from the point of reflection to the sending end. Further, a little consideration will show that the phase of the reflected wave when it reached the generator will depend on the time it takes to travel from the irregularity to the generator, or what amounts to the same thing, on the distance from the irregularity to the generator.

For a clearer analysis of the effect of an irregularity on a line, let us resolve the current entering the line into two components, one the current that would enter the circuit if there were no irregularity, and the other the reflected current that is present due to the irregularity. Let us consider the case where the distance from the sending end to the irregularity is such that in this distance there are, for a particular frequency, an even number of waves. Any current that may be reflected will reach the generator in phase with the component of current entering the line at that point, so the resultant line current will be the arithmetic sum of the two.

Now if the frequency is increased slightly, the wavelength will decrease so that there will no longer be an even number of waves in the distance from the irregularity to the generator. The reflected current reaching the sending end will not be in phase with the "smooth-line" component and the resultant current will consequently be less than before. If we continue to increase the frequency, we will again obtain a condition such that an even number of waves appear in the given distance, but now there is one more wave than previously. However, as before, the two component currents, being in phase, combine arithmetically to give the resultant current. In other words, the sending end current will vary with frequency. Therefore, since $Z = E/I$, if we measure the impedance of the circuit, we will find periodic variations such as appear in the curve of Figure 423. This is a typical impedance curve obtained by making measurements with the impedance bridge on a circuit having an impedance irregularity. Here points, M, N and O correspond to frequencies where the distance to the irregularity is such that the reflected current is exactly opposite in phase to the outgoing current at the sending end, and points P, Q and R to frequencies where the reflected current is exactly in phase with the outgoing current at the sending end.

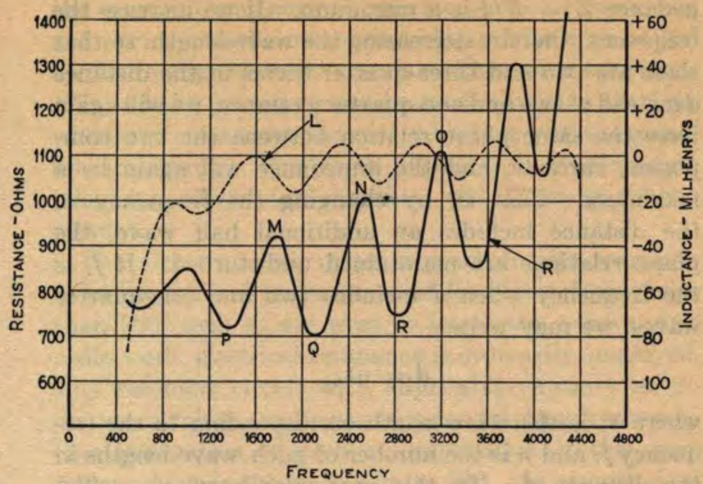


Fig. 423. Impedance Characteristic of Circuit with Single Irregularity

Such a curve permits a determination of the distance from the sending end to the irregularity by the use of the following formula:

$$d = \frac{W}{2(f_2 - f_1)} \tag{176}$$

where d is distance to the irregularity, W is the average velocity of propagation of the telephone currents over the particular line in question, which may be taken from Tables XII and XIII, and $f_2 - f_1$ is the average frequency interval between the adjacent "humps" of the impedance curve. If, in the case of a loaded circuit, we employ the number of "loads per second" for the term W, d will give the number of loading points between the sending end and the irregularity. If we employ miles per second for W, the distance will be in miles, etc.

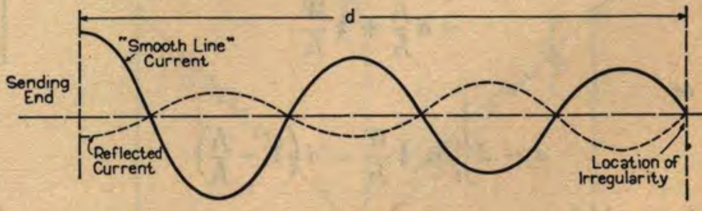


Figure 424

Equation (176) may be derived by referring to Figure 424 and setting up two equations with number of wave-lengths and distance as unknowns. Here the conditions are such that the reflected current is opposite in phase to the smooth-line component, so the impedance $Z = E/I$ is a maximum. If we increase the frequency, thereby decreasing the wave-length, so that there are two and three-quarter waves in the distance d instead of two and one-quarter as shown, we will again have the same phase relation between the two component currents, and the impedance will again be a maximum. That is, by changing the frequency so the distance includes an additional half wave, the phase relations are maintained undisturbed. If f_1 is the frequency when d includes two and one-quarter waves, we may write—

$$d = n\lambda_1$$

where λ_1 is the wave-length corresponding to the frequency f_1 and n is the number of such wave-lengths in the distance d. (In this particular case, $n = 2\frac{1}{4}$.) For the second case, where d includes an additional half wave-length, we may write—

$$d = (n + \tfrac{1}{2})\,\lambda_2$$

But

$$\lambda_1 = \frac{W}{f_1} \qquad \text{and} \qquad \lambda_2 = \frac{W}{f_2} \qquad \text{(from Equation 95).}$$

Substituting these values of λ_1 and λ_2 in the above equations, we have—

$$d = n\frac{W}{f_1}$$

$$d = (n + \tfrac{1}{2})\frac{W}{f_2}$$

Here we have two equations with two unknowns, namely d and n, and since we are not interested in n we will eliminate it; thus—

$$n = d\frac{f_1}{W}$$

Whence, we have—

$$d = \left(d\frac{f_1}{W} + \tfrac{1}{2}\right)\frac{W}{f_2}$$

$$= d\frac{f_1}{f_2} + \tfrac{1}{2}\frac{W}{f_2}$$

or

$$d - d\frac{f_1}{f_2} = \tfrac{1}{2}\frac{W}{f_2} = d\left(1 - \frac{f_1}{f_2}\right)$$

from which

$$d = \frac{W}{2f_2\left[1 - \frac{f_1}{f_2}\right]} = \frac{W}{2(f_2 - f_1)} \qquad (176)$$

To illustrate the use of Equation (176) let us consider a typical example.

Example: Calculate the distance to the irregularity on the 19-gage H-44 cable circuit whose impedance curve is shown in Figure 423.

Solution: On the R curve, we have "peaks" at 1000 and 3200 so that $(f_2 - f_1) = \frac{1}{3}(3200 - 1000) = \frac{1}{3} \times 2200 = 733$. Similarly from the L curve, $(f_2 - f_1) = \frac{1}{3}(3040 - 860) = 727$. Using the average value of $(f_2 - f_1)$ we have—

$$d = \frac{W}{2(f_2 - f_1)} = \frac{17640}{2 \times 730} = 12.1 \text{ loads, ans.}$$

As a matter of fact, the fault was located at the twelfth loading point.

It will be noted that the fault location secured in the above example is rather approximate. In modern practice considerably more accurate locations are desirable since clearing the trouble usually involves opening a cable splice. The standard instructions, covering fault location by means of impedance measurements, describe several refinements which may be applied to the basic technique described above in order to secure greater accuracy.

205. The Capacity Bridge

From the theory of the impedance bridge, it can be seen that it would be quite possible to measure capacity values with this device. However, since condensers are practically pure negative reactances it is more convenient to eliminate the variable resistance, thus simplifying the manipulation. Figure 425 shows the

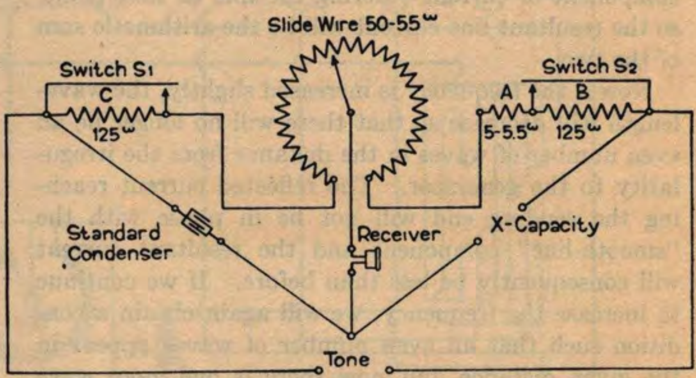


Fig. 425. Principle of the Capacity Bridge

impedance bridge when modified to measure capacities, in which form it is known as a "capacity bridge". Here, instead of having fixed ratio arms, we have variable arms and obtain balance by adjusting them until no sound is heard in the receiver. In this capacity bridge, the scale reading gives the ratio of the resistance arms, so the value of the unknown capacity is—

$$C_x = \text{Scale Reading} \times C_s$$

where C_s is the capacity value of the standard condenser.

206. Frequency Meters

As mentioned earlier, in alternating-current work we need to know the value of the frequency. There are several methods of determining this, the one employed in each case depending entirely upon the conditions. For frequency values lower than about 200 cycles, a simple vibrating reed device will give direct readings

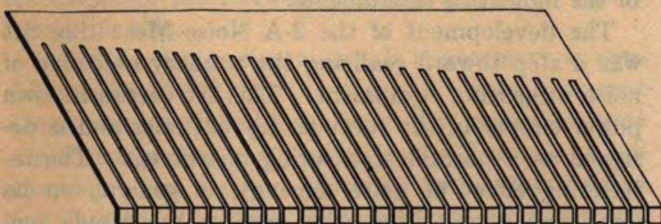


FIGURE 426

within a reasonable degree of accuracy. The theory of this instrument is illustrated by Figure 426. Here we have a comb shaped arrangement of metal reeds of varying lengths, each having a different natural period of vibration. A coil winding is connected to the circuit for which the frequency is to be determined and the core of this coil has long narrow pole pieces of the same length as the group of vibrating reeds. The reeds are

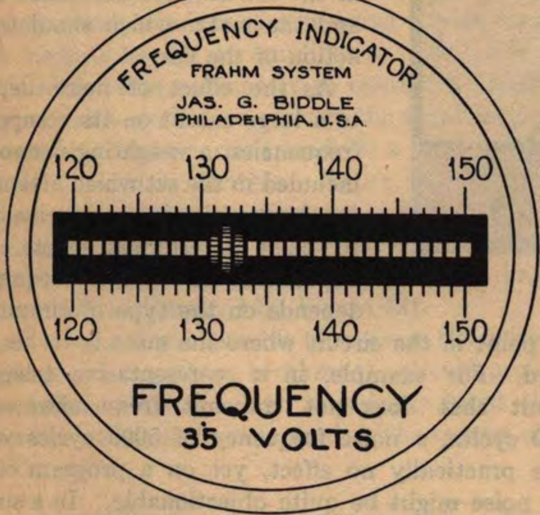


FIG. 427. VIBRATING REED FREQUENCY METER

inserted in the air gap between the pole pieces. There is a magnetic attraction, therefore, for each reed due to the alternating current in the coil, in the same way that there is magnetic attraction for the diaphragm of a telephone receiver. The particular reed which has a mechanical period of vibration corresponding to the current frequency will vibrate but other reeds will not respond to the magnetic pulses. The ends of the reeds are aligned under a scale as illustrated by Figure 427, and the frequency can be read by noting the long white line created by the reed vibrating with the greatest amplitude. This form of meter is employed in telephone offices for checking the 135-cycle ringing current.

