LOW-FREQUENCY LINES, FILTERS, AND IMPEDANCE-MATCHING DEVICES

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STUDY SCHEDULE NO. 22

For each study step, read the assigned pages first at your usual speed. Reread slowly one or more times. Finish with one quick reading to fix the important facts firmly in your mind, then answer the Lesson Questions for that step. Study each other step in this same way.

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Fundamentals of Power Transfer

THE general object of a radio communication system is to accept speech or music acoustic power, properly convert this to electric power, and via radio waves, transmit it to a distant point where it is reconverted to acoustic power for the listener.

The most important task of the communication system is to make the final acoustic output power a *faithful reproduction* of the original sounds. A secondary requirement is to perform the job cheaply, therefore *efficiently*, which means that the electrical circuits involved must be designed for minimum power loss consistent with high fidelity.

The over-all transmission process is composed of several stages. First, the acoustic power must be converted into electrical power by a microphone. The resultant electrical signal is then amplified and sent over wires to the radio transmitter, which may be anywhere from a few feet to many miles distant. At the transmitter, the audio-frequency power is converted to radio-frequency power, and this in turn is fed to an isolated antenna system by means of a high-frequency transmission line.

At any reception point, all these processes are duplicated in reverse. A receiving antenna reconverts the electromagnetic wave into electric current. This current is then carried by the "lead-in," or transmission line, to the receiver where it is amplified and "demodulated." Finally, the resulting audio power is further amplified and converted, by means of a loudspeaker, into acoustic power as a reproduction of the original speech or music.

In each of these stages of conversion of power from one form to another, in every amplifier along the line, and in every instance where power is transferred from one point to another, there is a possibility of signal distortion and loss of fidelity. Also, unless proper precautions are taken, inefficient coupling systems may waste and throw away the power which costs so much to obtain by amplification.

A very important subject of radio design is the study of proper interconnection between the various devices or stages, because improper matching of the output characteristics of one device with the input characteristics of the next is a source of both distortion and inefficiency.

In the complete communication system, wires, or lines, carry both audio-frequency and radio-frequency power. A low-frequency line is used to connect the microphone to the pre-amplifier, and another is used from the amplifier output to the transmitter itself. On the other hand, a high-frequency line carries power from the transmitter to the radiating antenna.

Low- and high-frequency lines are alike in that both are used to transfer power from a source, such as an amplifier or oscillator to some load such as a loudspeaker, or an antenna. ► The subject of radio-frequency lines and their own peculiar characteristics will be discussed in a later Lesson. In this Lesson we shall study the low-frequency line as a means of transferring audio-frequency power; and most important, we shall determine methods of impedance matching between any source and load, so that maximum fidelity and power transfer may be realized.

But first, let us examine some fundamentals which apply to all networks.

AUDIO FREQUENCY POWER AND IMPEDANCE

The knowledge of the amount of voltage fed into one end of a device and the amount coming out the other end is not sufficient to tell us what would happen if we changed either the source or the load, leaving the same device between them.

Our study of electrical circuits has taught us that knowledge of the voltage alone does not tell us what is in the circuit. We need to know *two* quantities to determine what is going on.

Remembering more of Ohm's Law, we realize that the two necessary quantities do not have to be current and voltage. For one of the quantities we can choose the *product* of voltage and current, which is *power*. For the other, we can take the *ratio* of voltage to current, which in a.c. circuits, is *impedance*. Hence, a knowledge of both power and impedance will suffice just as well.

In dealing with the transfer of power from a source to a load, we practically always have a loss, so from the standpoint of efficiency, we are vitally interested in the fraction of power that is transferred. In fact, this is frequently more important than the power levels, providing the equipment is properly designed to handle this power.

The use of simple fractions for determining the transmission of power, however, is not convenient. If we feed power to a device that transmits ¾ of it, and this output power is fed into another device that transmits ½ of its input, the final output power is ¾ x ½, or ⅜ of the original input. If more devices or networks are added, we must continue to multiply the fractions representing the transmission of each. This can be extremely awkward.

Use of Logarithms. You will recall that the use of logarithms reduces problems in multiplication to problems in addition. The logarithm of the product of any set of numbers is the sum of the logarithms of the individual numbers. Hence, if we express the transmission of a network, not by the fraction of the power transmitted, but by the logarithm of this fraction, the effect of transmitting power through a succession of devices can be found by *adding* the logarithms appropriate to each device.

However, if we take the ratio of power output to power input, we will get a negative logarithm whenever we have a loss, because the output will then be less than the input. When dealing with losses, as we will be in this Lesson, it is standard practice to invert the ratio so as to obtain a more convenient positive number. Now we find the logarithm of the ratio of input power to output power, which is called the *attenuation* when losses are being dealt with.

The fundamental unit of attenuation is called the "bel." This, however, is a large unit, and results in inconveniently small numbers. For practical work, the bel is divided by 10, giving a new unit, the *decibel,* which makes the number representing a given attenuation ten times as large.

As a matter of definition, then, *the attenuation in decibels (db) is equal to ten times the logarithm of the ratio of input to output power.*

Expressed in equation form, the attenuation:

$$\text{db} = 10 \times \text{logarithm} \ \frac{\text{Power Input}}{\text{Power Output}}$$

or simply:

$$\text{db} = 10 \log \frac{P_1}{P_o} \ *$$

Familiarity with the use of db has led us to drop the long phrase, "The attenuation of such-and-such a network is so many db," and say simply, "The network has so many db *loss.*" If the network contains an amplifier so that the output is greater than the input, we take the ratio of the power output to the power input, and say it has so many db *gain.*

Acoustic Loudness. The use of decibels (logarithms) to express the loss or gain of a system has more justification than arithmetic convenience. The response of the human ear, for example, is such that equal db increases in acoustic power result in equal increases of apparent loudness. A tone, successively produced at relative power levels of 1, 2, 4, 8, or 16, etc., times the original power level, produces equal steps up the ladder of loudness. These figures represent equal steps of 3 db increase in power.

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*Since $P_1 = E_1^2/R_1$ and $P_o = E_o^2/R_o$ we find that, if R_1 equals R_o, then $P_1/P_o = E_1^2/E_o^2$. Thus, if we want to use a voltage ratio rather than a power ratio, for equal resistance, this formula becomes

$$\text{db} = 10 \log E_1^2/E_o^2$$
$$\text{db} = 20 \log E_1/E_o$$

In a like manner we can show that

$$\text{db} = 20 \log I_1/I_o$$

Under average sound level conditions, a change of about 2 db is the least that can be noticed by the ear. Steps of 1 db in a pure tone can be detected in the absence of background noise, but the complex sounds of speech and music require a 3 db change in level before the difference in loudness is noticed.

LOAD MATCHING

An ideal power source may be represented by a *constant-voltage* generator having a fixed internal *series resistance* (or impedance). Fig. 1 shows this representation of a 30-volt battery with an internal resistance of 2 ohms. The constant voltage is the open-circuit voltage, or the voltage appearing at the terminals without a load.

Conditions for Maximum Power. If we now consider that a load resistance of R ohms is connected across the output terminals as in Fig. 1, by Ohm's

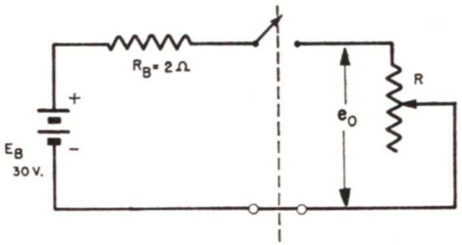


FIG. 1. A 30-volt battery with an internal resistance of 2 ohms, connected to a variable load. The current can be found by dividing the battery voltage by the total resistance (R_B + R).

Law we would find that the output voltage e_o is equal to the current multiplied by the load resistance R.

► Since the power delivered to the load resistance R is given by e_o^2/R, we can plot a curve of power output versus load resistance. This curve is shown in Fig. 2. Notice that, *in this example,* the maximum amount of power is delivered to the load when the load resistance is 2 ohms. This is important. The curve demonstrates that *for maximum power transfer to a load, the load resistance should be exactly equal to the source or generator internal resistance.*

The curve in Fig. 2 is not symmetrical about the peak point. We can make

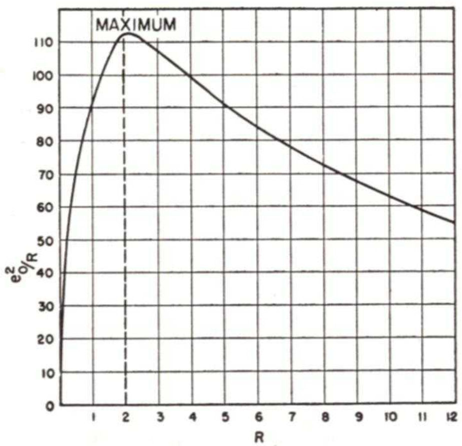


FIG. 2. How power into a load varies with the load resistance.

it symmetrical and thus easier to read for small values of load resistance if we plot the power, not against the load resistance itself, but against the ratio of the load to internal resistance of the source as shown in Fig. 3.

We can further generalize this transmission curve by plotting at each point, not the actual power, but the relative power drop in db below the maximum available power. This curve is shown in Fig. 4.

Fig. 4 shows how important resistance matching can be. For example, if we use no matching devices and connect a 50-ohm microphone directly to an amplifier with an input resistance of 400 ohms, we can expect a loss of 4 db. This is determined by noting that the ratio of the amplifier input resistance R_L to the microphone resistance R_G is equal to 8, and the curve crosses this value at the 4 db loss point.

To carry the illustration further, suppose the amplifier in the example just given has an *output* impedance of 500 ohms, and we have connected it to a 100-ohm line to feed a transmitter modulator. The ratio of source impedance to load impedance (or load to source, whichever is greater) happens to be 5. From the curve we see this impedance ratio corresponds to a loss of 2.5 db.

Our over-all mismatching losses amount to the sum of the individual losses, or simply 4 plus 2.5, which is 6.5 db. This is equivalent to a power

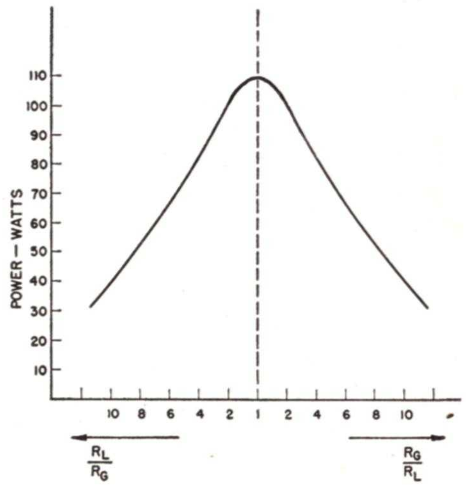


FIG. 3. Variation of the power in the load with the source-load resistance ratio.

loss ratio of over 4 to 1. *By mismatching we have reduced our power to ¼ of what it could be!*

Transfer of Maximum Voltage. There are some devices, however, with which impedance matching is not desirable. One such example is the grid circuit of a class A amplifier. The driving power required is negligible, for the grid input resistance is very high. When the class A stage is driven by a

preceding amplifier, there is no point in matching for maximum power, since the first stage can deliver much more power than the second requires for full output. The two stages simply are

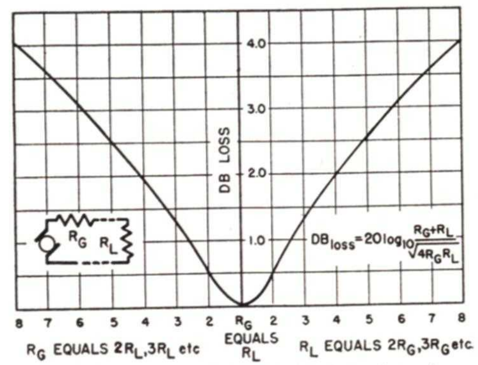


FIG. 4. The curve of Fig. 3, replotted to show the power drop in db below the maximum available.

coupled in such a manner as to present the appropriate *voltage* to the second. In fact, it is driving *voltage,* not power, that is important in this case.

The exact manner in which output voltage and output power each vary with load impedance in the circuit of Fig. 1 is illustrated in Fig. 5. The curves shown, however, do not apply to this circuit alone. The power and voltage curves will have the same shapes for any constant-voltage, fixed-resistance source. It is apparent from Fig. 5 that both maximum power and maximum voltage cannot be realized with the same load.

Fidelity Requirements. A second condition in which a mismatch is sought intentionally is associated with high fidelity. If the fidelity of a device changes with loading, it is not the maximum power output that is desired but the greatest output *consistent with low distortion.* We must therefore match for maximum undistorted power output.

A triode amplifier normally is operated into a load resistance of twice its internal plate resistance. From Fig. 4 we see that this means there would be a loss of ½ db. Although this condition of operation represents a specific mismatch, some power must be sacri-

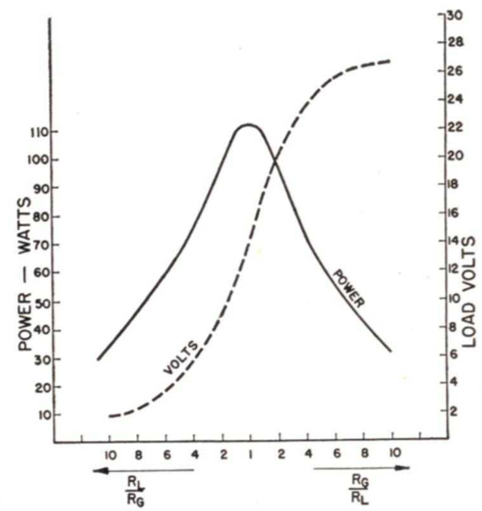


FIG. 5. Comparison of load power and load voltage for various source-load resistance ratios.

ficed to keep distortion within tolerable limits. The basic reason for such behavior is that the triode is not accurately represented by the simple equivalent constant-voltage generator with fixed internal resistance. Actually, the tube's internal resistance varies somewhat with signal level.

Transformers for Load Matching

How is impedance matching obtained in practice? An incandescent lamp can be designed to match a given battery resistance by proper choice of filament diameter and length. It is neither practical nor desirable to design all pieces of communication equipment to one standard resistance value, so it becomes necessary to use some means of changing the apparent resistance of a source or a load. This is accomplished by the use of transformers.

Transformer Turns Ratio. You will recall that if the secondary winding of a transformer has N times as many turns as the primary, it steps up the voltage N times. In other words, the voltage ratio is the same as the turns ratio. Let us look at Fig. 6. Here, the primary voltage is represented as e_p, and the secondary voltage as e_s. Since the turns ratio is N, actually e_s is equal to N times e_p.

In practice, transformers are designed to be highly efficient so that their losses are negligible compared to the input and output power. We can assume, then, that the secondary output power is equal to the primary input power.

