"*Spirit of Electricity.*"

Principles of Electricity

applied to

Telephone and Telegraph Work

A Training Course Text

Prepared for Employees of the

Long Lines Department

AMERICAN TELEPHONE AND TELEGRAPH COMPANY

November, 1938

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PREFACE

THIS book is the outgrowth of certain less formal notes used for training purposes in the Long Lines Department in earlier years. Its origin may be traced to a set of such notes bearing the title, "Elements of Electricity Applied to Telephone and Telegraph Work". These were issued in 1922 in response to a long-standing demand by Long Lines employees for text material that would demonstrate the fundamental principles of electricity and magnetism by applications to the electrical circuits and apparatus with which they worked.

Since that time technical developments in communications work have called for revisions at varying intervals, and the demand for the book has increased steadily. By 1928 the annual requirements justified its printing for the first time. Another edition, somewhat revised, was printed in 1929 and, at the instance of the Department of Operation and Engineering, was made available to all Bell System Companies and regularly carried in stock by the Western Electric Company. A more extensive revision was made in 1930.

The present edition attempts to bring within the scope of the text some of the more important technical advances that have been made since 1930. This has resulted in a substantial increase in the size of the volume. Seven chapters have been added and ten other chapters have been materially revised or enlarged.

Because of the variety and scope of the subjects covered, the book is necessarily rather voluminous, although every effort has been made to treat each subject taken up as briefly as is consistent with a reasonably adequate presentation of the theory and fields of application involved. On the other hand, the book is not intended to be in any sense a complete treatise on electrical communication. Its subject is electrical theory and such descriptions of telephone or telegraph equipment and circuits as have been included are employed primarily for illustrating some of the applications of this theory to practice. The purpose has been to cover the essential general principles of simple electrical theory and to illustrate each principle briefly by one or more of its outstanding applications, rather than to duplicate the field of the technical instructions and specifications to be found in every telephone office.

The use of higher mathematics has been avoided entirely, and even the more elementary branches have been used as sparingly as possible. A general knowledge on the part of the reader is assumed of only those branches of mathematics ordinarily taught in High Schools, including Algebra, Geometry, Logarithms and Trigonometry. There is a slight departure, however, in the chapters dealing with the solution of alternating-current circuits and with transmission theory. Here it has been thought desirable to make use of Vector Notation. Though this may involve the introduction of certain simple mathematical concepts with which some readers are not familiar, the great simplification that is thus effected more than justifies the time spent in learning these new concepts.

For anyone who has difficulty in following the derivation of formulas or in solving illustrative problems, there is available a booklet of mathematical notes which explains in a brief and simple manner the essentials of all branches of mathematics used in the text. In some cases it may be advisable for the reader to review these notes along with his study of this text, taking up each item as he needs it. A knowledge of the more elementary principles of Physics and Mechanics is also assumed, but for anyone wishing to review these subjects hurriedly, an Appendix is included giving the important fundamental definitions and concepts.

A word of caution is perhaps needed regarding the circuit drawings, tables, and other statistical data included at various points in the text. The circuit drawings are used primarily as a means of illustrating the principles under discussion. While they are reasonably representative of current practice, they may or may not conform in detail with any actual situation. Similarly, the tables and other data represent the best information available at this time, but they are subject to change and are not intended as a substitute for current data as issued in formal instructions.

C. F. Myers,
Supervisor of Instruction,

L. S. Crosby,
General Personnel Supervisor.

32 Sixth Avenue,
New York City,
November, 1938.

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Principles of Electricity

applied to

Telephone and Telegraph Work

CHAPTER I

ELEMENTARY DEFINITIONS AND OHMS LAW

1. Introductory

Electricity is well adapted to transmitting from one place to another and delivering in convenient form quantities of energy, either large or small. It enables the power engineer to harness the energy of an isolated waterfall and to transmit it to some distant city where it may be utilized in the form of heat, light, mechanical work, or to change the state of certain chemicals. Likewise, it enables the telephone engineer to transmit the human voice thousands of miles without loss of intelligibility. Although in accomplishing such feats as these electricity is used with precision, our knowledge of its exact nature is limited to the now generally accepted, though not entirely complete, electron theory.

According to this theory the electron is the smallest possible charge of electricity, just as an atom is the smallest possible chemical particle of any substance. Electrons are all identical and each has a definite negative charge and a definite mass. By means of various ingenious methods some of which involve isolating individual electrons, these values have been carefully measured. The charge is found to be about 4.79×10^{-10} electrostatic units and the mass about $\frac{1}{1850}$ part of that of the hydrogen atom. These values are so infinitesimally small as to be almost meaningless to anyone except a trained physicist.

All substances are made up of electrons and corresponding positive particles called protons. Much less is known about protons than about electrons; it has been determined, however, that they are of approximately the same size as electrons but nearly two thousand times as heavy. The most widely accepted theory of the structure of the atom postulates that it is made up of a nucleus consisting of one or more protons surrounded by an equal number of electrons, the number and arrangement of the positive and negative charges being different for each chemical element. The simplest and lightest element, hydrogen, is made up of a nucleus consisting of a single proton around which a single electron revolves in certain fixed orbits. Heavier elements have nuclei consisting of a number of protons held together and partially neutralized by about half as many electrons and the remaining half of the electrons of the atom revolve about the nucleus in various orbits. In every case the number of electrons revolving in orbits about the nucleus gives the **atomic number** of the element and the number of protons contained in the nucleus gives the **atomic weight**. The electron has the properties of a wave as well as of a corpuscle and it is therefore not entirely accurate to think of it as being located at a point in space. Nevertheless some idea of its status as a part of the atom-model that we have been considering may be obtained if we note that its magnitude and space relationships relative to the atom as a whole are comparable to those of the earth in relation to the solar system. Meanwhile we must remember that the atom itself is almost inconceivably minute.

Electrons are attracted toward the atom nucleus and are repelled by one another with tremendous forces relative to their magnitude—forces infinitely greater than the gravitational forces with which we are familiar. For this reason the electrons in the atoms are for the most part held permanently in place in fixed orbits around the atom nucleus, but in the atoms of many materials one or more of the electrons farthest out from the nucleus is attached rather loosely and may by various means be drawn away from the atom altogether. When this happens to a number of the atoms making up a substance, as for instance a piece of metal, it contains less than its normal quota of electrons and is said to be positively charged. At the same time something else must be negatively charged or contain more than its normal number of electrons, for those taken away from the original substance must of course go somewhere. The means of bringing about such a condition are too numerous to mention, although we will consider several of them in later chapters.

The electron theory explains the flow of current in a conductor as being merely a stream of electrons moving along the conductor from atom to atom in a definite direction under the influence of an outside applied force or pressure. Substances whose atoms have loosely attached outer electrons are good conductors while substances to whose atoms all electrons are tightly bound contain normally very few free electrons and are therefore poor conductors, or good insulators.

While in our study of vacuum tubes in a later chapter we will deal with electrons as such, for most of our purposes we will not need to be familiar with all the details of electron theory, nor to know exactly what electricity is. Though we cannot observe it any more than we can actually see the force of gravity, we can observe its effect on other things about us. In this way we associate it with skilfully constructed mechanisms that are set in motion at the throw of a switch or the touch of a button, and with forms of energy that may be conveyed from place to place and changed from one state to another. We learn the conditions under which certain chemicals, or work performed in a mechanical way, can produce energy in an electrical form and how, through means of intelligently controlling it, we may employ it for practical purposes. In so far as this text is concerned, **our chief interest in electricity lies—first, in the many convenient ways in which it can be produced; second, in the means of transmitting it; and third, in the simple methods by which it may in turn produce active forces.**

In what follows, then, we will be principally concerned with the study of the more important laws of electrical circuits which have been deduced from observation and with certain of the practical applications of these laws. For a proper understanding of the electrical quantities with which we will deal it is desirable that the reader have a general knowledge of the more fundamental physical quantities and for the benefit of any one who may wish to refresh his memory regarding these matters a brief review of elementary physics may be found in Appendix 1.

2. The Electrical Circuit

An electrical circuit in its simplest form consists of a source of electromotive force and a continuous conducting path through a resistance from the positive terminal to the negative terminal. The source of electromotive force may be direct or alternating. If direct, the positive and negative terminals remain unchanged, but if alternating, their polarity is changed or reversed at periodic intervals. Accordingly, the study of electricity is usually divided into two parts; first, that dealing with circuits having sources of direct electromotive force, commonly called **direct current circuits**; and second, that dealing with circuits having sources of alternating electromotive force, commonly called **alternating current circuits.**

3. Electrical Pressure or Electromotive Force

The flow of electricity through a circuit is analogous in many respects to the flow of water through a closed system of pipes. Figure 1 shows a simple electrical

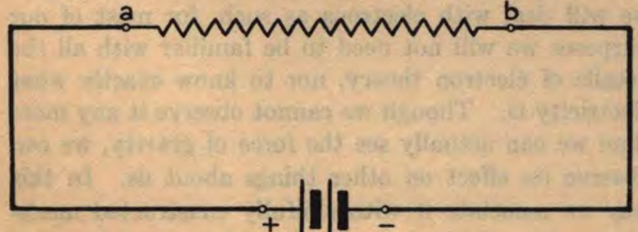


Fig. 1. Simple Electrical Circuit

circuit consisting of a battery connected to a resistance *ab*. Figure 2 shows a simple water circulating system. In the water mechanism, the pump creates a difference in pressure between the points *a* and *b*. This difference in pressure, or "pressure head", will cause water to

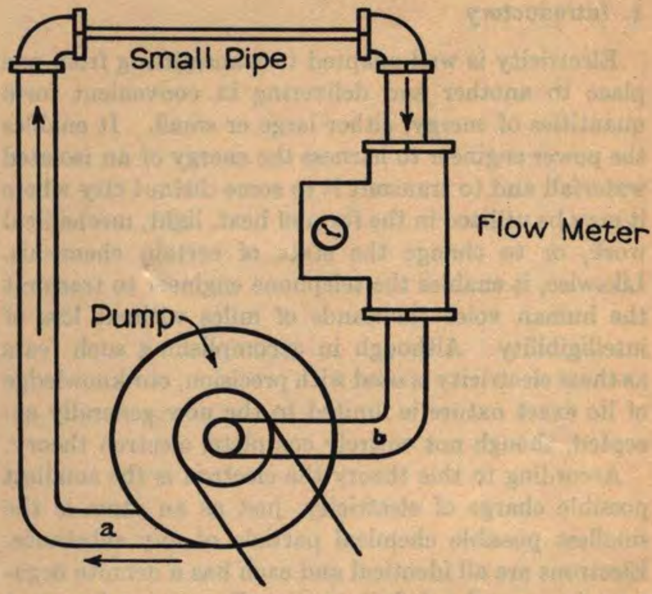


Fig. 2. Water Circulating System Analogous to Simple Electrical Circuit

flow from the outlet pipe *a*, through the small pipe to the flow meter, and return to the low pressure side of the pump at *b*. The amount of water that will flow will depend upon this difference in pressure and upon the nature of the small pipe. In the electrical circuit, the battery supplies the electrical pressure or electromotive force which causes electricity to flow from the high potential side of the battery. The amount of electricity that will flow depends upon this electromotive force and the nature of the resistance.

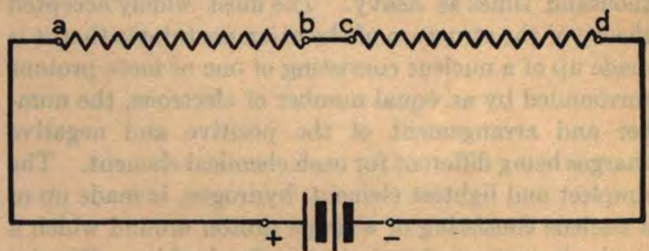


Fig. 3. Series Circuit

If a differential pressure gage were connected between the points *a* and *b* in the water system, it would register the difference in water pressure in some suitable unit such as "difference of head in feet". The electromotive force of the electrical circuit, on the other hand, is measured in terms of a unit called the **volt.**

It may be noted at this point that the terms electromotive force and difference in potential are commonly used synonymously. There is technical distinction, however, in that an electromotive force is always established by a battery or other primary source of electrical energy, whereas a difference in potential exists between any two points of a conductor through which current is flowing.

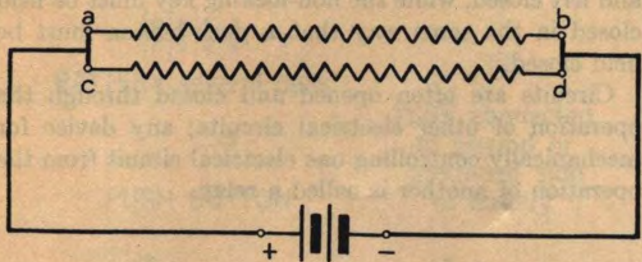


Fig. 4. Parallel Circuit

4. Resistance

In Figure 2, if the small pipe is made longer the flow of water will be decreased although the pump maintains the constant difference in pressure between the points *a* and *b*. Also, if the small pipe is decreased in size the flow of water will likewise be decreased. Though there is no simple unit for measuring this resistance to flow of water in a pipe, it is analogous to an electrical resistance in many respects. The unit of electrical resistance is called the **ohm** and is defined as **the resistance offered to electrical flow by a column of mercury one square millimeter in cross-section and 106.3 centimeters long at a temperature of zero degrees Centigrade.**

5. Current

In our water circulating mechanism we can describe the **rate of flow,** or the current, as the amount of water being circulated in gallons per second. In electrical work the current is expressed in amperes. **The measure of one ampere is the current which when passed through a solution of nitrate of silver between two silver plates under fixed conditions will cause a deposit due to electrolytic action of 0.001118 gram of silver per second.**

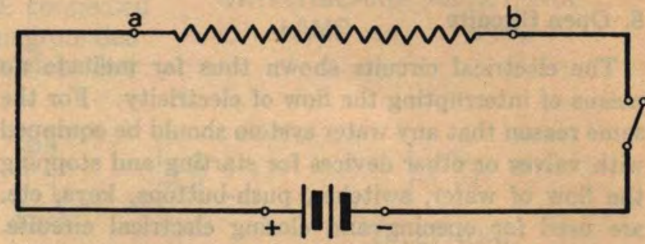


Figure 5

6. The Volt

The volt has been named as the unit of electrical pressure but its size has not been defined. **A source of electromotive force is said to have one volt of electrical pressure when it will establish a current of one ampere in a resistance of one ohm.**

7. Series and Parallel Circuits

A simple circuit may contain any number of resistances. Figure 3 shows such a circuit with two resis-

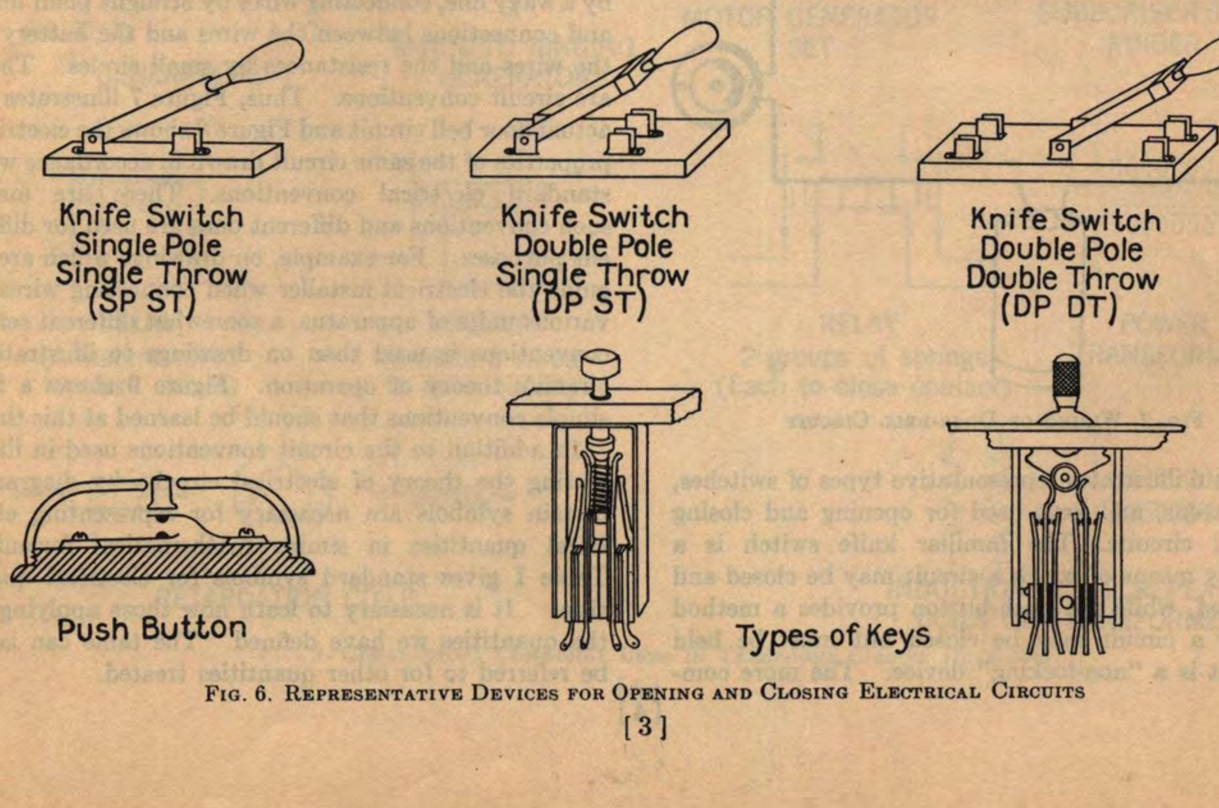


Fig. 6. Representative Devices for Opening and Closing Electrical Circuits

tances which when connected as shown are said to be "in series". Figure 4 shows another circuit with the same resistances connected "in parallel". Any number may be so connected in either case.

The current from a battery in a parallel circuit will divide between the various resistance branches but in a series circuit, as in the flow of water in a single pipe, it cannot divide and must be identical at every point. In other words, it must have an unchanged value in all parts of the circuit from the positive to the negative terminal of the battery.

8. Open Circuits

The electrical circuits shown thus far indicate no means of interrupting the flow of electricity. For the same reason that any water system should be equipped with valves or other devices for starting and stopping the flow of water, switches, push-buttons, keys, etc. are used for opening and closing electrical circuits. Figure 5 shows a circuit opened by means of a switch. Its metallic continuity is interrupted by the switch and when so interrupted there is no flow of electricity. This protects the source of electromotive force against unnecessary losses since when the circuit is open it cannot absorb any energy.

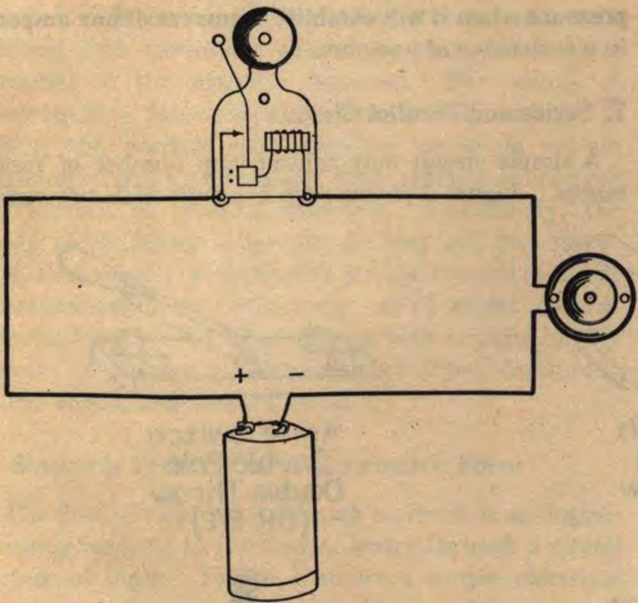


Fig. 7. Wiring of Door-bell Circuit

Figure 6 illustrates representative types of switches, push-buttons, and keys used for opening and closing electrical circuits. The familiar knife switch is a device by means of which a circuit may be closed and left closed, while the push-button provides a method whereby a circuit may be closed but must be held closed; it is a "non-locking" device. The more common designs of circuit closing apparatus used in telephone and telegraph work are called **keys.** When the circuit to be opened or closed does not carry an excessive current, these will perform the corresponding functions of the knife switch and the push-button; that is, they may be either locking or non-locking. The locking key may be operated or closed and left closed in the same way that the knife switch may be closed and left closed, while the non-locking key must be held closed in the same way that a push-button must be held closed.

Circuits are often opened and closed through the operation of other electrical circuits; any device for mechanically controlling one electrical circuit from the operation of another is called a **relay.**

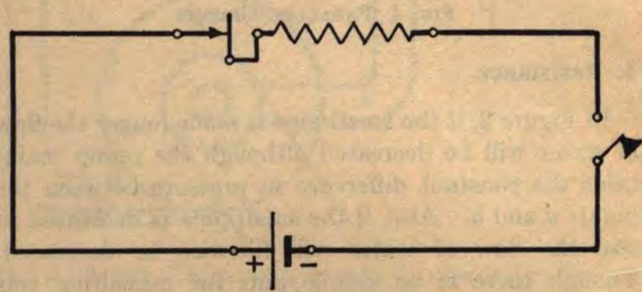


Fig. 8. Conventional Drawing for Circuit of Fig. 7

9. Electrical Symbols and Circuit Conventions

In the foregoing circuit diagrams we have represented the battery with a long and a short line, a resistance by a wavy line, connecting wires by straight plain lines, and connections between the wires and the battery or the wires and the resistances by small circles. These are circuit conventions. Thus, Figure 7 illustrates an actual door bell circuit and Figure 8 shows the electrical properties of the same circuit drawn in accordance with standard electrical conventions. There are many such conventions and different ones are used for different purposes. For example, on drawings which are to guide the electrical installer when connecting wires to various units of apparatus, a somewhat different set of conventions is used than on drawings to illustrate a circuit's theory of operation. Figure 9 shows a few simple conventions that should be learned at this time.

In addition to the circuit conventions used in illustrating the theory of electrical circuits by diagrams, certain symbols are necessary for representing electrical quantities in simple mathematical formulas. Table I gives standard symbols for electrical quantities. It is necessary to learn now those applying to the quantities we have defined. The table can later be referred to for other quantities treated.