The vibrating reed meter depends for its operation upon mechanical resonance, but for higher frequencies than 200, such as are used in telephone, carrier, and radio work, electrical resonance is ordinarily employed. Any resonant circuit with adjustable capacity or inductance values can be used for determining frequency. One illustration of such an application is the wave meter which is shown in Figure 428. In circuit A, if

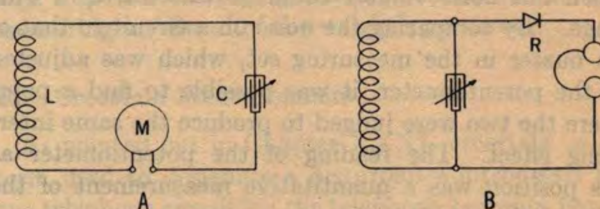


FIG. 428. WAVE METERS

an inductance L and a condenser C are so adjusted as to give resonance, the meter M will give a maximum reading. In circuit B, the resonant condition will obtain when maximum sound is heard in the telephone receivers which are connected in series with a rectifying device, R.

Instead of using the resonant circuit in the manner illustrated by Figure 428, it may be used in connection

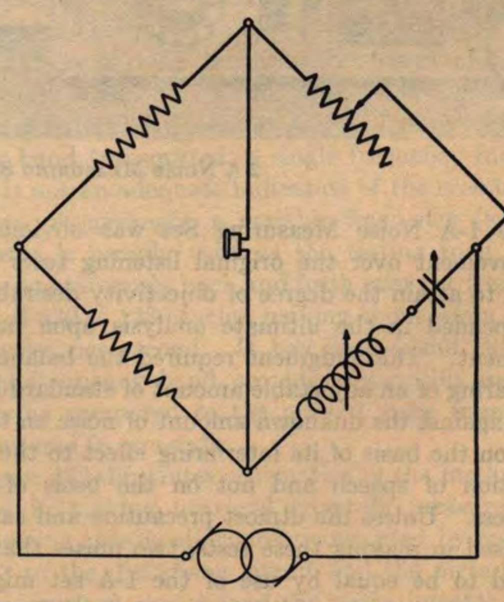
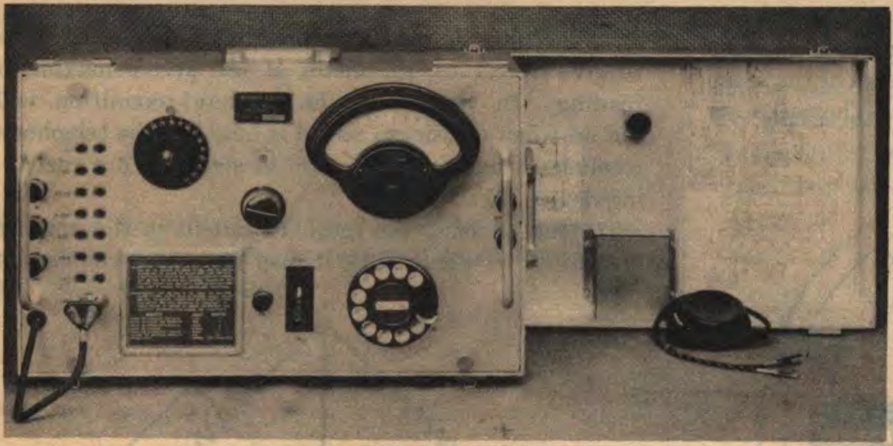


FIG. 429. IMPEDANCE BRIDGE ARRANGED FOR MEASURING FREQUENCY

with the impedance bridge, as illustrated by Figure 429. Here when the variable capacity and the variable inductance are resonant to the impressed frequency, it is possible to secure a bridge balance by means of the variable resistance arm alone. This scheme may be employed for calibrating vacuum tube oscillators.

207. Noise Measurements

From the early days of telephone service, some means of measuring the noise on telephone circuits has been in use. This at first consisted of merely listening on the circuit with a telephone receiver, which obviously resulted in a wide variety of ideas as to the amount of noise that was present. This led to the use of the 1-A Noise Measuring Set which consisted of a "buzzer" to produce a noise, and a potentiometer with which this noise volume could be varied over a wide range. By comparing the noise on a circuit to that of the buzzer in the measuring set, which was adjusted by the potentiometer, it was possible to find a point where the two were judged to produce the same interfering effect. The reading of the potentiometer at this position was a quantitative measurement of the noise on the circuit.



2-A Noise Measuring Set

The 1-A Noise Measuring Set was obviously an improvement over the original listening tests but it failed to attain the degree of objectivity desirable, for it depended in the ultimate analysis upon personal judgment. This judgment required the balancing or comparing of an adjustable amount of standard buzzer noise against the unknown amount of noise on the circuit, on the basis of its interfering effect to the interpretation of speech and not on the basis of equal loudness. Unless the utmost precaution and care was exercised in making these tests, two noises that were judged to be equal by use of the 1-A set might be found to have quite different interfering effects on a transmitted conversation.

These objections to the 1-A set were overcome to a large extent by the development of the 1-A Noise Amplifier. By using this amplifier in conjunction with a standard transmission measuring set, the ear was eliminated as the basis for determining the magnitude of the noise. This amplifier employed "frequency weighting" which was based on the information available as to the relative interfering effects of single frequency tones. While this amplifier was quite an improvement it fell short of being the ideal in other ways, such as portability and dynamic characteristics of the indicating instruments.

The development of the 2-A Noise Measuring Set was a step toward realizing the primary objective of noise measuring equipment. This set carries its own power supply in the form of dry batteries and is designed as a portable measuring instrument. The interfering effect of noise depends, in general, on its frequency composition, the relative magnitude and spacing of the various frequencies, the time duration of the noise, the type of circuit on which it occurs, and the person who is listening to it. These factors obviously complicate the design of a satisfactory noise measuring set. It is desirable to employ a meter to give the indication of noise; but before such a method can be satisfactorily used, the interfering effect of different noises on speech as heard by the ear must be determined, and then circuits must be incorporated in the set between the noise input and the meter which simulate the action of the ear.

As the effect of noise depends to a large extent on its component frequencies, a weighting network is included in the set which attenuates the frequencies in an inverse ratio to their interfering effects. The importance of various frequencies depends on the type of circuit and the point in the circuit where the noise is to be measured. For example, in a representative telephone circuit that does not transmit frequencies above 3000 cycles, a noise frequency of 5000 cycles would have practically no effect, yet on a program circuit this noise might be quite objectionable. In a similar manner, the subscriber's telephone subset which includes an induction coil and a condenser, affects the frequencies transmitted primarily in the low-frequency range. Therefore, the weighting that should be made in the measuring set would be different depending upon whether the measuring set was to be used at a subscriber's receiver or at a central office.

In the event the noise was of a single frequency, the weighting networks would insure that its measured effect would be the same as that determined by the average ear. The noise, however, usually comprises many frequencies, and the action of the ear must be considered in combining these different frequencies. The copper-oxide rectifier, which converts the amplified noise currents into direct current to operate the meter, is designed to combine the weighted noise currents in such a manner that the sum is approximately equal to the square root of the sum of the squares of all the weighted components. It has been found that the ear combines most ordinary noises in a very similar manner to this.

Following these adjustments, it is necessary to consider the effect of the time interval the noise is present on the circuit. It has been found that most telephone circuit noises must exist for at least 0.2 second to appreciably affect the ear. In order to incorporate this time factor in the noise measuring set, the meter has been given such dynamic characteristics that a noise duration of 0.2 second produces about the same deflection as a continuous noise of the same intensity. Noises of shorter duration produce proportionally smaller deflections.

The complete 2-A Noise Measuring Set includes (1) an input circuit which presents the proper impedance to the circuit to which the set is to be connected; (2) weighting networks for frequency discrimination so that for each type of measurements, the noises of different frequencies will be measured in accordance with their interfering effects; (3) a graduated potentiometer which provides a means of measuring a wide range of noise magnitudes; (4) an amplifier to raise the level of the noise currents sufficiently to operate a meter; (5) a copper-oxide rectifier to convert the amplifier output currents into direct current; and (6) a noise indicating meter calibrated in decibels. Figure 430 shows schematically this circuit arrangement. When measuring noise on ordinary types of telephone circuits at the central offices, the weighting network used has the characteristics illustrated by Figure 431.

As it is sometimes desirable to measure the noise on a telephone circuit without breaking the circuit, provision is made for bridging a high impedance input circuit across the circuit to be measured. The output of this

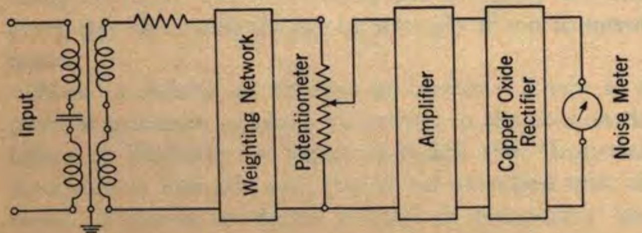


Fig. 430. Circuit of 2-A Noise Measuring Set

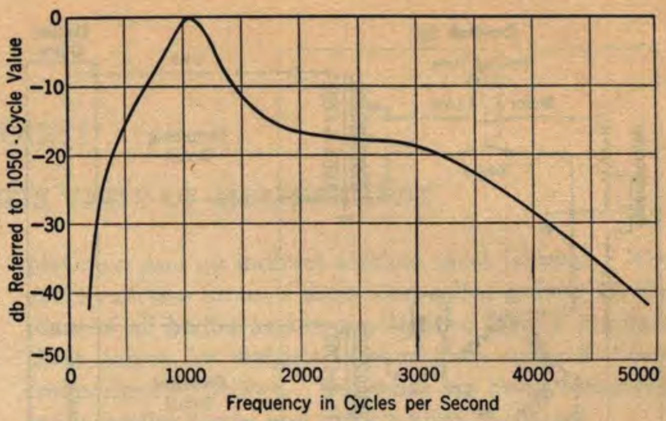


Fig. 431. Characteristic of Weighting Network of 2-A Noise Measuring Set

high impedance bridging circuit connects to the noise measuring set ahead of the weighting network.