The input power for the primary is merely the product of the voltage e_p times the current i_p, or simply:

$$P\ (\text{pri}) = e_p \times i_p$$

Likewise, for the secondary, the output power is the output voltage e_s multiplied by the output current i_s, or:

$$P\ (\text{sec}) = e_s \times i_s$$

Now we already know that the output voltage e_s is equal to the primary voltage *multiplied* by the turns ratio N. If this is so, then the secondary current i_s must be equal to the primary current i_p *divided* by N—otherwise, the power in the primary and secondary windings would not be the same.

So we can say:

1. The secondary *voltage* e_s is equal to the primary voltage e_p *multiplied* by N, or:

$$e_s = N \times e_p \qquad (1)$$

2. The secondary *current* i_s is equal to the primary current i_p *divided* by N, or:

$$i_s = i_p/N \qquad (2)$$

Reflected Impedance. The question of impedance now comes up. Just what is the impedance seen looking back into the secondary terminals 3 and 4 in Fig. 6?

By Ohm's Law we know the impedance to be the voltage divided by the current. We can write this:

$$Z_s = e_s/i_s \qquad (3)$$

We can substitute in equation (3) the equivalents for e_s and i_s from equations (1) and (2). Doing so, we get for the secondary impedance:

$$Z_s = e_s/i_s = \frac{N \times e_p}{i_p/N} = N^2 \times e_p/i_p \qquad (4)$$

The "e_p/i_p" of this last equation, however, is really the primary impedance Z_p. So we find the secondary impedance equation simplifies to:

$$Z_s = N^2 \times Z_p \qquad (5)$$

► This is one of the most important characteristics of transformers. Expressed in words, equation (5) means that *looking into the secondary terminals, the impedance is equal to the primary source impedance multiplied by the square of the turns ratio.*

We can, of course, rearrange equation (5) to read:

$$Z_p = Z_s/N^2 \qquad (6)$$

which means *looking into the primary, the impedance is equal to the secondary load impedance divided by the square of the turns ratio*—which is another way of saying the same thing.

Let us take a practical example to investigate further this impedance-changing characteristic of a transformer. Suppose we have a step-up transformer with a turns ratio N of 10, and we have placed a load resistor of 10,000 ohms across the secondary. What is the impedance looking into the primary?

We have then: $N = 10$; $Z_s = 10{,}000$ and by our equation (6):

$$Z_p = Z_s/N^2 = \frac{10{,}000}{10 \times 10} = 100 \text{ ohms}$$

In other words, *the 10,000-ohm secondary load resistance has been made to appear in the primary as only 100 ohms!*

To turn the problem around, let us imagine we use the same transformer with turns ratio of 10, but that we load the primary with a 100-ohm resistor. What impedance will we find looking back into the secondary?

We now have: $N = 10$; $Z_p = 100$ and equation (5) shows:

$$Z_s = N^2 \times Z_p = 10 \times 10 \times 100 = 10{,}000 \text{ ohms.}$$

The action of the transformer has made the primary impedance of 100 ohms appear as 10,000 ohms in the secondary!

It is obvious that this transformer with a turns ratio of 10 could be used nicely to match a 100-ohm source or generator to a 10,000-ohm load. By proper choice of turns ratio, transformers can be used to match impedances over a very wide range.

For using our basic transformer equation (5) in actual problems, we can put it into the following convenient form by making a few simple algebraic changes.

$$N = \sqrt{Z_s/Z_p} \qquad (7)$$

This last equation form simply says: *for a given impedance match, the transformer turns ratio needed is equal to the square root of the secondary to primary impedance ratio.*

As an example of the use of this statement, if we have a load resistance 9 times that of a source, then we must use a transformer with a turns ratio $N = \sqrt{9}$ or 3.

Should the load resistance be smaller than the source resistance, then of course, a step-down transformer would be required.

Transformers in Practice. It is not usually necessary to fabricate a special transformer for each impedance-matching case. Suppose we wish to connect a 50-ohm microphone to a 500-ohm load. This is a 10-to-1 impedance ratio, and without a transformer would result in a 4.8 db loss. An ideal transformer would have a turns ratio

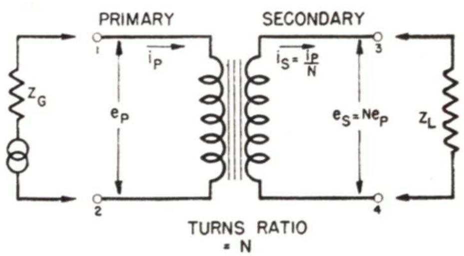


FIG. 6. The equivalent circuit of a transformer working between a generator with an impedance Z_G and a load with an impedance Z_L.

of $N = \sqrt{10}$, or 3.16. We could use a standard 3-to-1 transformer, however, thus making the 500-ohm load look like $500/3^2 = 55.6$ ohms, instead of exactly 50 ohms. According to the curve in Fig. 4, this would result in negligible loss. In fact, a 4-to-1 transformer would make the microphone see 31.26 ohms and still give a net loss of only ¼ db!

Just how much the turns ratio can be varied without serious loss is illustrated in Fig. 7, where the insertion loss of a transformer is shown for various ratios of the actual turns ratio N_A to the desired turns ratio N_D.

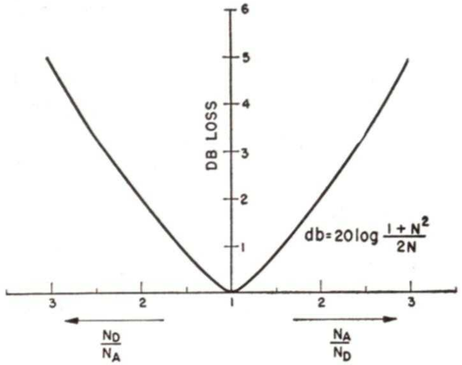


FIG. 7. Curve showing loss from use of improper turns ratio N_A instead of ideal ratio N_D.

Suppose we have the problem of connecting four 500-ohm sources in series and feeding a 500-ohm line as load. Obviously, the impedance ratio is 4, and we need a step-down transformer. The ideal transformer, therefore, would have a turns ratio of $N_D = \sqrt{4} = 2$.

If we should try a transformer with a turns ratio $N_A = 1.5$, we would then have the ratio of desired to actual turns ratio: $N_D/N_A = 2/1.5 = 1.3$. The curve in Fig. 7 shows that the mismatch would result in a loss of less than ½ db.

Had we endeavored to substitute a transformer with turns ratio $N_A = 2.5$, then we would have had the ratio $N_A/N_D = 1.25$, which still corresponds to a loss of less than ½ db.

OTHER PROPERTIES OF TRANSFORMERS

Up to this point, the turns ratio is the only property of a transformer we have discussed. There are, however, several other characteristics which must be considered in the use of transformers in communication systems.

Frequency Response. Most communication-type transformers, such as amplifier interstage, microphone-to-amplifier, and amplifier-to-line are intended to be used over a frequency spectrum of several octaves. How well a transformer operates over such a wide range depends upon the design and construction.

Let us look at the response curve for a typical 3-to-1 interstage transformer in Fig. 8. The secondary voltage is relatively constant from about 100 cycles to approximately 5000 cycles, then rises to a slight peak at 8000 cycles, and then drops quickly at still higher frequencies. Below 100 cycles, the output also decreases at a rapid rate. These sudden dips in response are caused by inherent design limitations in the transformer itself.

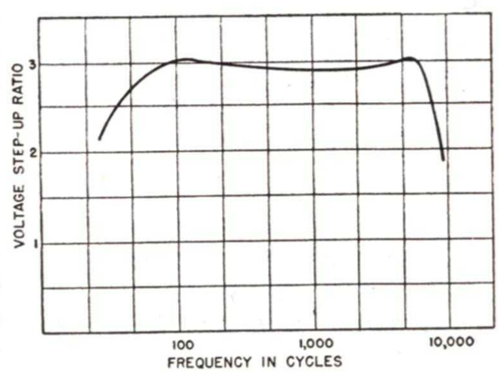


FIG. 8. This curve shows the variation of output voltage with frequency for a typical 3-1 interstage audio transformer.

In Fig. 9 is shown the equivalent circuit of such an interstage transformer as it would operate between a triode driver and the grid of a following amplifier.

The triode driver output voltage is represented as e_g and its plate resistance as r_p. The transformer primary winding is L_1; the secondary is L_2. The inevitable leakage inductance which is not completely coupled by the magnetic flux is indicated as L_3. Capacitor C represents the distributed capacity between the closely-spaced wires of each winding.

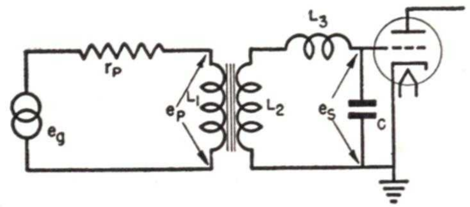


FIG. 9. The equivalent circuit of a triode amplifier, transformer coupled to the grid of a following amplifier stage.

The undesirable "extra" reactances of the distributed capacity C and leakage inductance L_3 actually form a series-resonant circuit shunted across the transformer secondary. Their effect is to build up a peak grid voltage e_s at their natural resonant frequency near 8000 cycles. Above this resonance, however, L_3 chokes back more and more current while the by-pass action of C is increased. This divider action results in a very sharp drop in the voltage e_s actually applied to the amplifier grid. This effect is quite apparent in the curve of Fig. 8.

► We see then, that *leakage inductance and distributed capacity together determine the high-frequency response of a transformer.*

At low frequencies, we find another factor becoming important. The primary inductance L_1 is effectively in series with the generator (triode plate) resistance r_p. At medium frequencies, the inductive reactance of L_1 is quite large, so that most of the generator voltage is developed across the primary, and very little is lost across the generator resistance. As the frequency is reduced, however, the reactance of L_1 becomes lower and lower, and finally, almost all of the generator voltage is wasted across the generator resistance r_p instead of being applied to the transformer.

► This is a characteristic trait. *The response of any transformer will decrease as soon as the primary reactance falls to values comparable to the generator resistance.*

In a properly designed transformer, enough wire and iron is used to make the inductive reactance of each winding at least twenty times as great as the load impedances with which the transformer is designed to work. If these conditions are not met, then the transformer cannot perform properly as an impedance-matching device, and the wide-range frequency response may be seriously diminished.

It is for these reasons that transformers should never be used with impedances higher than those recommended; and why transformers are designed and sold on an ohm-to-ohm rather than a turns ratio basis.

Effects of Direct Current. When *high-fidelity* transformers are used with single-ended stages, it is almost imperative that no d.c. be allowed to flow through the windings, otherwise, core saturation will produce distortion. (In push-pull operation this presents no problem.) Fig. 10 shows how the d.c. flow through the transformer primary can be eliminated. The plate choke through which the plate current flows is L; the customary blocking condenser is C. Only the audio-frequency components flow through the transformer, which still serves to match the tube impedance to a low-impedance loudspeaker or line.

COMMERCIAL TRANSFORMERS

Commercial transformers are often made with both the primary and secondary windings divided into two equal sections so that eight terminals may

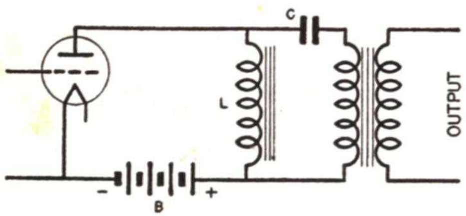


FIG. 10. How impedance coupling is used to keep direct current from flowing in a transformer winding, thereby preventing saturation and preserving the low-frequency response.

be brought out, four on the primary, and four on the secondary as in Fig. 11. This is done so that the same transformer can be used for matching a number of different impedances.

Suppose each section of the primary has sufficient inductance to operate properly with a 500-ohm source. Now if we connect sections 1 and 2 in series, the total primary then will operate with a 2000-ohm source. This is true because the impedance varies as the *square* of the turns, and since we have *doubled* the primary turns by using both halves in series, the source impedance can be multiplied by a factor of *four*.

If we connect the primary windings in *parallel* instead of in series, the proper source impedance will not be changed but will remain 500 ohms. The

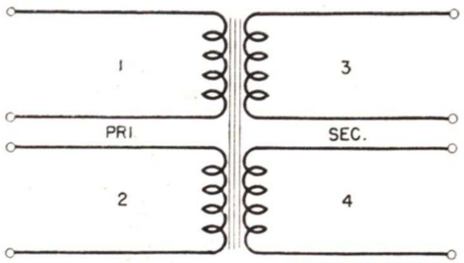


FIG. 11. A commercial transformer arrangement which permits the matching of a number of different impedances.

only result of a parallel connection is a reduction of the effective resistance of the primary.

The same reasoning applies to the secondary or load side of the transformer. If each section of the secondary of Fig. 11 is designed to work into a 125-ohm load, then by connecting them in series, thereby doubling the effective turns, we will enable the total secondary to work into a 4 × 125 = 500-ohm load.

The transformer arrangement of Fig. 11 provides a variety of source-to-load impedances. The exact combinations available, and the necessary winding connections are shown in Table 1.

It should be clearly understood that, although this transformer is made to

TABLE 1

Load Impedance Pri.	Load Impedance Sec.	Primary terminal connections	Secondary terminal connections
500	125	1 parallel 2	3 parallel 4
500	500	"	3 series 4
2000	125	1 series 2	3 parallel 4
2000	500	"	3 series 4

work from 500 to 125 ohms (as one example from the table), it can also be used to work from 125 to 500 ohms. This is accomplished, obviously, by simply interchanging primary and secondary connections to the external circuit.

Tapped Transformers. Quite often when a source feeds a load that may change impedance from time to time—such as a power amplifier feeding first one, then two, then three banks of loudspeakers in parallel—a tapped secondary is used for impedance matching. See Fig. 12. Naturally, fewer turns are used on the secondary as the load impedance is decreased.

Feeding Grid Circuits. As mentioned earlier, class A grid circuits require negligible driving power, and it is the highest *voltage*, not power, that is important. For this reason, grid-driving transformers ordinarily are made with the turns ratio as high as possible without too great a sacrifice of high-frequency response in the increased dis-

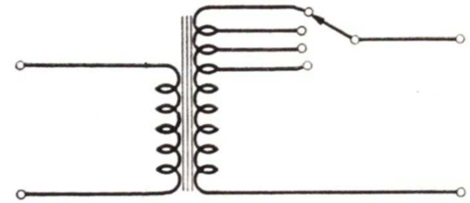


FIG. 12. A tapped transformer can be used in cases where the load impedance is changed from time to time.

tributed capacity. Transformers in the high-quality class usually have step-up ratios of 2-1 to 5-1 for a 10,000-ohm plate to a grid; 5-1 to 15-1 for a 500-ohm line to a grid, and 10-1 to 25-1 for a 200-ohm microphone to a grid.

Transmission Lines

So far we have discussed the problems of delivering power from a source to a load and found that the impedances must be matched for minimum transmission power loss. We have discovered, too, the advantages of the transformer as a theoretically "lossless" impedance-changing device. But nothing has been said about the wires that are used to make the connections.