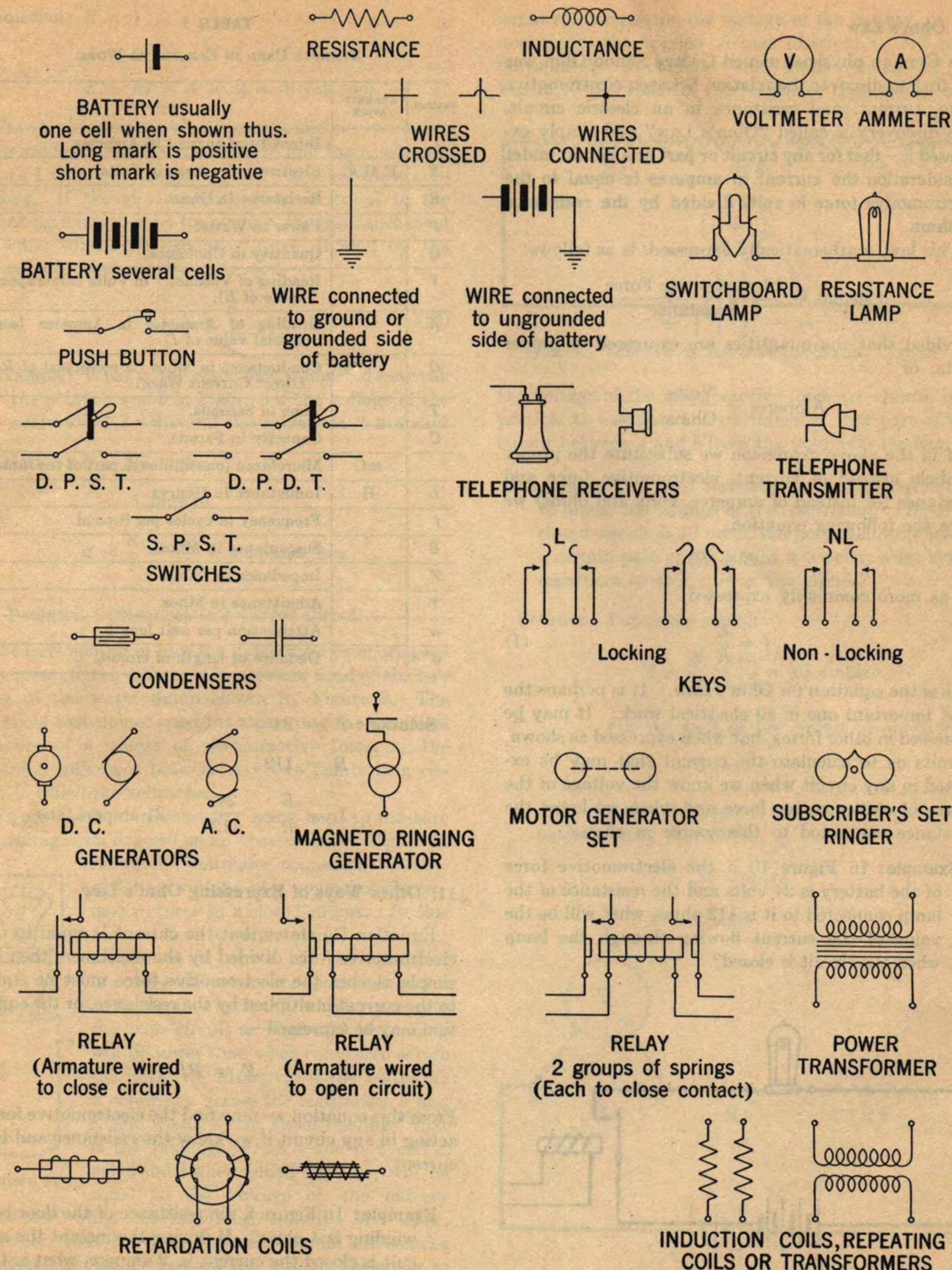


FIG. 9. CONVENTIONS COMMONLY USED IN TELEPHONE CIRCUIT DRAWINGS

10. Ohm's Law

A German physicist named George Simon Ohm was the first to discover the relation between electromotive force, current, and resistance in an electric circuit. The discovery is called "Ohm's Law" and simply expressed is—**that for any circuit or part of a circuit under consideration the current in amperes is equal to the electromotive force in volts divided by the resistance in ohms.**

This law, mathematically expressed, is as follows:

$$\text{Current} = \frac{\text{Electromotive Force}}{\text{Resistance}}$$

provided that the quantities are expressed in proper units, or

$$\text{Amperes} = \frac{\text{Volts}}{\text{Ohms}}$$

If in the above expression we substitute the proper symbols instead of current, electromotive force and resistance (or instead of amperes, volts and ohms) we have the following equation:

$$I = E \div R,$$

or, as more commonly expressed,

$$I = \frac{E}{R} \tag{1}$$

This is the equation for Ohm's Law. It is perhaps the most important one in all electrical work. It may be expressed in other forms, but when expressed as shown, permits us to calculate the current that may be expected in any circuit when we know the voltage of the source of electromotive force and when we know the resistance connected to this source in ohms.

Example: In Figure 10 if the electromotive force of the battery is 24 volts and the resistance of the lamp connected to it is 112 ohms, what will be the value of the current flowing through the lamp when the circuit is closed?

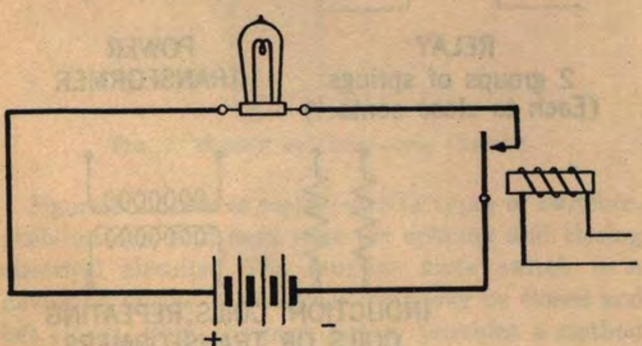


FIGURE 10

TABLE I

SYMBOLS USED IN ELECTRICAL WORK

SYMBOL	ABBREVIATION	STANDS FOR
I		Intensity of current in Amperes.
E	E.M.F.	Electromotive force in Volts.
R		Resistance in Ohms.
P		Power in Watts.
Q		Quantity in Coulombs.
V		Reading of Voltmeter in Volts (some special value of E).
A		Reading of Ammeter in Amperes (some special value of I)
G		Conductance in Mhos (is reciprocal of R in Direct Current Work).
T		Time in Seconds.
C		Capacity in Farads.
	mf.	Microfarad (one millionth part of the farad).
L	H	Inductance in Henrys.
f		Frequency in cycles per Second.
B		Susceptance in Mhos.
Z		Impedance in Ohms.
Y		Admittance in Mhos.
α		Attenuation per unit length.
d		Distance or length of circuit.

Solution: $E = 24$

$R = 112$

$$I = \frac{E}{R} = \frac{24}{112} = .21 \text{ ampere, ans.}$$

11. Other Ways of Expressing Ohm's Law

Equation (1) states that the current is equal to the electromotive force divided by the resistance; then by simple algebra the electromotive force must be equal to the current multiplied by the resistance, or the equation may be expressed—

$$E = RI \tag{2}$$

From this equation we may find the electromotive force acting in any circuit if we know the resistance and the current.

Example: In Figure 8 the resistance of the door bell winding is 4 ohms. If during the instant the circuit is closed the current is .2 ampere, what is the voltage of the dry cell?

Solution: $R = 4$

$I = .2$

$E = RI = 4 \times .2 = .8$ volt, ans.

The third case is one where current and electromotive force are known and it is desired to find the resistance. Ohm's Law may likewise be stated to cover these conditions. If the electromotive force is equal to the resistance multiplied by the current, the resistance must be equal to the electromotive force divided by the current or, algebraically expressed—

$$R = \frac{E}{I} \qquad (3)$$

Example: What is the resistance connected between the points *a* and *b* in Figure 5 if the voltage of the battery is 1.3 volts and the current is .5 ampere?

Solution: $E = 1.3$ volts

$I = .5$ ampere

$$R = \frac{E}{I} = \frac{1.3}{.5} = 2.6 \text{ ohms, ans.}$$

12. Potential Differences in a Closed Circuit

We have spoken of how the differential pressure gage may measure the difference in pressure head of the two sides of the water pump shown by Figure 2. The electrical instrument used for measuring the electrical pressure of a source of electromotive force, or the potential difference between any two points in a circuit, is called the **voltmeter.**

Figure 11 shows a voltmeter being used to measure the voltage of a dry cell on an open circuit. Figure 12 shows the voltmeter connected to measure the voltage of a source of electromotive force in a closed circuit. In this case we have a simple circuit with three resistances in series. If the voltmeter is connected across the points *a* and *b* as shown in Figure 13, which represents the same circuit as Figure 12, its reading will be lower than when connected across the battery. Moreover if the voltmeter is connected across the resistances *b* and *c*, and *c* and *d*, the three readings, that is the readings across *a* and *b*, *b* and *c*, and *c* and *d* when added together, will be equal to the voltage of the battery (measured while the circuit is closed). We learn, therefore, that **the sum of the potential differences measured across all parts of the circuit, beginning at the positive pole of the battery and returning to the negative, is equal to the voltage of the battery,** or we might say, the applied voltage distributes itself proportionately throughout the series circuit. If in Figure 13 the value of the resistance from *a* to *d* and

FIGURE 11

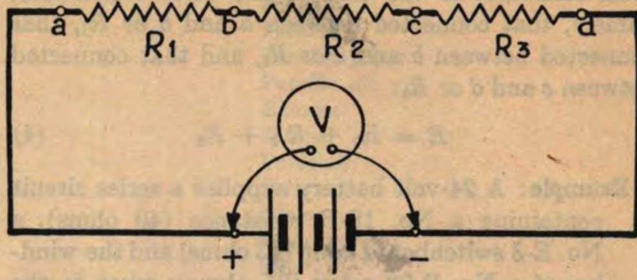


FIG. 12. VOLTAGE OF BATTERY ON CLOSED CIRCUIT

the voltage of the electromotive force are known, it is possible to calculate the resistance of that part of the circuit between *a* and *b* from the voltmeter reading.

Example: The total resistance of a series circuit is 15 ohms, the voltage of the electromotive force on closed circuit is 10 volts, the potential drop across a certain part of the circuit is 3 volts; what is the resistance of this part of the circuit?

Solution: For entire circuit:—

$$I = \frac{E}{R} = \frac{10}{15} = .67 \text{ ampere.}$$

For the part of the circuit in question—

$E = 3$ volts.

I of series circuit is same in any part of circuit as for entire circuit, therefore,

$I = .67$ ampere

$E = 3$ volts

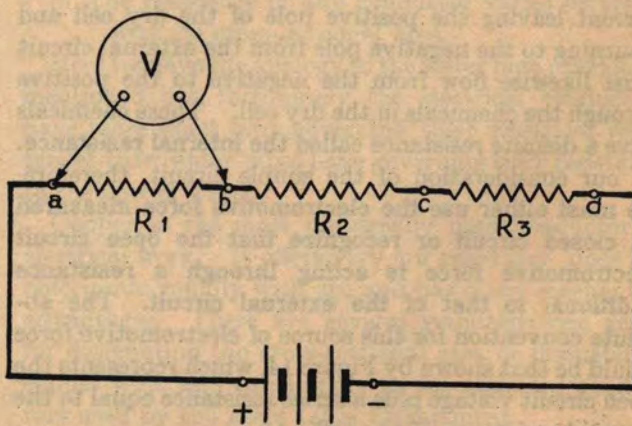


FIG. 13. VOLTAGE DROP ACROSS ONE RESISTANCE OF CLOSED CIRCUIT

$$R = \frac{E}{I} = \frac{3}{.67} = 4.5 \text{ ohms, ans.}$$

The total resistance of a series circuit is equal to the sum of all the individual resistances. In Figure 12 the total resistance is the sum of the three resistances; namely, that connected between *a* and *b* or R_1, that connected between *b* and *c* or R_2, and that connected between *c* and *d* or R_3:

$$R = R_1 + R_2 + R_3 \qquad (4)$$

Example: A 24-volt battery supplies a series circuit containing a No. 18-B resistance (40 ohms), a No. E-3 switchboard lamp (43 ohms) and the winding of a No. B-22 relay (95 ohms); what is the total resistance of the circuit and what current will flow through the switchboard lamp?

Solution: $R = R_1 + R_2 + R_3$

$$= 40 + 43 + 95 = 178 \text{ ohms, ans.}$$

$$E = 24 \text{ volts}$$

$$I = \frac{E}{R} = \frac{24}{178} = .13 \text{ ampere, ans.}$$

13. Internal Resistance

If a dry cell, as shown in Figure 11, is placed in a closed circuit like that of Figure 3 and its voltage again measured with a voltmeter, a reading will be obtained which will be somewhat less than the reading on open circuit. This means that the electromotive force of the dry cell depends to some extent upon the value of the current it is furnishing. As the current is increased the electromotive force is decreased. This is due to a potential drop within the cell itself, which is merely a drop across a resistance in the same way that the potential measured across the terminals *ab* in Figure 12 is a drop across a resistance, excepting that in this case the resistance is inside the dry cell. Any electrical current leaving the positive pole of the dry cell and returning to the negative pole from the external circuit must likewise flow from the negative to the positive through the chemicals in the dry cell. These chemicals have a definite resistance called the **internal resistance.** In our consideration of the simple circuit, therefore, **we must either use the electromotive force measured on closed circuit or recognize that the open circuit electromotive force is acting through a resistance additional to that of the external circuit.** The absolute convention for this source of electromotive force would be that shown by Figure 14, which represents the open circuit voltage plus a series resistance equal to the internal resistance of the cell.

The ordinary dry cell has an internal resistance averaging about one ohm, but this greatly increases with the aging of the cell. In the telephone central office where storage batteries are used almost exclusively, the internal resistance is negligible for most direct current considerations.

14. Electrical Power

In the simple circuits we have thus far considered we have only dealt with resistance, electromotive force, and electrical current, but each of these circuits is actually converting energy from chemical to heat or some other form. They, therefore, have a definite power consumption or represent a definite transfer of power to some external device.

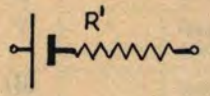


FIGURE 14

The scientific unit for work is the joule, equal to about $\frac{3}{4}$ths of one foot-pound, and the scientific unit for rate of doing work or power is the watt, which is equal to about $\frac{3}{4}$ths of one foot-pound per second. The electrical units have been so derived as to facilitate convenient calculations in transforming expressions for power and energy from mechanical to electrical units. **In the electrical circuit if we multiply the electromotive force in volts by the current in amperes we have an expression for the power in watts.** The watt may, therefore, be defined as an electrical unit as well as a mechanical unit and is the power expended in a circuit having an electromotive force of one volt and a current of one ampere.

Because the watt is the connecting relation between mechanical units and electrical units, its value in terms of horsepower should be committed to memory (see Appendix I), and the following formula should be considered second only to Ohm's Law in importance:

$$P = EI \qquad (5)$$

A somewhat more convenient form for determining the power expended in any given resistance is—

$$P = I^2R \qquad (6)$$

This latter equation is apparent from Ohm's Law, which states that $E = IR$ and we may, therefore, substitute IR for E in Equation (5), which gives us I^2R.

Example: In Figure 12, what is the power expended in the resistance between terminals *a* and *b* if the potential difference is equal to 10 volts and the resistance is 5 ohms?

Solution: $P = EI$, and

$$I = \frac{E}{R} = \frac{10}{5} = 2,$$

then $P = EI = 10 \times 2 = 20$ watts, ans.

15. Quantity of Electricity

In Figure 2 we may say that the amount of water that will pass through the small pipe in a given interval of time is a definite number of gallons; thus the gallon is a unit of quantity of water. The amount of electricity that flows through an electrical conductor in one second when the current intensity (or rate of flow) is one ampere is called a **coulomb.** This is the unit for measuring quantities of electricity.

TABLE II

RESISTIVITY OF VARIOUS METAL CONDUCTORS AT 0°C.
(Compared to pure copper of same length and cross section)

KIND OF METAL	TIMES THE RESISTANCE OF PURE COPPER
Silver	.941
Copper (pure)	1.000*
Copper (annealed)	1.018
Copper (hard drawn)	1.025
Gold	1.423
Aluminum	1.679
Magnesium	2.788
Zinc	3.449
Tungsten (hard drawn)	3.474
Nickel	4.442
Iron (pure)	5.673
Platinum	6.301
Tin	6.730
Steel (soft)	7.564
Tantalum	9.359
Lead	12.692
German Silver	21.218
Steel (hard)	29.294
Mercury	60.301
Cast Iron (hard)	62.692

* Resistivity of pure copper 1.56 microhms per cm^3.

16. Properties of Electrical Conductors

A column of mercury was used to define the standard unit of resistance, the ohm. Other metals could have been used for this fixed standard but their dimensions would have been different from that of mercury. Dr. Ohm investigated the conducting properties of various kinds of metals and called those offering very high resistance to the flow of electricity "poor conductors"

TABLE III

INSULATING MATERIALS
(Given in the order of their approximate insulating properties)

Dry air
Shellac
Paraffin
Paraffin paper
Paraffin oil
Ebonite
Rubber
Porcelain
Sulphur
Glass
Mica
Silk
Varnish
Dry paper
Celluloid
Dry wood
Slate
Fiber
Distilled water
Alcohol

and those offering comparatively little resistance to the flow of electricity "good conductors". There is another classification for material having extremely high resistance, in fact so high as to give an open circuit for all practical purposes. These are called **insulators.**

Table II shows a few conductors in the order of conductivity. Those offering the least resistance are at the top of the list. Table III shows a list of materials which are commonly used as insulators. There are many other good insulators but they are not all adaptable for use as such in practice.

In addition to the law showing the relation between electromotive force, current, and resistance, Ohm investigated the properties of conductors and established in addition to their relative values the following laws:

a. The resistance of any conductor varies directly with its length.
b. The resistance of any conductor varies inversely with its cross-sectional area.

Here we have the analogy to the water pipe previously mentioned but fortunately the electrical conductors have more exact laws governing their electrical resistances than water pipes have governing their resistance to the flow of water.

Copper is the most universally used conductor in electrical work. It offers very low resistance, does not deteriorate rapidly with age and has many mechanical advantages. There are several standard wire gages for designating the cross-sectional area or diameter of copper wire, and three apply to the standard conductors used by the Long Lines Department.

Table IV shows the standard gages of wire used by the Long Lines Department and their resistance values.

TABLE IV

ELECTRICAL PROPERTIES OF COPPER CONDUCTORS STANDARDIZED BY LONG LINES DEPARTMENT

CONDUCTORS	NO.	SIZE		WEIGHT	RESISTANCE		INDUCTANCE	CAPACITY
		Gage	Diameter in Inches	Lbs. per Wire Mile	*Ohms per Loop Mile	Ohms per 1,000 feet (single wire)	Henrys per Loop Mile	Mfs. per Loop Mile
Open Wire (12-inch spacing)	8	B.W.G.	.165	435	4.02 (use 4)	.381	.00337	.00915
	10	N.B.S.G.	.128	264	6.68	.632	.00353	.00871
	12	N.B.S.G.	.104	174	10.12 (use 10)	.959	.00366	.00837
Cable (side circuits of standard quadded cable)	10	A.W.G.	.102	168	10.55	.999	.001	.062
	13	A.W.G.	.072	82.6	21.15	2.003	.001	.062
	16	A.W.G.	.051	41.2	42.41	4.016	.001	.062
	19	A.W.G.	.036	20.5	85.01	8.05	.001	.062
	22	A.W.G.	.025	10.2	170.44	16.14	.001	.062

* These resistance values are for 20° C or 68° F; add 2/10 of 1% per degree Fahrenheit for higher temperatures.
Note: A.W.G. is American Wire Gage and is same as B. & S. which is Brown and Sharpe Gage. B.W.G. is Birmingham Wire Gage and N.B.S.G. is New British Standard Gage.

Simple rules for remembering the approximate constants of the **cable conductors** are as follows:

a. Five sizes of cable conductors are standard for the Long Lines Department and all are A.W.G. (or B and S).

b. The largest size is #10 A.W.G. Add three gages for successive smaller sizes,—thus #10, #13, #16, #19 and #22.

c. The diameter of #10 A.W.G. is slightly greater than one-tenth inch and its resistance is slightly greater than ten ohms per loop mile.

d. Smaller sizes double resistance by the addition of each three gages beginning with #10 as a base.

e. In cables, conductors are slightly longer than the cable lengths due to the spiraling effect. This will average about 5%.

f. Three sizes of conductors are standard for open wire; 104 (#12 N.B.S.G.), 128 (#10 N.B.S.G.) and 165 (#8 B.W.G.)

g. A #10 is the nearest A. W. Gage to 104 (#12 N.B.S.G.) but is slightly smaller.

CHAPTER II

THE SOLUTION OF D.C. NETWORKS

17. Parallel Circuits

Figure 15 shows two resistances connected in parallel. If we apply Ohm's Law to either of these, we shall find that the current in it must be equal to the potential measured across the particular resistance divided by its value in ohms; and for this particular circuit, the potential measured across either resistance is the potential of the battery. The battery is in reality supplying two currents, one through the resistance *ab* and the other through the resistance *cd*. These two currents are united and flow together in the conductors connecting the poles of the battery with the junctions of the two resistances. Likewise, for any circuit having two resistances connected in parallel, the current supplied to the combination must be greater than the current supplied to either of the resistances. If we think of the combination of resistances in Figure 15 as equivalent to a single resistance that might be substituted in their stead, we may say accordingly that the value in ohms of two resistances in parallel is less than that of either resistance taken singly.

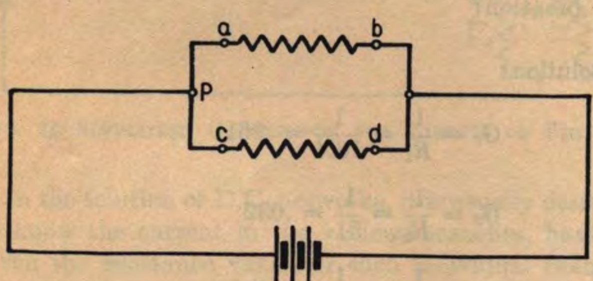


FIG. 15. TWO RESISTANCES IN PARALLEL

We may make calculations for determining the current in a parallel circuit such as is shown by Figure 15, but these are more complicated than for a simple series circuit having more than one resistance, such as is shown in Figure 3. The solution of a parallel circuit is accomplished with the aid of Kirchoff's Laws in addition to Ohm's Law.

18. Kirchoff's First Law

Kirchoff's First Law states that at any point in a circuit there is as much current flowing to the point as there is away from it. This applies regardless of the number of branches that may be connected to the point in question. The law can be interpreted by its application to point P in Figure 15. If I is the current being supplied by the battery to the combination of the two resistances in parallel, and I_1 and I_2 are the respective currents through the two parallel resistances, then—

$$I = I_1 + I_2 \qquad (7)$$

If we apply Ohm's Law to the entire circuit and let R represent the value of **the combined resistances in parallel,** we have—

$$R = \frac{E}{I} \quad \text{or} \quad R = \frac{E}{I_1 + I_2}$$

But

$$I_1 = \frac{E}{R_1} \quad \text{and} \quad I_2 = \frac{E}{R_2}$$

Therefore,

$$R = \frac{E}{\dfrac{E}{R_1} + \dfrac{E}{R_2}}$$

But in this latter equation, the E's can be cancelled and the equation written—

$$R = \frac{1}{\dfrac{1}{R_1} + \dfrac{1}{R_2}}$$

and if we simplify this compound fraction by simple algebra—

$$R = \frac{R_1 R_2}{R_1 + R_2} \qquad (8)$$

This gives an equation for calculating the combined value of two parallel resistances. Expressed in words it may be stated as follows: **To obtain the combined resistance of any two resistances in parallel, divide their product by their sum.**

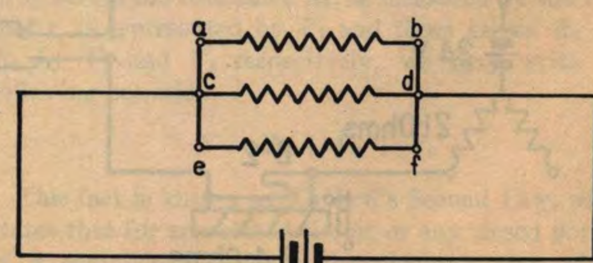


FIG. 16. THREE RESISTANCES IN PARALLEL

Example: What is the combined resistance of the inductive and non-inductive windings of a type-B relay used in a local A-board cord circuit if the inductive winding measures 16.4 ohms and the non-inductive winding measures 22 ohms?