208. Crosstalk Measurements

As pointed out in Chapter XXXI, crosstalk is the term used to designate a disturbance introduced into one telephone circuit by the telephone currents flowing in another. This disturbance may be caused by inductance, capacitance, or resistance coupling between the two circuits, or by combinations of these three general types of coupling.

The amount of crosstalk induced from one circuit into another is, of course, determined by the degree of coupling between the two circuits. A measurement of this coupling in terms of loss is therefore a measurement of the crosstalk. If the loss is low, the crosstalk is obviously large, and vice versa. While this loss could be measured with a standard transmission measuring set, using a calibrated amplifier in the receiving side if the loss was beyond the measuring range of the set, the results would only be for single frequency values. As the crosstalk may occur at practically any frequency in the band transmitted, a single frequency measurement is not an adequate indication of the crosstalk intensity. Accordingly, a variable frequency (warbler) oscillator is usually used as the testing tone, which sweeps continuously back and forth over the frequency range of 830 to 1230 cycles, making about seven sweeping cycles per second. It has been found that this testing frequency band provides reasonably accurate results as compared to the normal voice frequencies encountered in crosstalk.

Figure 432 illustrates schematically the method employed in making near-end crosstalk measurements. The testing tone is alternately connected by means of Key 3 to the disturbing circuit and the metering (or loss measuring) circuit, and the potentiometer (crosstalk meter) in the measuring circuit is adjusted until

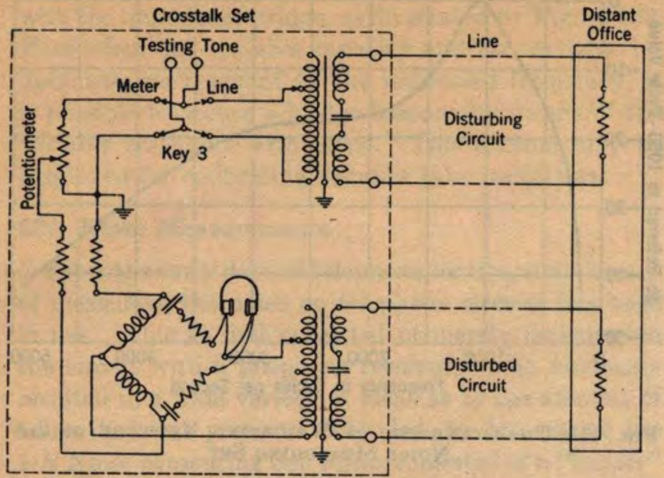


FIG. 432. CROSSTALK MEASURING SET ARRANGED FOR NEAR-END CROSSTALK MEASUREMENT

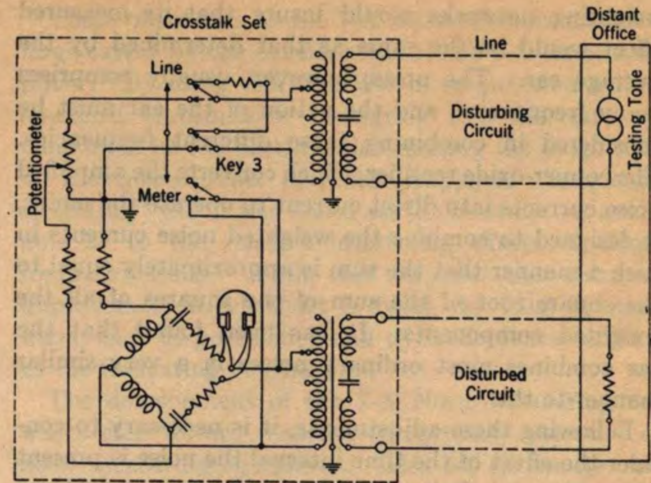


FIG. 433. CROSSTALK MEASURING SET ARRANGED FOR FAR-END CROSSTALK MEASUREMENT

the volume of the tone heard in the receivers is the same for both positions of the key. When the volume of the testing tone is the same over both paths, the loss injected by the potentiometer is obviously equal to the coupling loss between the two circuits. The setting of the potentiometer then gives a quantitative measurement of the near-end crosstalk from the disturbing to the disturbed circuit. By having this potentiometer calibrated directly in Crosstalk Units and Decibels, a direct reading is obtained.

It is important to note that in making near-end crosstalk measurements, any line noise on the disturbed circuit is present in the receivers when listening to the testing tone for both positions of Key 3. Therefore, such noise has no effect on the measurements.

In making far-end crosstalk measurements, the testing tone must be applied at the far-end of the disturbing circuit. The circuit arrangement used is illustrated in Figure 433. By operating Key 3, the disturbing circuit is connected to the calibrated potentiometer (crosstalk meter) where sufficient loss may be added to attenuate the testing tone to the same magnitude as that present in the receivers when Key 3 is operated to terminate the disturbing circuit. The loss measured by the potentiometer is then the measured far-end crosstalk. (As was pointed out in Article 195, it is necessary to add the line loss of the disturbing circuit to this value in order to get the true far-end crosstalk.)

In making crosstalk tests at frequencies above the voice band, some provision must be made in order that the crosstalk frequencies, which are above the audible range, may be heard in the regular telephone receiver. To accomplish this a detector-amplifier is usually employed, which functions as a heterodyne detector in that the presence of the high-frequency currents in its input is made known by an audible frequency in the receiver. This audible frequency is produced by beating the incoming high frequency with that of an oscillator included within the detector-amplifier itself. The arrangements illustrated in Figures 432 and 433 for near- and far-end crosstalk measurements, respectively, are used with the addition of a detector-amplifier in making crosstalk tests at these higher frequencies. The input test tone is supplied at a single frequency. A series of tests at various frequencies is normally made to determine the relative amount of crosstalk throughout the frequency band being considered.

APPENDIX I

PHYSICAL QUANTITIES AND THEIR UNITS OF MEASUREMENT

1. Mechanical Quantities

In addition to the general physical quantities, **force, work** and **energy** the essential ones belonging to the mechanical group are **time, length, area, volume, mass, density, speed, acceleration, and power.** The combined group consists of two classes, viz. (a) those that are fundamental, and (b) those that may be defined from their relations to the fundamental ones.

In dealing with any one physical quantity we need first to know its definition or exact nature and next the method whereby it may be measured. Its measurement will always require a comparison with some unit which is either a fixed standard or may be derived from some fixed standard or standards. In addition to this it may sometimes require the fixing of a positive or negative sign to the numerical size of the quantity.

2. The Three Fundamental Units

There are three units that are fundamental to all mechanical measurements. We may either use these directly to measure any quantity in which we are interested or to derive other units with which the quantity may be measured. These fundamental units, which we must preserve as standards and which must be remembered at all times because they cannot be produced mathematically, are **the unit of time, the unit of length and the unit of mass.**

We daily express periods of time by comparison with familiar units such as second, minute, hour, day and year and have vivid conceptions of their size or greatness, for example: 10 minutes, 2 hours, 3 years, etc. We likewise express distances by comparing them with simple units of length such as inch, foot, yard, mile, etc.

Though the fundamental units of time and length are too familiar to require discussion, the conception of the third fundamental unit or **the unit of mass** is easily confused with the conception of force, and the every day term **weight** may be wrongly taken to mean **mass:**

Mass is defined as amount of matter. There is a piece of platinum carefully preserved in the Standards Office at Westminster which is called the "Imperial Avoirdupois Pound", and this is our standard unit of mass. There is no direct method of comparing the mass of bodies of other material with this piece of platinum and an indirect method must be used. The earth exerts a force of attraction called **gravity** on the mass of all bodies and in practice we merely compare these forces (or **weights**) rather than make a direct comparison of masses. But what are being compared are in reality forces and not amounts of matter.

From the same units with which we express distances, that is, from the units of length, we may derive mathematically units for area or volume. For area, such units are the square inch, the square foot, the square mile, etc. and for volume they are the cubic inch, cubic foot, cubic yard, etc.

3. Density and Specific Gravity

Density is defined as the mass per unit volume of a substance. To measure it we employ the unit of volume (length $\times$ length $\times$ length) and the unit of mass, i.e., mass divided by volume. In practice we seldom use density in the absolute sense but use instead **specific gravity.** The specific gravity of any material is the ratio of the weight of a given volume of that material to the weight of an equal volume of water. For example, the specific gravity of cast iron, a cubic foot of which weighs approximately 500 lbs., is 7.7 since a cubic foot of water weighs approximately $62\frac{1}{2}$ lbs. Likewise the specific gravity of the electrolyte (diluted sulphuric acid) which is ordinarily used in storage batteries is about 1.200, which means that any unit of its volume will weigh 1.2 times as much as the same volume of water.

4. Velocity or Speed

We may express **speed** or **velocity** as the time rate of traveling **distance.** If any body in motion, such as a train, continues its motion without change for one unit of time, such as the hour, the distance is a quantitative measure of its speed. For example, a train's speed may be thirty miles per hour which means if it keeps moving for one hour with the same speed as at the instant observed, it will cover a distance of thirty miles. This does not mean, however, that the train will keep moving for one hour or that it will eventually move the distance of thirty miles; it merely means that if the conditions under which the train is moving at the instant observed are unchanged during a one hour period the train will traverse a distance of thirty miles.

5. Acceleration

If the train at any time should either acquire more speed or "slow down" it would be accelerated. **Acceleration** is defined as the time rate of changing speed. A train is positively accelerated when getting up to speed and negatively accelerated when slowing down. This quantity is an ideal example of a measurement requiring more than a mere numerical comparison. If in one minute's time the train should **increase** its speed one mile per minute, it would have one unit's **positive** acceleration or an acceleration of **plus** one mile—per minute—per minute; if in the same length of time it **decreased** its speed one mile per minute, it would have **negative** acceleration of one mile—per minute—per minute.

6. Force

Mass, such as the piece of platinum already mentioned, is drawn toward the center of the earth by the force of gravity. If it were free to fall, the force of gravity would give to it an acceleration of about 32 feet—per second—per second; that is, at the end of the first second from the time it started to fall it would have a velocity of 32 feet per second and at the end of the next second a velocity of 64 feet per second, etc. That influence which tends to set any body in motion or to change the direction or speed of any body already in motion is called **force.** The **unit** of **force** is that force which the influence of gravity exerts on the standard pound of mass when at sea level. It is called the **pound of force.**

When the forces acting upon a body in any one direction are equal to those acting in the opposite direction, the body is in equilibrium or at rest. When these forces become unbalanced, the body is set in motion and work is performed.