If the connecting leads are sufficiently short, then nothing need be said except that they are wires. When, however, we begin to speak of connecting a studio microphone in one city to a transmitter in another city, or a transmitter in a building to an antenna out in a field nearby, we must look into the details of the transmission of energy along the wires.

WHAT IS A TRANSMISSION LINE?

Any two-conductor connector between pieces of equipment will be called generally a "transmission line." Transmission lines may have several different forms.

The Open-Wire Line. The most elementary line is a pair of parallel wires strung several inches apart. The "open-wire" line is used extensively for power transmission where insulation is a problem. Such a line, however, usually is not satisfactory for communication purposes, as the separation between the wires leads to excessive pickup of stray noise, 60-cycle hum, and "cross-talk" (or interference) from a neighboring pair.

The Twisted Pair. Stray pickup is reduced by decreasing the wire spacing to the minimum allowed by the insulation, and twisting the two wires together to form a "twisted pair." Close spacing and twisting tends to produce cancellation of pickup in one wire by equal and opposite pickup in the other.

► Notice that the open wire and twisted pair are "balanced" transmission lines. That is, neither of the two wires connect to ground, therefore, the capacity and leakage from both wires to ground will be the same. For this reason, as we will see shortly, we need to use transformers that are balanced to ground to couple into and out of these lines. We do this by connecting to ground the center tap of the transformer winding which is connected to the line.

The Coaxial Line. A still quieter line can be made by shielding. If one wire is grounded (hence, an unbalanced line), then the shield itself can be one conductor, and only one wire need be run inside the shield. This arrangement of a cylindrical shield conductor with another conductor running down its center is called a "coaxial" line.

CHARACTERISTICS OF A TRANSMISSION LINE

The longer the line is extended, and the higher the signal frequency becomes, the more important is the role played by the electrical properties of a transmission line in the efficiency and fidelity of the transmission.

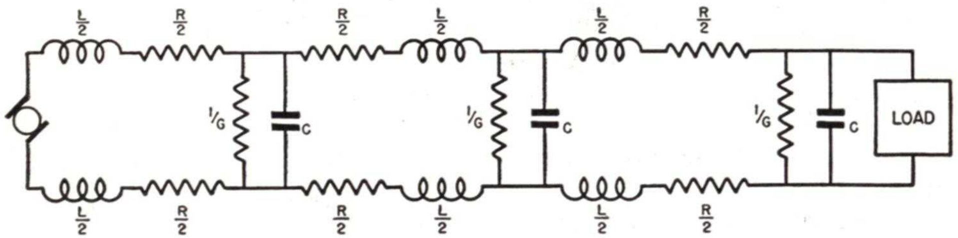


FIG. 13. The "lumped parameter" equivalent circuit of a transmission line.

In audio-frequency transmission, it is mainly a question of transmitting all signal frequencies with the *same* attenuation, and with the least possible loss. We will discuss some of these problems here.

Radio-frequency lines and lines carrying video signals are more complex, and involve such factors as frequency-phase shift and signal velocity. These subjects will be reserved for a later Lesson.

Approximate Equivalent Line. Since there is a magnetic field around any current-carrying conductor, it is reasonable to expect the wires in a transmission line to have some self-inductance; and, of course, all wires have some resistance. Furthermore, since the wires are closely spaced, there is considerable capacity between the wires of a line. Also, as no insulation is perfect, there is some leakage resistance or conductance between the wires.

All of these parameters*—inductance, resistance, capacity, and conductance—are not "lumped" in one place; instead, they are distributed more or less evenly throughout the entire length of the line. These are commonly called "distributed constants."

*Any of the factors or variables that determine the behavior of a device, a circuit, or a mathematical equation are commonly referred to as the parameters of that device, circuit, or equation. In discussions of a radio circuit, the word "parameter" is generally used in reference to the inductance, capacitance, and resistance of the circuit.

The calculation of the performance of a long line with distributed constants requires complex mathematics. For our purpose, however, we can make the problem easier by using what is called a "lumped parameter" approximate equivalent of a transmission line. Such an equivalent is made by theoretically dividing the line into a great number of elementary sections, and placing small inductances, capacities, and resistances in each section. A lumped parameter approximation is shown in Fig. 13. In this case, the line is divided into three sections. As the number of sections is increased, the approximation becomes more nearly the same as the actual "distributed constants" of the line. However, reasonably accurate results can be obtained with only three or four sections.

Since each wire has some resistance, one-half of the total (loop) resistance is placed in each side of the line. This resistance is represented by the sum of the resistors R/2 in the individual sections. Similarly, the inductance associated with each wire is represented by the small inductances L/2. Capacity between the wires is pictured as the small capacitors C, and the leakage resistance (conductance) is shown as the resistors 1/G.

This representation of a transmission line is approximately true at all frequencies of the spectrum.

Simplified Equivalent Lines. For audio-frequency work, we find we can simplify the line still further. For one thing, the effect of the inductance at low frequencies is usually quite small compared to the ohmic resistance of the line. With good insulation, the leakage resistance, also, is so high that its effect can be neglected.

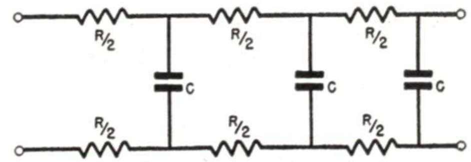


FIG. 14. The line in Fig. 13, for audio-frequency work, can be reduced to this simplified resistance-capacitance network.

► The complicated line of Fig. 13 then, can be reduced to the approximation in Fig. 14. This is a good representation of audio lines over 1000 feet long.

Short Lines. If we go still further and consider audio lines shorter than 1000 feet, we find we can neglect the capacity between the wires without too much error. For extremely short audio lines then, we reduce the approximation to that of Fig. 15A, which can be drawn simply with one-half the total loop resistance R/2 in each wire as in Fig. 15B.

Now let us study this approximation to a short audio line and see how it will perform in delivering power.

BEHAVIOR OF SHORT LINES

Suppose we have a power amplifier and we wish to locate a public-address speaker a few hundred feet away. The

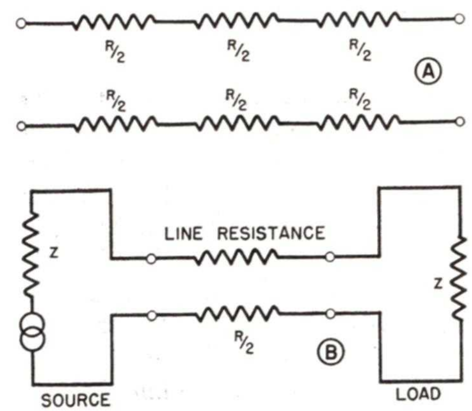


FIG. 15. For very short lines, the capacity may be neglected as shown at A. This is equivalent to considering the line loop resistance as the only parameter between the source and load, as shown at B.

line connecting the two would perform like the approximation in Fig. 15B.

Ordinarily the amplifier impedance Z is matched and equal to the speaker (or load) impedance Z. It is apparent, however, that the total load on the source with the line interposed is not

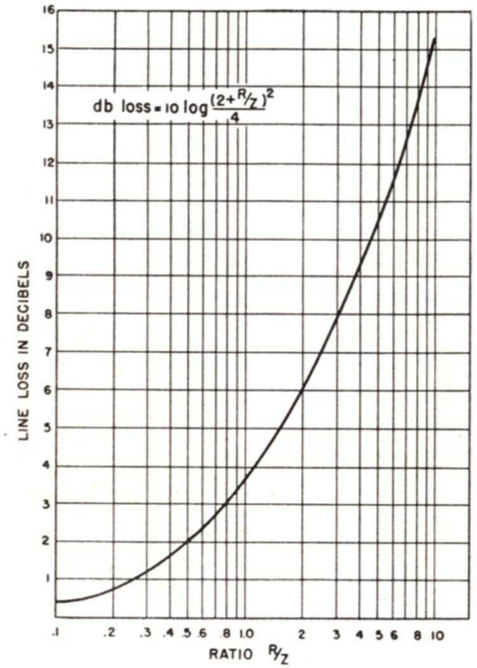


FIG. 16. Chart showing how line loss in db is determined by the ratio of line resistance R to the source-load impedance Z.

the load impedance alone, but the line resistance as well. In other words, the total load on the amplifier is now R + Z. This results in a loss of power due to the mismatch between the source impedance Z and its total load R + Z, as well as a further loss arising from the fact that part of the power delivered is wasted in the line resistance R.

Not only is the source impedance Z mismatched because it sees a load R + Z, but the load, also, sees a total impedance of R + Z, and it, too, is mismatched. Theoretically, this can lead to distortion and considerable loss of power.

► It is obvious that if the line resistance R is kept very small compared to the source and load impedance Z, then the mismatch and consequent power loss will be reduced to acceptable limits.

It is important, therefore, to use a load impedance that is high compared to the resistance of the line.

Determining Losses. For convenience in determining the loss in lines with various values of load impedance, the curve in Fig. 16 is given. Here, the total line loss in db is plotted against the ratio of line resistance R to the common source and load impedance Z.

To use this curve, it is necessary to know the total (loop) resistance of the line under consideration. The resistance per 1000 feet of line (2000 feet of wire) for various common sizes of copper wire is given in Table 2.

TABLE 2

Wire Size B & S Gauge	R—ohms per 1000 loop-feet
10	2.0
12	3.2
14	5.2
16	8.2
18	13.0
20	20.7
22	33.0
24	52.4

As an example of the use of the wire table and the curve of Fig. 16, let us find the loss due to a 200-ft. line of #24 wire, connecting a 50-ohm source to a 50-ohm load.

From the wire table, we find a 1000-ft. line of #24 copper wire has a total resistance R of 52.4 ohms. A 200-ft. line is 1/5 as long, so it has a total resistance R of 10.48 ohms. Since we are using a source-load impedance Z of 50 ohms, we have an R/Z ratio of

10.48/50 = 0.209. Inspection of the curve in Fig. 16 shows that an R/Z ratio of 0.209 corresponds to a loss of approximately 0.8 db.

As another illustration, let us say we wish to connect an amplifier to a 15-ohm speaker 250 feet away. What is the smallest wire that can be used for the line without making the loss exceed 2 db?

From the loss curve we find for a loss of 2 db, the ratio R/Z = 0.5.

As we are using an impedance Z = 15 ohms, we see that the line resistance R cannot exceed:

$R = Z \times 0.5 = 15 \times 0.5 = 7.5$ ohms.

Now, a resistance of 7.5 ohms for 250 feet of line means a limit of 30 ohms per 1000 loop-feet. From the wire table, we see that #20 wire is the smallest that can be used.

With short audio-frequency lines, if the line resistance is low compared to the source-load impedance—and it must be for efficient transmission—then the source sees a load that is essentially that of the original load. Hence, if we have a line which terminates at the far end in a 500-ohm load, we say we wish to couple into a "500-ohm line," and an amplifier feeding such a line would need a plate-to-500-ohm output-coupling transformer.

For another case, if a signal from a 50-ohm microphone comes into the studio on a line, we say that a 50-ohm line must be coupled to the mixer.

By using transformers at each end, a line can be operated at any desired impedance level. This is sometimes an advantage, because a higher source-load impedance Z will allow the use of smaller wire, or what is more important, will result in less loss for wire of a given size.

► Line impedance, however, should not be made too high. If the impedance is too great, then noise pickup and cross-talk from other lines may become objectionable. Worst of all, with very high-impedance source and load, the capacity of the line may no longer be neglected. Since the capacity may pass as much current as the high-impedance load, serious losses of the higher frequencies may result.

All these factors mean that the impedance level decided upon for short line operation must be a compromise. In practice, lines are operated with impedances from 50 to 200 ohms for low-level signals, and from 200 to 500 ohms for high-level signals. The higher impedance for higher power is permissible because noise pickup and cross-talk are less objectionable, and reducing the current increases the efficiency by reducing the I^2R loss.

Long Transmission Lines

Capacity Effects. As we attempt to use longer and longer lines for the transmission of audio-frequency currents, we find the inherent capacity of the line itself becomes more and more important. In lines extending more than about one mile, the capacity cannot be neglected. That is, the simple line approximation in Fig. 15A is no longer sufficiently true, and we must go back to the circuit in Fig. 14.

► Now let us look at Fig. 14 a moment. Certainly, the three sections made up of resistance and capacity resemble a typical R-C low-pass filter. We should expect then, as the frequency is in-

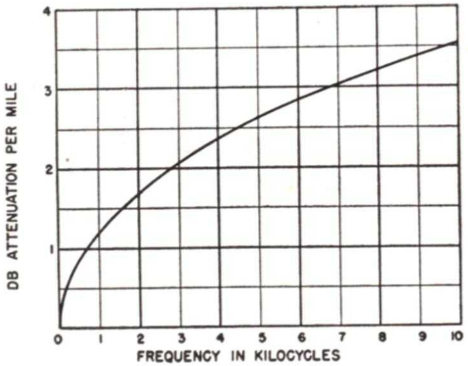


FIG. 17. Illustrating the high-frequency attenuation in one mile of ordinary telephone line.

creased, and the reactances of the shunt capacities C correspondingly decreased, that there will be more voltage drop across the series resistor R/2 and thus the voltage across the load at the end of the line would become smaller and smaller. This is exactly what happens. For indeed, *the long transmission line behaves exactly like a resistance-capacitance low-pass filter!*

When one mile of standard telephone cable is properly terminated at both ends, its attenuation in db varies with frequency in the manner shown in Fig. 17. A line 2, 5, or 20 miles long would have an attenuation 2, 5, or 20 times that shown in the figure.

In standard broadcast a.m. transmitters, all frequencies from about 50 to 8000 cycles are generally transmitted. In a high-fidelity line connecting the studios with the transmitters, these frequencies must not have a variation in attenuation of more than 2 db.

A transmission line with a characteristic like Fig. 17 is not a high-fidelity line. To make it so, some means of improving the high audio-frequency response must be found.

Loading of Lines. It has been found that when special inductance coils are inserted in series with the line at a number of separate points, not only is the high audio response considerably improved, but also, the over-all attenuation of the line is substantially reduced at the same time!

This insertion of extra inductance in a line is called "line loading," and the coils themselves are called "loading coils."

The exact reasons why loading improves performance of a line can be outlined only by complex mathematics. We can visualize what happens, however, if we go back to our approximate equivalent line in Fig. 14. Suppose that in each fictitious section of the line we place an inductance in series with the resistors. We then have the network shown in Fig. 18.

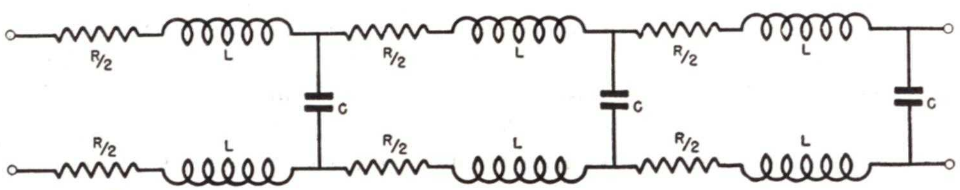


FIG. 18. The equi· ·nt circuit of a loaded transmission line, obtained by a·ding loadi·. coils L to the network in Fig. 14.