Solution:

$$R = \frac{R_1 R_2}{R_1 + R_2} = \frac{16.4 \times 22}{16.4 + 22} = 9.4 \text{ ohms, ans.}$$

Figure 16 shows a circuit having three resistances in parallel. A formula similar to (8) can be worked out for combinations of this kind, or calculations can be made to obtain the combined resistance of *ab* and *cd* and this value then combined with *ef*. But for problems involving more than two resistances in parallel, it is usually simpler to use the conductance method.

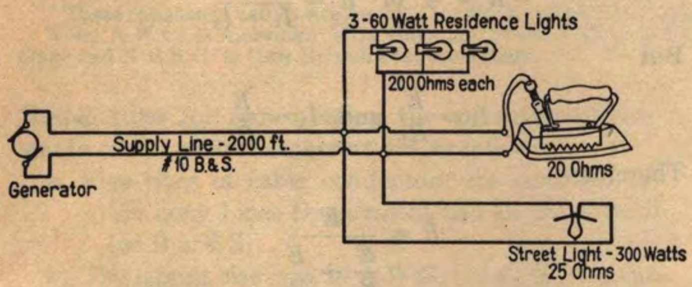


Fig. 17. Small Electrical Power System

19. Conductance

Conductance is defined in direct current work as the reciprocal of resistance. It is expressed by the symbol G, and for any single resistance—

$$G = \frac{1}{R} \tag{9}$$

For a combination of resistances in parallel, such as is shown by Figure 16, the conductance of the combina-

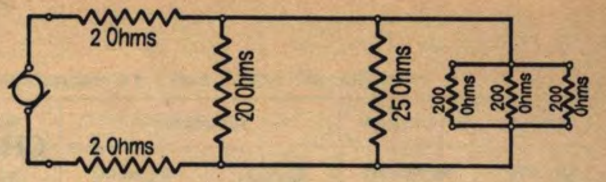


Fig. 18. Convention for Circuit of Fig. 17

tion is equal to the sum of the individual conductances, or—

$$G = G_1 + G_2 + G_3 \tag{10}$$

In a circuit having a number of resistances in parallel, it is often of advantage to solve for the total conductance of the circuit and then find its total resistance by taking the reciprocal of the total conductance.

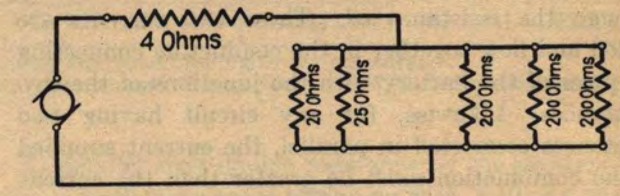


Figure 19

Example: If a B-3 relay has an inductive winding of 16.4 ohms, a non-inductive winding of 31 ohms, and these are shunted by an 18-U resistance (of 100 ohms), what is the resistance of the combination?

Solution:

$$G_1 = \frac{1}{R_1} = \frac{1}{16.4} = .061$$

$$G_2 = \frac{1}{R_2} = \frac{1}{31} = .032$$

$$G_3 = \frac{1}{R_3} = \frac{1}{100} = .010$$

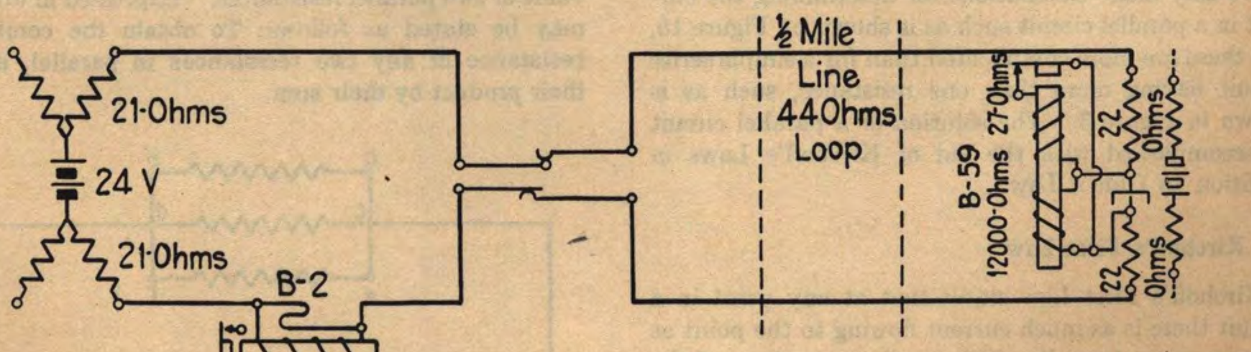


Fig. 20. Cord Circuit and Switching Trunk of Local Telephone Connection

$$G = G_1 + G_2 + G_3$$
$$= .061 + .032 + .010 = .103$$
$$R = \frac{1}{G} = \frac{1}{.103} = 9.7 \text{ ohms, ans.}$$

20. Direct Current Networks

Several resistances may be connected in such manner as to form very complicated networks. In practice many circuits are of this type. For example, Figure 17 illustrates a 110-volt power distribution line supplying a residence and a street light. We may represent the electrical characteristics of such a circuit by the network shown by Figure 18, and can further simplify this network as shown by Figure 19. Power supply systems are usually complicated networks of this sort.

In the same way, many telephone circuits may be analyzed by drawing their equivalent network diagrams. Figure 20 represents an A-board local cord circuit connected to a local switching trunk having ½ mile of 19-gage cable. The equivalent network is shown by Figure 21.

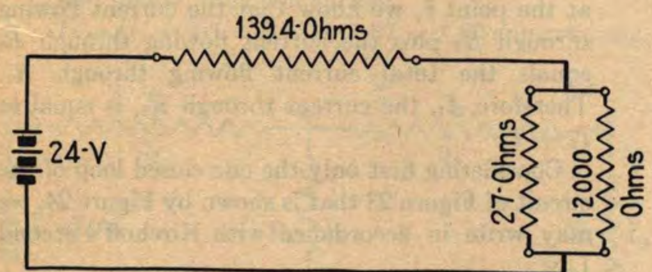


FIG. 21. SIMPLIFIED CONVENTION FOR CIRCUIT OF FIG. 20

In the solution of D.C. networks, it is usually desired to know the current in the various branches, having given the resistance values of each individual branch and the voltage of the source or sources of E.M.F.

Example: What is the value of the current through each winding of the B-59 relay in Figure 20?

Solution: We must first find the total current through both windings and have: $I = \frac{E}{R}$ where E is 24 volts and

$$R = 139.4 + \frac{R_1 R_2}{R_1 + R_2} = 139.4 + \frac{27 \times 12000}{27 + 12000}$$
$$= 139.4 + 26.9 = 166.3 \text{ ohms}$$

Then $I = \frac{24}{166.3} = .144$ ampere.

But the potential drop V across the two windings is equal to the current times the combined resistance of the two windings, or

$$V = I \times \frac{R_1 R_2}{R_1 + R_2}$$
$$= .144 \times 26.9$$
$$= 3.88 \text{ volts.}$$

Then, applying Ohm's Law to each winding independently, we have—

$$I_1 = \frac{V}{R_1} = \frac{3.88}{27} = .144 \text{ ampere, ans.}$$

and

$$I_2 = \frac{V}{R_2} = \frac{3.88}{12000} = .00032 \text{ ampere, ans.}$$

21. Kirchoff's Second Law

When a current flows through a resistance there is always a difference in potential between the ends of the resistance, the value of which depends upon the current flowing and the value of the resistance. This difference in potential is commonly called the IR **drop** since it is equal to the product of the current and the resistance. This IR drop acts in the opposite direction to, or opposes, the E.M.F. which drives the current through the resistance.

In a closed circuit, such as is shown in Figure 22, the sum of the IR drops across the three resistances must be equal to the impressed E.M.F. Thus if the

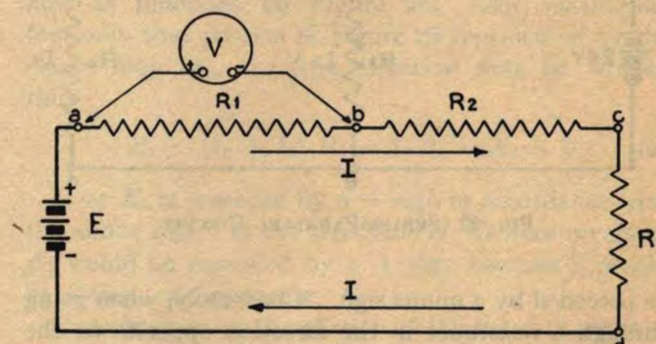


FIGURE 22

drop across the resistance R_1, as measured by the voltmeter, is represented by V_1 and those across R_2 and R_3 by V_2 and V_3 respectively, we may write the following equation—

$$E = V_1 + V_2 + V_3 \qquad (11)$$

This fact is known as Kirchoff's Second Law, which states that for any closed circuit or any closed portion of a complicated circuit, the algebraic sum of the E.M.F.'s and the potential drops is equal to zero.

In the case of Figure 22, Kirchoff's Second Law may be written as follows:

$$E - R_1I - R_2I - R_3I = 0 \qquad (12)$$

In solving any network problem, the first thing to do is to draw a good diagram. When the problem is to be solved by Kirchoff's Laws, the next step is to assign letters to all the unknowns in the circuit and to put arrows on the circuit diagram to indicate the **assumed** directions of current flow. If Kirchoff's First Law is applied at the junction points of a network, the number of unknowns may be kept down. Thus, if three wires meet at a point, and I_1 and I_2 have already been assigned to the currents in two of them, the third current may be designated as their sum or difference, depending upon the assumed direction of current flow. That is, instead of using a third unknown I_3, we will have $(I_1 + I_2)$ or $(I_1 - I_2)$. This will eliminate one equation. However, at least as many equations as there are unknowns must be written.

In the practical applications of Kirchoff's Laws, the correct use of algebraic signs is fundamentally important. When one sign has been given to the electromotive force in the direction of current flow, the opposite sign must be given to the IR drops in the direction of current flow. In other words, when going through a resistance in the same direction as the current flow, there is a drop in voltage and this voltage should

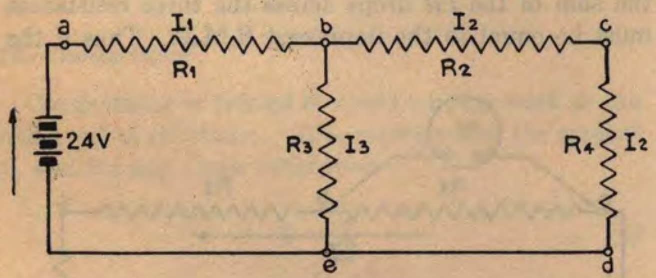


Fig. 23. Series-Parallel Circuit

be preceded by a minus sign. Conversely, when going through a resistance in the direction opposite to the current flow, there is a rise in voltage which should be preceded by a plus sign. We may for convenience accept the clockwise direction as positive, or accept as positive **all E.M.F.'s which tend to make a current flow in a clockwise direction, and as negative all potential drops due to this flow of current as well as any E.M.F.'s in the circuit tending to make current flow in the opposite direction.** It is immaterial whether the directions of current flow assumed are actually correct as long as they are consistent throughout the network. The signs of the answers will show whether or not the assumed directions are correct. When the value of a current found by solving the equations is preceded by a − sign, it merely means that the actual

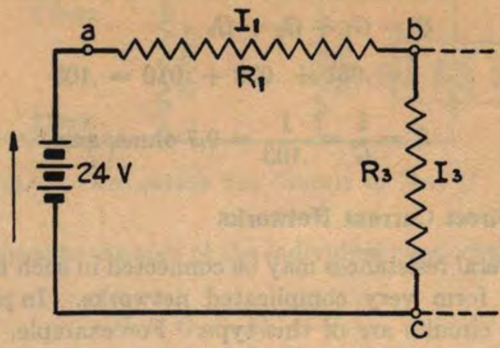


Figure 24

direction of flow is opposite to the direction which was assumed.

Example: Find the current values in each branch of Figure 23, if the resistance of $R_1 = 5$ ohms, $R_2 = 10$ ohms, $R_3 = 15$ ohms, and $R_4 = 20$ ohms, and the voltage $E = 24$ volts.

Solution: We may first assume that the direction of current flow is clockwise through both branches of the network. Applying Kirchoff's First Law at the point b, we know that the current flowing through R_3 plus the current flowing through R_2 equals the total current flowing through R_1. Therefore, I_1, the current through R_1, is equal to $I_2 + I_3$.

Considering first only the one closed loop of the circuit of Figure 23 that is shown by Figure 24, we may write in accordance with Kirchoff's Second Law:

$$E - R_1(I_2 + I_3) - R_3I_3 = 0 \qquad \text{(a)}$$

and for the closed loop shown by Figure 25—

$$E - R_1(I_2 + I_3) - R_2I_2 - R_4I_2 = 0 \qquad \text{(b)}$$

We thus have two independent equations containing two quantities which are unknown, namely, I_2 and I_3. Substituting the known values of E, R_1, R_2, R_3, and R_4, these equations may be written as follows:

$$24 - 5(I_2 + I_3) - 15I_3 = 0 \qquad \text{(a)}$$

$$24 - 5(I_2 + I_3) - 10I_2 - 20I_2 = 0 \qquad \text{(b)}$$

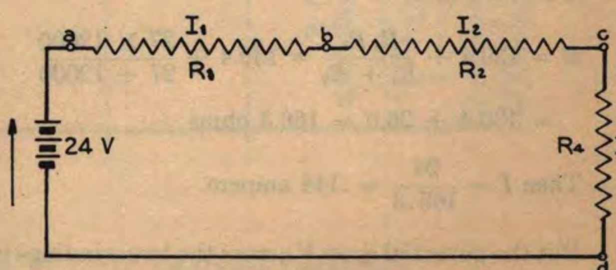


Figure 25

Simplifying, the equations become—

$$24 - 5I_2 - 20I_3 = 0 \qquad \text{(a)}$$

$$24 - 35I_2 - 5I_3 = 0 \qquad \text{(b)}$$

Multiplying equation (a) by seven, and subtracting equation (b) from it, we have—

$$\begin{array}{rl} 168 - 35I_2 - 140I_3 = 0 & \text{(a)} \\ 24 - 35I_2 - 5I_3 = 0 & \text{(b)} \\ \hline 144 - 135I_3 = 0 & \end{array}$$

or,

$$I_3 = \frac{144}{135} = 1.07 \text{ amperes, ans.}$$

Then, substituting this value in Equation (b), we have—

$$24 - 35I_2 - 5\,(1.07) = 0$$

or,

$$I_2 = \frac{24 - 5.35}{35} = \frac{18.65}{35} = 0.53 \text{ ampere, ans.}$$

and finally, the main current,

$$I_1 = (I_2 + I_3) = 0.53 + 1.07 = 1.60 \text{ amperes, ans.}$$

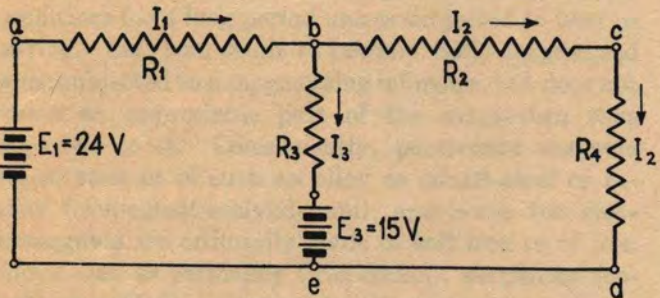


FIG. 26. CIRCUIT CONTAINING MORE THAN ONE SOURCE OF E.M.F.

22. Networks Containing More Than One Source of E.M.F.

If a network contains more than one source of E.M.F. it may be solved by either of two distinct methods. The first of these is to solve for current values in each branch of the circuit considering only one E.M.F. at a time, and to add or subtract as the case may be, the current values thus obtained for the individual branches. If we have, for example, a network such as is shown by Figure 26, containing the sources of E.M.F., E_1 and E_3, we may imagine that E_3 is omitted and the solution under this condition would be similar to that for Figure 23. The current values thus obtained would be those due to the E.M.F. designated as E_1. Those due to the E.M.F. designated as E_3 could be solved by assuming E_1 short-circuited and solving the network

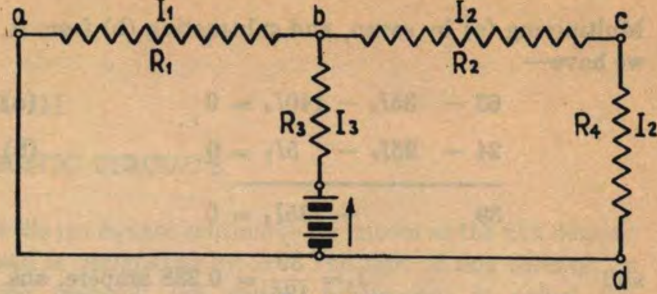


FIGURE 27

shown by Figure 27. The values obtained for branches 1 and 3 in the two cases would be subtracted and the values for branches 2 and 4 would be added.

The second method is to apply Kirchoff's Laws, taking all E.M.F.'s into consideration in each equation. This is the more general method. In this case, we have two sources of E.M.F. which oppose one another. In applying Kirchoff's Second Law to this network, we would consider the main source of electromotive force, or E_1, as positive, and the second source, or E_3, as negative. A good rule to remember in this connection is: The potential due to the battery electromotive force rises in passing through a battery from the − to the + terminal and should, therefore, be preceded by a + sign. Conversely, in passing through a battery from the + terminal to the − terminal, the potential due to the battery drops and should be preceded by a − sign.

In this case, we may assume the directions of current flow as indicated on Figure 26. Now, considering first only that portion of Figure 26 represented by the closed loop *abe*, the first equation may be written thus—

$$E_1 - (I_2 + I_3)\,R_1 - I_3 R_3 - E_3 = 0 \qquad \text{(a)}$$

Here E_3 is preceded by a − sign in accordance with the above rule. If the terminals of E_3 were reversed, E_3 would be preceded by a + sign because it would be aiding, not opposing, E_1.

Considering now the portion of the network without the *be* branch, we find the loop similar to that shown in Figure 25 and we may write the following equation—

$$E_1 - (I_2 + I_3)\,R_1 - I_2 R_2 - I_2 R_4 = 0 \qquad \text{(b)}$$

Substituting the known values of E_1, E_3, R_1, R_2, R_3, and R_4, these equations may be written as follows:

$$24 - (I_2 + I_3)\,5 - 15I_3 - 15 = 0 \qquad \text{(a)}$$

$$24 - (I_2 + I_3)\,5 - 10I_2 - 20I_2 = 0 \qquad \text{(b)}$$

Simplifying, the equations become—

$$9 - 5I_2 - 20I_3 = 0 \qquad \text{(a)}$$

$$24 - 35I_2 - 5I_3 = 0 \qquad \text{(b)}$$

Multiplying (a) by seven, and subtracting (b) from it, we have—

$$63 - 35I_2 - 140I_3 = 0 \quad \text{(a)}$$

$$24 - 35I_2 - 5I_3 = 0 \quad \text{(b)}$$

$$39 \qquad - 135I_3 = 0$$

and

$$I_3 = \frac{39}{135} = 0.288 \text{ ampere, ans.}$$

Substituting this value of I_3 in the simplified form of equation (a) above, we have—

$$63 - 35I_2 - 40.3 = 0$$

from which

$$I_2 = \frac{22.7}{35} = 0.648 \text{ ampere, ans.}$$

Then, $(I_2 + I_3) = .648 + .288 = .936$ ampere, ans.

CHAPTER III

MAGNETS AND MAGNETIC CIRCUITS

23. Nature of Magnetism

Magnetism is a peculiar property of iron, nickel, and cobalt and is most pronounced in iron and certain of its alloys. As with electricity, our knowledge of its exact nature is incomplete. Our study here will be confined to those laws concerning the magnetic properties of materials learned through observation and to the behavior of such materials under practical conditions.

The early Greeks were familiar with a natural stone that would attract bits of iron. It was a form of iron ore, now known as magnetite, and the power of attraction possessed by it was called "magnetism". It was later learned that this magnetic property could be artificially given to steel or iron by means of an electrical current.

Magnets as we know them may be classed as **permanent magnets** and **electromagnets.** A hard steel bar when magnetized becomes a **permanent** magnet because it tends to retain its magnetism under normal conditions for a long period unless subjected to heat or jarring. Soft iron tends to become easily magnetized when subjected to a magnetizing influence, but does not retain an appreciable part of the magnetism thus imparted to it. Consequently, **permanent magnets** are of **steel** or of such an alloy as **cobalt-steel** or **re-alloy** (iron-cobalt-molybdenum), and cores for **electromagnets** are ordinarily made of **soft iron** or of iron alloys such as **permalloy** (iron-nickel), **perminvar** (cobalt-iron-nickel), or **permendur** (iron-cobalt).

24. Permanent Magnets

Figure 28 represents a rectangular steel bar magnet which will attract bits of iron brought near to either end, and will exert a force of either repulsion or attraction upon other magnets in its vicinity. The influence of a magnet may be detected in the space surrounding the magnet in various ways, and is found to vary inversely as the square of the distance from the magnet. To account for this phenomenon, the magnet is said to establish a magnetic field, which is represented by the curved lines in Figure 28. These curved lines are merely a convention for illustrating the effect of the magnet. They are commonly known as "lines of magnetic induction". All the lines as a group are referred to as the "flux", and designated by the symbol ϕ. The unit of flux is one line and is called the "maxwell". The flux per unit area, or the number of maxwells per square centimeter, is known as the flux density and is designated by B. The unit of flux density, or one maxwell per square centimeter, is called the "gauss".

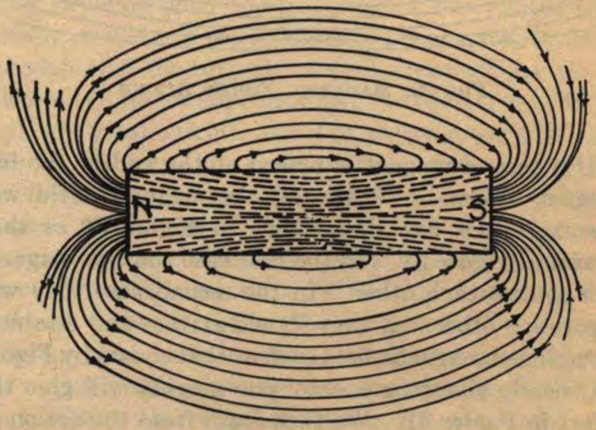


FIG. 28. MAGNETIC FIELD AROUND BAR MAGNET

The lines of magnetic induction, previously mentioned, are thought of as passing through a magnet from the south to the north pole, leaving the magnet at its north pole and reentering the magnet at its south pole. This is the significance of the arrows shown on the lines of magnetic induction in Figure 28.