7. Work

Work is done when a force moves a body **in the direction of the force.** It is measured by the product of the force and the distance through which it acts. If a pound of force acts upon a body through a distance of one foot, one **foot-pound** of work is performed. If a six pound hammer is raised from the floor to a bench three feet in height, six times three, or eighteen foot-pounds of work is performed. If the vertical distance between the two floors of a building is 12 feet and a man weighing 150 pounds ascends from one floor to another he performs 12×150 or 1800 foot-pounds of work. Here it should be noted that the force acts through the vertical distance only and not through any horizontal distance he may travel while ascending flights of stairs. The distance must be measured parallel to the direction of the force.

Work in its scientific sense is measured by the result and not by effort or exertion. A man may attempt to move a weight until he is fatigued, but he performs no **mechanical work** unless he succeeds in elevating **the weight** against the force of gravity, changing its position in a given plane against the friction and inertia tending to hold it at rest, or overcoming any other resisting forces that may be acting upon it.

8. Power

In the same sense that speed is the time **rate** of traversing distance, **power** is the time **rate** of **doing work.** Its unit of measurement, therefore, not only requires the fundamental units of foot and pound for its derivation but must include time also. If a machine is capable of performing 33,000 foot-pounds of work, we have no conception of the size of the machine. It may be a small machine capable of performing this amount of work in fifteen days or a large machine capable of performing the work in three seconds, but a machine that can perform **33,000** foot-pounds of **work in one minute** is rated at **one horse power.**

9. Energy

Energy is ability to do work or is stored work and may therefore be expressed in the same units as work. A suspended weight or a fly wheel in motion are said to have stored mechanical energy by virtue of their ability to perform useful work.

Energy also exists in forms other than mechanical. A strong acid may have the ability to dissolve metals; a solution may have the ability to generate an electric current when in contact with two dissimilar metals. Food, when digested, permits man to perform useful tasks. These are examples of ability to perform work, but the **energy** is in a **chemical** and not a **mechanical state.**

Another very common form of chemical energy is that of coal which when burned gives off large quantities of heat. Heat is within itself a form of energy and may be made to perform work. It may be applied to a boiler in such manner as to generate steam under pressure and the steam may be used to operate an engine or turbine. The amount of **heat** required to raise the temperature of one pound of water one degree Fahrenheit is the **British Thermal Unit.**

The following have approximately equal energy values, viz.: 10,000,000 foot-pounds:

1. One pound of coal.
2. One day's supply of food for adult man of average muscular vigor.
3. Medium sized 24-volt storage battery fully charged.
4. Amount of water in average railroad water tank when elevated from ground level.

Energy may be converted from one form to another but **can never be produced or destroyed.** This does not mean, however, that one pound of coal could be made to pump sufficient water to fill the railroad water tank mentioned in the foregoing. The best known devices for converting heat or chemical energy into mechanical energy are very inefficient and only a small part of the coal's energy value can be made to produce useful work, the greater portion being lost by heat radiation or conduction rather than utilized by conversion.

10. The Metric System of Measurements

The system of measurements used by all civilized countries except Great Britain and the United States is called the **metric system.** It is based upon multiples of ten or upon the decimal plan, in a manner similar to that of our money system in this country. On account of its marked advantages over our awkward pound, foot, acre, quart, bushel, etc., it has been standardized for all scientific work, and consequently the electrical units are based upon this more convenient system. In the same manner that our money system greatly facilitates calculations by being based on the multiples of ten (namely, mil, cent, dime, dollar and eagle) the metric system provides units for measurement of all physical quantities which facilitate simple calculations involving these quantities.

A system of prefixes is used in the metric system to indicate multiples of ten or decimal parts. This is illustrated in Table I, which gives the metric units for measurements of length.

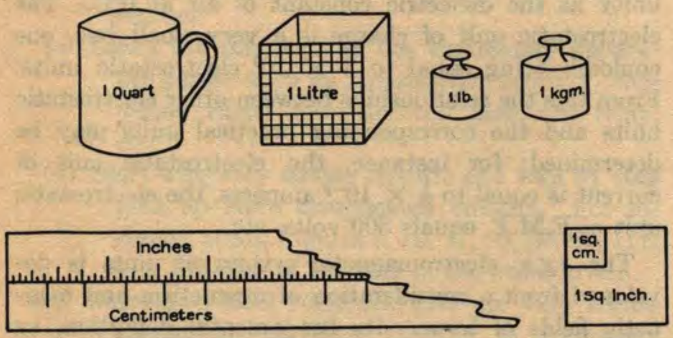


Fig. 1. Comparison of English and Metric Units

The measurement of volume and liquid capacity is based upon the litre, which is **a cube** with each dimension ten centimeters in length. It is, therefore, equal to 1000 cubic centimeters. It can be remembered as being a little larger than our quart.

The measurement of **mass** is based upon the **gram,** a very small unit and approximately equal to the mass of one cubic centimeter of water. It can be remembered as equal to one-fifth of our five-cent piece. In other words, a U. S. five-cent piece weighs 5 grams. The kilogram or 1000-gram weight is more commonly used. It can be best remembered as being a little larger than two pounds.

TABLE I

Metric Units of Length

UNIT	DECIMAL
1 Millimeter	.001 meter
1 Centimeter	.01 meter
1 Decimeter	.1 meter
1 Meter	
1 Dekameter	10. meters
1 Hectometer	100. meters
1 Kilometer	1000. meters

There is only one system for measuring **time,** and the **second** has the same meaning in the scientific system as in our system.

Temperature is based upon the Centigrade scale instead of the Fahrenheit wherever the metric system is used. Zero degrees on the Centigrade scale corresponds to freezing point of water at sea-level atmospheric pressure (14.7 lbs. per square inch) or 32 degrees on the Fahrenheit scale; 100 degrees on the Centigrade scale is the temperature of boiling water and corresponds to 212 degrees on the Fahrenheit scale. All intermediate degrees are proportional, which means that one degree Fahrenheit is equal to five-ninths degree Centigrade.* Figure 2 illustrates the thermometer showing both scales.

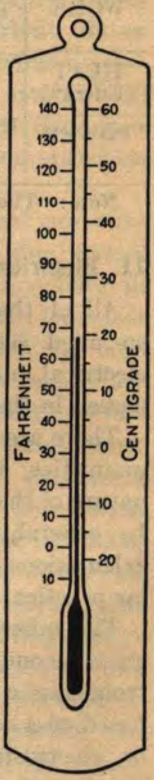


Fig. 2. Comparison of Fahrenheit and Centigrade Scales

Table II gives the conversion factors between the English and metric systems for the units of various physical quantities. Included in this table are several units that have not been discussed in the preceding pages, such as watt, dyne and joule, but the significance and use of such units is explained in the body of the text. Particularly important from our point of view are the relations between the watt or kilowatt and the horse power, and between the joule and the foot-pound.

* For converting temperatures from Fahrenheit to Centigrade the above relations may be expressed as a formula—

$$C^\circ = (F^\circ - 32)\ 5/9$$

and for converting Centigrade to Fahrenheit—

$$F^\circ = (C^\circ \times 9/5) + 32$$

TABLE II

Relation Between Metric and English Units

QUANTITY	Conversion from Metric to English: Name	Abbreviation	Value	English equivalent	Conversion from English to Metric: Name	Metric Equivalent
TIME	Second	sec.	(Same in both systems)		second	(Same in both systems)
LENGTH	centimeter meter kilometer	cm. m. km.	.01 meter — 1000 meters	.39 inch 1.094 yards .625 mile	inch foot yard mile	2.54 cm. .305 meter .915 meter 1.609 kilometers
MASS	gram kilogram	gm. kg.	— 1000 grams	.035 ounce (avoir.) 2.204 pounds	ounce pound	28.35 grams .454 kilogram
AREA	square-centimeter square-meter	sq. cm. sq. m.		.155 sq. in. 1.2 sq. yd.	square-inch	6.45 sq. cms.
VOLUME	cubic centimeter litre	cu. cm.	.001 litre 1000 cu. cms.	.061 cu. in. 1.06 quarts	cubic-inch cubic-foot gallon	16.39 cu. cms. 28.3 litres 3.8 litres
FORCE	dyne (seldom used) gram (of force) kilogram (of force)		.00102 gram 980 dynes 1000 grams	.035 ounce (of force) 2.20 pounds (of force)	pound (of force)	.454 kilogram (of force)
WORK	erg (seldom used) joule		1 dyne-centimeter 10,000,000 ergs	.738 ft. lb.	foot-pound	1.35 joules
HEAT (Energy)	gram-calorie large-calorie	gm.-cal.	gm. of water 1° C 1000 calories	3.087 ft.-lbs. 3.963 B.t.u.'s	British thermal unit	1055 joules 255 gram-calories
POWER	watt kilowatt	kw.	joule per second 1000 watts	.74 ft.-lb. per sec. 1.34 horse power	horse power	746 watts or ¾ kw. approximately

Note:—The decimal fractions in the above table are not carried further than is essential for the applications made in this text.

11. Electrical Units

All of the practical units as well as some of the so-called scientific or absolute units for measuring electrical and magnetic quantities are defined and discussed in the body of the text.

There are three systems of units for measuring such quantities, the utility of each depending upon the nature of the calculations or measurements to be made. In general, except for certain kinds of theoretical calculations and laboratory experimentation, the familiar practical system is to be preferred.

The other two systems are the c.g.s. (centimeter-gram-second) electrostatic system and the c.g.s. electromagnetic system. The first of these systems is based on a consideration of electric charges and electric or electrostatic fields. As is well known electric charges or so-called "static electricity" may be produced on the surface of substances such as amber, rubber or glass by rubbing them vigorously with dry silk or fur. Such charges are accompanied by an electric or "static" field of force which, acting in the space around a charged body, causes it to attract non-charged bodies such as small bits of paper and to attract or repel other charged bodies, depending on whether the other bodies are charged with electricity of the opposite or the same sign. An electric field is always present wherever electricity is moving or tending to move. The electrostatic unit of charge or quantity of electricity is that possessed by each of two bodies which repel or attract each other with a force of one dyne when the bodies are one centimeter apart. This definition depends upon the arbitrary selection of unity as the dielectric constant of air at 0°C. The electrostatic unit of charge is a very small one, one coulomb being equal to 3×10^9 electrostatic units. From this the relationships between other electrostatic units and the corresponding practical units may be determined; for instance, the electrostatic unit of current is equal to $\frac{1}{3} \times 10^{-9}$ amperes, the electrostatic unit of E.M.F. equals 300 volts, etc.