If we make the further assumption that the inductive reactance of the loading coils is very high compared to the ohmic resistance of the line, then we can neglect the resistances and redraw our circuit without them. Finally, as the approximate equivalent of a *loaded* transmission line, we have the network of Fig. 19.

► But look at Fig. 19 again. It is a network of series inductances and shunt capacities. Indeed, similar circuits are used to eliminate the a.c. ripple from rectified power supplies. *By the addition of loading coils, the transmission line has been made to resemble an inductance-capacitance low-pass filter!*

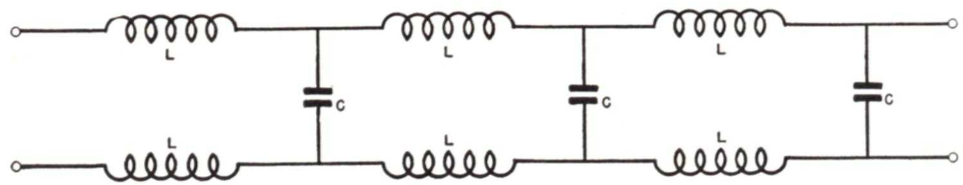


FIG. 19. If the ohmic resistance is negligible, then the equivalent of a loaded line is a low-pass filter.

It is a characteristic of low-pass filters, when properly terminated at source and load ends, to transmit with very little attenuation all frequencies up to a certain critical point. Above this point, usually called the "cut-off frequency," the attenuation suddenly begins to rise. The response of a loaded transmission line looks very much like Fig. 20.

Compare the transmission of a loaded line in Fig. 20 to that of an unloaded line in Fig. 17. *Below the cut-off frequency,* the loaded line actually has become a high-fidelity line. In addition, the transmission losses within this "pass-band" have been very drastically reduced.

Of course, the losses *above* the cut-off frequency are even worse in a loaded line than they were without loading. In practice, however, it is found that the smaller the loading coils happen to be, and the greater the number of coils used along the line, the higher the cut-off frequency becomes. This means that the fidelity of a line can be increased to any degree, and improvement is limited only by the cost of the increasing number of loading coils needed for higher fidelity.

Telephone companies always load their long lines such as city-to-city and leased-broadcast wires. For chain-broadcasting service, radio stations may lease telephone lines with any degree of fidelity.

Equalization of Lines. For local pickup—that is, where a broadcast station picks up a local event and sends it through short local telephone lines to the station—it is possible to lease either loaded high-fidelity lines, or ordinary lines which have no frequency-response correction at all. As it is cheaper to lease ordinary lines, they are quite commonly used, and in such cases, another form of frequency-distortion correction is utilized.

Since the ordinary telephone line is decidedly capacitive, and its transmis-

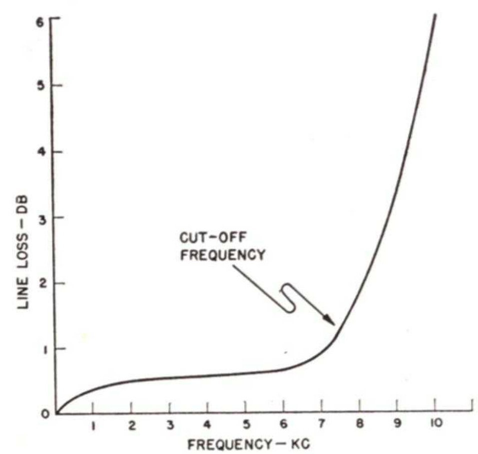


FIG. 20. Below the cut-off point a loaded line has high-fidelity response.

sion, therefore, *falls off* with rising frequency, any network which has an impedance *rising* with frequency can be used at the termination to correct the frequency response. The parallel-tuned circuit operated below its natural resonant frequency is such a network.

The Shunt Equalizer. Suppose we shunt the tuned circuit of Fig. 21A across the termination of an ordinary telephone line. What will be the result?

The natural response of an ordinary line is similar to curve 1 in Fig. 21B. Note how transmission falls steadily as frequency is increased.

Curve 2 shows the variation in impedance of a parallel resonant circuit which is *tuned slightly above* the highest frequency to be transmitted.

As the tuned circuit has low impedance at very low frequencies, it effectively "shorts out" and reduces the line output voltage. As frequency is increased, however, the tuned-circuit impedance also increases, and more and more of the line output is made available. Power delivered by the line, therefore, is reduced at every point, but the reduction is much more severe at

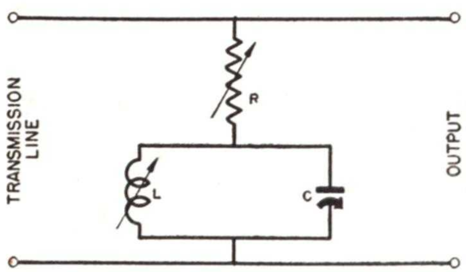


FIG. 21A. A typical shunt equalizer.

the low frequencies. The net effect is shown by the curve 3. The corrected frequency response is much flatter.

This method of achieving high fidelity by adding a corrective network is called "equalizing" a line. When an equalizer of any type is connected *across* the line, it is called a "shunt equalizer."

In Fig. 21A the equalizer is shown with all components variable. This is common practice, for it is found no two telephone lines are exactly alike, and also, the fidelity requirements for dif-

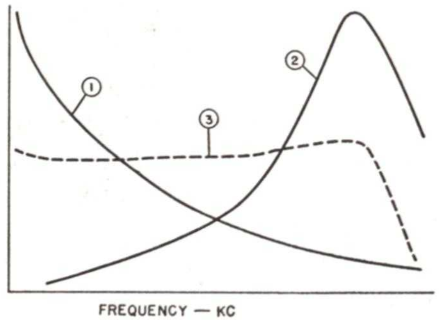


FIG. 21B. Equalizer impedance variation (2) makes an ordinary line (1) into a high-fidelity line (3).

ferent occasions may not be identical.

Two shunt equalizers used by one broadcast station are as follows:

No. 1: L = 0.005 henrys
C = 0.171 μf
Natural resonant frequency 5489 c.p.s.
Highest desired transmission frequency is 5000 c.p.s.

No. 2: L = 0.0022 henrys
C = 0.171 μf
Natural resonant frequency 8400 c.p.s.
Highest desired transmission frequency is 8000 c.p.s.

Adjusting an Equalizer. In adjusting an equalizer to a line, a very simple procedure is followed. First, the line is properly terminated at both source and load ends with impedance-matching transformers or other devices. Then the equalizer is adjusted by varying the inductance or capacity so that the resonant impedance peak

will be from 5% to 10% *higher* in frequency than the highest frequency to be transmitted.

Next, an audio-frequency signal generator is attached to the line source, and a vacuum tube voltmeter or a db power level meter is connected to the line output.

With the equalizer in place, various frequencies from 50 to 8000 cycles (or whatever the high-frequency limit happens to be) are fed into the line source *with the input voltage kept constant.*

The voltage or power delivered at the line output is measured at every frequency point by the voltmeter or db meter. When these readings are plotted against frequency, the result is a graph of the corrected line transmission response.

If the response curve is not as flat as desired, then the height and breadth of the equalizer impedance peak may be altered by changing the variable resistor value. The frequency of the peak also may be changed by adjusting the equalizer inductance or capacity.

After each adjustment, a new complete response curve should be run.

The Series Equalizer. If any unexpected dips or "valleys" occur in the corrected response curve for a given line, it is possible to eliminate these by inserting an additional equalizer. This should be a "series-type," placed in series with the line, instead of a parallel equalizer. One type of series equalizer is a variable resistor, a capacitor and an inductance, connected as a series-resonant circuit, and inserted in series with the line.

Since the series resonant circuit will present a high impedance to all frequencies but one, it is possible to "flatten out" sudden dips in the curve. The amount of correction and the frequency at which it occurs can be varied by adjustment of the resistor, inductance, and capacity values.

R-C Equalizers. It is also possible to equalize the response of a transmission line by means of an R-C network. When the line to be equalized can be represented as shown in Fig. 14, then the equalizing network of Fig. 22A can be used.

Notice that the equalizer is an R-C high-pass filter, and the response of the line with the equalizer would be similar to the curves in Fig. 21B. However, there is no resonant peak as shown in curve 2. For this reason, there is no resonant step-up, and the attenuation of an R-C equalizer is more than that of an L-C equalizer. Therefore, the L-C equalizer is more generally used.

The type of R-C equalizer shown in Fig. 22B is more useful than the basic type, for it permits equalization of lines when the value of the load resistance connected to the end of the line affects the frequency response. The values of R_1, R_2, and C_1 depend on the constants of the line, the load

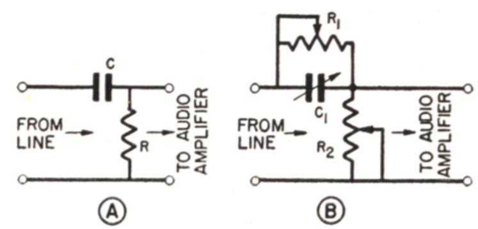


FIG. 22. R-C networks can be used to equalize the frequency response of a transmission line.

resistance, the amount of attenuation permissible, and the response desired. They are generally variable controls so that individual lines may be equalized.

Use of Leased Lines. To prevent undue cross-talk and interference with other services, the telephone company will not now permit a level in excess of 4 VU (about 2.5 milliwatts) in a leased line although at one time about 12 milliwatts were permissible.

This power limitation means that the losses of a long line cannot be supplied by higher power input. Instead, they must be compensated for by amplification at the receiving end. If the line loss is great, however, then the amplifier must be very sensitive. This may lead to trouble with noise pickup and other disturbances.

In practice, it is, therefore, desirable to break the line into sections of, say, ten miles each, and to insert amplifiers at each point to bring up the signal to the 4 VU level before feeding it into the next section of line. In this manner, the signals can be maintained near the maximum allowable level all along the line, and noise and cross-talk will be minimized. Each ten-mile section is separately equalized.

For special purposes, a "quiet line" can be requested. This is usually a twisted pair located in the center of a bundle of pairs with a common lead sheathing. Lines can be ordered from the telephone company which have very low noise levels, as much as 50 to 65 VU below the reference level.

LONG TRANSMISSION LINES

Up to this point we have considered relatively short transmission lines and lines not more than a few miles in length. We found, too, that within certain limits we can use any pair of matched impedances that we want, at source and load ends, as long as we are satisfied with the frequency response.

With extremely long lines, however, another electrical property becomes important and must be considered.

Characteristic Impedance. Since a transmission line has shunt capacitance, it can draw input current even if the receiving end is not loaded. The resistance of the line then, will cause some power loss when this current flows. A transmission line obviously has an impedance of its own which will depend upon the length of the line, since a longer line will have higher shunt capacitance.

If the line is extremely long, because of dissipation along the line, very little power will reach any load placed at the far end. If practically no power reaches the distant load, then the line input impedance will be independent of the termination.

► This results in the fact that very long lines (known as infinite lines) have an impedance which depends only upon the line construction. This impedance is called the "characteristic impedance" of the line.* It varies somewhat with frequency, and is determined mainly by the size of the wires, their separation, and the insulation between them.

Impedance Matching. In the earlier part of this Lesson we discovered that maximum power is delivered to a load when the load impedance is exactly equal to the generator impedance. To deliver power over a long transmission line, we have the problem at the source end of matching the source impedance to the characteristic impedance of the line. Similarly, at the load or receiving end, the characteristic impedance of the line must be matched to the load itself.

► In other words, *both the source and load impedances must be equal to the line impedance, if we are to obtain the greatest power transfer.*

The standard telephone cable is made of twisted #19 AWG wire, having a loop resistance of 88 ohms per mile, and behaving as if it had a loop capacity of 0.054 microfarads per mile. For audio frequencies, the inductance and leakage conductance can be neglected. The characteristic impedance of such a line at 1000 cycles is about 570 ohms. This means the line is best used to match a 570-ohm source to a 570-ohm load in the voice audio range.

Although the line will work to some degree with other impedances, its efficiency is greatly reduced by the mismatch or "reflection loss."

And there can be reflections, too. Long lines improperly terminated quite often have "echoes" due to waves of reflected energy. These echoes make speech very unpleasant, and in the transmission of pictures, actually will produce "ghost" images!

Reflections of electrical waves on lines is a subject of particular importance in radio-frequency lines. We will, therefore, study them in detail in another Lesson.

*It has been found that the characteristic impedance of a line can be determined by shorting the end of the line and noticing the impedance Z_1 of the line as seen by the signal source; then with the end of the line open, the input impedance Z_2 is again measured. The characteristic impedance Z_o may now be calculated from this formula.

$$Z_o^2 = Z_1Z_2, \text{ or } Z_o = \sqrt{Z_1Z_2}$$

Control of Audio Power

ATTENUATORS

So far, we have found ways of transmitting audio-frequency power over transmission lines, no matter whether they are only a few feet or several miles in length. We have found, also, that impedance matching plays an important part in the transfer of energy from any source to any load.

We have not discussed, however, any methods of controlling the amount of power that is being transferred.

It is obvious that the same amount of power will not be satisfactory for all applications. More power is required to fill a well-populated theater with sound than a relatively empty one. If the sound equipment is capable of driving the speakers to sufficient output in the first case, the power must be reduced when the audience is small, or the sound will be much too loud.

In radio broadcasting, the audio equipment must be capable of sufficient gain when a quiet-voiced speaker is "on the air," but the power must be reduced when a full orchestra is playing or the carrier will be over-modulated.

Signals transmitted over telephone lines must be kept below a certain power level to avoid cross-talk with other lines.

A slightly different example is the use of a monitoring speaker in a broadcast station. Only a very small part of the actual power available should be fed to the monitor speaker.

All these cases, and many more, require the use of volume controls or attenuators.

Construction of the L-Pad. The simplest way of attenuating the power reaching a load is to add resistance in series with the load as in Fig. 23A. However, this increases the load impedance as seen by the source, and there is considerable mismatch.

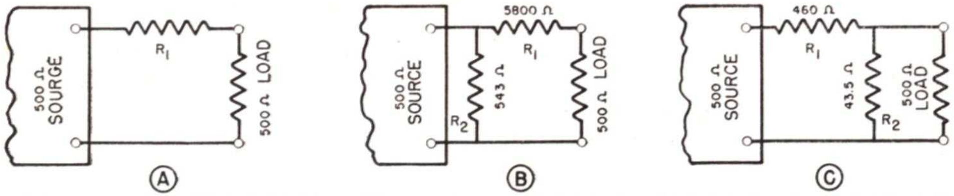


FIG. 23. This drawing shows both a right- and a left-handed L-pad. Both match load to source, but not source to load. At the resistance values given, the attenuation will be 22 db.

The original correct value of load resistance can be realized by placing an additional shunt resistance of the appropriate value across the input to the network. We then have the set-up in Fig. 23B. Such an arrangement of resistors is commonly called a "right-handed L-pad."