Lines of magnetic induction are always closed loops. Each line or loop may be thought of as acting within itself somewhat like a stretched rubber band in that it tends to become as short as possible. Yet each of these lines or loops has a repelling effect upon its neighbors, tending to keep them separated from each other. A vivid graphical demonstration, not only of the presence

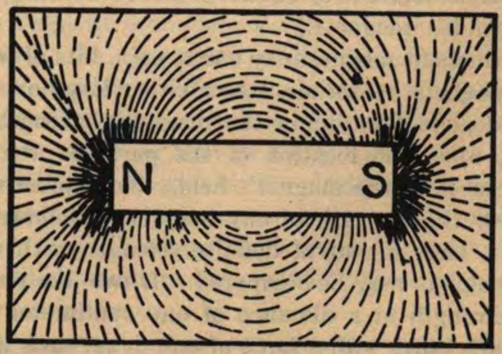


FIGURE 29

of the magnetic field but of the arrangement of the lines of magnetic induction, may be had by sprinkling iron filings upon a glass plate placed above a magnet. Figure 29 shows how the filings arrange themselves under such a condition.

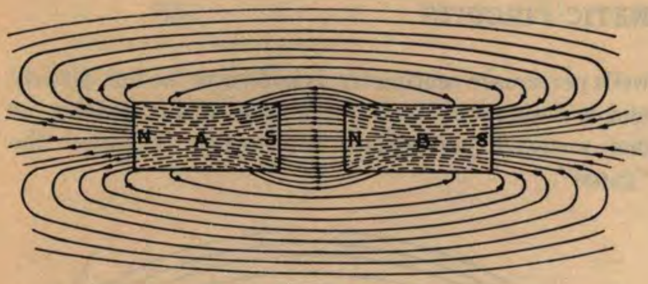


FIG. 30. MAGNETIC FIELDS AIDING

If a second magnet is placed at the end of the bar magnet shown in Figure 28, the magnetic field will become either like that shown in Figure 30 or that shown in Figure 31. In the first case the two magnets will attract each other. In the second case they will repel each other. If they should attract and establish a combined magnetic field such as that shown by Figure 30, merely changing ends of one magnet will give the effect in Figure 31. We then learn from the action of one magnet toward another that the two ends of any magnet are unlike. These two ends are called the **poles** and for convenience, the pole having one influence is called the **north pole** and that having the opposite influence is called the **south pole**. The distinction comes from the earth, which is itself a magnet. If a bar magnet is suspended so as to swing freely, that pole which tends to point toward the north is called the north seeking or north pole; the other is called the south pole. The needle of the surveyor's compass is an application of a bar magnet free to swing on its pivot and its north pole will point to the earth's magnetic pole located near the geographical north pole. (However, since the earth is itself a magnet, it may be noted that with this conventional definition, the pole nearest the geographical north is the earth's south magnetic pole inasmuch as it attracts unlike or north seeking poles of suspended magnets.)

A permanent magnet may exert, upon bits of iron or other magnetic materials, forces either large or small. These depend first, upon the magnet's strength and second, upon the location of the particles attracted with respect to the magnet's field. To express quantitatively the strength of any magnet, it follows that we must have a unit of definite strength with which other magnets may be compared. **If two like poles of equal strength at a distance of one centimeter apart repel each other with a force of one dyne, each is said to be a pole of unit strength or is called a "unit pole".** The unit pole is very small but if we can imagine one end of a small magnet sufficiently isolated from the other end and placed in the magnetic field about the magnet in Figure 28, it will tend to move in the path of the curved line nearest it and in the direction designated by the arrow if it is a north pole, and opposite to the direction designated by the arrow if it is a south pole. While the force one magnet exerts upon another depends upon the nature of their combined fields, there is an approximate law which states that the force of attraction or repulsion varies inversely as the square of the distance separating the poles in question and directly as the strength of the magnets. Expressed as an equation this law may be written—

$$f = \frac{m_1 m_2}{d^2}$$

Here f is in dynes, d in centimeters and m_1 and m_2 are the strength of the magnets in unit poles.

If the strength of the magnet in Figure 28 is doubled, the magnetic field will be strengthened in proportion, and may be represented by a more congested arrangement of lines of magnetic induction. The force that will be exerted upon a unit pole located at any point in the magnetic field will depend upon the **intensity** of the field, or the **extent to which the lines of magnetic induction at that particular point are crowded.** The unit of field intensity is that field which will exert a force of one dyne upon a unit pole. It is therefore usually expressed in dynes per unit pole and represented by the symbol H. It follows that if a pole of K units be placed in a magnetic field of intensity H, the force acting on the pole will be $f = K \times H$.

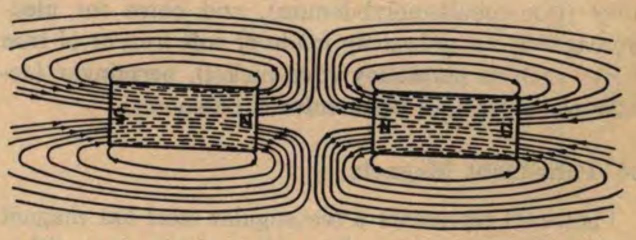


FIG. 31. MAGNETIC FIELDS OPPOSING

We have said in a preceding paragraph that the flux density B is the number of lines of magnetic induction passing through a unit area. By definition, unit flux density is one line of magnetic induction per square centimeter. We have also said that lines of magnetic induction are merely conventions for illustrating the effect of a magnetic field. Such a line may therefore be defined arbitrarily. In practice, it is usually defined as that magnetic induction per square centimeter in air, which exists in a magnetic field having an intensity of 1 dyne per unit pole. Thus in air the field intensity H and the flux density B have the same numerical value and are sometimes spoken of in the same units, that is, in lines per square centimeter or gausses. Such nomenclature leads to confusion, however, and it is better to think of field intensity only in terms of dynes per unit pole.

In Figure 28 we see that the magnetic field has greatest intensity nearest the poles. If we wish to create a field of greater intensity we can accomplish it by bending the magnet into the form of a horseshoe like that shown in Figure 32. Here each line emerging from the north pole returns to the south pole of the magnet through a much shorter distance than that represented by any one of the curved loops in Figure 28. If, with

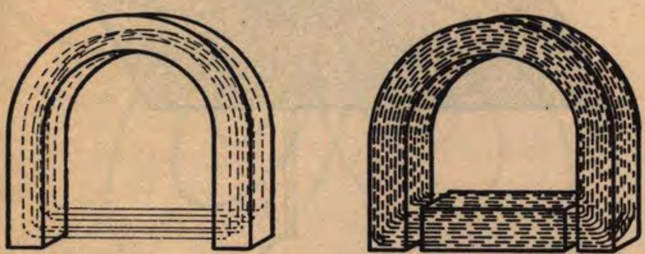


FIGURE 32

a unit pole or by other means, we should test the strength of the field between the two poles of a horseshoe magnet, we would find it more intense than that of a straight magnet of equal strength. We learn then that we not only shorten each line represented by a closed loop but, in so doing, create more lines. This gives us for a magnetic circuit an analogy to the electrical circuit. In the electrical circuit, if we have a conductor connected between the positive and negative poles of a battery and decrease the resistance by decreasing its length, we increase the current strength. In the case of the magnet, if we decrease the lengths of the paths from the north to the south pole by bending the magnet into the form of a horseshoe, we increase the number of lines of magnetic induction.

Again, if we insert between the poles of the horseshoe magnet in the space now filled with air, a piece of soft iron or other magnetic material, we greatly increase the number of lines of magnetic induction existing in the circuit formed by the magnet itself and the soft iron used for closing this circuit between the north and south poles. This is analogous to decreasing the resistance of an electrical circuit by substituting a conductor of lower resistance for one of higher resistance.

25. The Magnetic Circuit

As electric current is caused to flow in an electric circuit, so magnetic flux can be established in a magnetic circuit. Magnetic flux ϕ, or the total number of lines of induction existing in the circuit, then, is in some respects analogous to electric current.

The flux density B or the number of lines of induction per unit area may be written—

$$B = \frac{\phi}{A} \tag{13}$$

where A is the area taken at right angles to the direction of the flux and ϕ is the flux through and normal to this area.

It follows from the discussion of lines of magnetic induction being increased by the insertion of materials other than air in the magnetic field, that the flux density depends upon the materials of the completed magnetic circuit and the strength of the magnet, in the same sense that the current strength in any given cross-section of conductor depends upon the resistance of the closed electrical circuit and the electromotive force applied. We may then consider that there is a property of the magnetic circuit which is analogous to the resistance of an electrical circuit. This property is called **reluctance.** Likewise, there is a property of the magnet which is analogous to the electromotive force of a battery. This is called the **magnetomotive force** and is expressed in "gilberts". For the complete magnetic circuit, we may apply an equation identical in form to Ohm's Law which, in words, may be stated—**the flux for any given magnetic circuit is equal to the magnetomotive force of the magnet divided by the reluctance of the closed circuit.** Expressed mathematically, this may be written—

$$\phi = \frac{M}{R} \tag{14}$$

where the symbol for flux is ϕ, for magnetomotive force, M, and for reluctance, R. This may be compared to Ohm's Law as expressed by Equation (1),

$$I = \frac{E}{R}$$

In practice the above magnetic equation is seldom used in the form shown but from this relation we derive other equations dealing with flux density, field intensity and the magnetic properties of iron. These are treated along with the discussion of electromagnets, from which we shall learn more of the terms magnetomotive force and reluctance, as well as their respective units.

While we may see that in many respects the magnetic circuit is analogous to the electrical circuit, it is well to remember that the analogy is not complete since there are other respects in which the two circuits differ. The two more important of these to bear in mind are as follows:

(a) A magnetic circuit can never be entirely opened; a magnetic field must exist at all times in the vicinity of a magnet. For this reason the magnetic circuit would be more nearly analogous to the electrical circuit submerged in water. When the continuity of the metal conductors forming such an electrical circuit was broken, the circuit would be completed through the liquid across its gap. Though the current strength might be decreased in this way, the circuit could never be entirely opened; neither would the current be limited to the submerged metal conductors. There would be other flow surrounding the conductors but not of such great intensity as inside the metal conductors.

(b) Flux is not strictly analogous to current since current is rate of flow of electricity while the nature of flux is more nearly a state or condition of the medium in which it is established.

26. Electromagnets

If a straight vertical conductor carrying an electrical current pierces a cardboard as shown in Figure 33, there may be detected on the plane of the cardboard a magnetic field with lines of magnetic induction encircling the conductor. To illustrate further, if iron filings are sprinkled on the cardboard, they will form visible concentric circles as shown by Figure 34.

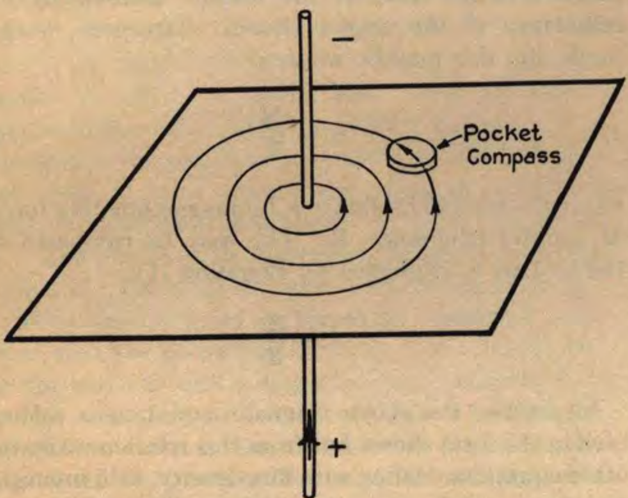


Fig. 33. Magnetic Field Around Current-Carrying Straight Conductor

Through such observations as these, we learn that **wherever an electrical current is flowing there is an established magnetic field, and the loops formed by the encircling lines are always in a plane perpendicular to the electrical conductor.**

If in either Figure 33 or 34 a compass is placed near the conductor, the needle will align itself tangent to some one of the many concentric circles. If the compass is moved slowly around the wire, the needle will revolve on its pivot and maintain its tangential relation. It will also be found that the direction of the lines with respect to the direction of current flow is that represented by the arrows in Figure 33.

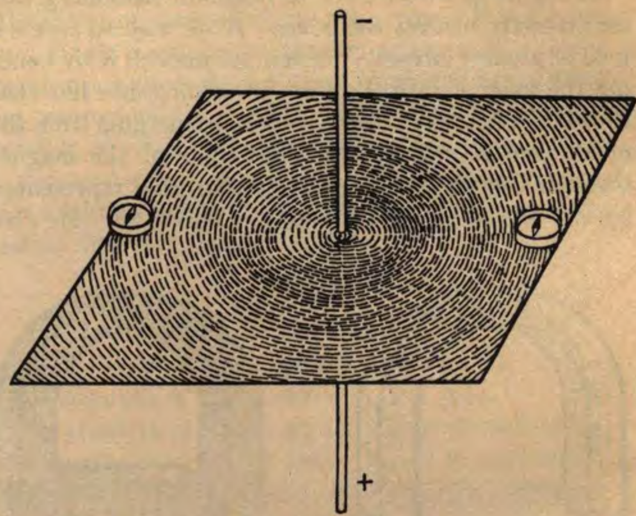


Figure 34

Though this magnetic effect is a positive one, under the conditions shown in the figures and even with a very strong current in the conductor, the magnetic field represented by the concentric circles is relatively weak. But if the electrical conductor is made to form a loop, the groups of lines forming concentric circles for every unit of the conductor's length can be imagined as arranging themselves as shown in Figure 35. The closed loops are no longer circular. They become more crowded in the space inside the loop of wire and less crowded in the space outside the loop of wire. Accordingly, the intensity of the magnetic field within the loop is increased. This may be more clearly seen by considering the single line which Figure 36 shows enclosed by imaginary boundaries both within and without the loop of wire. We may express the field intensity in terms of the cross-section of this imaginary bounding space. At the point p inside of the loop the intensity is such as to give one line for the area represented by the cross-section a, and at the point P outside of the loop the intensity is such as to give one line for an area represented by A.

If, instead of having an electrical circuit consisting of one loop of wire, we have a circuit consisting of several turns of wire, such as the winding on the spool shown in Figure 37, the intensity of the field is multiplied by the number of turns of wire. Thus, the value of the field intensity at any point for two turns would be twice that for a single loop; for three turns, three times that for a single loop; and for n turns, n times that for a single loop, providing the turns are sufficiently

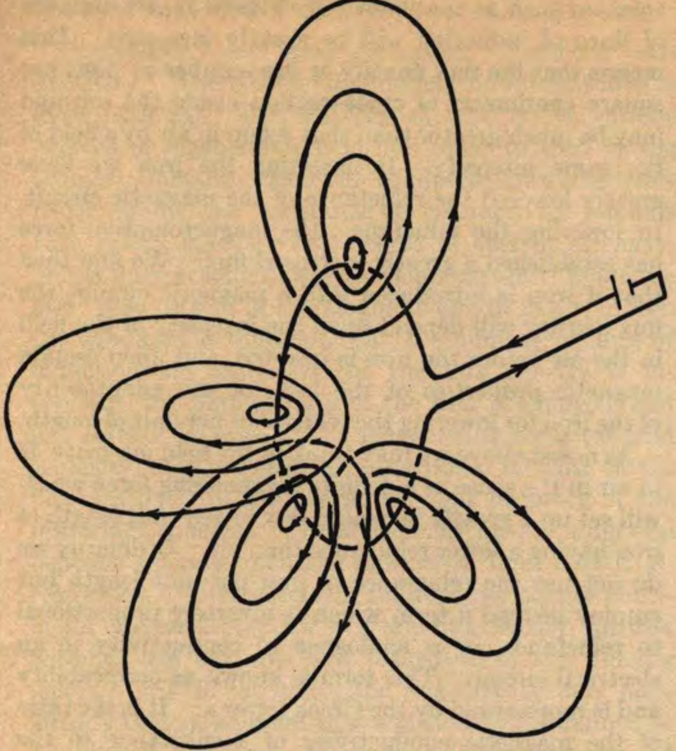


FIG. 35. MAGNETIC FIELD AROUND CURRENT-CARRYING LOOP

close together so that the flux leakage between successive turns is negligible.

Comparing Figure 37 with Figure 28, we find that the current in the coil of wire creates a magnetic field similar to that of the bar magnet. In Figure 33 the relation between direction of current flow and direction of lines of induction was shown by arrows. We use this same relation in Figure 35 and going one step farther, we may determine the north and south poles of the magnet formed by the coil shown in Figure 37. A simple way to remember the relation for any electrical winding is illustrated by Figure 38. Here if we

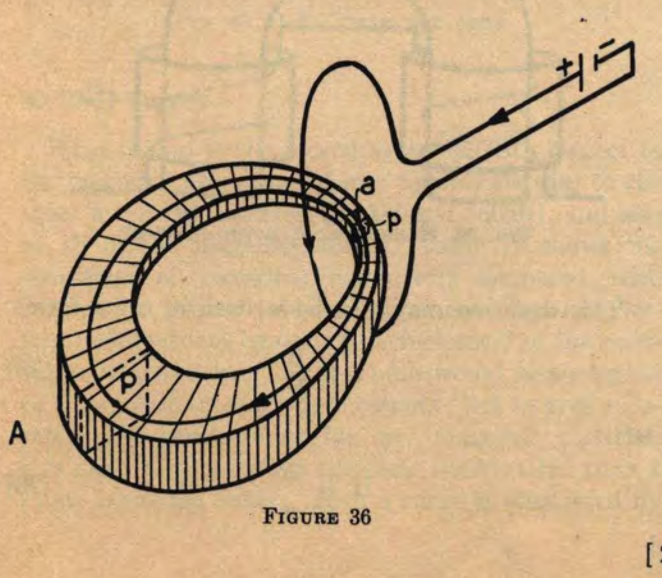


FIGURE 36

assume current flowing through a winding in the direction of "turn" for a right-hand screw, the lines leave the point of the screw which is the north pole and enter the slot which is the south pole.

In Figure 32 the number of lines in the magnetic circuit established by the horseshoe magnet was greatly increased by the insertion of a piece of soft iron between the north and south poles. Likewise, if in Figure 37 the spool shown has a soft iron core, the number of lines will be greatly increased. Further, if the core of the winding is bent in the shape of a horseshoe as shown in Figure 39, we have the customary electromagnet which is capable of exerting considerable force.

27. Relation Between Current and Field Intensity

If we increase the current strength in the winding shown by Figure 37, we will find that the intensity of the magnetic field is increased proportionately. Thus, the value of H, or the magnetic field intensity in air, is directly proportional to the current flowing in the winding. We may accordingly establish a definite relation between field intensity and electrical current for any given set of conditions.

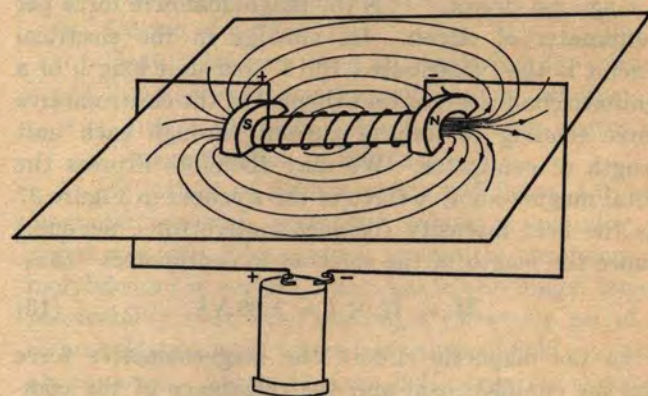


FIG. 37. MAGNETIC FIELD AROUND AIR-CORE SOLENOID

A winding such as that shown in Figure 37 is called a "solenoid". If such a solenoid is very long as compared to its diameter, and is constructed with one turn of wire to each centimeter of length, and the current in the winding is one ampere, the field intensity in the air on the inside of the solenoid can be shown to be equal to .4 × 3.1416 or 1.26 dynes per unit pole. The field intensity inside any long air-core solenoid would then be expressed by the following equation:

$$H = \frac{1.26NI}{l} \qquad (15)$$

where N is the total number of turns of the winding, I is the current in amperes and l is the length of the solenoid in centimeters. This equation is apparent since one turn of wire per centimeter for a current of one ampere gives 1.26 dynes per unit pole, and the intensity is increased proportionally to the current and to the number of turns of wire per centimeter of length.

In electrical practice, the term "ampere-turn" is frequently used, which merely means the product of N and I in Equation (15).

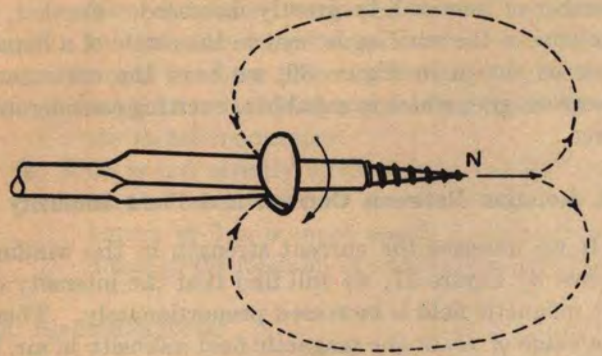


FIGURE 38

The field intensity H may be thought of as the force tending to produce magnetic flux in each unit length of a magnetic circuit. It is the **magnetomotive force per centimeter of circuit.** Its analogy in the electrical circuit is the "distributed EMF" per unit length of a uniform conductor, or that element of the electromotive force tending to force a current through each unit length of conductor. We may therefore express the total magnetomotive force of the solenoid in Figure 37 as the field intensity (magnetomotive force per cm.) times the length of the solenoid in centimeters—thus:

$$M = H \times l = 1.26NI \qquad (16)$$

In the magnetic circuit, the magnetomotive force for the complete coil and the reluctance of the complete circuit are not the most convenient quantities for practical calculations. The magnetic circuit, though analogous in many respects to the electrical circuit, does not take a definite form. The lines of induction may not be distributed equally throughout the cross-section of the circuit. There is always present the surrounding air which is an electrical insulator but is not a magnetic insulator. We, therefore, use another equation more frequently than Equation (14), which is likewise in the form of Ohm's Law, but is expressed in terms of quantities **per unit** of magnetic circuit rather than for the complete magnetic circuit.

28. Flux Density, Field Intensity and Permeability

In discussing field intensity we have thus far considered it only in connection with magnetic circuits in air. However, we have seen that if iron is inserted in a solenoid such as that shown by Figure 37, the number of lines of induction will be greatly increased. **This means that the flux density or the number of lines per square centimeter of cross-section** inside the solenoid may be much greater than that set up in air by a field of the same intensity. In inserting the iron we have greatly lowered the reluctance of the magnetic circuit. In lowering the reluctance, the magnetomotive force has established a greatly increased flux. We find then that if iron is introduced into a magnetic circuit, the flux density will depend upon the intensity of the field in the air before the iron is inserted, and upon certain magnetic properties of the iron, or the adaptability of the iron for lowering the reluctance per unit of length.