The c.g.s. electromagnetic system of units is developed from a consideration of magnetism and magnetic fields of force. Its fundamental definition, as given in the text, is that of the unit of magnetic pole strength, i.e., unit pole strength is that possessed by each of two magnetic poles that repel or attract each other with a force of one dyne when placed one centimeter apart. This definition depends upon the arbitrary selection of unity as the permeability of air at 0°C. The c.g.s. electromagnetic unit of E.M.F. is equal to 10^{-8} volts, while the electromagnetic unit of current equals 10 amperes. Other relationships between electromagnetic and practical units may be derived from these.

APPENDIX II

CIRCUIT DIAGRAM READING

A casual glance at a complicated telephone circuit drawing often leads the student to think that circuit reading is like a strange language acquired only by the most difficult study. He is confronted with a maze of apparatus parts connected by a complex entanglement of wires, and he despairs at the thought of memorizing hundreds of such circuits, each different from the others. Fortunately, this impression is more or less of an illusion. Though circuit reading is somewhat of a "knack", it depends for the most part upon skill to be acquired by learning the underlying principles step by step. Furthermore, while the best telephone men may diligently study these principles and apply them in practice, they never attempt to memorize circuit drawings. The large number of complicated circuits that are essential to the proper handling of telephone service can never be mastered in this way.

Accordingly, the following hints are given not as a key to every telephone circuit, but as a review of the underlying principles and as a recommended procedure for those not familiar with toll central office circuits:

(a) **Learn the principles of current flow including Ohm's and Kirchoff's Laws.**

(b) **Memorize the conventional circuit symbols for the commonly used units of telephone apparatus.**

(c) **Learn the functions of the elementary apparatus parts**—such as transmitter, receiver, condenser, ringer, induction coil, jack, key, relay, etc.

(d) **Start with very simple circuits and step by step take up more complicated circuits.** For instance, study Figures 7, 10, 17, 20, 80, 81, 111, 112, 113, etc. of the text in the order given.

(e) **Remember that voice currents are very feeble alternating currents ranging in frequency somewhere between 200 and 3000 cycles, that ringing currents are alternating currents of 20, 135, or 1000 cycles, and that the majority of relays are operated with direct currents.** On most drawings, the direct path of voice currents is shown by heavy solid lines and the path for direct or ringing currents, when not over the same conductors as voice currents, is shown by light lines.

(f) **Bear in mind that direct currents will not flow through condensers and will flow through retardation coils and that alternating currents will effectively flow through condensers but are greatly reduced in their flow through retardation coils. Also, that direct current will not flow from one winding to another of an induction coil, repeating coil or transformer, and that alternating currents will flow, or to be more exact, are induced across from one winding to another of these coils.**

(g) **Learn first the purpose of operation, or in other words, that which the circuit is designed to accomplish.** For example, in Figure 113 of the text we might say that the circuits shown are designed to permit local telephone connections where one operator can establish all connections and the circuit design must permit common battery single position operation as follows:

Subscriber takes receiver off hook and subset circuit in conjunction with line circuit must give operator signal by lighting lamp in face of switchboard in front of her. Operator must be able to answer subscriber by connecting her telephone set to some one of her cord circuits and connecting this cord circuit to calling party's line. Line circuit must contain relay features which will extinguish lamp when operator makes this connection. Cord circuit must permit operator to make similar connections to called party's line and to ring called party. Ringing current must not reach ear of operator or calling party. Operator must have signal associated with cord circuit that will tell her when called party has answered and when both parties are through talking and hang up. The system must be so designed as to provide direct current over line to subscribers' telephones for transmitter supply. The circuits must be designed to give an efficient voice-current path from one subscriber to the other and from each subscriber to operator. Keys must be provided to permit operator to disconnect her head set from, or connect her head set to, cord circuit at will, thereby permitting her to handle connections with other cord circuits while two particular subscribers are talking.

We thus have the detailed performances for which the circuit must be designed. These are requirements for mechanical and electrical features that should be provided in a somewhat automatic manner. Knowing these, or in other words, knowing the operating procedure for establishing connections between two common battery subscribers, each part of the circuit will stand out as having a specific purpose and make circuit reading comparatively simple.

(h) **Get an approximate mental picture of the layout of the various apparatus parts.** For example, in Figure 113, we may think of the transmitter, receiver and hook switch in the telephone set located on a table or desk at the subscriber's office or residence; the condenser, induction coil and bell located in the bellbox at the subscriber's office or residence; the protector and relay equipment of the subscriber's line circuit located in the central office terminal room; the jack and lamp located in the face of the switchboard in front of the operator; the cord circuit equipment located somewhere in the switchboard position, with the plugs, keys and supervisory lamps of this circuit in the key-shelf where the supervisory lamps are visible and the plugs and keys are readily accessible.

(i) **Do not attempt to read wiring diagrams.** These are intended to assist the wireman in making the proper connections of cable pairs and other wires to the apparatus terminals and are not intended for circuit study. Make sure that the circuit drawing is a schematic or theory drawing and not a wiring diagram.

(j) **Learn thoroughly the operation of a few important telephone circuits pertaining to your particular branch of telephone work.**

(k) **If the circuit is a difficult and unusual one, ask someone to explain it to you in preference to following a written explanation of it.**

APPENDIX III

CONSTRUCTION AND USES OF CURVES

The use of graphical charts or curves is usually the most convenient and effective method of presenting data where two interdependent variables are involved. Such charts and curves are in most cases more readily understood than the corresponding mathematical equations and are not only easier to follow but take up less space than tables.

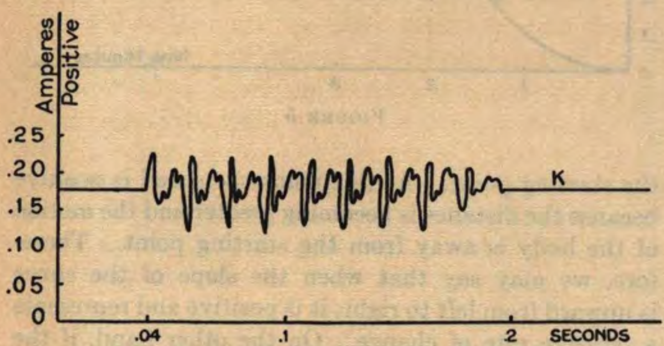


FIGURE 1

Any relationship between two variables, such for instance as $y = ax$, where y varies as x varies (or is said to be a "function of x") may be clearly pictured on a simple plane chart where one variable, say x, is plotted on a horizontal scale and the other variable, y, is plotted on a vertical scale. The two scales do not need to be alike, although each one must itself be uniform. Thus, Figure 1 shows the rather complex relationship between time and the current in the primary of a telephone subset induction coil when a certain vowel sound is spoken into the transmitter. Here we have a horizontal scale representing time, beginning at some instant designated as zero time and charted to the right in graduations of tenths of seconds and a vertical scale representing current values charted upward in graduations of .05 ampere. Thus curve K shows all

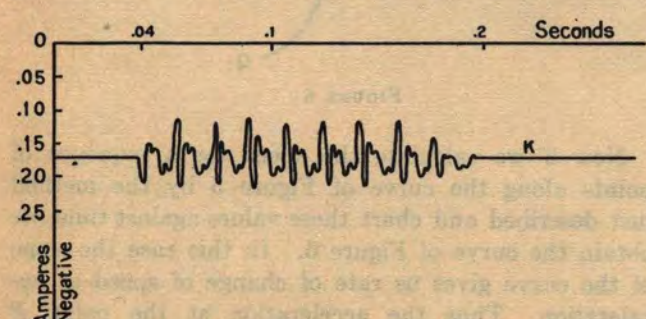


FIGURE 2

values of the current in the circuit for the interval of time considered and conveys a better and more complete idea of what is actually taking place in the circuit than could be expressed in words. During the interval of time between zero and .04 second, i.e., during the first .04 second considered, there is a steady current through the transmitter and induction coil primary of .18 ampere; but beginning at the instant represented by .04 second a vowel sound is spoken into the transmitter and its resistance is alternately lowered and increased causing fluctuations in the current value which are represented by the irregular portion of the curve between .04 and .2 second. The value of the current could of course be charted over any period of time but in this case we have sufficient values to show clearly the effect on the transmitter current of a certain spoken vowel.

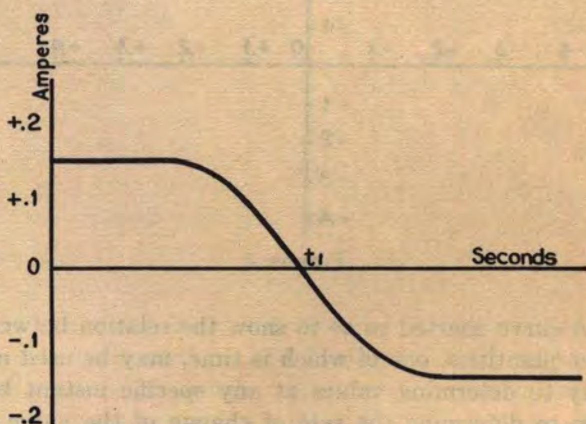


FIGURE 3

The chart of Figure 1 is adequate for showing the magnitude of a current varying with time. Frequently, however, cases are encountered where it is desirable to show not only the magnitude of the current but its direction as well. To take care of this a convention has been adopted that values plotted upward on the vertical scale shall be considered as positive and values plotted downward as negative. Similarly, values plotted to the right on the horizontal scale are positive and to the left, negative. Thus, if the battery causing the current in Figure 1, which we have considered positive, were reversed, we would have a current in the opposite or negative direction and our current-time diagram would be that of Figure 2. In the same way, to represent an alternating current we may use a current-time curve such as that shown by Figure 3, which beginning at zero time represents a sequence of positive current values decreasing to zero value after the time t_1, and followed by a sequence of negative current values (i.e., current in the opposite direction) increasing from zero at time t_1.

Obviously curves such as those illustrated and discussed above may be used for other purposes than charting the relation between current and time. They may quite as frequently be used to chart relations between voltage and time or, in fact, between any two quantities that are so related that one is a function of, or is dependent on the other. There are frequent occasions, also, for charting the relationship between two quantities both of which may be either positive or negative. For this purpose a diagram known as a "complete rectangular coordinate diagram" such as that illustrated by Figure 4 is used. The hysteresis curves shown by Figures 42 and 43 in Chapter III of the text are good examples of diagrams of this kind.