For the particular values of resistance given in the example, the load receives only 1/159 of the total power delivered by the source (hence, the L-pad attenuation is 22 db), and the generator still sees a load impedance of 500 ohms. This can be checked readily by using Ohm's Law to calculate the resultant resistance of R_2 in parallel with R_1, and the load in series.

It would be just as effective to shunt the load with a resistor such as R_2 in Fig. 23C, and then add the series resistor R_1 to bring the effective input impedance back to 500 ohms. This results in a "left-handed L-pad" with the same attenuation of 22 db.

One important point should not be overlooked: Although the *source* impedance is properly matched in Figs. 23B and 23C, in each case the impedance *as seen by the load* is very different from 500 ohms. In Fig. 23B, the load sees an impedance even greater than 5800 ohms; in Fig. 23C, the impedance presented to the load is less than 43.5 ohms!

► Such a mismatch is characteristic of L-pads; they cannot be used to match *equal* source and load impedances at the same time for any resistor values.

Symmetrical Attenuator Pads. In order to maintain the proper impedance match for both source and load when their impedances are equal, it is necessary to use what is called a "symmetrical" attenuator. The T-pad in Fig. 24, so called because of its resemblance to the letter "T," is a symmetrical attenuator. As seen from each end, the resistor network is exactly the same, and the input and output terminals may be interchanged without altering the performance in any way.

For the particular case in Fig. 24, the T-pad accurately matches a 500-ohm source to a 500-ohm load, and inserts an attenuation of 22 db. That the source actually sees an input impedance of 500 ohms with the load in place can be determined by finding the effective resistance of the series group (R_1 and the load) in parallel with R_2, then adding this to the value of R_1. Similarly, since it is a symmetrical network, the impedance looking back from the output terminals also will be 500 ohms with a 500-ohm source connected.

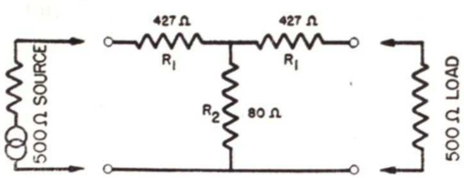


FIG. 24. Here is a symmetrical T-pad that matches both source and sink. The resistance values chosen here will cause a 22 db. loss.

Should it be desirable to match a 250-ohm source to a 250-ohm load with the same 22 db attenuation, then the circuit of Fig. 24 can be used if the resistors are halved in value.

The T-pad is used extensively in circumstances where the impedance values of source and load must not be disturbed, and yet a given amount of signal attenuation is necessary. The load resistor, for example, may represent a monitoring or indicating meter which would be overloaded if full power were applied, and yet impedance values cannot be changed without destroying the meter calibration.

Design of a T-Pad. The circuit of the T-pad has been generalized so that by solving relatively simple equations, an attenuator with any given db loss for use with any desired source-load impedance can be constructed.

The general form of the T-pad is given in Fig. 25. The two "arm" resistors, which are equal in value, are indicated as R_A; the "leg" resistor is shown as R_L.

To determine the actual values of resistances for a given attenuator, the source-load impedance Z and the desired attenuation db are substituted in the following equations:

$$K = \log^{-1}\frac{db}{20} \qquad *(1)$$

$$A = \frac{K-1}{K+1} \qquad (2)$$

$$L = \frac{2K}{K^2-1} \qquad (3)$$

$$R_A = AZ \qquad (4)$$

$$R_L = LZ \qquad (5)$$

As an example of the use of these equations, suppose we wish to design a T-pad having an attenuation loss of 30 db which will properly match a 500-ohm source to a 500-ohm load.

We have then: Z = 500 ohms

db = 30

Now, if we put the desired db value in equation (1), we will determine the reduction factor K. Equation (1) simply says that K is equal to "that number which has db/20 for its logarithm." We find then:

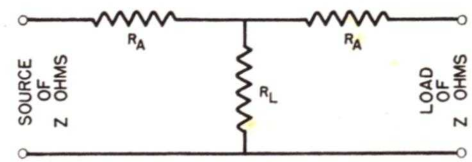


FIG. 25. The basic symmetrical T-pad.

$$K = \log^{-1}\frac{db}{20} = \log^{-1}\frac{30}{20} = \log^{-1} 1.5$$

and from our table of logarithms, we determine that the number which has 1.5 for a logarithm is 31.6, or simply,

$$K = 31.6$$

Next, taking this value of K and putting it into equation (2), we obtain the *arm* multiplying factor A:

$$A = \frac{K-1}{K+1} = \frac{31.6-1}{31.6+1} = \frac{30.6}{32.6} = 0.939$$

And similarly, by putting the value K = 31.6 in equation (3), we determine the *leg* multiplying factor L:

$$L = \frac{2K}{K^2-1} = \frac{2 \times 31.6}{31.6^2-1} = \frac{63.2}{999} = 0.063$$

Finally, by multiplying the *arm* factor A by the desired impedance value Z, as indicated in equation (4), we determine the resistance value of the T-pad series *arms:*

$$R_A = A \times Z = 0.939 \times 500 = 470 \text{ ohms}$$

Likewise, by inserting the values of the *leg* factor L, and the impedance Z, in equation (5), we obtain the resist-

*The symbol "log-1" indicates an anti-log. This means K is a *number* with a logarithm of db/20. If this were written just as "log," then K would be a *logarithm* for a number ratio.

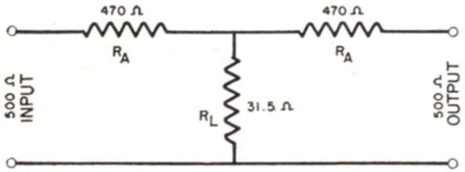


FIG. 26. A 500-ohm T-pad with an insertion loss of 30 db.

ance of the T-pad shunt *leg* resistor:

$$R_L = L \times Z = 0.063 \times 500 = 31.5 \text{ ohms}$$

The completed T-pad, having an input-output impedance of 500 ohms and an insertion loss of 30 db, will appear as in Fig. 26.

For convenience, the multiplying factors of a T-pad have been calculated for a number of different attenuation losses. These are given in the table of Fig. 27. Arm and leg resistances can be determined simply by multiplying the appropriate source-sink impedance Z by the factor values corresponding to the desired attenuation. This table will be of particular convenience in the design of decade attenuators to be studied later.

Construction of a Pi-Pad. The pi-pad sketched in Fig. 28 is another type of symmetrical attenuator pad.

Attenuation db	Arm factor A	Leg factor L
0	0	∞
1	0.057	8.66
2	.114	4.31
3	.171	2.83
4	.226	2.10
5	.280	1.65
6	.332	1.31
7	.383	1.12
8	.430	0.946
9	.476	.812
10	.520	.703
20	.818	.202
30	.939	.0633
40	.980	.0200

FIG. 27. Table of multiplying factors for the design of any T-pad.

This network is so named because of the resemblance to the Greek letter π, written as "pi" (pronounced as "pie").

The pi-pad of Fig. 28 is like the T-pad in Fig. 24, in that it has an attenuation loss of 22 db and a 500-ohm input and output impedance.

Since the pi-pad and T-pad have identical behavior, we can expect a definite numerical relation between them. We can, for example, set up a general pi network as in Fig. 29, and determine the various resistances for any attenuator, as follows:

► The resistance of the *arm* of the pi-pad is equal to the impedance Z *divided* by the T-pad *leg* factor L. Simi-

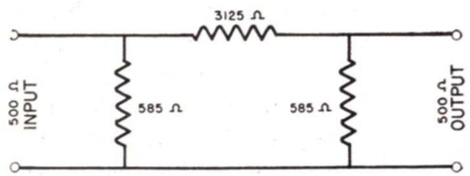


FIG. 28. A 500-ohm pi-pad with an insertion loss of 22 db.

larly, each *leg* of the pi-pad is equal to the impedance Z *divided* by the T-pad *arm* factor A.

In equation form, these are:

$$R_A \text{ (for the pi)} = Z/L \qquad (6)$$

$$R_L \text{ (for the pi)} = Z/A \qquad (7)$$

It should be noted that the factors A and L are the same as those outlined for the T-pad, and these are calculated exactly as shown before. However, the factors are used in division instead of multiplication. In other words, for a T-pad, use equations (1), (2), (3), (4), and (5); for a pi-pad use equations (1), (2), (3), (6), and (7).

In ordinary use, there is no preference between a T-pad and a pi-pad, for these networks are interchangeable, and the selection of a suitable pad is merely a matter of personal choice.

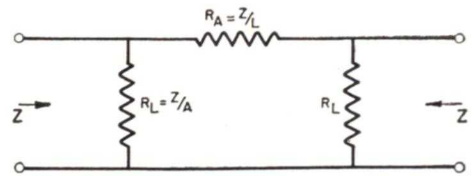


FIG. 29. The general form of the pi-pad attenuator.

Attenuator Pads in Tandem. The input impedance of a symmetrical pad is equal to its output impedance, thus the input of one pad can be used as the load resistance for a preceding pad. This can be kept up for a great number of times, and the total attenuation of the pads in tandem will be equal to the sum of the individual pad losses.

Thus, if we have three 500-ohm to 500-ohm pads, giving respectively 5, 8, and 4 db attenuation, we can connect these in any order, and the whole combination will have an input impedance of 500 ohms, and an attenuation of $5 + 8 + 4 = 17$ db when used with a 500-ohm load.

This is a very convenient feature of symmetrical pads, for it means that a great number of attenuation values can be realized with combinations of a relatively few separate attenuators.

Balanced Attenuator Pads. All of the attenuator pads discussed so far have been designed for use with unbalanced lines which have one wire grounded. With balanced lines, which maintain the wires at opposite potential with respect to ground, these attenuators would not be suitable, be-

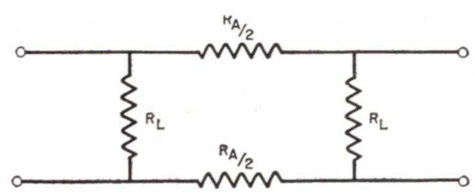


FIG. 30. A general form of balanced pi-pad.

cause the stray attenuator capacitance would upset the line balance and tend to increase noise pickup and cross-talk.

The T and pi-pads, however, can be balanced very easily. This is accomplished merely by splitting the series resistance arms in half and inserting half in the upper and half in the lower side of the line.

Thus, the pi-pad in Fig. 29 becomes the balanced pi-pad of Fig. 30. In the same manner, the T-pad of Fig. 25 is changed to the balanced version in Fig. 31. This latter network quite often is called an H-pad.

Notice also, as shown in Fig. 31, how balanced transformers T_1 and T_2 are used to couple between balanced lines

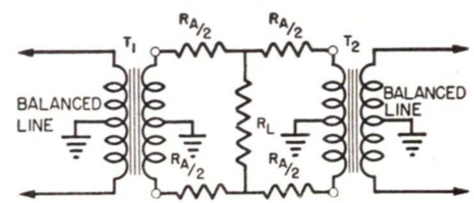


FIG. 31. The balanced T-pad becomes an H-pad.

and balanced attenuator pads. Because the capacity to ground and leakage to ground of the two lines is the same, the center taps of the transformer windings can be connected to ground without affecting the balance.

VARIABLE ATTENUATORS

The attenuation introduced by an L-pad, a T-pad, or a pi-pad, obviously, can be made adjustable by using variable resistors for the arms and legs. The T-pad of Fig. 25, for example, could be made with variable R_A and R_L values.

► If this is done, however, the relations outlined in the table of Fig. 27 must be maintained or the variable pad will not have constant input and output impedances.

And why is a constant impedance characteristic so important? For illustration, suppose we are driving three speakers with a single power amplifier, and that the power to each speaker is controlled by an individual variable T-pad. If these pads have constant impedances regardless of the attenuation setting, then any one control can be adjusted independently of the other two. Should the pad impedances change with attenuation, however, then adjusting the power to one speaker would affect the output of the other speakers as well.

Tapped Attenuators. A reasonable, practical design of a variable attenuator is made possible by the fact that attenuation does not need to be continuously variable since the ear cannot detect level changes of less than about 2 db. Hence, if the variation

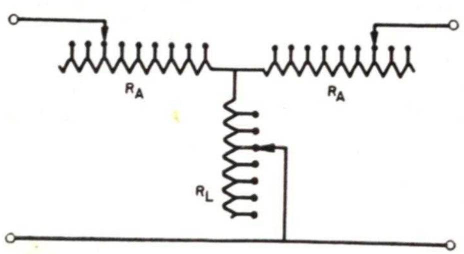


FIG. 32. The usual style of tapped T-pad variable attenuator.

were made in steps of 1½ db, it would sound like a continuous change. This allows us to use a set of fixed resistors and a three-gang switch as shown in Fig. 32.

One advantage of using step attenuation in the design of a variable T-pad is that such an arrangement can be calibrated accurately.

In addition, the gang switch can be made with large wiping contacts which

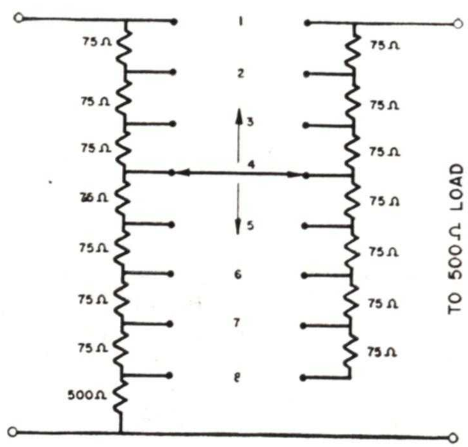


FIG. 33. An approximate T-pad using only two variable resistors.

keep contact noise at a minimum. This is in contrast to a rheostat or a potentiometer which seldom can be reset to a given point, and usually has a high degree of noise from contact variation.

An Approximate Variable T-Pad. To increase the amount of attenuation in this pad, the value of R_A must be increased, and the value of R_L must be decreased. For this reason, then, it is possible to combine one of the R_A's with R_L, so that the simpler construction of Fig. 33 can be used.

In the figure, the right-hand resistance forms the right arm of the T-pad, the upper part of the left-hand resistance forms the left arm, and the lower part forms the leg.

Switch position	Input Resistance	db Loss
1	336	1.9
2	432	3.5
3	522	5.0
4	605	6.5
5	680	7.9
6	748	9.3
7	808	10.8
8	861	12.4

FIG. 34. Input resistance and db loss of the approximate T-pad for each switch point.

This arrangement is not a true T-pad, but only an approximation since it is not possible to make the attenuator have exactly constant values of input and output resistance at all settings. As shown in the table of Fig. 34 it varies from 336 to 861 ohms.

While 336 to 861 ohms seems a wide range for matching a 500-ohm source and load, the mismatch loss is only ¼ db at the extreme positions.

It will be noted that the use of this particular attenuator involves a loss of 1.9 db at the minimum setting. This is called the "insertion loss" of this particular device.