As noted above we may think of the field intensity H in air in the sense of a definite **magnetizing force** which will set up a greatly increased flux in any unit length of iron having a lower reluctance than air. Ordinarily we do not use the reluctance of iron per unit length but employ instead a term which is inversely proportional to reluctance, or is analogous to conductivity in an electrical circuit. This term is known as permeability and is represented by the Greek letter μ. It is the ratio of the magnetic conductivity of a substance to the magnetic conductivity of air. Using this ratio, we may express the flux per unit cross-section in the form of an equation as follows:

$$\frac{\phi}{A} = H \times \mu \qquad (17)$$

or

$$B = H \times \mu \qquad (18)$$

where B is the conventional symbol for flux density.

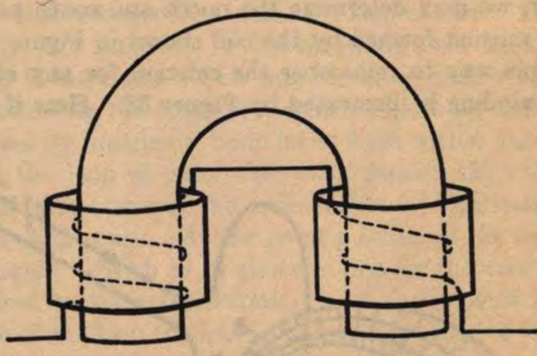


FIG. 39. HORSESHOE ELECTRO-MAGNET

This equation may also be written in other forms:

$$\mu = \frac{B}{H} \qquad (19)$$

or

$$H = \frac{B}{\mu} \qquad (20)$$

29. Magnetic Properties of Iron

Permeability has been compared to electric conductivity. There is one distinction, however, which is most essential. The stability of iron under various degrees of magnetization is not equal to that of the ordinary metallic electrical conductor. In the electrical circuit the resistance or conductivity remains very nearly fixed for any degree of current strength, unless there is some change in temperature. While the same may be said of the magnetic circuit in air, in iron the condition is different. As the number of lines of induction are increased (or the flux density is increased), the permeability of the iron is changed, and any further increase in the magnetizing force (or field intensity) may not mean a proportional increase in the flux density. In simpler terms, that property of the iron which enables it to establish more lines of induction depends entirely upon the number of lines that it already has. After a certain number per square centimeter of cross-section, or a certain flux density, the iron becomes less effective and regardless of any further increase in field intensity, the flux density may have already become so great that additional lines cannot be established any more readily than if the core were of air. This condition is called the "saturation point" of the iron.

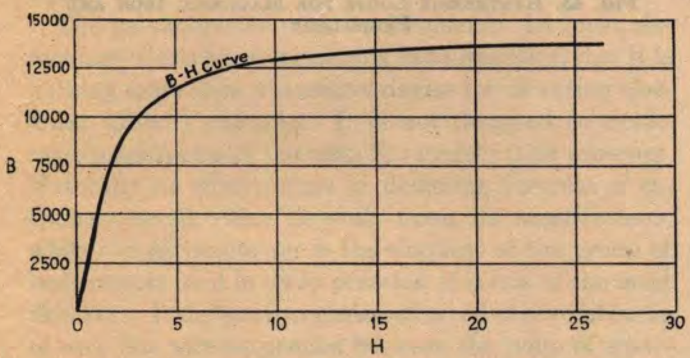


Fig. 40. B-H Curve for Iron

30. B-H Curves

What is said in the preceding article with respect to the magnetic properties of iron applies likewise to the other magnetic materials (nickel and cobalt), and also to all of the magnetic alloys. Table II shows the resistance of electrical conductors compared with copper. A similar table could be compiled for electrical conductivity by taking the reciprocal of the resistance values shown. Such a table would be analogous to a magnetic table for permeability; but to give accurately the permeability for any magnetic material, it is necessary to show a complete curve rather than a single tabulated value. Such a curve is illustrated by Figure 40 which is taken for a magnetic iron used by the Western Electric Company in the manufacture of certain relays and other telephone apparatus. This curve was determined after the iron had been annealed for three hours at a temperature of 900° C. Every

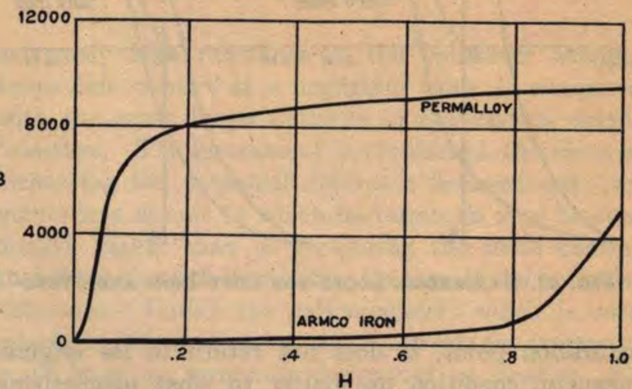


Fig. 41. B-H Curve for Permalloy

magnetic material has some such curve. A magnetization curve will ordinarily depend upon many things, such as—

(a) Whether cast iron, wrought iron, steel or an alloy of these with other metals.
(b) Degree of purity.
(c) Heat treatment used in preparing the metal.
(d) Previous magnetic history; that is, whether or not it has been subject to a high degree of magnetization in the past.

At low values of field intensity (H below 1.0) the magnetic material, permalloy, which is an alloy of nickel and iron (plus a small amount of chromium or molybdenum in certain cases) has a very much higher permeability than iron, making it extremely useful in communication work where low values of field intensity are common. Figure 41 gives B-H curves for a standard permalloy and a standard iron for low values of H; it will be noted that the magnetic flux for a given magnetizing force is very much greater in the permalloy than in the iron over the range covered.

31. Hysteresis

If a piece of iron is subjected to an increasing magnetizing force until the saturation point is reached and then the magnetizing force is decreased to zero and established in the opposite direction until the saturation point is again reached, and if the magnetizing force is again decreased to zero and again increased until the cycle is completed, the relations between flux density and field intensity for all parts of the cycle may be represented by a curve such as one of those shown by Figure 42. This is called the "hysteresis loop". Here it is seen that after the iron has once reached the

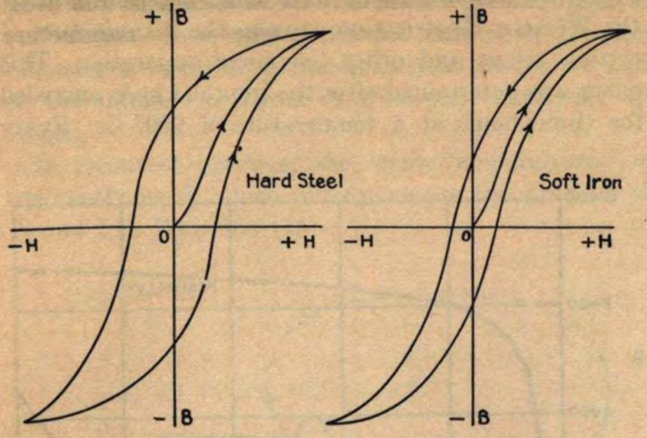


Fig. 42. Hysteresis Loops for Soft Iron and Steel

saturation point, it does not return to its original magnetic condition no matter to what magnetizing forces it may be subjected. For example, an inspection of the hysteresis loop shows that iron will retain a certain degree of magnetization after the magnetizing influence has been reduced to zero. This is particularly true of hard steel and is the reason that all permanent magnets are made of hard steel or a material having similar characteristics. The two curves of Figure 42 illustrate the difference in the hysteresis loops of hard steel and soft iron. The fact that soft iron has a narrow hysteresis loop makes it adaptable for the cores of electromagnets. We may note here also that the hysteresis loop for permalloy is very much narrower than that for soft iron at low values of magnetizing force. This is illustrated in Figure 43 where the hysteresis loop for permalloy and iron are compared.

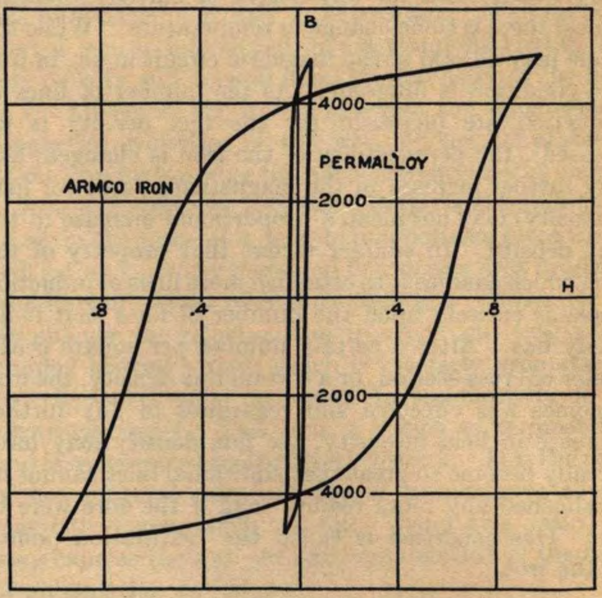


Fig. 43. Hysteresis Loops for Magnetic Iron and Permalloy

CHAPTER IV

ELECTRICAL MEASUREMENTS IN DIRECT-CURRENT CIRCUITS

32. The Measuring Instruments

We have been discussing such electrical quantities as the volt, the ampere, the ohm and the watt, but little has been said about the electrical instruments that are used to measure these quantities. The more commonly used group of instruments includes the galvanometer, the voltmeter, the ammeter, the Wheatstone bridge (including a galvanometer) the megger, and the wattmeter. At this stage of our study it is important that we learn the fundamental principles of these measuring instruments and the distinction between instruments designed for different purposes, but it is not important that we study long descriptions of their construction or those details of design pertaining only to their manufacture. They are ordinarily sealed at the factory and are seldom repaired by the field maintenance man. Let us, therefore, concern ourselves with the intelligent and skillful use of them and only with those principles of their operation that are essential to this.

The galvanometer may be considered the most elementary of electrical measuring instruments in that it is nothing more than a sensitive device for detecting electrical (direct) currents. It is not designed to determine magnitudes of currents but merely their presence. Naturally its effectiveness in detecting currents of extremely small value depends upon its sensitiveness; while the galvanometer is the simplest of the group of instruments used in daily practice, it is one of the most delicate. It ordinarily consists of a coil of several turns of very fine wire suspended between the poles of a permanent horseshoe magnet and held in a neutral position by the torsion of fine suspension fibres, or other equally delicate means. The suspended coil carries a light needle which stands at the center of a fixed scale when the coil is in its neutral position with respect to the permanent magnet. A very small current through the suspended coil will set up a magnetic field that will tend to align itself with the field of the permanent magnet and thereby cause a deflection of the needle from its neutral position on the fixed scale.

The voltmeter (of the revolving coil type) is a galvanometer more ruggedly designed, having an extremely high resistance and with the scale so calibrated as to read the potential impressed on the terminals of the instrument. Of course the voltmeter deflection is caused by a very small current flowing through the high resistance winding, but in most simple circuits the extremely high resistance of the voltmeter winding keeps this current at a negligible value as compared with the much larger currents in the various circuit branches. The instrument is considered, therefore, as measuring the potential difference between any two points in a circuit to which its terminals may be connected, rather than as measuring the small current that flows through its winding because of the potential difference. Unlike the galvanometer, which is used merely to detect the presence of current, the voltmeter must have its terminals designated as positive and negative (unless it is a "zero center scale" type). We have already illustrated the use of this instrument in determining the E.M.F. of a battery or a drop in potential between two points in a series circuit.

The ammeter is likewise an application of the galvanometer principle, likewise usually more rugged in design, and likewise has a calibrated scale. But in this case the scale is calibrated to measure the value of the current that flows through its winding instead of the electromotive force across its terminals. We recall that in Figure 2 a flowmeter determined the gallons-of-water-per-minute being pumped through a water circuit. The ammeter is analogous to this flowmeter. It is inserted directly in the path of the current and the entire flow goes through the instrument (or through a calibrated shunt associated with the instrument). For the same reason that a water flowmeter should not be so constructed as to retard appreciably the flow of the water by causing a drop in head, the ammeter must, unlike the voltmeter, have a very low resistance. We may, therefore, think of the ammeter as a "flowmeter" designed with very low resistance so that it will not cause an appreciable readjustment of current or voltage values in any network when it is inserted to measure the current in any one branch.

Later we are to discuss the precautions to be taken in the use of instruments but just here it might be stated that wherever the ammeter is used there should be a consciousness that Ohm's Law is never failing in that it applies to every circuit branch, and that the current which will flow through the ammeter will be very large if an appreciable potential is connected across its terminals without other resistance in the circuit. As an illustration, if an ammeter has an internal resistance of .005 ohm and an electromotive force of one volt is connected to its terminals, the current through it in accordance with Ohm's Law will be 200 amperes. This may be considerably in excess of the maximum current value for which the instrument is designed. It is well to remember, therefore, that the ammeter is an instrument that will cause a short-circuit when connected across points in a circuit having a considerable difference in potential, while the voltmeter is on the other hand for most practical purposes an open circuit, and unless connected to points having potentials higher than its greatest scale reading, it cannot be damaged from excess current values. In the language of the electrician, **the ammeter must always be inserted and never connected across.**

Voltmeters and ammeters are manufactured for different ranges of voltage and current values and one instrument often has several scales. Instruments for measuring small values are prefixed with milli, meaning one-thousandth, or micro, meaning one-millionth. Thus, we have milliammeter, millivoltmeter, etc. It is obvious that **an instrument must not be used when the value of the voltage or current to be measured is likely to be greater than the maximum scale reading.**

If a voltmeter measures at any given instant the E.M.F. across any electrical circuit (either branch or mains) and an ammeter at the same instant measures the current in the same circuit (either branch or mains), the product of the two readings is, from the formula $P = EI$, equal to the power in watts supplied to the circuit. Meters are designed with both ammeter and voltmeter terminals to read this product, or the power in watts directly. These are called wattmeters.

There are two remaining instruments in the commonly used group. These are the Wheatstone bridge and the megger. The Wheatstone bridge is simply a network of resistances which can be used in connection with the galvanometer for measuring an unknown electrical resistance by an accurate comparison method. The megger is a combination of a magneto source of electromotive force and a sensitive meter, calibrated to read values of very high resistances connected across its terminals. A more detailed description and the practical use of these instruments is given in connection with the discussion of the various methods of measuring resistance.

33. Measurements of Resistance

There are numerous methods for measuring electrical resistance and the one which is most practical depends upon—

(a) the magnitude of the resistance to be measured;
(b) the conditions under which it is to be measured;
(c) the degree of accuracy required.

Probably the most difficult resistance measurements are those of extremely low values. Examples of these are: the internal resistance of an ammeter (or the resistance of an ammeter shunt); the resistance of an electrical connection such as the connection between cells of a storage battery; the resistance of an electrical bond, such as bonds used to prevent electrolysis and connected between railroad rails and water pipes or from one railroad rail to another.

Where very low resistances are to be measured accurately, it is usually a complicated laboratory process. Fortunately, we have but few such cases in our work, though there are cases where the presence of low resistance values is to be determined but not necessarily with a great degree of accuracy. For example, in the case of a connection between the cells of a storage battery, we may desire to know whether the resistance of the connection is greater than it should be. Were this to be accurately measured, the measurement would be a difficult one to make but it can usually be determined for practical purposes by some simple test such as touching the two sides of the connection with the terminals of a telephone receiver and listening for a click due to the potential drop caused by the resistance. It follows that we may confine our attention here to the practical methods used for measuring either those resistance values which are appreciable, such as the ones that are important in simple circuits, or those resistance values which are extremely high, such as the insulation resistance of cable or open wire conductors.

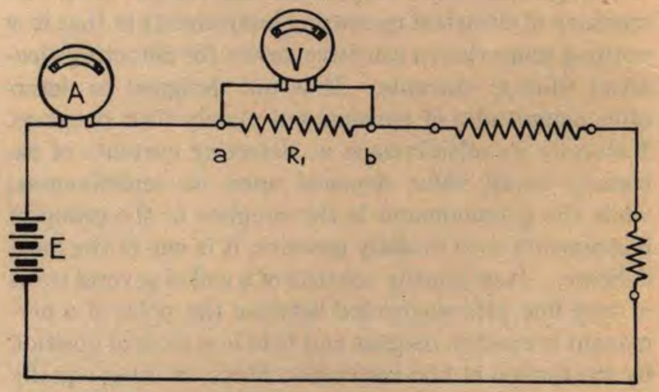


Fig. 44. Voltmeter-Ammeter Resistance Measurement

34. Voltmeter-Ammeter Method

Figure 44 shows a simple series circuit. Let us assume that it is desired to determine the value of the resistance R_1. We have learned that if a voltmeter is connected across the terminals a and b as shown, it will measure the potential drop across the resistance. But if, at the same instant this reading is taken, an ammeter is so inserted as to read the value of the current flowing through the resistance R_1, we will have not only an E.M.F. reading but a current reading as well and from the two, the value of the resistance may be calculated by Ohm's Law.

Example: In Figure 44, the voltmeter reading is 5 volts and the ammeter reading is .5 ampere; what is the value of resistance R_1?

Solution:

$$R_1 = \frac{V_1}{I_1} = \frac{5}{.5} = 10 \text{ ohms, ans.}$$

35. Drop in Potential Method

If in Figure 45 it is desired to determine the value of the resistance R_1, the "drop in potential method" can be used if a second resistance R_2 of known value is inserted in series and the voltage drops across both R_1 and R_2 are measured. Since the two resistances are in series, the same current is flowing through both and from Ohm's Law:

$$I = \frac{V_1}{R_1}, \quad \text{and also} \quad I = \frac{V_2}{R_2}$$

Therefore,

$$\frac{V_1}{R_1} = \frac{V_2}{R_2}$$

which may be written, either—

$$\frac{V_1}{V_2} = \frac{R_1}{R_2}$$

or

$$R_1 = R_2 \frac{V_1}{V_2} \qquad (21)$$

Example: If in Figure 45 the value of R_2 is 10 ohms and the drop across it is 12 volts, what is the value of R_1 which has a drop of 8 volts?

Solution:

$$R_1 = R_2 \frac{V_1}{V_2} = 10 \times \frac{8}{12} = 6.67 \text{ ohms, ans.}$$

36. Insulation Measurements

The application of the drop in potential method which has greatest importance in telephone and telegraph work is its special adaptation to insulation measurements.

If the series circuit in Figure 45 contains no resistance other than R_1 and R_2, it is not necessary to measure the drop across R_1 because it will be equal to the potential of the battery minus the drop across R_2. The formula for this special case may then be written—

$$R_1 = R_2 \frac{E - V_2}{V_2} \qquad (22)$$

where E is the E.M.F. of the battery.

If R_1 is very high in value such as a "leak" due to poor insulation, it can be measured using formula (22) but instead of using a second known resistance, the voltmeter itself may be inserted in series with the bat-

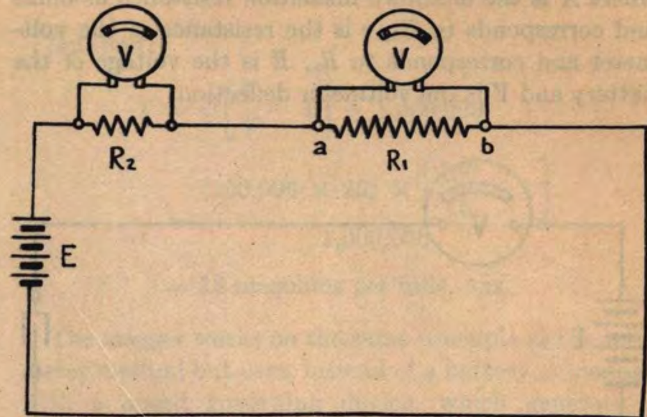


Fig. 45. Drop in Potential Resistance Measurement

tery and R_1 as shown in Figure 46. The reading V_2 then applies to the drop across the **voltmeter's own resistance** which, as has been previously stated, is very high. But since the resistance being measured is very high, this gives greater accuracy than if a known resistance R_2 having a lower value were inserted and a drop of lower value measured across it. As a matter of fact, voltmeters used for measuring insulation are especially designed to have abnormally high internal resistance; the ones used in the standard testboard testing circuits have a resistance of 100,000 ohms.

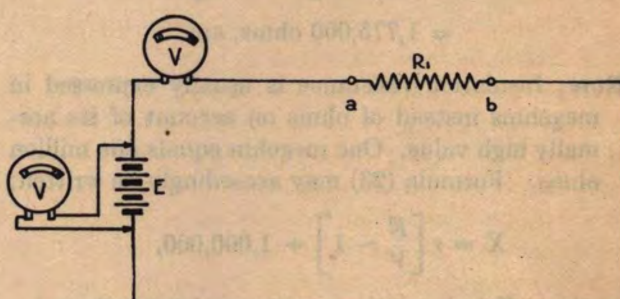


Fig. 46. Insulation Measurement

Figure 47 shows the drop of potential method with series voltmeter for measuring the insulation of a condenser. Figure 48 shows a "leak" between two cable conductors and Figure 49 a "leak" between an open wire and ground, both being measured in the same manner.

For this application formula (22) is ordinarily written—

$$X = r \frac{E - V}{V}$$

or

$$X = r\left[\frac{E}{V} - 1\right] \qquad (23)$$

where X is the unknown insulation resistance in ohms and corresponds to R_1, r is the resistance of the voltmeter and corresponds to R_2, E is the voltage of the battery and V is the voltmeter deflection.

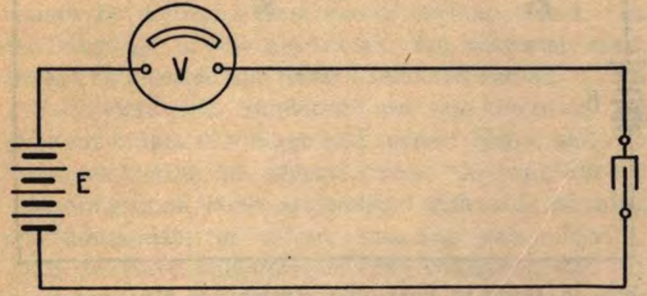


FIGURE 47

Example: The voltmeter shown in Figure 47 has a resistance of 100,000 ohms. If it reads 8 volts as shown and 150 volts when connected directly across the battery terminals, what is the insulation resistance of the condenser?

Solution:

$$X = r\left[\frac{E}{V} - 1\right]$$

$$= 100{,}000\left[\frac{150}{8} - 1\right]$$

$$= 1{,}775{,}000 \text{ ohms, ans.}$$

Note: Insulation resistance is usually expressed in megohms instead of ohms on account of its normally high value. One megohm equals one million ohms. Formula (23) may accordingly be written:

$$X = r\left[\frac{E}{V} - 1\right] \div 1{,}000{,}000,$$

where X is insulation resistance in megohms instead of ohms.