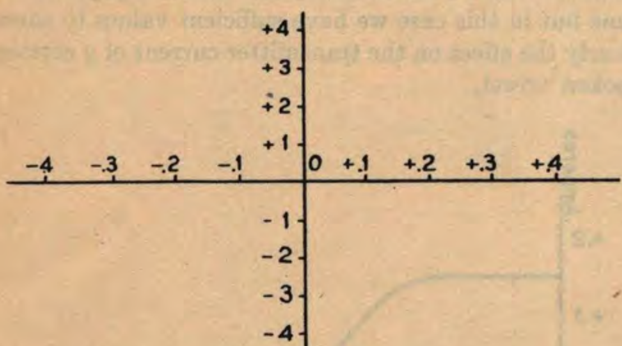


FIGURE 4

A curve charted so as to show the relation between two quantities, one of which is time, may be used not only to determine values at any specific instant but also to determine the **rate of change** of the value at any instant. The curve of Figure 5, for instance, which represents the relation between distance covered by some moving object and the time required to cover this distance, will tell us how far the object is from the starting point at any instant and also the speed or **rate of change of distance** at that instant. This speed or rate of change of distance is defined as the **slope** of the curve at the designated instant. Thus, the speed at which the moving object of Figure 5 is traveling .2 minute after it leaves the starting point may be determined by finding the slope of the curve at the point P, where a vertical line drawn upward from the .2 minute graduation on the time scale intersects the curve.

The slope may be determined by drawing through the point on the curve a straight line **tangent** to or having the same direction as the curve at the point and constructing on this straight line a right triangle of any convenient size but having one leg, as *bc*, vertical and the other, *ab*, horizontal; then the value of *bc* measured on the vertical scale of the chart divided by the value of *ab* measured on the horizontal scale is the slope of the curve at the point P. This quantity is distance divided by time and therefore is a measure of rate of change of distance or, in more useful terms, the **speed** at which the body is moving .2 minute after it leaves

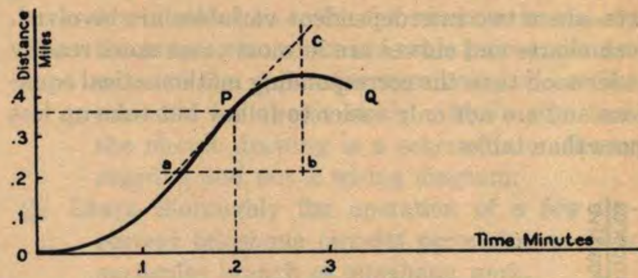


FIGURE 5

the starting point. At this point the speed is positive because the distance is becoming greater and the motion of the body is away from the starting point. Therefore, we may say that when the slope of the curve is upward from left to right, it is positive and represents a positive rate of change. On the other hand, if the body ceases to move away from the starting point and returns toward it, it may be said to have negative speed. Such a condition is illustrated by point Q of Figure 5 and here the slope of the curve is downward from left to right and is negative.

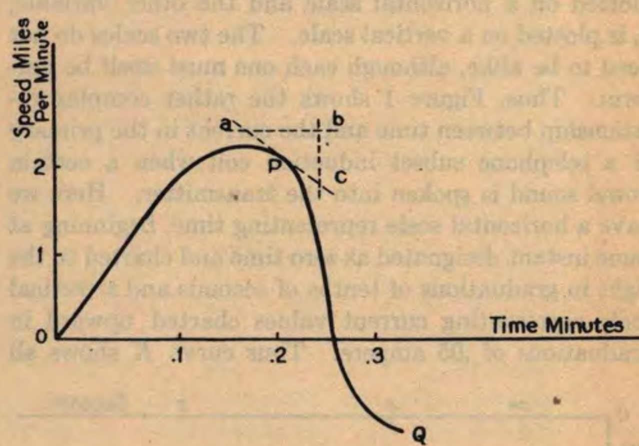


FIGURE 6

Now if we calculate the speed for a sequence of points along the curve of Figure 5 by the method just described and chart these values against time, we obtain the curve of Figure 6. In this case the slope of the curve gives us rate of change of speed or **acceleration.** Thus the acceleration at the point P of Figure 6, or .2 minute after the body starts to move, may be measured by drawing a tangent to the curve at the point, constructing a triangle as shown, and dividing the length of the line *bc* as projected on the speed scale by the length of the line *ab* as projected on the time scale. As before, we have positive acceleration meaning increasing speed, which is represented by positive slope, and negative acceleration meaning decreasing speed or "slowing down", which is represented by negative slope. At point *P* the acceleration is negative since the body is slowing down prior to reversing its direction of motion and returning toward the starting point.

APPENDIX IV

WAVE MOTION FREQUENCY SCALES

1. Vibratory Motion

Manifestations of vibratory or wave motion are common in all nature and many forms take place about us continuously. Perhaps the most obvious form is the water wave which everyone has observed to be a form of vibratory motion. By means of condensations and rarefactions of the atmosphere, which are as truly a form of wave motion as the water wave, we experience the sensation called sound. By virtue of still another form of wave motion consisting of a vibration in the mysterious substance which fills all space, called "ether", we recognize the color of some brilliant object. To go further, if this vibration is of a certain particular frequency, the color registered by the retina of the eye may be red but if nearly twice this particular frequency, we are conscious of violet color instead of red. In an electrical conductor, we may think of an electrical current flowing first in one direction and then in the other, as another example of wave phenomena not altogether different from that of light, and although belonging to an entirely different category, similar in some respects to that of sound or even water waves.

Audible sound is defined as a disturbance in the atmosphere whereby a form of wave motion is propagated from some source with a velocity of approximately 1075 feet per second, the transmission being by means of alternate condensations and rarefactions of the atmosphere in cycles having a fundamental frequency ranging somewhere between 16 and 32,000 cycles per second. In Figure 1, we have a scale arranged by frequencies corresponding to periodic recurring octaves. It shows the musical staff notation, and the frequencies of the notes vary as numbers whose logarithms can be shown as a simple progressive scale. Figure 2 shows another sound scale as a simple table of frequencies again arranged by graduations corresponding to octaves or logarithms. Here, however, frequencies are shown by tens, hundreds and thousands instead of as notes of the musical scale. This figure also extends the frequencies to the absolute limits of human hearing or the so-called "limits of audibility". Vibrations coming within the limits of this scale are referred to as **audio-frequencies,** and an alternating electrical current having a frequency, for example, of 1000 cycles, would be carrying the equivalent of a monotonous pure tone lying somewhere between C″ and C‴ of the musical scale.

Figure 3 illustrates a scale of ether vibrations which transmit light, radiated heat, and the electromagnetic waves of wireless telephony and telegraphy. This scale is likewise arranged by graduations proportional to octaves. The number shown opposite each graduation corresponds to frequency, and the wave-lengths at the important division points are given in Angstrom units. As noted on the drawing, this is a very small unit equal to one ten-millionth of a millimeter. The division points are rather arbitrarily chosen in most cases and are intended only to group the various classifications in a general way. Except in the case of visible light, there is naturally no sharp dividing line between the different classes of radiation.

2. Alternating-Current Frequency Scales

All alternating-current frequencies that have any general application in practice, can be grouped and

	C_1	C	D	E	F	G	A	B	c	d	e	f	g	a	b	c′
Relative Pitch	$\frac{1}{2}$	1	$\frac{9}{8}$	$\frac{5}{4}$	$\frac{4}{3}$	$\frac{3}{2}$	$\frac{5}{3}$	$\frac{15}{8}$	2	$\frac{9}{4}$	$\frac{5}{2}$	$\frac{8}{3}$	3	$\frac{10}{3}$	$\frac{15}{4}$	4
Frequencies	32	64	72	80	$85\frac{1}{3}$	96	$106\frac{2}{3}$	120	128	144	160	$170\frac{2}{3}$	192	$213\frac{1}{3}$	240	256

	d′	e′	f′	g′	a′	b′	c″	d″	e″	f″	g″	a″	b″	c‴	c^{IV}
Relative Pitch	$\frac{9}{2}$	5	$\frac{16}{3}$	6	$\frac{20}{3}$	$\frac{15}{2}$	8	9	10	$\frac{32}{3}$	12	$\frac{40}{3}$	15	16	32
Frequencies	288	320	$341\frac{1}{3}$	384	$426\frac{2}{3}$	480	512	576	640	$682\frac{2}{3}$	768	$853\frac{1}{3}$	960	1024	2048

FIG. 1. MUSICAL SCALE

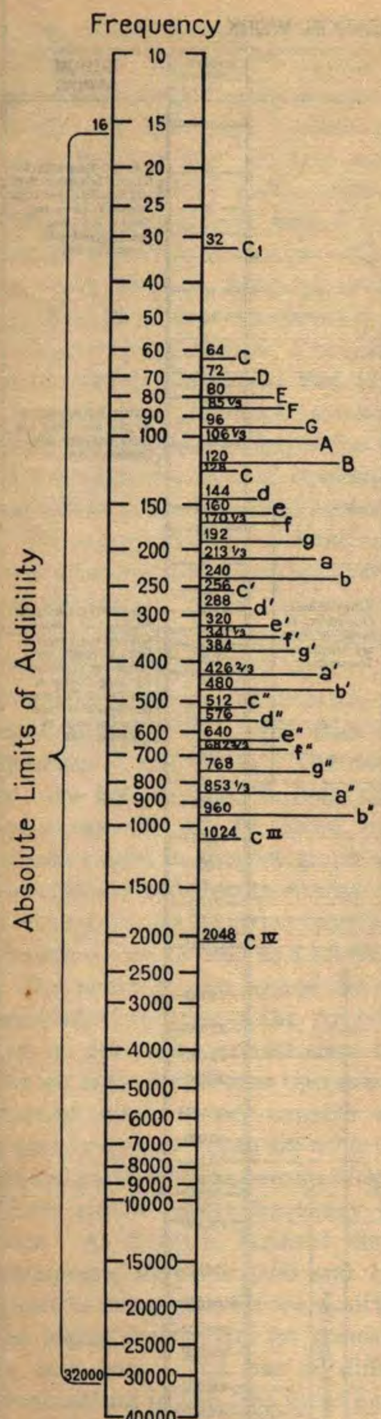


Fig. 2. Logarithmic Musical Scale

graphically pictured similarly to the foregoing scales for sound and light. Thus, we have represented in Figure 4 a chart of the alternating-current frequencies used in both power and communication work. This chart naturally includes many of the same frequency values shown in the foregoing scales, because frequencies identical in value with those of sound are commonly used in telephone circuits, and frequencies used in wireless telephony and telegraphy are also often used for various wire applications, such as carrier. A study of Figure 4 gives us the entire field of use of electricity in the alternating-current form, so ordered that we see at a glance the wide range of frequency values and their adaptability to different utilitarian applications.