Decade Attenuator Pads. Suppose we have need for an attenuator adjustable in 1 db steps from 0 to 50 that maintains perfect impedance matching at all settings. We could, of course, design a conventional T-pad having 50 steps, but this would be expensive and awkward to use.

A better method would be to build two attenuators, one having 4 steps of 10 db each, and the other having 10 steps of 1 db each. Then, if these are constant-impedance pads, they can be connected in tandem so that the separate attenuations will add. Thus, one knob controls the tens, and the other controls the units of attenuation. Such an arrangement is called a "decade attenuator."

The design for such a decade attenuator is shown in Fig. 35. The particular resistance values shown were determined by taking the difference between the corresponding R_A and R_L values of Fig. 26. As pictured, the decade pad is designed to work between a source of 1 ohm and a load of 1 ohm. For any other common source and load value Z, simply multiply the values shown by Z.

The Bridged-T. A modification of

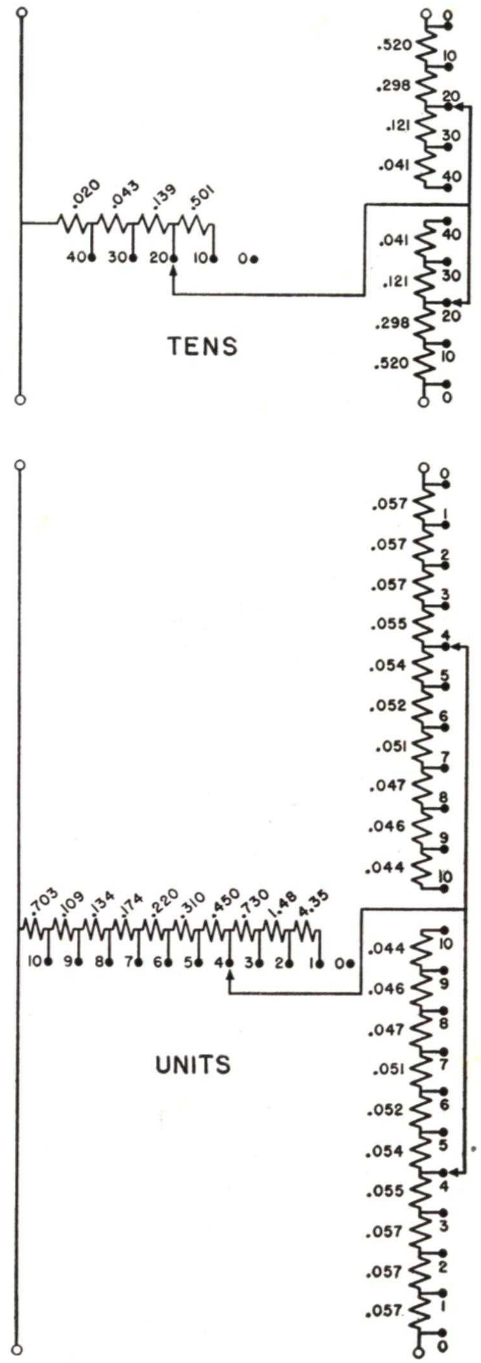


FIG. 35. A decade attenuator adjustable in 1 db steps from 0 to 50 db. This is designed for a source-load impedance of 1 ohm. For other values, multiply the figures shown by the desired impedance in ohms.

the T-pad, of considerable importance because it requires only two variable resistors, is shown in Fig. 36. This is known as the "bridged-T." It is a symmetrical attenuator, having two fixed arms bridged from input to output by a variable one, and a variable leg.

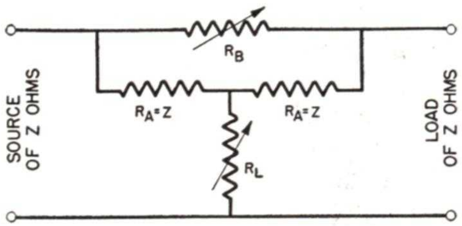


FIG. 36. The bridged-T is a true symmetrical pad with only two variable resistors.

The resistances of the fixed arms R_A are always equal to the input-output impedance Z. The bridge and leg resistors, R_B and R_L, are related to the impedance Z in this way:

$$R_B = B \times Z$$
$$R_L = L \times Z$$

The multiplying factors, B and L, can be determined from the table in Fig. 37.

The Potentiometer. If we have the problem of taking the signal from a 600-ohm line, and feeding it to the grid of a class A audio amplifier through a volume control and the necessary transformer, we may do it on the basis of circuits already discussed as outlined in Fig. 38. This is, however, an unnecessarily expensive and complicated way of getting the desired results.

Attenuation db	Bridge-arm factor B	Leg factor L
0	0	∞
1	0.122	8.20
2	.259	3.86
3	.413	2.42
4	.585	1.71
5	.778	1.285
6	.995	1.005
7	1.239	0.808
8	1.512	.661
9	1.818	.550
10	2.162	.462
20	9.000	.111
30	30.62	.0326
40	99.00	.0101

FIG. 37. Multiplying factors for design of the bridged-T.

Since the current drawn by a class A grid is negligible, the tube input impedance at audio frequencies is very high. It is possible, therefore, to use a potentiometer to feed the grid. With-

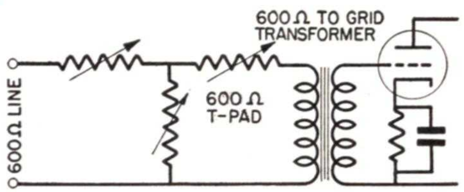


FIG. 38. This shows how a T-pad (as volume control) can be used with a grid input transformer to couple a 600-ohm line to the grid of an amplifier.

out grid current flow, the potentiometer input resistance will be independent of the setting, and always equal to the full resistance value. Thus, a simple, cheap, and fully satisfactory means of controlling the signal is made by using a step-up transformer and potentiometer as shown in Fig. 39.

Potentiometer control is ordinarily used at relatively high signal levels to avoid excessive noise. The potentiometer frequently is continuously variable, but it can be made as a tapped bank of resistors as in Fig. 40.

How a Potentiometer Attenuator is Designed. As an example of use of the potentiometer design equation of Fig. 40, suppose we are to construct a 100,000-ohm potentiometer giving up to 20 db attenuation in 2 db steps.

This requires 11 resistors with a total resistance Z = 100,000 ohms. See

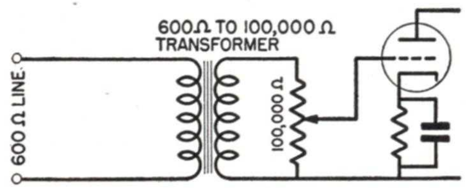


FIG. 39. A potentiometer (with a step-up transformer) makes a simpler volume control when feeding a class A grid.

Fig. 41A, and consider the resistors as being labeled R_1, R_2, etc., from the bottom up.

The bottom resistor R_1 is computed first, and since we wish the attenuation to be 20 db at this point, the equation of Fig. 40 is used as follows:

$$R = \frac{Z}{\log^{-1} db/20} = \frac{100{,}000}{\log^{-1} 20/20}$$

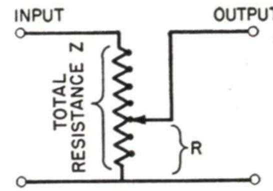
$$= \frac{100{,}000}{\log^{-1} 1} = \frac{100{,}000}{10} = 10{,}000 \text{ ohms}$$

The value of the resistor R_2 (*second from bottom*) is calculated next. As the attenuation is to be 18 db at this point, we proceed:

$$R = \frac{100{,}000}{\log^{-1} 18/20} = \frac{100{,}000}{\log^{-1} 0.9}$$

$$= \frac{100{,}000}{7.94} = 12{,}600 \text{ ohms}$$

Now it is important to note that *R represents the total resistance from the 18 db point to ground.* To find the value of R_2, therefore, it is necessary to subtract the value of R_1 from this last computed resistance R.



$$R = \frac{Z}{\log^{-1} \frac{db}{20}}$$

FIG. 40. The general form and design equation for the potentiometer attenuator.

Then we have as the actual value of R_2:

$$R_2 = R - R_1$$
$$R_2 = 12{,}600 - 10{,}000 = 2600 \text{ ohms.}$$

In a similar manner, for the 16 db point we find R = 15,800 ohms. But R_3 is not known until the values of R_1 and R_2 are subtracted from R. We determine R_3 then:

$$R_3 = R - R_1 = 15{,}800 - 12{,}600$$
$$= 3200 \text{ ohms (approximately).}$$

By this process, all the required resistors can be determined. The com-

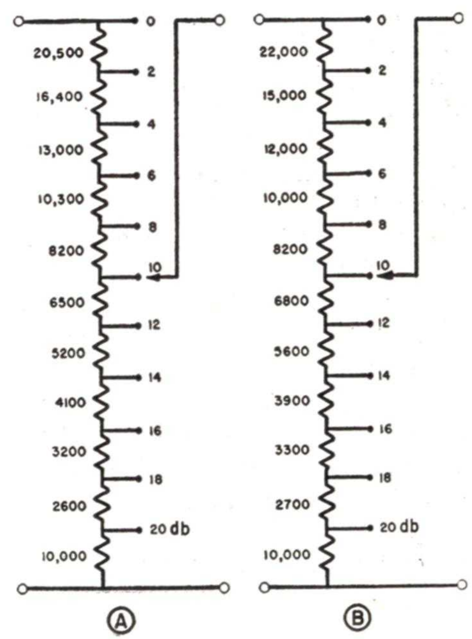
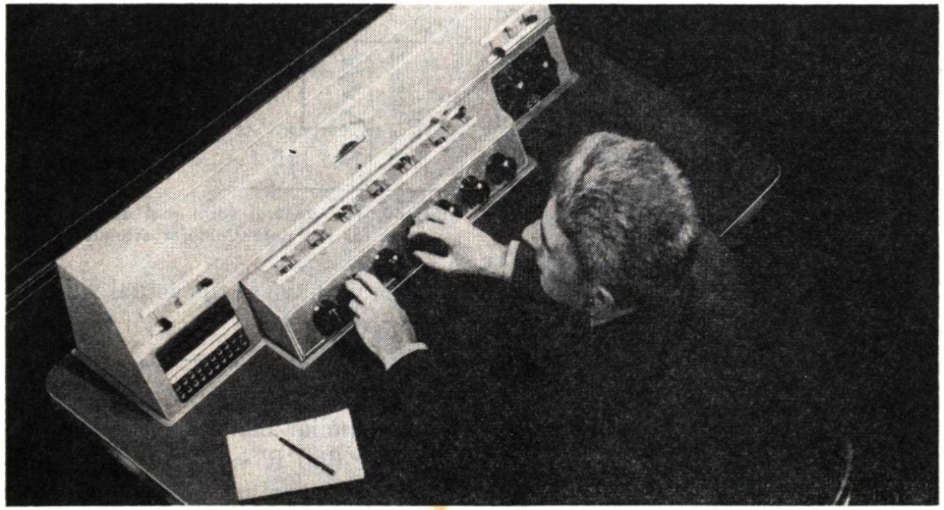


FIG. 41. Potentiometers giving attenuation up to 20 db in 2 db steps. Exact resistors are shown in A; standard 10% tolerance resistors have been substituted in B.



Courtesy Western Electric Company

Convenience and flexibility are integral parts of 25A speech input console design. In spite of the large number of controls within easy reach of the operator, the equipment has a neat clean-cut appearance. Operator is controlling balance between two microphone pickups by manipulation of two of the four microphone mixers.

pleted potentiometer will appear as in Fig. 41A. As we desired, the over-all resistance Z is 100,000 ohms.

Usually, it is not practical to use special-valued resistors, and it is customary to substitute the nearest value of standard resistors. If this is done with resistors having 10% tolerance, the potentiometer in Fig. 41A is changed to that in Fig. 41B.

Although such practice invokes some error, with 10% tolerance this error can never exceed 0.4 db. With resistors of closer tolerance, the maximum error is even less.

For rapid potentiometer calculation, the ratio R/Z has been computed for a number of attenuation values. These are tabulated in Fig. 42. For any given db point, the resistance R is determined by multiplying the over-all resistance Z by the fraction appearing in the R/Z column.

Attenuation db	R/Z	Attenuation db	R/Z
0	1.00	22	0.0795
2	0.795	24	.0631
4	.631	26	.0501
6	.501	28	.0398
8	.398	30	.0316
10	.316	32	.0251
12	.251	34	.0199
14	.199	38	.0158
16	.158	36	.0126
18	.126	40	.0100
20	.100	—	

FIG. 42. Design table for rapid potentiometer calculation.

As before, with the exception of the bottom resistor R_1, true resistor values are not given until all previously determined resistances are subtracted from each value of R.

How Audio Signals Are Mixed

In modern broadcasting, very few programs use one microphone only. Suppose a dialogue is being carried on between two actors, and a background of music is to be maintained. When the actors pause in speaking, the music should be brought in somewhat louder; when they resume talking, the music must again be reduced to the background.

This means at least two microphones are necessary—one for the speakers, another for the music. The output volume of each microphone must be variable and independent of the other.

To accomplish this, two volume controls are used, one in each microphone circuit. The combined outputs are then fed into a common speech amplifier which has a master volume control for regulation of the combined level.

Circuits have been designed to take the signals from any number of sources, mix them with independent control of each, and regulate the common output level. Such units are called "mixers."

The two most important requirements of a mixer are freedom from extraneous noise as the controls are adjusted, and constant impedance as seen from the output.

A Series Mixer. A simple series mixer is outlined in Fig. 43. It is obvious that if the output impedance of any one of the controls (commonly called "faders") *should change* very much with its setting, the combined output levels of all the controls would be affected. Hence, they are designed as constant-impedance pads so this will not happen.

In the figure, the impedance of each fader is seen to be 50 ohms, the four connected in series giving a total impedance of 200 ohms. The master control, therefore, sees a 200-ohm source, and a 200-ohm variable pad must be used at this point.

Parallel Mixing. If the signal source were 400 ohms in impedance in-

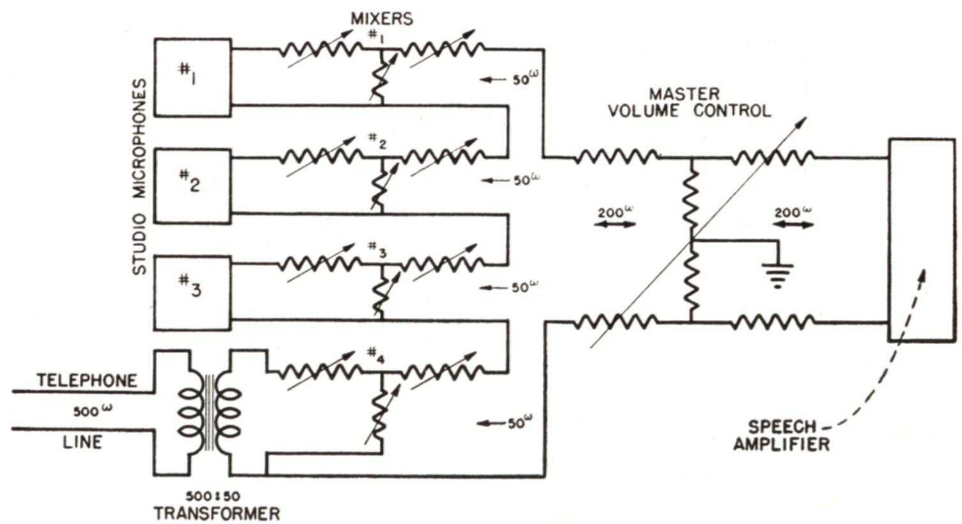


FIG. 43. A series mixer which allows independent control of the signals from four sources and has a master attenuator for adjustment of the combined level.

stead of 50, then connecting them in series would result in an output impedance of 1600 ohms. This normally is too high for low noise pickup and cross-talk. In that case, it would be more desirable to use a parallel mixer as in Fig. 44.