In the standard testboard circuits, dry cell batteries are ordinarily used for insulation testing batteries and

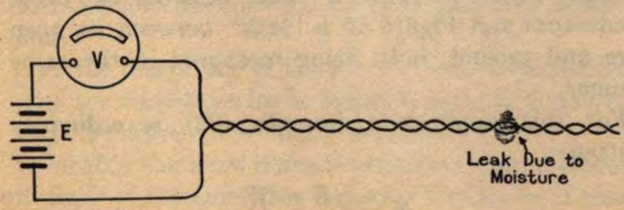


FIG. 48. METALLIC INSULATION TEST

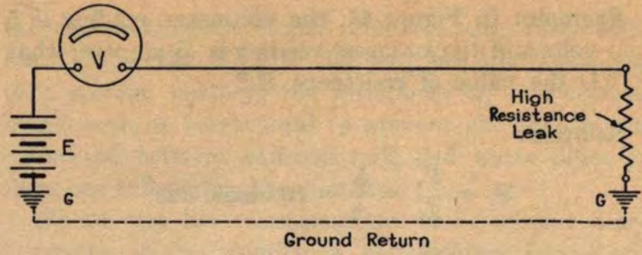


FIG. 49. TEST FOR INSULATION OF SINGLE WIRE

these are wired metallic; that is, they have neither the positive nor negative terminal grounded. In some cases, however, it is necessary to measure insulation across a pair of wires such as cable conductors using a

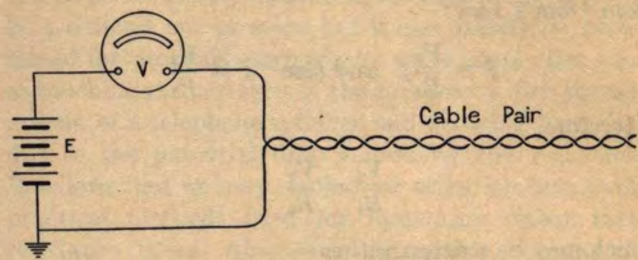


FIGURE 50

permanently grounded testing battery (such as a 120-volt telegraph battery tap). Figure 50 shows such a test, but here the "metallic" or "mutual" insulation is not distinct from the "wire to ground" insulation on account of one conductor being grounded. It is necessary in this case to take two readings, one with one conductor grounded and the other with the second conductor grounded. Figure 51 shows a reversing key wired in the testing circuit to facilitate this test. While neither of the two readings gives the mutual resistance, it may be calculated from these readings and readings for the single conductors to ground.

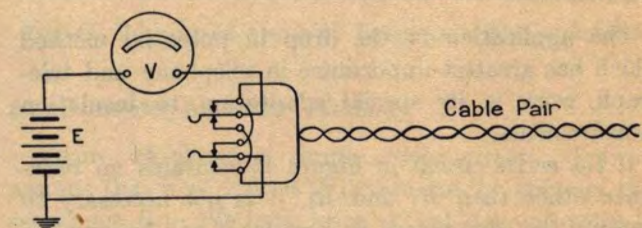


FIGURE 51

When determining the insulation resistance of cable or open wire pairs, the value is ordinarily expressed as the average resistance per mile rather than a value for the entire circuit. In calculating the resistance per mile value, it is assumed that each mile of wire (or circuit) has a concentrated leak and that these are all equal as shown in Figure 52. It is further assumed that the series resistance of the wire is negligible compared with the insulation resistance. Formula (23) when expressed for X equal to the resistance per mile instead of the entire circuit becomes—

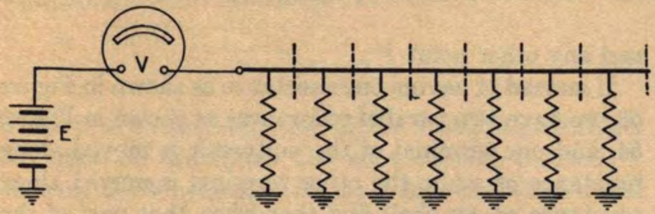


FIGURE 52

$$X = rl\left[\frac{E}{V} - 1\right] \text{ in ohms} \qquad (24)$$

or

$$X = rl\left[\frac{E}{V} - 1\right] \div 1{,}000{,}000 \text{ in megohms,}$$

where l is the length of the circuit in miles in both cases.

Example: Assume the wire shown in Figure 52 to be 20 miles long and the voltmeter to have a resistance of 100,000 ohms. What is the insulation of the wire to ground in megohms per mile if the battery E.M.F. is 150 and the voltmeter reading is 15?

Solution:

$$X = rl\left[\frac{E}{V} - 1\right] \div 1{,}000{,}000$$

$$= \frac{(100{,}000 \times 20) \times \left[\frac{150}{15} - 1\right]}{1{,}000{,}000}$$

$$= 18 \text{ megohms per mile, ans.}$$

The megger works on the same principle as the series meter method but uses, instead of a battery, a magneto with a speed governing device, which generates a constant E.M.F. It also includes a meter calibrated to read megohms directly, instead of volts. The generator potential is much higher than the battery potentials used for telephone testing, 400 volts being ordinarily used on the more common types. The internal resistance of the meter (and generator) are extremely high and the generated voltage cannot sustain any appreciable current, thereby making the instrument safe for practical work.

CHAPTER V

ELECTRICAL MEASUREMENTS IN DIRECT CURRENT CIRCUITS—(*Continued*)

37. Theory of the Wheatstone Bridge

The Wheatstone bridge has been described as a network of resistances that may be used in connection with a galvanometer to measure unknown resistance values. An analysis of its theory is the next step in order after the study of the potential drop method of measuring resistance.

In Figure 53, the voltmeter has one terminal permanently connected to a and the other terminal may

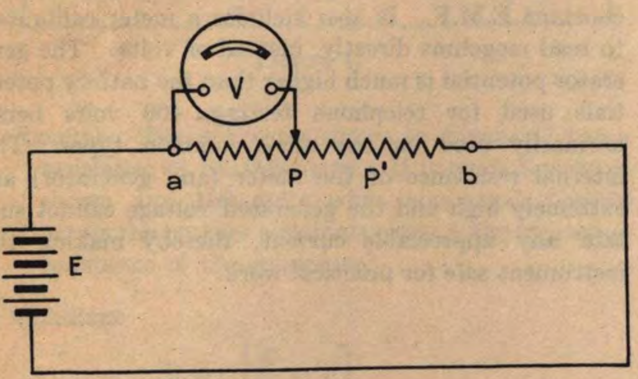


FIGURE 53

be moved along the resistance ab. The voltmeter reading will be zero when both terminals are at a, and will gradually increase as the point P is moved toward b. We shall find that the potential drop measured between the points a and P is always proportional to that part of the resistance between a and P, or we may write:

$$\frac{aP}{aP'} = \frac{V}{V'}$$

where V' is the potential drop measured between a

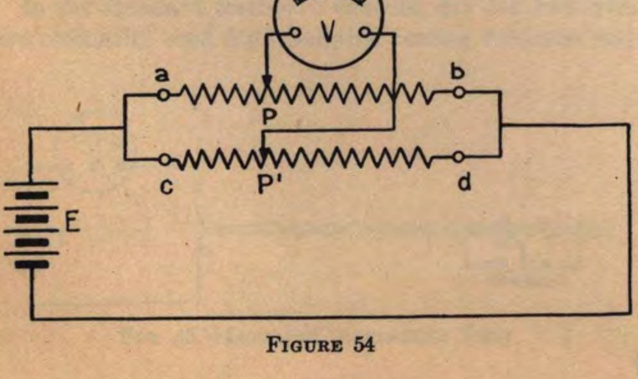


FIGURE 54

and any other point P'.

If instead of having one resistance as shown in Figure 53, we have two parallel resistances as shown in Figure 54, and one terminal of the voltmeter is moved along resistance ab while the other terminal is moved along resistance cd, we shall find that when that part of the resistance ab between the points a and P is proportional to that part of the resistance cd between the points c and P', the difference in potential between the points P and P' will be zero and there will be no reading of the voltmeter. Mathematically, this may be expressed:

$$\frac{aP}{cP'} = \frac{ab}{cd}$$

Likewise, we may develop a similar expression for the remaining part of the resistance:

$$\frac{Pb}{P'd} = \frac{ab}{cd}$$

From these two relations, we may write:

$$\frac{aP}{cP'} = \frac{Pb}{P'd} \tag{25}$$

This merely means that potential drop always distributes itself proportionally along one or more resistances. In Figure 55 let us assume that the resistances

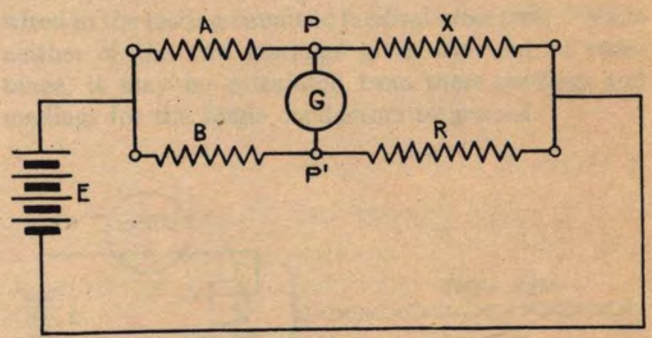


FIGURE 55

represented by the branch A, the branch B, and the branch R, are known, and that the resistance shown as the branch X is unknown. Inasmuch as the meter connected between the points P and P' is merely being used to determine that these two points have the same potential, a galvanometer can be employed instead of a voltmeter, and will be preferable in that it is more sensitive. If there is no deflection in the galvanometer

$R1 = R2 = R3 = Rx \quad \frac{R1}{R2} = \frac{Rx}{R3}$

needle, we may write the same relation as was given by Equation (25), namely:

$$\frac{A}{B} = \frac{X}{R} \quad \text{or} \quad \frac{A}{X} = \frac{B}{R}$$

This equation can be expressed:

$$X = \frac{A}{B}R \tag{26}$$

which is the usual equation of the Wheatstone bridge.

Figure 56 illustrates the conventional method of showing the Wheatstone bridge. It is almost identical

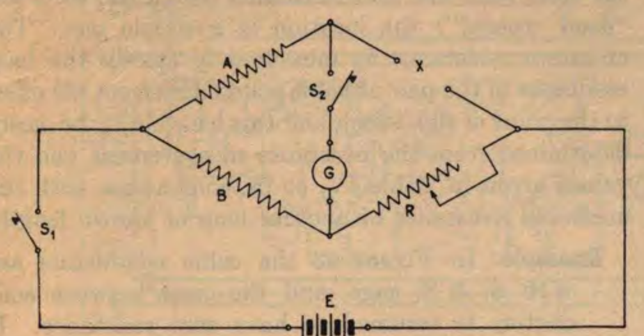


Fig. 56. Standard Wheatstone Bridge Convention

to the arrangements shown by Figure 55, but has the resistances connected in a diamond shaped diagram. S_1 is a switch for disconnecting the battery when not in use, and S_2 is a similar switch for disconnecting the galvanometer. Binding posts are shown for connecting the unknown resistance to be measured, which is usually designated as X. The resistances A and B are called the **ratio arms** of the bridge and the resistance R is variable so that for any unknown resistance X, the value of R may be adjusted to obtain a perfect balance, or to bring the galvanometer needle to the stationary or zero point on the scale. Though Figure 56 shows the resistance branch R as variable and the arms A and B as fixed, a balance could also be obtained by changing the ratio A/B in formula (26) instead of varying the value of R.

In all forms of the Wheatstone bridge it is permissible to reverse the connections for the battery and the galvanometer in so far as these connections concern the theory of operation. Thus, in Figure 56 the galvanometer and dry cell could be interchanged without in any way affecting the operation of the bridge.

There are many commercial types of Wheatstone bridge testing instruments. There are four in particular that are used extensively in telephone and telegraph work. One is a small portable type bridge, which is also used sometimes at local test desks. The others are especially designed bridges for use in connection with toll testboards and are known as the "#12001 Bridge", the "Wheatstone Bridge per KS-3011", and the "Wheatstone Bridge per KS-5411". The dial and circuit arrangements of the three last named bridges are shown by Figures 58, 59, and 60 and the theory is shown by Figure 57. The portable bridge has both the battery and galvanometer mounted inside the case and is arranged for making simple resistance measurements, Varley loop tests, and Murray loop tests.

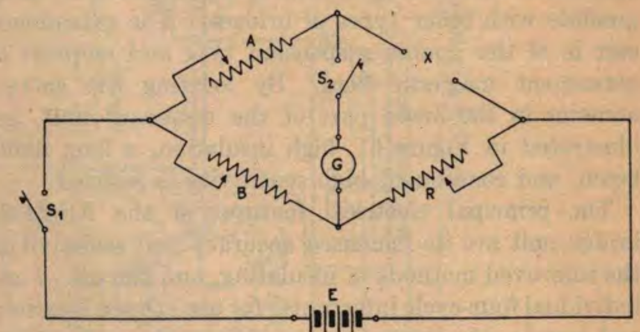


Figure 57

The #12001 bridge was designed for the #4 testboard testing circuit. The galvanometer is mounted in a separate case which in turn mounts inside the bridge case in such manner that it may be conveniently removed. All connections are brought out to binding posts so that various testing combinations can be secured with the particular key arrangement of the #4 testboard testing circuit. The galvanometer is equipped with resistance shunts and has a coil resistance of 288 ohms. Dials are used for the A and B arms and for the variable balancing arm, but no sliding contacts are exposed.

The Wheatstone bridge per KS-3011 is designed for use in both the #4 and #5 toll testboards and is of a type similar to the #12001 bridge but is somewhat more accurate. It employs a reflection type galvanometer having a lamp and scale instead of a needle, which permits the detection of smaller current values than is possible with the preceding type of bridges. The ratio arms are controlled by a single dial which gives the ratio A/B directly for nine values as follows:

$$1000,\ 100,\ 10,\ 1,\ \tfrac{1}{4},\ \tfrac{1}{9},\ \tfrac{1}{10},\ \tfrac{1}{100},\ \tfrac{1}{1000}$$

In addition, one position of the dial designated as M-1000 is for use in making open location measurements and Murray loop tests, as outlined in Article 40. The rheostat arm has a total resistance of 9999 ohms and is adjustable by means of four inverted dials in steps of one ohm. The resistances in the rheostat arm are accurate to $\frac{1}{10}$ of 1 per cent and in the ratio arms to $\frac{1}{20}$ of 1 per cent. Three additional dials are provided for use in connection with open location tests.

The Wheatstone bridge per KS-5411 is for use primarily in toll cable offices, but may be used for trouble location tests on open wire. It provides for certain improved methods of operation and greater accuracy in determining and locating faults than is possible with former types of Wheatstone bridge testing units.

In general, the operating principles of the KS-5411 bridge are the same as those for the KS-3011 bridge and other similar types. Each KS-5411 bridge is individual to the testboard position it occupies and cannot be associated with testing circuits in other positions, as is possible with other types of bridges. The galvanometer is of the double suspension type and employs a permanent magnetic field. By locating the galvanometer in the lower part of the testboard unit, as illustrated in Figure 61, high insulation, a long light beam, and consequent high sensitivity is realized.

The principal electrical features of the KS-5411 bridge unit are its increased accuracy and sensitivity, the improved methods of insulating, and the use of an individual four-cycle interrupter for use in open location and certain other tests.

Dial setting ratios, as shown below, are provided in the bridge circuit for use as required:

$$1000,\ 100,\ 10,\ 1,\ \tfrac{1}{4},\ \tfrac{1}{10},\ \tfrac{1}{100},\ \tfrac{1}{1000}$$

In addition, two positions of the dial designated as M-1000, and M-10,000 are provided for use in making Murray loop tests and open location tests.

38. Simple Loop Tests or Plain Resistance Measurements

In the telephone and telegraph plant, the Wheatstone bridge is used extensively in locating faults in both cable and open wire conductors. The simplest test of this kind is the location of a cross between two wires. Figure 62 shows the Wheatstone bridge connected to the office end of a cable pair which has its conductors crossed together some distance from the office. If the cross itself has zero resistance (i.e., if the wires are "dead crossed") the location is a simple one. The unknown resistance as measured is merely the loop resistance of the pair of cable conductors from the office to the point of the defect, and this length may be easily determined from the resistance measurement and the values given in Table IV, or by comparison with the measured resistance of another loop of known length.

Example: In Figure 62 the cable conductors are #19 B. & S. gage, and the cross between conductors is assumed to have zero resistance. If the value of X, as measured by the Wheatstone bridge, is 55 ohms, how far is the cross from the telephone office?

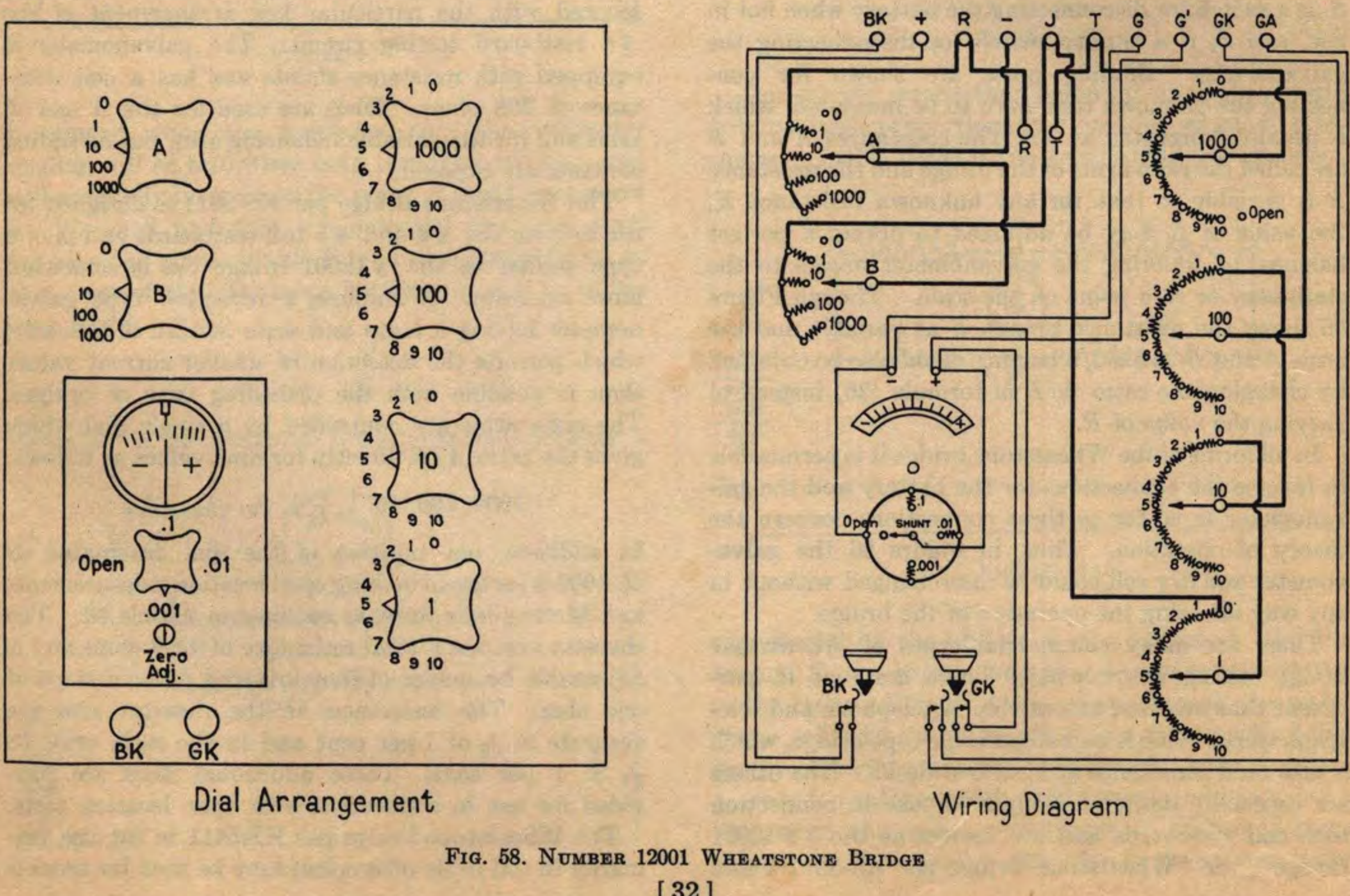
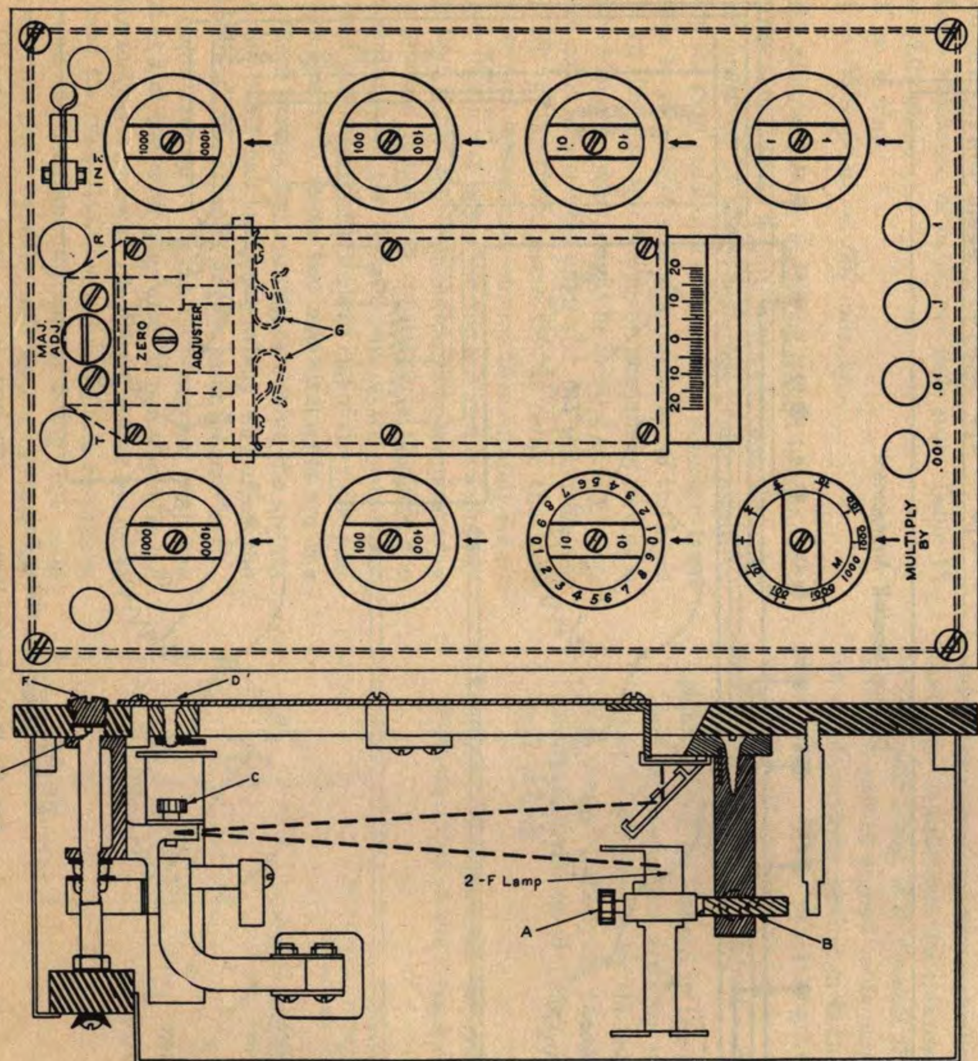
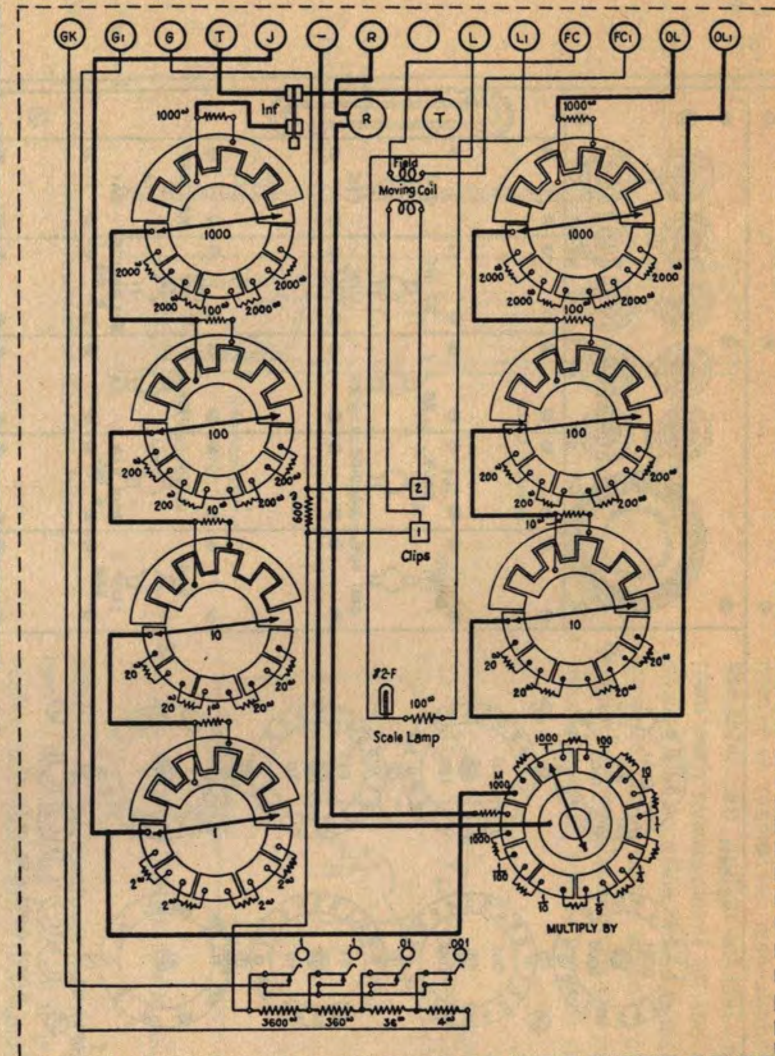