This chart is divided into four columns, representing four different fields of use of electricity. The first column is headed "Power Work", the second "Telegraph", the third "Telephone", and the fourth "Telephone Signaling". Adjacent to each column, either on the right or left, is a scale of frequencies which is again arranged by octaves with the actual frequency values shown. In other words, it is a logarithmic rather than an arithmetic scale, which shows frequencies from 10 cycles upward.

In the "Power" column, we find a bracketed note opposite 25 cycles which covers a frequency often used in power work when alternating-current railways are involved and lighting is a secondary consideration. But 60 cycles is indicated as the standard frequency used most extensively in power work. Each standard has its advantages. Alternating-current machinery designed for 25 cycles need not be designed to operate at high speeds, and in some cases permits better power control than 60 cycles. On the other hand, 25 cycles is not desirable for lighting inasmuch as the ordinary lamp filament will sufficiently cool between positive and negative current peaks of the cycle (or in the $\frac{1}{50}$th second interval) to cause the light to "flicker". The 60-cycle system is ideally adapted to lighting and for most applications is well adapted to power work.

Near the bottom of this column is a note explaining that lightning frequencies range from a few hundred thousand cycles to the higher wireless frequencies. Of course, all exposed power and communication lines, as well as radio receiving and transmitting circuits, are subject to lightning hazards, and protection design is an application to circuits involving these frequencies.

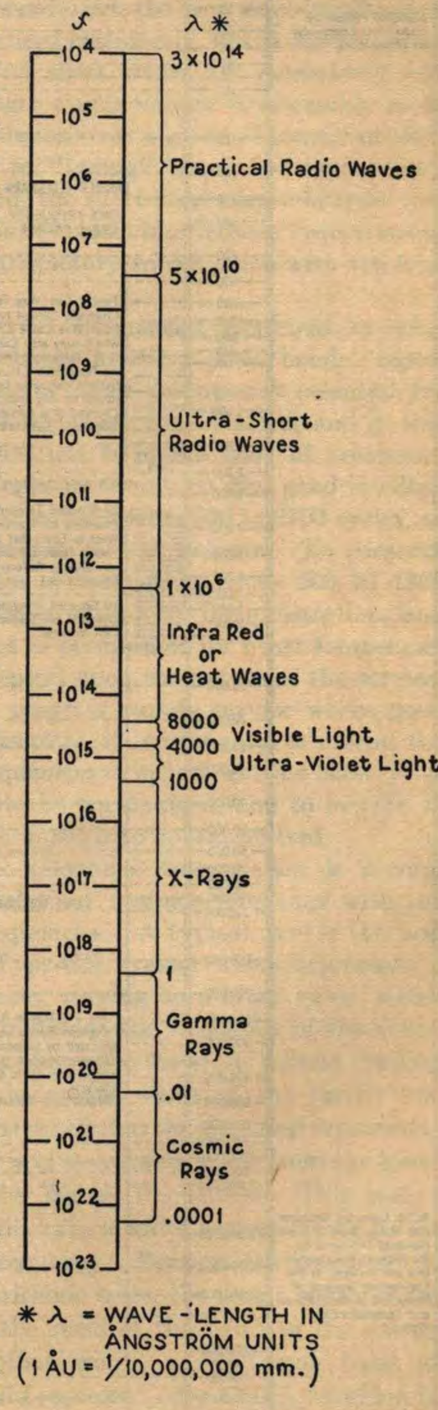
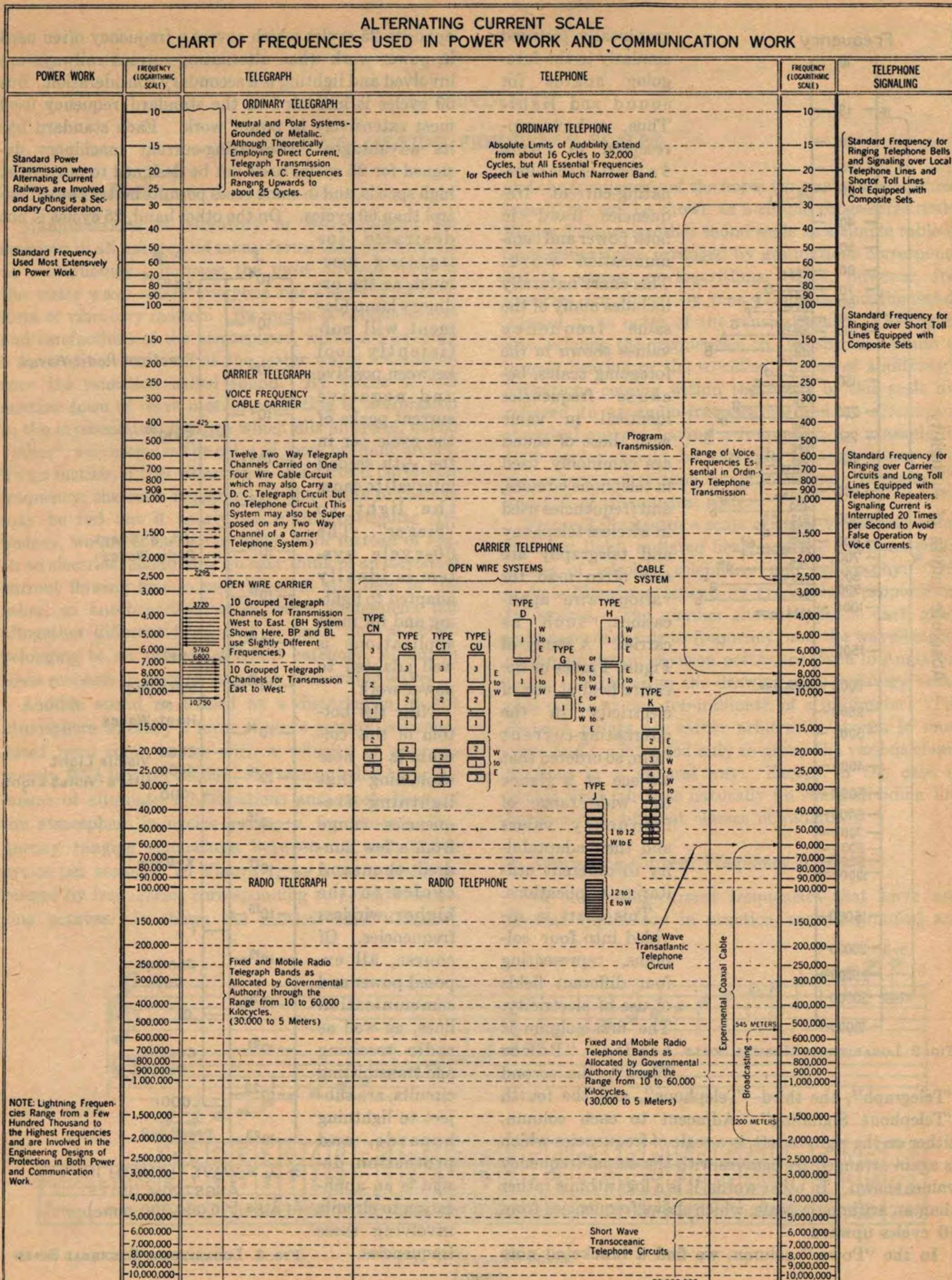


Fig. 3. Logarithmic Ethereal Scale



ALTERNATING CURRENT SCALE
CHART OF FREQUENCIES USED IN POWER WORK AND COMMUNICATION WORK
POWER WORK
FREQUENCY (LOGARITHMIC SCALE)
TELEGRAPH
TELEPHONE
TELEPHONE SIGNALING
Standard Power Transmission when Alternating Current Railways are Involved and Lighting is a Secondary Consideration.
Standard Frequency Used Most Extensively in Power Work.
NOTE: Lightning Frequencies Range from a Few Hundred Thousand to the Highest Frequencies and are Involved in the Engineering Designs of Protection in Both Power and Communication Work.
ORDINARY TELEGRAPH
Neutral and Polar Systems-Grounded or Metallic. Although Theoretically Employing Direct Current, Telegraph Transmission Involves A. C. Frequencies Ranging Upwards to about 25 Cycles.
CARRIER TELEGRAPH
VOICE FREQUENCY CABLE CARRIER
Twelve Two Way Telegraph Channels Carried on One Four Wire Cable Circuit which may also Carry a D. C. Telegraph Circuit but no Telephone Circuit (This System may also be Super posed on any Two Way Channel of a Carrier Telephone System.)
OPEN WIRE CARRIER
10 Grouped Telegraph Channels for Transmission West to East. (BH System Shown Here, BP and BL use Slightly Different Frequencies.)
10 Grouped Telegraph Channels for Transmission East to West.
RADIO TELEGRAPH
Fixed and Mobile Radio Telegraph Bands as Allocated by Governmental Authority through the Range from 10 to 60,000 Kilocycles. (30,000 to 5 Meters)
ORDINARY TELEPHONE
Absolute Limits of Audibility Extend from about 16 Cycles to 32,000 Cycles, but All Essential Frequencies for Speech Lie within Much Narrower Band.
Program Transmission.
Range of Voice Frequencies Essential in Ordinary Telephone Transmission.
CARRIER TELEPHONE
OPEN WIRE SYSTEMS
CABLE SYSTEM
TYPE CN
TYPE CS
TYPE CT
TYPE CU
TYPE D
TYPE G
TYPE H
TYPE J
TYPE K
E to W
W to E
1 to 12 W to E
12 to 1 E to W
RADIO TELEPHONE
Long Wave Transatlantic Telephone Circuit
Experimental Coaxial Cable
Broadcasting
545 METERS
200 METERS
Fixed and Mobile Radio Telephone Bands as Allocated by Governmental Authority through the Range from 10 to 60,000 Kilocycles. (30,000 to 5 Meters)
Short Wave Transoceanic Telephone Circuits
22,000,000
Standard Frequency for Ringing Telephone Bells and Signaling over Local Telephone Lines and Shortor Toll Lines Not Equipped with Composite Sets.
Standard Frequency for Ringing over Short Toll Lines Equipped with Composite Sets
Standard Frequency for Ringing over Carrier Circuits and Long Toll Lines Equipped with Telephone Repeaters. Signaling Current is Interrupted 20 Times per Second to Avoid False Operation by Voice Currents.