Here the four sources of 400 ohms have a parallel output impedance of 100 ohms. A step-up input transformer matches this 100 ohms to a 100,000-ohm potentiometer used as master control at the input of the amplifier.

Reducing Switch Clicks and Noise. All resistive mixers attenuate the signals to some degree, and they should not be used for low-level signals. Instead, the signal from each microphone should be pre-amplified before going into the mixer. In this way, the gain needed after the mixer is reduced, and so contact noise and switch clicks, are kept at a minimum.

Electronic Mixers. If both an amplifier and a fader are needed for each channel, it is logical to combine these two elements. This can be done by feeding each channel through an individual potentiometer attenuator to separate triode amplifiers, and mixing the signals in the outputs of the triodes. Such an arrangement is called an "electronic mixer." Fig. 45 shows one type.

We can see that although the individual grid circuits are kept isolated, the triode output signals are mixed as they are developed across the common plate load resistor R_L. Since the gain of a triode depends on the plate load, the "decoupling" resistors R_D prevent the triode plate resistances from paralleling R_L too much.

POWER LEVEL INDICATOR

In order to know how much audio-frequency power is being fed to a line, to a modulator, or to any other device, it is necessary to have an indicating meter. A special high-resistance a.c. voltmeter, commonly called a "power level indicator," is designed for this purpose.

A vacuum tube voltmeter may be used as a level indicator. More often, the instrument is a copper oxide bridge rectifier type as shown in Fig. 46.

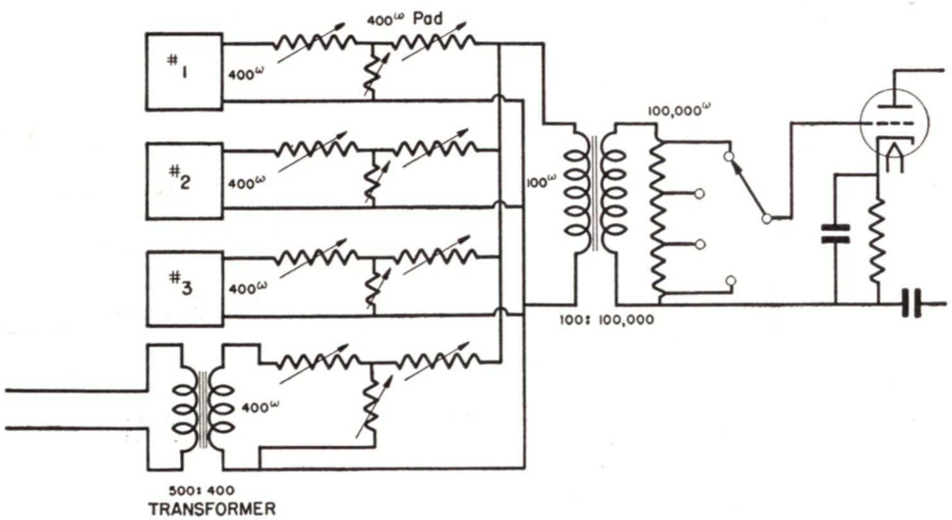


FIG. 44. A parallel mixer recommended for high impedance sources. A potentiometer is used for the master volume control.

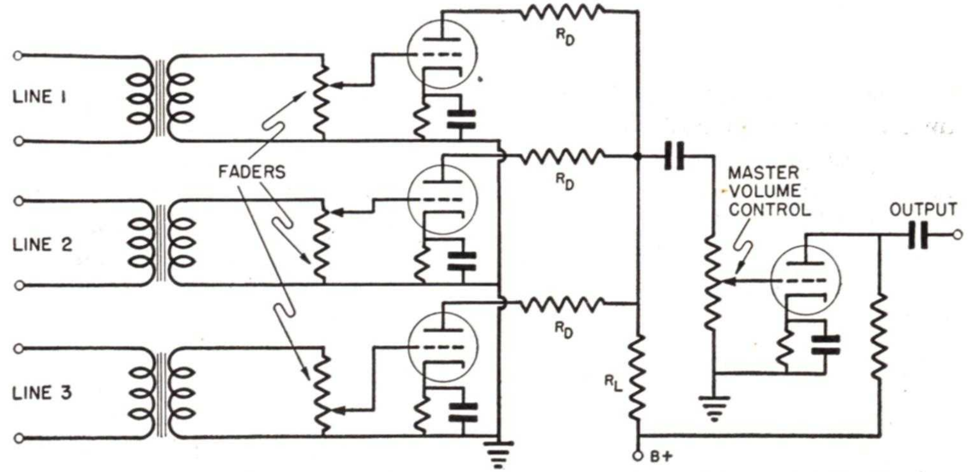


FIG. 45. An electronic mixer using potentiometers for faders and for master control.

Since we know by Ohm's Law that a given voltage across a definite impedance corresponds to a certain power, these meters usually are calibrated directly in decibels from —10 to +6 db. Volts are seldom indicated. By means of the attenuator, the range of this meter can be extended another 30 db without disturbing the input impedance.

The jewelled meter movement is well damped (dead beat), so that rapid fluctuations in signal will not affect the reading.

Substantially correct readings are obtainable up to about 10,000 cycles.

Reference Levels. Since decibels measure only the ratio of one power level to another, a basic power reference level must be used. Although there are several reference levels, the usual standard is the Volume Unit, which is 1 milliwatt of power in a 600-ohm line. A VU meter, thus, will read zero when it is connected across a 600-ohm line in which the power level is 1 milliwatt, +20 db when the power is 100 milliwatts, or —20 db when the power is .01 milliwatt.

Obviously, if a power-level meter is used with a source impedance different from the 600 ohms for which it is designed, the db readings will not be correct. A correction chart, however, usually is supplied with the meter.

We will study more about power-level meters in the Lessons on the audio systems of broadcast stations.

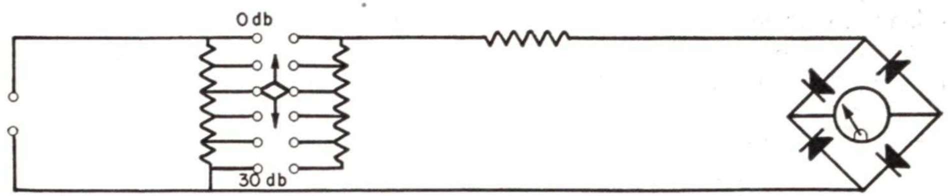


FIG. 46. A copper-oxide-rectifier-type a.c. voltmeter commonly calibrated directly in decibels for use as a power level indicator. The attenuator extends the range 30 db.

Electrical Filters

In many radio- and audio-frequency applications it is not desirable that all frequencies pass through the equipment with equal ease. In carrier telephony (wired wireless), for instance, many low-frequency carriers are separately modulated with speech, and then transmitted over the same transmission line. At the receiving end, these modulated carriers are separated, detected, and amplified by individual receivers. It is obvious that each receiver must be capable of accepting only one carrier with its side bands, and rejecting all others.

Another case of the use of filters is in single-side-band radio transmission. Here, after the carrier has been modulated, a single side band must be separated from the carrier and the remaining side band.

In a television transmitter, vestigial side band transmission is used. This means that one side band of the television picture is transmitted, and the other is partly suppressed.

All modulators and detectors produce distortion frequencies which must be separated from the desired modulation or detection products.

The common rectified a.c. power supply is a more simple example. In this instance, all rectified a.c. components must be suppressed, and only the direct current (which is zero frequency) be allowed to pass.

There are many more applications in which it is necessary to discriminate between different frequencies which may try to flow in a certain apparatus. Special circuits have been devised for accepting or rejecting bands of frequencies. These are called "electric wave filters."

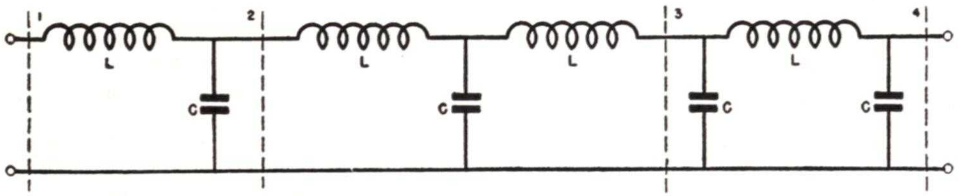


FIG. 47. A low-pass ladder filter.

THE "LADDER" FILTER

Wave filters can be made in a number of forms. The most common is the "ladder" filter shown in Fig. 47. This is a "low-pass" filter because it attenuates low frequencies very much less than high frequencies. Such action should be apparent, for, if current of increasing frequency is sent through the network, the inductive reactance of the series arms becomes higher and "chokes back" the current. In addition, more and more of the current is by-passed through the shunt capacitors as the frequency increases.

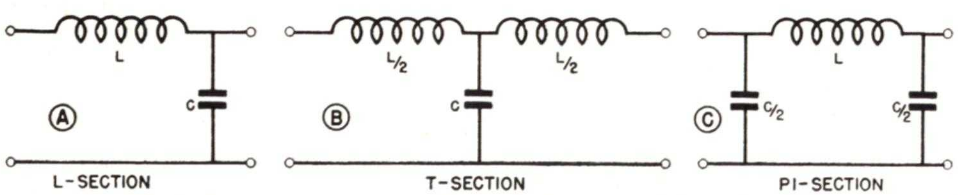


FIG. 48. Illustrating how a ladder filter can be broken into L, T, or pi-sections.

Use of Filter Sections. The design of a ladder filter is much easier if it is broken into sections. For instance, we can take the portion between points 1 and 2 of Fig. 47 and have an L-section as in Fig. 48A; or if we wish, we can consider that part between points 2 and 3 as the T-section of Fig. 48B. Again, we might use the pi-section between points 3 and 4 which is shown in Fig. 48C. It makes no difference how we divide up a ladder filter, for if proper values are used, when several sections of the same type are connected together, the same ladder filter is the result.

The Cut-Off Frequency. The behavior of all the sections in Fig. 48 is the same. We already know that such sections will pass low frequencies much better than high ones. At first thought, we might expect the attenuation to become gradually worse as the frequency is increased. This is not the case.

Because of phase shifts within the sections, if proper input and output impedances are used, low frequencies up to a certain point will be passed with very little attenuation; above this critical point which is called the "cut-off frequency," the attenuation suddenly begins to rise as the frequency is continuously increased. A typical attenuation characteristic for a single section is given in curve 1 of Fig. 49.

When two or more sections are connected, the resultant attenuation is equal to the sum of the losses in each section. Thus, the attenuation of two sections is given by curve 2, three sections by curve 3, and so on. A complete ladder filter with any degree of attenuation can be built up in this way.

High-Pass Filters. If we should interchange the inductances and capacitors in a low-pass filter, it will become a high-pass filter. Typical sections of a high-pass filter are given in Figs. 50A, 50B, and 50C.

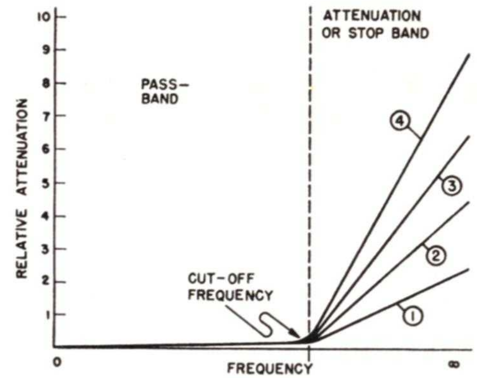


FIG. 49. Curves showing relative attenuation of low-pass filters with one, two, three and four sections.

The high-pass action should be expected, for the series capacitors offer

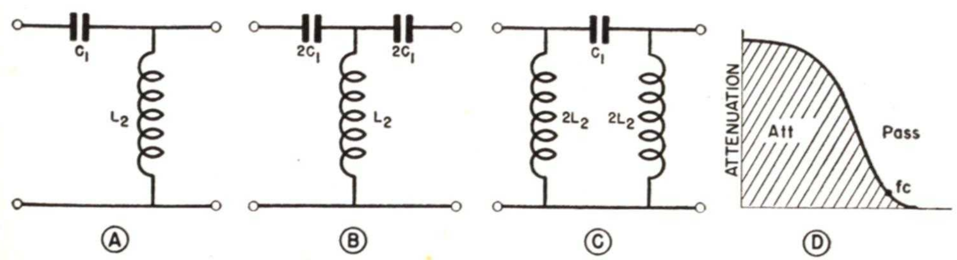


FIG. 50. Typical sections and the attenuation characteristic of a high-pass filter.

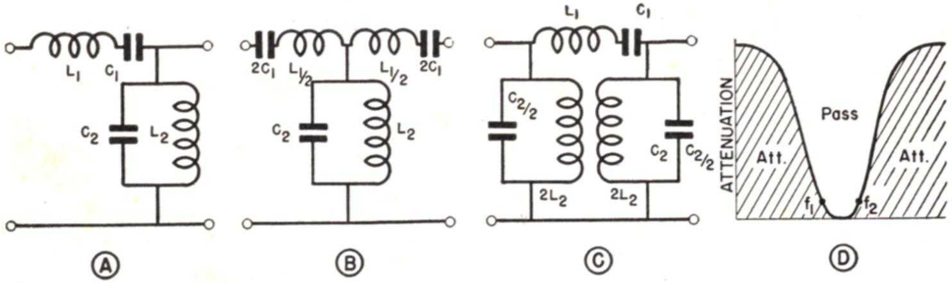


FIG. 51. Band-pass filter sections and their attenuation characteristic.

less and less reactance as the frequency goes up, while the current shunted by the inductances correspondingly decreases. Fig. 50D shows the typical attenuation characteristic of a high-pass filter.

Other Filter Types. By using more complex sections, it is possible to construct filters that will pass a band of frequencies, attenuating frequencies both higher and lower than the desired band. These are called "band-pass" filters. Typical sections, and the attenuation characteristic are in Fig. 51.

Still another type, the "band-elimination" filter is made in a similar manner. As the name implies, this filter presents high attenuation to a given band of frequencies but passes all others with little loss. See Fig. 52. More about these filters will be given in another Lesson.