Fig. 58. Number 12001 Wheatstone Bridge

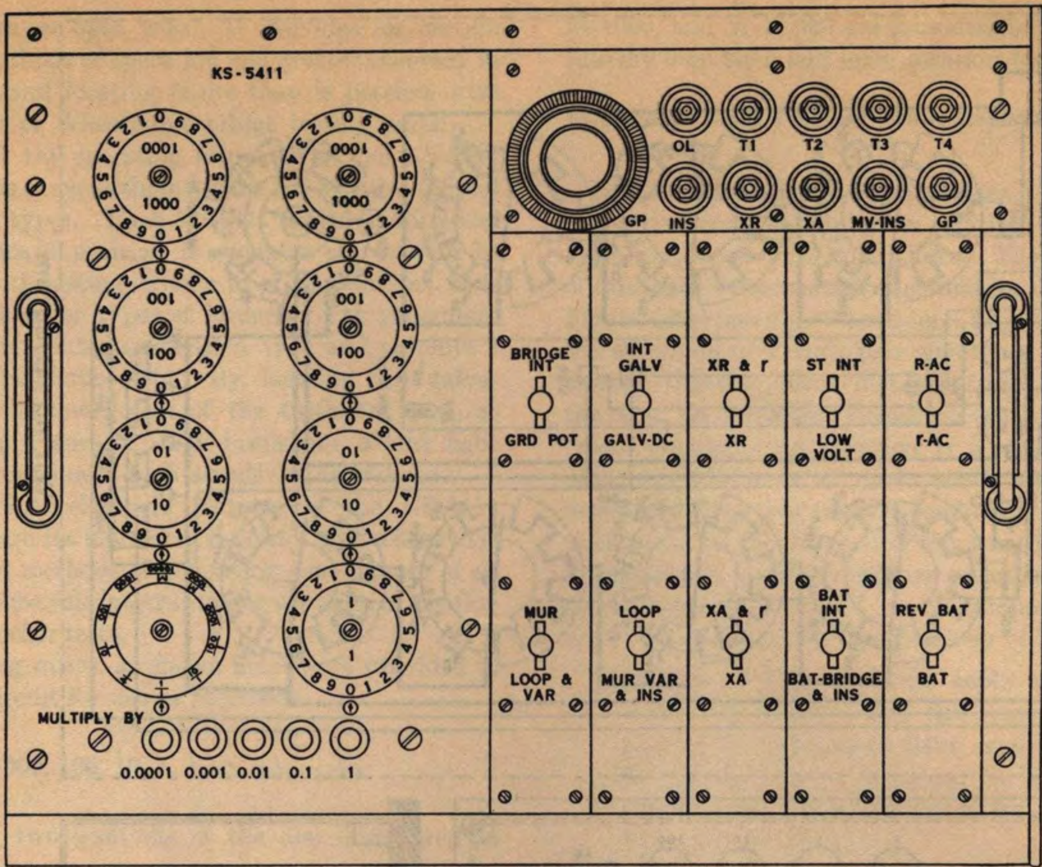


Dial And Galvanometer Arrangement

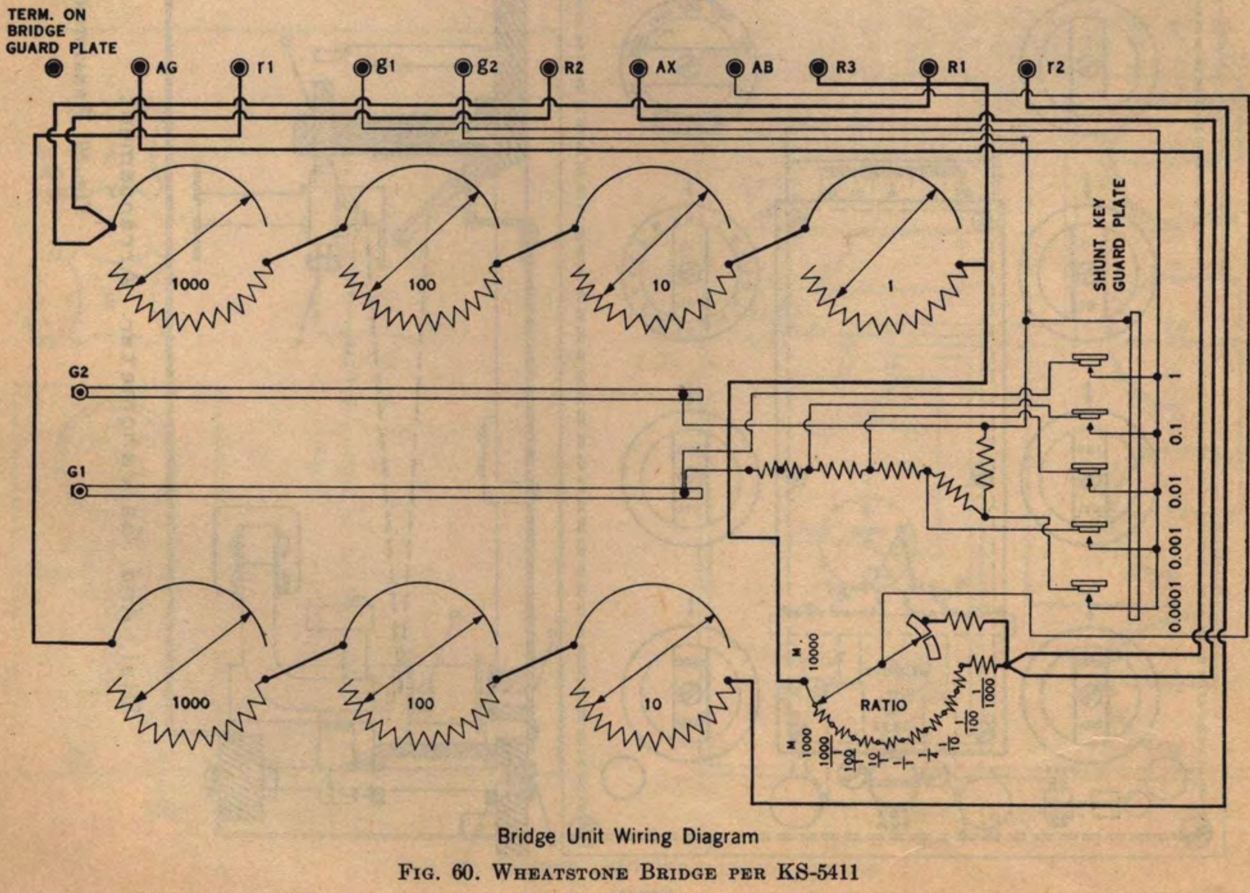


Wiring Diagram

Fig. 59. Wheatstone Bridge per KS-3011



Dial and Keyshelf Arrangement



Bridge Unit Wiring Diagram

FIG. 60. WHEATSTONE BRIDGE PER KS-5411

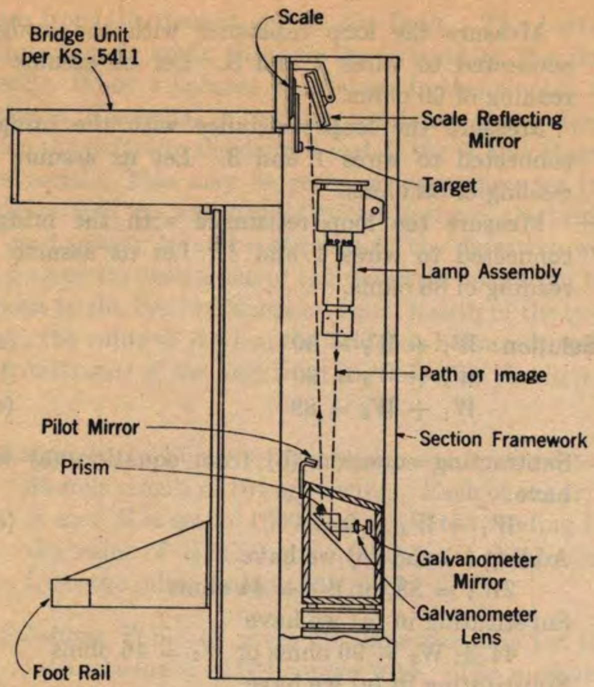


FIG. 61. OPTICAL ARRANGEMENT OF KS-5411 BRIDGE

Solution: Let d = distance in miles
Loop resistance of cable per mile from Table IV is 85.01 ohms
$X = 55$ ohms
$d = 55 \div 85.01 = .657$ mile, ans.

If the cross shown in Figure 62 should itself have a resistance value, which is quite often the case, the value of X as measured by the Wheatstone bridge would be equal to the loop resistance of the cable conductors from the office to the defect plus the resistance of the cross itself. In this case the defect, on account of having a definite resistance value, might be located at any intermediate point between the office and .657 mile from the office. It is, therefore, necessary in using the loop method to make two measurements to accurately determine the location of any cross when it is not definitely known that it has zero resistance. The simplest way to determine if it has zero resistance is to make one measurement with the distant end of the cable pair open, and another with the distant end of the cable pair short-circuited. **If these two measurements are the same, which means that opening or closing the distant end of the cable pair does not in any way affect the measurement, the cross is known to have zero resistance** (dead crossed). If the measurement with the distant end of the cable pair crossed is lower in value than the measurement with the distant end of the cable pair open, the cross itself has some definite resistance value, and the location, instead of being .657 mile away, is some point between .657 mile away and the office. One way to determine the exact location in this case is to make loop measurements from each end of the cable pair, and to calculate an imaginary location from each measurement on the assumption of a zero cross. The location, when calculated from the measurement made at the office end, will be too far away, and when calculated from the measurement made at the distant end, will be too near the office. The actual

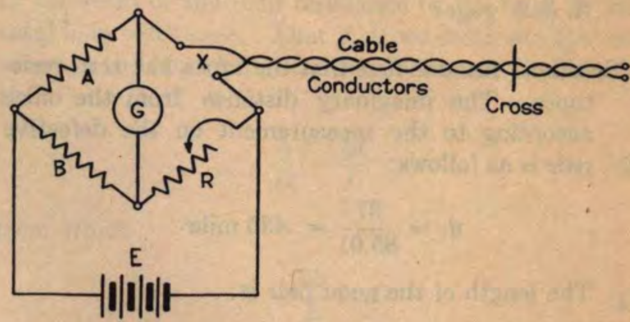


FIG. 62. LOOP RESISTANCE MEASUREMENT TO CROSS

location is the mean, or point half way between the two. Of course, in practice it is oftentimes not convenient to transfer the Wheatstone bridge to the distant end of the circuit in order to make measurements from that end. A substitute for this method, which amounts to the same thing, is to connect the distant end of the defective pair to a good cable pair as shown in Figure 63. This permits testing in both directions from the same office. If the exact length of the good pair is not known, it can always be determined by making a measurement with the distant end crossed.

Note: It will be learned in the next article that the quickest and most accurate method of locating a cross in practice is by the metallic Varley. But the theory underlying the foregoing should be thoroughly mastered before taking up the later type of test.

Example: In Figure 63 the good cable pair when short-circuited at the distant end has a loop resis-

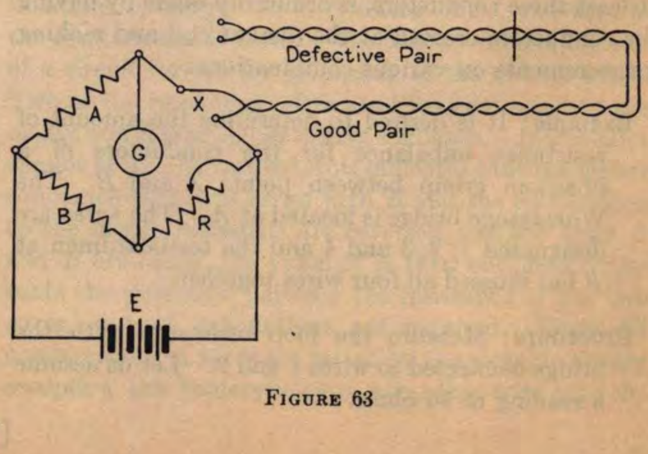


FIGURE 63

tance of 63 ohms. When connected to the defective pair as shown, the measurement from the office to the cross over the good pair and the distant end of the defective pair is 108 ohms. The measurement from the office to the cross on the defective pair is 37 ohms. What is the distance in miles from the office to the defect, and what is the resistance of the cross if the cable pairs are 19 B. & S. gage?

Solution: Assume first that the cross has zero resistance. The imaginary distance from the office according to the measurement on the defective pair is as follows:

$$d_1 = \frac{37}{85.01} = .435 \text{ mile}$$

The length of the good pair is:

$$l = \frac{63}{85.01} = .741 \text{ mile}$$

The imaginary distance from the distant end is:

$$l - d_2 = \frac{108}{85.01} - .741 = .529 \text{ mile}$$

Then the actual cross is at point half way between .435 mile from the near office and .529 mile from the distant office or .741–.529 = .212 mile from the near office. The actúal location is therefore

$$\frac{.435 + .212}{2} = .323 \text{ mile, ans.}$$

The resistance of the cross caused an error in the single measurement location of .435 − .323 = .112 mile of cable pair. This expressed as resistance is .112 × 85.01 = 9.5 ohms, ans.

Another application of the simple loop resistance measurement is to determine any inequality in the resistance of individual conductors, or as is commonly expressed, to locate "resistance unbalances" due to such causes as defective splices in cable pairs or defective sleeve joints in open wire. This test, requiring at least three conductors, is ordinarily made by having the conductors crossed at the distant end and making measurements on various combinations:

Example: It is desired to determine the amount of resistance unbalance for the conductors of a phantom group between points *A* and *B*. The Wheatstone bridge is located at *A*. The wires are designated 1, 2, 3 and 4 and the testboardman at *B* has crossed all four wires together.

Procedure: Measure the loop resistance with the bridge connected to wires 1 and 2. Let us assume a reading of 90 ohms.

Measure the loop resistance with the bridge connected to wires 2 and 3. Let us assume a reading of 90 ohms.

Measure the loop resistance with the bridge connected to wires 1 and 3. Let us assume a reading of 88 ohms.

Measure the loop resistance with the bridge connected to wires 1 and 4. Let us assume a reading of 88 ohms.

Solution: $W_1 + W_2 = 90$ (a)

$W_2 + W_3 = 90$ (b)

$W_1 + W_3 = 88$ (c)

Subtracting equation (b) from equation (a) we have:

$W_1 - W_3 = 0$ (d)

Adding (c) and (d) we have

$2W_1 = 88$, or $W_1 = 44$ ohms

Substituting in (a) we have

$44 + W_2 = 90$ ohms or $W_2 = 46$ ohms

Substituting in (c) we have

$44 + W_3 = 88$ ohms or $W_3 = 44$ ohms

And since $W_1 + W_4 = 88$, we likewise have

$44 + W_4 = 88$ or $W_4 = 44$ ohms

Thus we learn W_2 has a resistance unbalance of 2 ohms, ans.

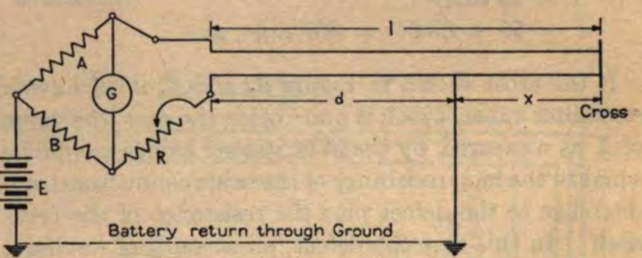


Fig. 64. Grounded Varley Test

39. Varley Loop Tests

A Wheatstone bridge may be used to locate a defect due to a grounded open wire or cable conductor as well as a defect due to a cross between conductors. There are two recognized methods of making tests of this kind. One is known as the Varley loop test and is the more generally used; the other is known as the Murray loop test. Figure 64 shows the theory of the Varley loop test.

In comparing this figure with Figure 56, we can recognize a Wheatstone bridge circuit with the connections made in a little different way. The variable resistance R is in series with the resistance d of the defective wire from the office to the fault. The resistance X of Figure 56 becomes in Figure 64 the series resistance l of the good wire from the office to the distant end, plus the resistance x of the defective wire from the distant end to the fault. The battery connection is made through the ground to the fault itself. When a balance is obtained in this circuit, the value of R is equal to the loop resistance of the circuit from the defect to the distant end, if the A and B arms are equal. This may be seen by inspection, for it is evident that the adjustment of the R arm of the bridge is used merely to add resistance to the defective wire and since the resistance of the defective wire from the bridge to the fault balances an equal length of the good wire, the value of R when the bridge is balanced equals the resistance of the loop from the defect to the distant end.

Example: In Figure 64 the bridge is connected to a 30-mile circuit of 104 open wire. Each of the arms A and B is set at 1000 ohms. If the reading for the value of R is 22 ohms, how far is the ground from the office making the test?

Solution: Table IV gives 10 ohms per mile for the loop resistance of 104 copper wire. The measurement of 22 ohms represents the resistance of the loop from the defect to the distant end. This distance is therefore $22/10 = 2.2$ miles. If the circuit is 30 miles long, the defect is located $30 - 2.2$ or 27.8 miles from the measuring office, ans.

The above example assumes that the two wires of Figure 64 are alike, and that the loop resistance values per mile given in the table are correct under all conditions. Although the first assumption will usually be true in practice, unit resistance values may vary appreciably due to temperature differences. In either event, it is still possible to locate the fault, by making an ordinary loop resistance measurement on the pair, in addition to the Varley measurement.

Thus, referring again to Figure 64, it will be seen that when a Varley balance is obtained, with the bridge ratio arms equal—

$$R + d = l + x \qquad \text{(a)}$$

or

$$d = (l + x) - R \qquad \text{(b)}$$

This, of course, is true regardless of whether the good and the defective wires are of the same make-up.

Similarly, if the loop resistance L is measured, we have—

$$L = l + x + d \qquad \text{(c)}$$

from which

$$d = L - (l + x) \qquad \text{(d)}$$

Now, adding (b) and (d), we get

$$2d = L - R \qquad \text{(e)}$$

In other words, **the loop resistance from the measuring end to the fault is equal to the loop resistance measurement minus the Varley measurement.** Since we do not know the unit resistance value of the two wires, we still do not know the distance to the fault. It is obvious, however, that if the wires are of uniform make-up throughout their whole length, the ratio of the distance to the fault to the total length of the line will be equal to the ratio of the loop resistance to the fault to the total loop resistance. That is, if we designate the distance to the fault k and the total length of the line, D, then—

$$\frac{k}{D} = \frac{2d}{L} \qquad \text{(f)}$$

from which

$$k = \frac{2d}{L} \times D \qquad \text{(g)}$$

or, applying (e) above—

$$k = \frac{L - R}{L} \times D \qquad \text{(h)}$$

Example: Assume as in the above example that the total circuit length is 30 miles and that the Varley reading is 22 ohms with the ratio arms equal. If a loop resistance measurement gives 300 ohms, what is distance of the ground from the measuring end?

Solution:

$$k = \frac{300 - 22}{300} \times 30 = 27.8 \text{ miles, ans.}$$

A modification of the Varley test may be used for accurately measuring resistance unbalances, which is in some respects preferable to the method of combination loop measurements described in the foregoing article. It is called the metallic Varley, and is shown by Figure 65-A. In making this test, all wires are short-circuited at the distant end in the same manner as when making a series of loop tests for the various combinations of wires. At the testing office, one wire of the combination is used for the battery return, instead of a circuit formed by grounding at the distant office. Two of the remaining wires are then connected to the bridge and R is adjusted to give a balance. If a balance cannot at first be secured, this indicates that the higher resistance wire is in series with R and the connections to the bridge terminals are reversed. If the arms A and B are equal, the value of R then obtained represents the difference between the resistance of the two wires, and no calculations are required. When all combinations of wires are tested by the metallic Varley excepting the battery return wire, this wire may be interchanged with any one of the others and included in the tests.