FIGURE 4

The second column of Figure 4 headed "Telegraph", covers four classes of telegraph transmission. First, the more common system designated "ordinary telegraph", including the familiar grounded neutral and polar systems and metallic cable systems; second, the voice-frequency carrier current telegraph system; third, the superposed carrier current telegraph; and fourth, the radio telegraph. The first application, or the more common telegraph system which makes and breaks a direct-current circuit is usually thought of as a direct-current system, but considering the rapidity of the makes and breaks and the wave-like nature of the telegraphic pulses, we have an approximation to an alternating-current transmission that we could describe as a telegraphic band of frequencies beginning at zero and shading into about 25 cycles.

The other applications make and break a steady flow of alternating current rather than direct current. The voice-frequency cable carrier system makes use of a series of frequencies beginning at 425 cycles and extending to 2295 cycles or higher. By this means twelve two-way telegraph channels are secured over a 4-wire cable circuit, with a separation in frequency between channels of 170 cycles. The use of the 4-wire circuit permits transmitting in both directions at the same frequency. It will be noted, however, that the frequencies used in this telegraph application are within the ordinary voice range so that a cable circuit used for a voice-frequency carrier telegraph system cannot at the same time be used as a telephone circuit.

Due to the greater cost of the wire facilities, it is not economical to operate the voice-frequency carrier system on open wire circuits since this would not permit the use of the facilities as telephone circuits. Open wire circuits, however, are capable of transmitting much higher frequencies than are cable circuits. This feature makes possible the superimposing on an open wire telephone circuit higher frequency carrier telegraph systems. As shown in Figure 4, these carrier systems use frequencies between 3300 and 11,250 cycles, ten frequencies being used for transmitting West to East and ten higher frequencies for transmitting East to West in each case. The use of different frequencies for transmitting in the two directions makes possible the application of the system on an ordinary 2-wire circuit.

The radio telegraph employs frequencies from about 10,000 cycles to more than 60,000,000 cycles. The wave-lengths corresponding to any radio frequency (or any ethereal vibration whatsoever) are calculated from the formula—

$$\lambda = \frac{300,000,000}{f}$$

where λ is the symbol for wave-length in meters and f is the frequency. This is evident from the known speed of ethereal waves, which is 300,000,000 meters per second. The number of wave-lengths in one second (or frequency) times the length of one wave must equal the distance traveled in one second, or 300,000,000 meters.

Wave-lengths are assigned for radio telegraph and radio telephone communication through Government or International Conference regulation and are accordingly subject to change from time to time. In the earlier days of the radio art, the long wave-lengths (low frequencies) were used exclusively, but in modern practice both long and short waves are extensively employed. The longer waves require considerably more power for transmission over a given distance but they are less subject to "fading" than the short waves. On the other hand, the short waves are relatively free from certain kinds of "static interference" which sometimes prevents satisfactory transmission with the long waves.

Alternating-current frequencies employed in **telephone transmission** are in every case "bands" representing the range of voice frequencies essential for intelligibility in a telephone conversation, and in this respect the application is unlike that of telegraph. The ordinary telephone circuit, to give good intelligibility, must employ a band from 200 to 2700 cycles, as shown by the bracket in this column. To preserve quality, the range of frequencies from 500 to 1800 should be transmitted with very little distortion, and the distortion that is permissible for other frequencies will, of course, depend upon the nature of the service. For example, in program supply service where good quality is very essential, it is necessary to extend the entire band of frequencies to include at least 3000 cycles if only speech is to be transmitted, and to include at least 5000 cycles if music is to be transmitted.

Carrier current telephone transmission is accomplished by "modulating" a single frequency with the band of voice frequencies. A typical case is the first channel of the Type-CN system which represents a 7600-cycle frequency varying to a lower value, which variation would represent the frequency of the voice. Thus, each carrier channel is shown as a band shading away from the designated value of the carrier frequency. It will be noted that the graphical representation of the bands is of decreasing width from the lower frequencies to the higher frequencies. This means that each band will have a width corresponding to the range of voice frequency. Because the frequency is plotted on a logarithmic scale, the scalar width of the band represents the ratio of the band to the carrier frequency. In other words, a 3000-cycle band at 30,000 cycles would represent only a 10% variation in the carrier frequency, while a 3000-cycle band with a

9000 carrier frequency would represent a 33⅓% variation.

Frequencies used for radio telephony extend through the same range as those employed in radio telegraphy. At the lower frequency end, they overlap the carrier frequencies and they extend upward far beyond the scale of the chart. Because they are subject to frequent change, no attempt is made to show the detailed allocations for various purposes. However, the ordinary (long wave) broadcast band is indicated as extending through the range between 550 and 1500 kc. The chart also shows the allocations for the short-wave transoceanic telephone circuits and for the single long-wave circuit between New York and London.

While **telephone signaling** is essentially a part of telephone communication service, it is, nevertheless, distinct from the transmission of the human voice. The fourth column of Figure 4 shows three telephone signaling systems. The standard ringing current for ringing telephone bells and signaling over local telephone lines, as well as over short toll lines not equipped with composite sets, has a comparatively low frequency of approximately 20 cycles, and is ordinarily referred to as "20-cycle ringing".

The 20-cycle current would naturally fall within the band of frequencies discussed in connection with "ordinary telegraph", which as has been explained, is essentially direct current, but shades into the alternating scale to about 25 cycles. Consequently, it would not be practicable to separate in the telephone office a 20-cycle ringing current from a telegraph current, and a ringing current of this frequency transmitted over composited telephone lines would interfere with the telegraph service. Conversely, telegraph currents would interfere with signaling. For this reason a second frequency, well outside the ordinary telegraph range of frequencies, is used for ringing over the shorter composited circuits. The standard frequency value assigned for this use is 135 cycles.

On long voice-frequency telephone circuits equipped with a number of telephone repeaters, there are advantages to be gained in eliminating the special apparatus which is required at each repeater point to relay the 135-cycle current around the telephone repeater circuit. This is necessary because the repeater set is ordinarily not designed to "pass" the 135-cycle frequency. Of course, the repeater is designed to amplify the band of frequencies representing the actual voice, and any ringing current within this range of frequencies would not only pass through the repeater but would at the same time be amplified. Such an arrangement would not require relaying equipment for the ringing current. This third ringing frequency is 1000 cycles interrupted 20 times per second, the latter feature being necessary in order that conversation over the circuit, which will ordinarily contain some 1000-cycle notes, will not operate the ringer. This type of signaling is also used on carrier telephone circuits for similar reasons.

In concluding the foregoing explanation of the various frequency values and frequency bands in general use, it should be remembered that this description is neither complete nor fundamentally descriptive of the distinction between various alternating-current applications. Though the frequency scale illustrates in one sense characteristic differences in the currents, it should be remembered that other characteristic differences such as energy values might illustrate an equally wide diversification. Furthermore, the use of a single frequency does not mean that no consideration must be given to the nature of other frequencies which may indirectly become involved in the same application. To illustrate, a 60-cycle, 3-phase power line when unbalanced has a residual current flow of 180 cycles, and the same power system with imperfect sine wave form may carry harmonics which give a band of frequencies extending through the voice-frequency range, resulting in serious inductive interference with communication circuits paralleling the power lines. Another case is radio telegraph transmission. While employing a single wave-length, this must have a sufficiently accurate sine wave form so that it will not radiate harmonics which would interfere with radio communications em-

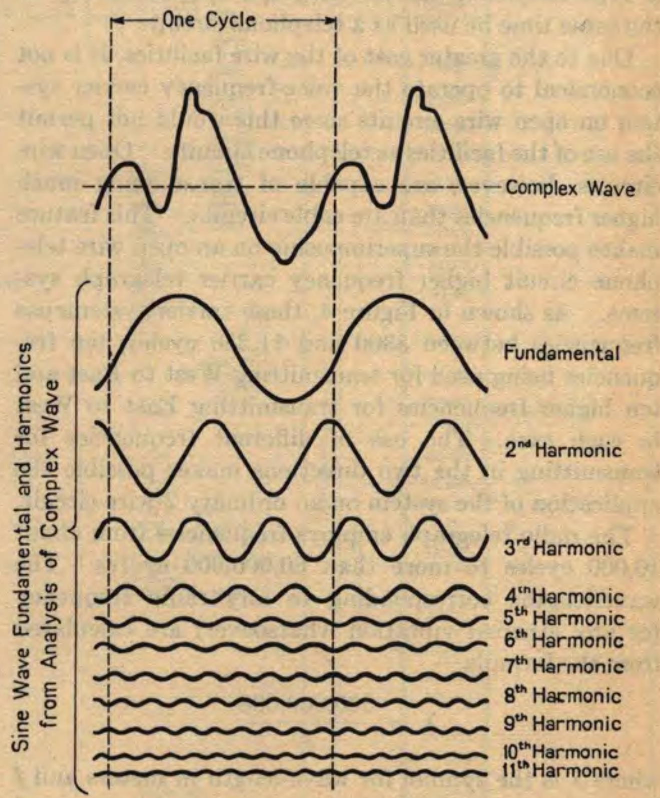


Fig. 5. Analysis of Sound Wave by Fourier's Theorem

ploying other wave-lengths. The application of alternating currents of a single frequency or a single band of frequencies, therefore, may and actually does in the majority of instances, involve the suppression of other frequencies as well as the use of the individual frequency.

3. Wave Analysis

It is known that the characteristic wave forms for vowel sounds are complex waves rather than sine waves of a designated frequency, and that alternating currents which represent the tones of the human voice are seldom, if ever, a simple sine wave. The same may be said for many other cases of alternating-current transmission. We would be hopelessly involved if we should attempt to deal with any current wave shape that might be encountered as distinct and apart from the sine wave, which is the basis of most alternating-current circuit calculations. Fortunately, for all practical cases, **any steady state alternating-current wave form, regardless of how irregular, may be considered as a series of sine wave currents being simultaneously transmitted.** That is to say, any irregular current wave shape can be analyzed or broken up into a series of sine wave shapes. This series can further be restricted to a fundamental frequency and multiples or harmonics of this fundamental frequency. For example, Figure 5 shows an irregular wave so analyzed, and by actually adding the successive and corresponding values of all the sine waves, we can chart the original and irregular wave shape.

The practical treatment of any irregular wave shape for electrical transmission, is then, divided into two steps:

a. **Analyze the wave shape into a fundamental sine wave and its harmonics.**

b. **Deal with each of these component sine waves individually.**

The actual mechanical analysis, such as that shown by Figure 5, is quite difficult, and while the analysis can be made from an "oscillogram", this is for the most part a combination of a laboratory process and involved mathematical calculations. For most purposes, however, we need be concerned with the concept only, rather than with the actual analysis.

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