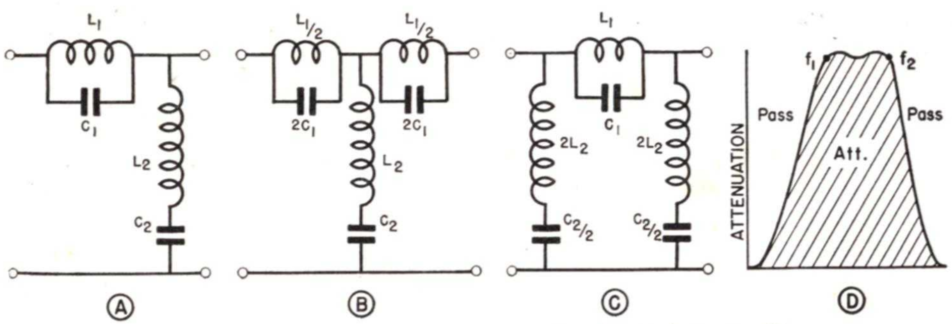


FIG. 52. Sections and the attenuation curve for a band-elimination filter.

Lesson Questions

Be sure to number your Answer Sheet 22RC.

Place your Student Number on *every* Answer Sheet.

Most students want to know their grade as soon as possible, so they mail their set of answers immediately. Others, knowing they will finish the next Lesson within a few days, send in two sets of answers at a time. Either practice is acceptable to us. However, don't hold your answers too long; you may lose them. Don't hold answers to send in more than two sets at a time or you may run out of Lessons before new ones can reach you.

1. How must the impedance of the load be related to the impedance of the source to obtain a maximum transfer of power?
2. What is the required turns ratio of an impedance-matching transformer to couple a 500-ohm line to a 4500-ohm load?
3. Why will an audio line, 150 feet long, working between a low-impedance source and load, require no equalization?
4. What causes increased attenuation of the high audio frequencies in a transmission line?
5. Why is it not permissible to feed more than a level of 4 VU into a telephone line?
6. When using a long transmission line, what impedance terminations are essential at the sending and receiving ends?
7. Why are grounded center-tap transformers frequently used to terminate both ends of a program wire line?
8. When can a potentiometer be used as an attenuator without changing the impedance match?
9. What is a Volume Unit?
10. Sketch a one-section, low-pass, T-type filter.

THE VALUE OF REVIEW

Man has acquired so much new knowledge in recent years that it has become impossible for one person to know even a small fraction of the available information. Educational authorities realize this fact, and the colleges of today consider a man well-educated if he knows the elementary ideas *and knows where to find other information when he wants it.*

Radio, along with the other fields of endeavor, has outgrown the memorizing ability of the human mind. Also, radio is such a comprehensive field that occasionally you cannot recall important facts previously studied. Review is obviously the solution to this problem.

Time spent in review several weeks or months after a book is studied will be far more profitable than an equivalent amount of extra time spent on the book initially, for your mind has then had a chance to file and store away the information secured from the first study. Each review results in more information being transferred from the textbook to your mind, and soon, with no conscious attempt to memorize, you will find yourself able to recall an amazing number of valuable facts.

J. E. SMITH

K = Antilog $\frac{db}{20}$, db = 20 log K

Table XXVIII. K Factors for Calculating Attenuator Loss

dB	K	dB	K	dB	K	dB	K
.05	1.0058	9.5	2.9854	29.0	28.184	49.0	281.84
.1	1.0116	10.0	3.1623	30.0	31.623	50.0	316.23
.5	1.0593	11.0	3.5481	31.0	35.481	51.0	354.81
1.0	1.1220	12.0	3.9811	32.0	39.811	52.0	398.11
1.5	1.1885	13.0	4.4668	33.0	44.668	54.0	501.19
2.0	1.2589	14.0	5.0119	34.0	50.119	55.0	562.34
2.5	1.3335	15.0	5.6234	35.0	56.234	56.0	630.96
3.0	1.4125	16.0	6.3096	36.0	63.096	57.0	707.95
3.5	1.4962	17.0	7.0795	37.0	70.795	58.0	794.33
4.0	1.5849	18.0	7.9433	38.0	79.433	60.0	1000.0
4.5	1.6788	19.0	8.9125	39.0	89.125	65.0	1778.3
5.0	1.7783	20.0	10.0000	40.0	100.000	70.0	3162.3
5.5	1.8837	21.0	11.2202	41.0	112.202	75.0	5623.4
6.0	1.9953	22.0	12.589	42.0	125.89	80.0	10,000
6.5	2.1135	23.0	14.125	43.0	141.25	85.0	17,783
7.0	2.2387	24.0	15.849	44.0	158.49	90.0	31,623
7.5	2.3714	25.0	17.783	45.0	177.83	95.0	56,234
8.0	2.5119	26.0	19.953	46.0	199.53	100.0	10^5
8.5	2.6607	27.0	22.387	47.0	223.87		
9.0	2.8184	28.0	25.119	48.0	251.19		

(B) Combining or Dividing Network

$$R_B = \left(\frac{N-1}{N+1}\right) Z$$

Loss = 3 db Per output

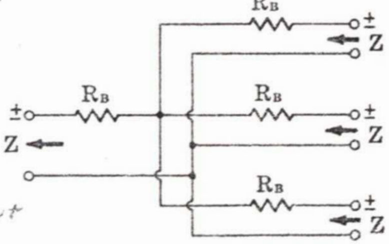


Fig. 105

where,

R_B is the resistance of the building-out resistors in ohms,
N is the number of circuits fed by the source impedance,
Z is the source impedance in ohms.

(C) T-Type Attenuator (Between Equal Impedances)

$$R_1 \text{ and } R_2 = \left(\frac{K-1}{K+1}\right) Z$$

$$R_3 = \left(\frac{K}{K^2-1}\right) 2Z$$

(See Inside Back Cover)

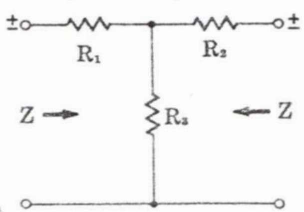


Fig. 106

(D) H-Type Attenuator (Balanced-T Attenuator)

Calculate the values for R_1, R_2, and R_3 as for an unbalanced T-attenuator (Fig. 106). Then halve the values of R_1 and R_2, as shown in Fig. 107. The tap on R_3 is exactly in the center.

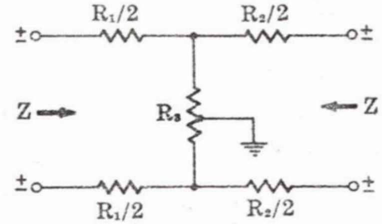


Fig. 107

(E) Taper Pad (T-Type Attenuator Between Unequal Impedances)

$$R_1 = Z_1\left(\frac{K^2+1}{K^2-1}\right) - 2\sqrt{Z_1Z_2}\left(\frac{K}{K^2-1}\right)$$

$$R_2 = Z_2\left(\frac{K^2+1}{K^2-1}\right) - 2\sqrt{Z_1Z_2}\left(\frac{K}{K^2-1}\right)$$

$$R_3 = 2\sqrt{Z_1Z_2}\left(\frac{K}{K^2-1}\right)$$

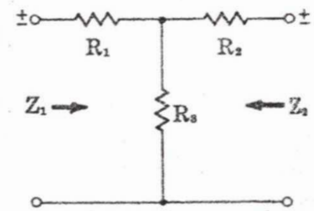


Fig. 108

where,

Z_1 is the larger impedance.

(F) Bridged-T Attenuator (Unbalanced)

$$R_1 = Z$$

$$R_5 = (K-1)Z$$

$$R_6 = \left(\frac{1}{K-1}\right)Z$$

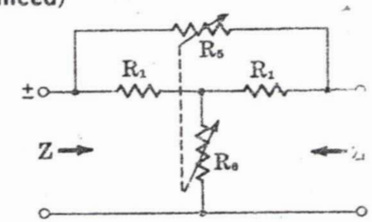


Fig. 109

R_5 and R_6 are connected to a common shaft, and each varies inversely in value with respect to the other.

(G) Balanced Bridged-T Attenuator

Calculate the values for R_1, R_5, and R_6 as for an unbalanced bridged-T attenuator (Fig. 109). Then halve the values as shown in Fig. 110.

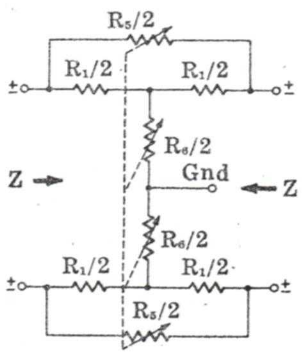


Fig. 110

(H) L-Type Attenuators

An L-type attenuator can supply an impedance match in only one direction. If the impedances it works out of and into are unequal, it can be made to match either—but not both—impedances. The arrows in the following illustrations indicate the direction of impedance match.

Between equal impedances and with the impedance match in the direction of the series arm:

$$R_1 = Z\left(\frac{K-1}{K}\right)$$

$$R_2 = Z\left(\frac{1}{K-1}\right)$$

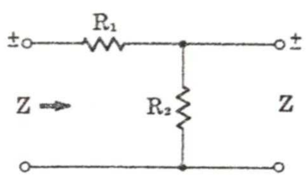


Fig. 111

Between equal impedances and with the impedance match in the direction of the shunt arm:

$$R_1 = Z\ (K-1)$$

$$R_2 = Z\left(\frac{K}{K-1}\right)$$

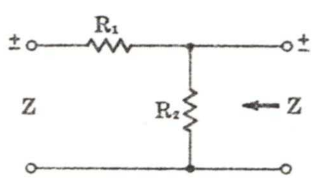


Fig. 112

Between unequal impedances and with the impedance match toward the larger value:

$$R_1 = \left(\frac{Z_1}{S}\right)\left(\frac{KS-1}{K}\right)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{1}{K-S}\right)$$

where,

$$S = \sqrt{\frac{Z_1}{Z_2}}$$

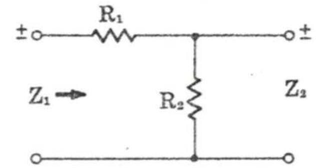


Fig. 113

Between unequal impedances and with the impedance match toward the smaller value:

$$R_1 = \left(\frac{Z_1}{S}\right)(K-S)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{K}{KS-1}\right)$$

where,

S equals $\sqrt{\frac{Z_1}{Z_2}}$

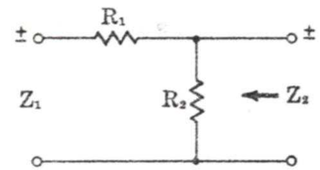


Fig. 114

(I) Pi-Type Attenuator (Between Equal Impedances)

$$R_1 = Z\left(\frac{K+1}{K-1}\right)$$

$$R_2 = \left(\frac{Z}{2}\right)\left(\frac{K^2-1}{K}\right)$$

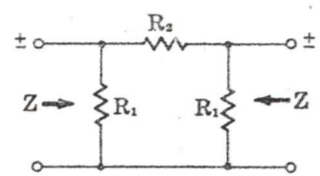


Fig. 115

(J) Pi-Type Attenuator (Between Unequal Impedances)

$$R_1 = Z_1\left(\frac{K^2-1}{K^2-2KS+1}\right)$$

$$R_2 = \left(\frac{\sqrt{Z_1 Z_2}}{2}\right)\left(\frac{K^2-1}{K}\right)$$

$$R_3 = Z_2\left(\frac{K^2-1}{K_2 - 2\frac{K}{S} + 1}\right)$$

where,

S equals $\sqrt{\frac{Z_1}{Z_2}}$

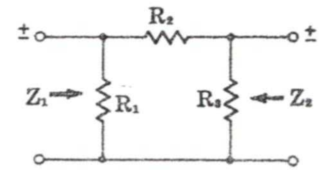


Fig. 116

(K) O-Type Attenuators

Calculate the values for a pi-type attenuator (Figs. 115 and 116), then halve the values for the series resistors as shown in Figs. 117 (balanced) and 118 (unbalanced).

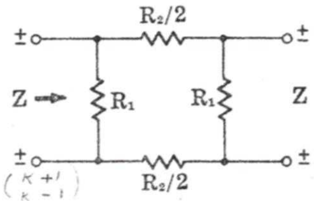


Fig. 117

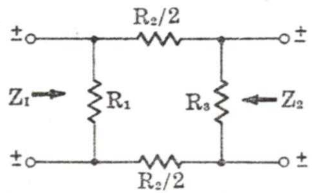


Fig. 118

(L) U-Type Attenuator

For impedance match in the direction of the series arms:

$$R_1 = \left(\frac{Z_1}{2S}\right)\left(\frac{KS - 1}{K}\right)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{1}{K - S}\right)$$

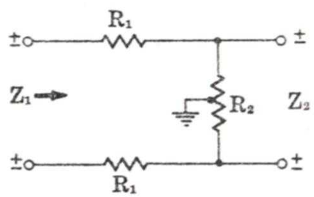


Fig. 119

For impedance match in the direction of the shunt arm:

$$R_1 = \left(\frac{Z_1}{2S}\right)(K - S)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{K}{KS - 1}\right)$$

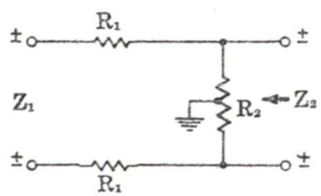


Fig. 120

where,

The arrows indicate the direction of the impedance match,

S equals $\sqrt{\frac{Z_1}{Z_2}}$

(M) Lattice-Type Attenuator

$$R_1 = \left(\frac{K - 1}{K + 1}\right) Z$$

$$R_2 = \left(\frac{K + 1}{K - 1}\right) Z$$

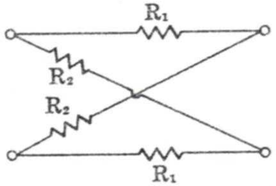


Fig. 121

(N) Ladder-Type Attenuator

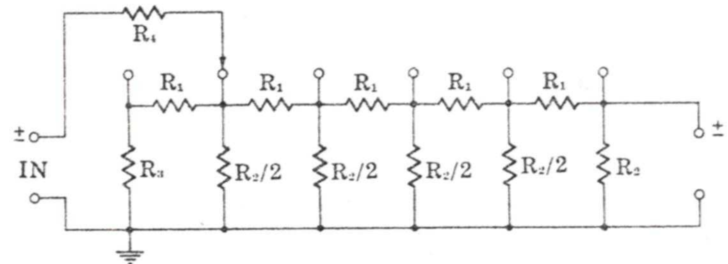


Fig. 122

$$R_1 = \left(\frac{K^2 - 1}{2K}\right) Z$$

$$R_2 = \left(\frac{K + 1}{K - 1}\right) Z$$

$$R_3 = \frac{R_2 \times Z}{R_2 + Z}$$

$$R_4 = \frac{Z}{2}$$

$$Z_{in} = Z_{out}$$

where,

K depends on the loss per step—not on the total loss.