A similar test requiring only three wires is commonly used in testboard work for locating crosses, particularly those having high resistance. As noted in the preceding article, the location of a cross having resistance by the use of loop resistance measurements involves certain difficulties. By using a good third wire of the same gage as that of the pair in trouble, and connecting the bridge for a metallic Varley measurement as shown in Figure 65-B, the resistance of the cross is removed from the "balanced" circuit of the bridge and placed in the battery circuit. Here it has no effect on the measurement, providing its resistance is not so high that the current supplied to the bridge is insufficient for its satisfactory operation. As may be seen from the diagram of connections, when the bridge is balanced with equal values in the ratio arms A and B, the resistance of the good third wire, plus the resistance of one wire of the crossed pair from the distant end to the fault, is equal to the resistance of one wire from the fault to the measuring end plus the resistance, R, in the rheostat arm of the bridge; or, we may write—

$$l + (l - d) = d + R$$

from which—

$$d = \frac{2l - R}{2}$$

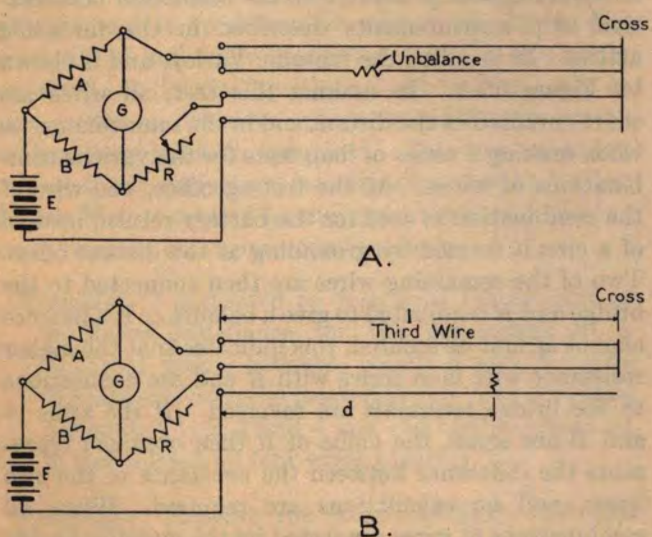
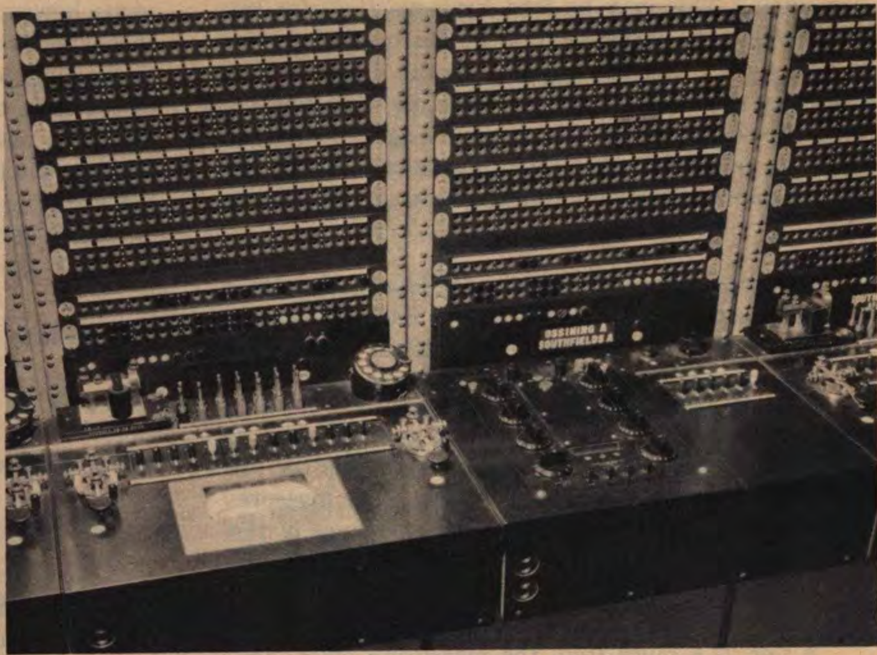


Fig. 65. Metallic Varley Tests

In locating a cross by this method in practice, it is only necessary to make a Varley measurement as described above and a loop resistance measurement on the pair consisting of the good third wire and one wire of the crossed pair, shorted together at the distant end. Then the loop resistance of the crossed pair from the measuring end to the fault may be obtained directly by subtracting the Varley reading from the loop resistance reading.



Testshelf of No. 5 Testboard Showing Voltmeter and Wheatstone Bridge Testing Arrangement

The Varley test may also be used for locating a cross between one wire of a circuit and some other wire, of different characteristics such as one wire of an iron circuit. The procedure here is to ground the wire of the second circuit, cross the first circuit at the distant end, connect the bridge to it and locate the ground by the Varley method described above, which is equivalent to locating the cross.

40. Murray Loop Tests

The theory of the Murray loop test is similar to that of the Varley. But instead of setting the arms A and B to have equal values and using the adjustable dials R to compensate for the difference in wire resistance between the good wire connection and the defective wire connection, the arm B is eliminated altogether and the variable resistance arm is connected in its place as shown in Figure 66. In this arrangement, the ratio of the reading R to the setting of the arm A is equal to the ratio of the resistance of the defective wire from the measuring office to the ground to the resistance of this same wire from the ground to the distant office plus the resistance of the good wire, or expressed mathematically—

$$\frac{R}{A} = \frac{l-d}{l+d}$$

This, of course, assumes that the defective and good wires have the same series resistance per mile, as would ordinarily be the case where for any given circuit being tested the defective wire's mate is used.

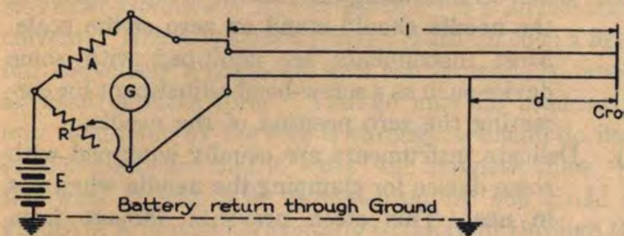


FIG. 66. MURRAY TEST

Example: In Figure 66, the arm A is set at 1000 ohms, and the bridge is balanced by varying the arm R. If the value of R is 634 ohms and the length of the circuit under test is 65 miles, what is the distance from the testing office to the fault?

Solution: The simple bridge relation gives

$$\frac{R}{A} = \frac{l-d}{l+d}$$

or,

$$\frac{634}{1000} = \frac{(65-d) \times \text{res. per mile}}{(65+d) \times \text{res. per mile}}$$

If the resistance per mile of each wire is the same, this factor will cancel and we have—

$$\frac{634}{1000} = \frac{65-d}{65+d}$$

which gives by cross multiplying

$$634\,(65+d) = 1000\,(65-d),$$

or

$$1634d = 23790$$

from which

$$d = 14.56 \text{ miles}$$

$$l - d = 50.44 \text{ miles, ans.}$$

There are a number of other standard tests made with the Wheatstone bridge and with these as well as with the tests that have been described, the procedure in practice is somewhat more involved than the simple theory might indicate. There are in nearly all practical tests various complicating factors such as temperature variations, effect of loading coils, short lengths of cable, irregular facilities, etc., which must all be considered if accurate locations are to be made.

For instance, in the majority of toll cables, parts of each section are aerial and other parts are underground. There is usually a considerable temperature difference between the two types of facilities, as a result of which the resistance of the wires in the two types of facilities is different. It is therefore necessary to apply correction factors to the measurements taken in order to accurately locate a fault. The details of how these various factors are taken care of in practice present a rather complete study in themselves, however, and their consideration is necessarily beyond the scope of this book. The intent here has been only to treat a few of the outstanding testing methods in a more or less theoretical way, with a view to establishing the general principles upon which all testing work is based.

NO. 5 PRIMARY TESTBOARD

41. Precautions to be Taken in the Use of Measuring Instruments

The intelligent use of the electrical measuring instruments described in this chapter and in Chapter IV depends upon skill and care, as well as upon an understanding of the theory of each test made. There are numerous precautions to be taken, some of which are to be learned only through experience, but a few cardinal ones are listed here and may be studied accordingly. In the use of all testing instruments there is one important motive,—the most accurate results possible should be secured without damaging the instruments.

a. **Precautions should be taken against dropping or jarring** electrical measuring instruments, particularly galvanometers, voltmeters, ammeters, and wattmeters. Instruments of this class have revolving coils, either suspended in a delicate manner or equipped with jewelled bearings. Jarring may permanently damage the instruments or impair their accuracy.

b. Electrical measuring instruments should be **kept free from dampness.**

c. **Instruments should never be connected to circuits where the values to be measured are likely to be greater than the highest scale reading,** and the low resistance coils of a Wheatstone bridge should not have E.M.F.'s impressed across them sufficiently high to cause currents greater than the carrying capacity of these coils. In general, for ordinary testing, the *A* and *B* arms should not be set upon low values if the voltage of the testing battery is comparatively high.

d. **Milliammeters and millivoltmeters are designed for measuring millivolts and milliamperes, not volts and amperes.**

e. **Ammeters must always be inserted in series** and never across any branch or any part of an electrical circuit.

f. In any Wheatstone bridge test, **the galvanometer and the battery key should not be closed excepting when the tester is actually endeavoring to obtain a bridge balance.** If the galvanometer is equipped with a shunt, **the lowest resistance shunt should be bridged across the galvanometer when beginning to obtain a balance.** When an approximate balance is obtained in this way, the next lowest value shunt should be bridged, and thus by degrees the galvanometer coil may be connected into the circuit.

g. In making Wheatstone bridge measurements, **precautions should be taken against extraneous sources of E.M.F. such as ringing current, cords with battery, telegraph legs, etc.** being connected to the circuit at the distant end or at some intermediate point while a measurement is being made.

h. For accuracy when reading scales of electrical instruments **the eye should be directly over the needle,** that is, the line of vision should be perpendicular to the scale. Some instruments are equipped with small mirrors beneath the needle in order to guard against reading the scale at an angle. When the image of the needle in the mirror is directly beneath the needle, the eye is perpendicular to the scale. An error caused by the eye not being perpendicular to the scale is called error due to parallax.

i. When an indicating instrument is not connected, the **needle should stand on zero of the scale.** Most instruments are equipped with some device such as a screw-head adjustment for correcting the zero position of the needle.

j. Delicate instruments are usually equipped with some device for **clamping the needle when not in use.** This often prevents damage from jarring. Instruments not equipped with a clamping device may have their coils "dampened" by short-circuiting their terminals. Care must be taken that the short-circuit is removed before using the instrument.

k. Errors are often encountered in Wheatstone bridge measurements on long cable and open wire circuits due to foreign potentials caused by induction, ground potentials, etc. To correct for these, it is often necessary to reverse the polarity of the testing battery and make a second measurement. The average of the two measurements may be recorded as the correct one.

l. In making Wheatstone bridge measurements, precautions should be taken that the circuit under test is absolutely cleared of all bridged or other apparatus not permanently associated with the circuit and essential for giving simple continuity.

m. **Beware of loose connections.** Make sure that all connections to binding posts or elsewhere have zero resistance. Do not use high resistance wires for leads.

n. **Make a mental estimate of the value you expect to read** from your knowledge of the conditions. This will often prevent mistakes due to errors that are obvious, and will usually prevent the improper use of the instruments.

CHAPTER VI

THE DIRECT-CURRENT DYNAMO-ELECTRIC MACHINE

42. Induced Electromotive Force

Chapter III describes how lines of magnetic induction exist around any wire in which there is an electrical current. Not only does a current establish such a field, but conversely a magnetic field can be made to create an electromotive force. Voltage may be induced in any conductor by moving it through a magnetic field in such a manner that it "cuts" the magnetic lines. If the wire indicated in cross-section by the circle in Figure 67 is moved horizontally to the right through the magnetic lines having a direction vertically downward, it may be considered that the wire displaces or "stretches" the lines, which may be thought of as possessing a certain elasticity. This finally causes them to wrap themselves around the conductor, as shown. Referring to Figure 38 in Chapter III and applying this figure conversely to our new conditions, we find that a magnetic field which loops around a conductor in a clockwise direction, gives rise to a current flowing **into** the conductor as seen in cross-section. This is illustrated in Figure 67-D and E.

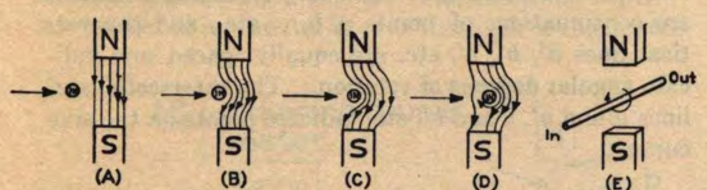


FIG. 67. WIRE MOVING THROUGH MAGNETIC FIELD

This rule, stated in another way, is called the **right-hand rule** for remembering the induced E.M.F. relation. It is illustrated in Figure 68. The forefinger of the right hand represents the direction of the lines of magnetic induction (flux—north to south); the thumb, when pointed perpendicular to the forefinger, represents the direction in which the conductor moves; and the second finger, when perpendicular to both the forefinger and the thumb, gives the direction of the induced E.M.F., or the direction of current flow. If a galvanometer is connected to the conductor, as in Figure 69, it will be found that the effect is more noticeable when the conductor is moved swiftly. From these and other similar experiments we learn that the law for induced E.M.F. may be stated as follows:

When any conductor is made to cut lines of magnetic induction there will be an E.M.F. induced in it, and the direction of the E.M.F., the direction of the flux, and the direction of the motion of the conductor have a perpendicular relation as shown by the right-hand rule. The amount of induced E.M.F. depends upon the rate of cutting magnetic lines, or the number of lines cut per second. In the established system of electrical and magnetic units, an E.M.F. of one volt is induced when a conductor cuts 100,000,000 lines per second.

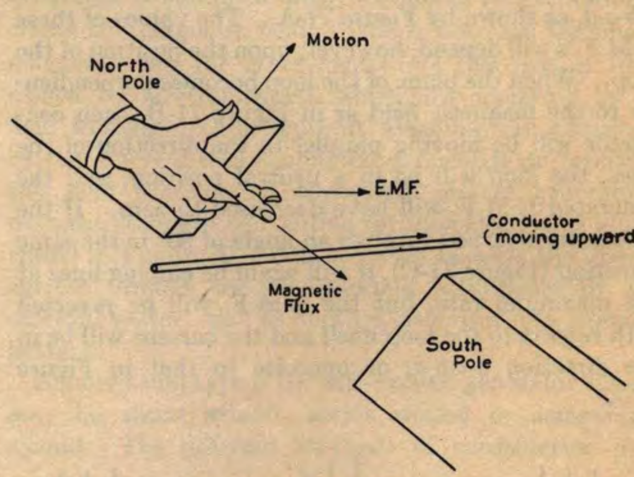


FIG. 68. RIGHT-HAND RULE

43. E.M.F. Induced in a Revolving Loop

Instead of a single conductor cutting lines of magnetic induction, we may have a loop of wire revolving in the

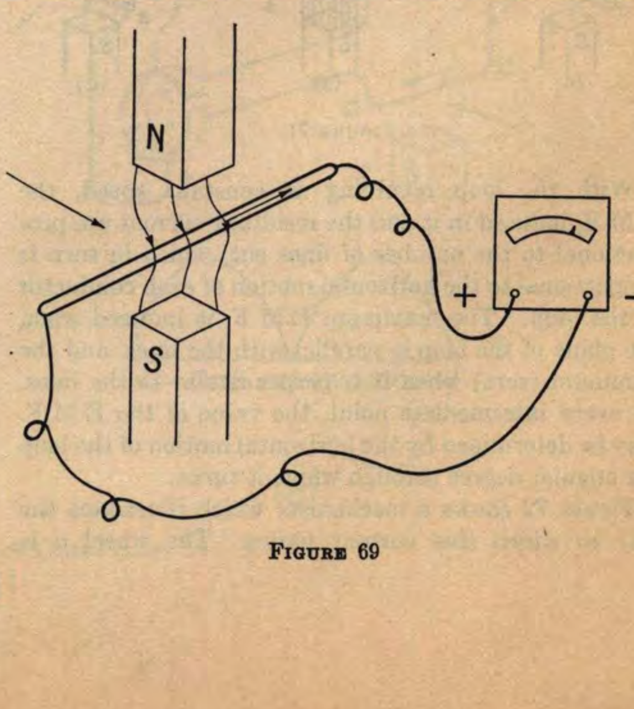


FIGURE 69

magnetic field between the poles of a magnet, as shown in Figure 70. In this case, the conductor nearest the south pole moves to the left while the conductor nearest the north pole moves to the right, and the E.M.F. induced has a different direction in the two conductors.

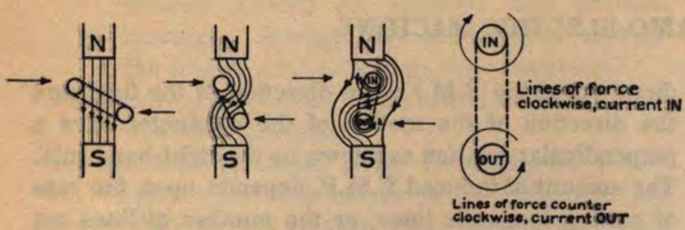


FIG. 70. LOOP ROTATING IN MAGNETIC FIELD

But, because the loop is complete, these E.M.F.'s will aid in causing a continuous current in the direction *a-b-c-d*, as shown by Figure 71-A. The values of these E.M.F.'s will depend, however, upon the position of the loop. When the plane of the loop becomes perpendicular to the magnetic field as in Figure 71-B, each conductor will be moving parallel to the direction of the lines, the loop will be in a neutral position, and the generated E.M.F. will have decreased to zero. If the loop is then turned through an angle of 90° in the same direction (Figure 71-C), it will again be cutting lines at the maximum rate, but the E.M.F. will be reversed with respect to the loop itself and the current will be in the direction *d-c-b-a*, or opposite to that in Figure 71-A.

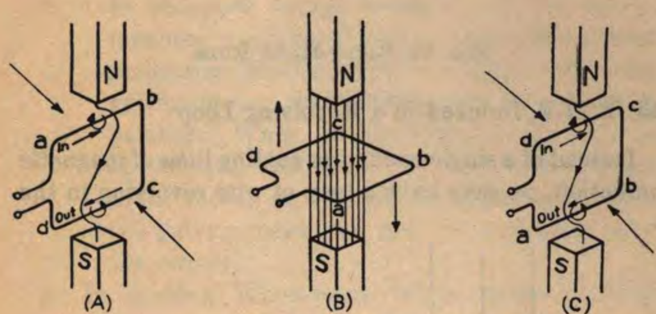


FIGURE 71

With the loop revolving at constant speed, the E.M.F. induced in it and the resultant current are proportional to the number of lines cut, which in turn is proportional to the **horizontal** motion of each conductor of the loop. The maximum E.M.F. is induced when the plane of the loop is parallel with the lines, and the minimum (zero) when it is perpendicular to the lines. At every intermediate point, the value of the E.M.F. may be determined by the horizontal motion of the loop per angular degree through which it turns.

Figure 72 shows a mechanism which illustrates the way in which this current varies. The wheel *d* is rotating at a constant speed, causing the attached pin *a* to slide in the slot *o*, moving the bar *b* (with the pencil *e* attached) vertically between the guides *gg*. When the horizontal component of the motion of the pin *a* is a maximum, that is, when the motion is in an entirely horizontal direction, the pencil *e* is at either its highest or lowest position, depending upon whether the motion of *a* is from left to right or right to left. When the horizontal motion of *a* is zero, *e* is midway between its extreme high and low points. If *f* represents a strip of paper which is being moved horizontally to the right at a constant speed while the wheel *d* is also rotated at a constant speed, the pencil *e* will draw a curve as shown. This curve will indicate a positive maximum (or highest point) when the horizontal motion of *a* to the right is a maximum; and will indicate center or zero points when the horizontal motion of *a* is zero. If the pin in this mechanism represents one conductor of a loop of wire revolving in a vertical magnetic field, the position of the pencil *e* with respect to the mid-point of its travel represents the E.M.F. induced in the conductor. This analogy is apparent since the induced E.M.F. in each loop is proportional to the horizontal motion of the loop. The curve not only represents maximum and zero points but shows all intermediate values of the induced E.M.F. as well.

Such a curve is called a **sine wave.** It is the **fundamental** wave form in alternating-current circuits of all kinds. A sine wave may be actually plotted by the method shown in Figure 73, where the horizontal lines are continuations of points *a*, *b*, *c*, etc., and the vertical lines *a′*, *b′*, *c′*, etc. are equally spaced and indicate **angular degrees of rotation.** The intersections of lines *a* and *a′*, *b* and *b′*, etc. indicate points on the sine curve.

44. Principle of the Direct-Current Generator

The revolving loop or **armature** shown in Figure 71

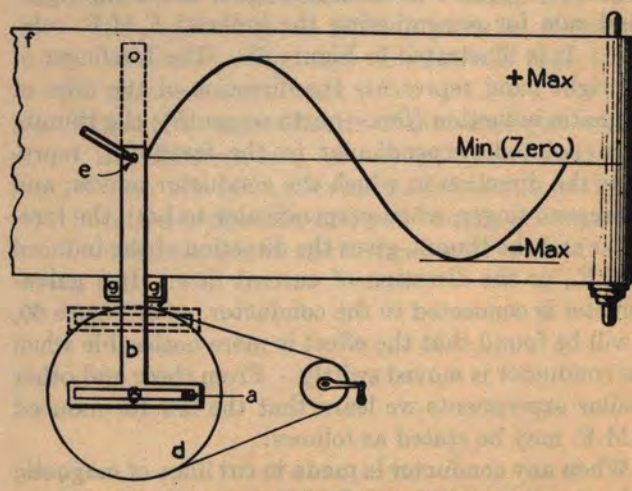


FIG. 72. MECHANISM FOR DRAWING SINE WAVE

may be connected to slip-rings, as shown in Figure 74-A. In this case the resulting E.M.F. between the two terminals or **brushes** will reverse in direction as the loop revolves, giving rise to an **alternating** E.M.F., one **cycle** of which is plotted in the figure. If it is desired to produce a unidirectional E.M.F., it is necessary to

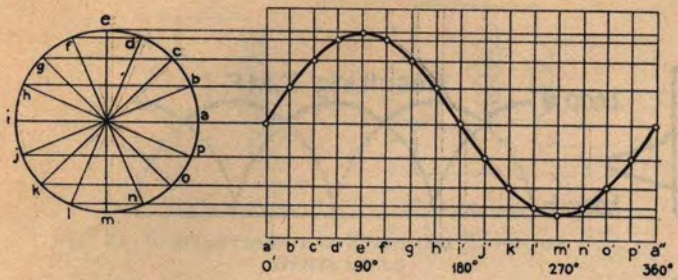


Fig. 73. Graphical Construction of Sine Wave

devise some means for reversing the connections to the loop at the same time that the current in the loop reverses. This is done by means of the **commutator** shown in Figure 74-B. This commutator effects the reversing of the connections to the armature leads just as the E.M.F. or current is reversed, changing the negative half-cycle to a positive pulsation. The resultant E.M.F. then consists of two positive pulsations per revolution of the loop, as shown.

Generators may be constructed with more than one loop, as in Figure 75 in which two loops and four commutator segments are shown. The resultant E.M.F. is represented by the full lines at the right of the figure. Comparing Figure 74-B with Figure 75, it may be seen that an increase in the number of loops causes a smaller fluctuation in the armature E.M.F. An armature wound with many turns therefore produces a practically continuous non-pulsating E.M.F., causing a **direct** current.

A standard generator armature consists of a large number of loops or turns which may be wound in several ways, depending upon the number of poles and the speed and voltage desired. The actual winding, however, consists of one continuous conductor, from which taps are brought out and connected to commutator segments, instead of a large number of separate loops.

In Figures 71, 74 and 75, we have assumed that the generator is equipped with permanent magnets which create the magnetic field. This is the case for small magnetos, but for other generators this field is furnished by electromagnets which are energized by a **field winding.** Direct-current machines are classified by the different means adopted to energize or "excite" this field winding. A **separately excited generator,** with the standard convention for indicating it, is illustrated in Figure 76. It is so called because the direct current through the field winding is furnished by an external source, such as another generator or a storage battery.

A more usual type is the **self-excited generator** which may be **shunt wound, series wound** or **compound wound.** The different methods of construction are

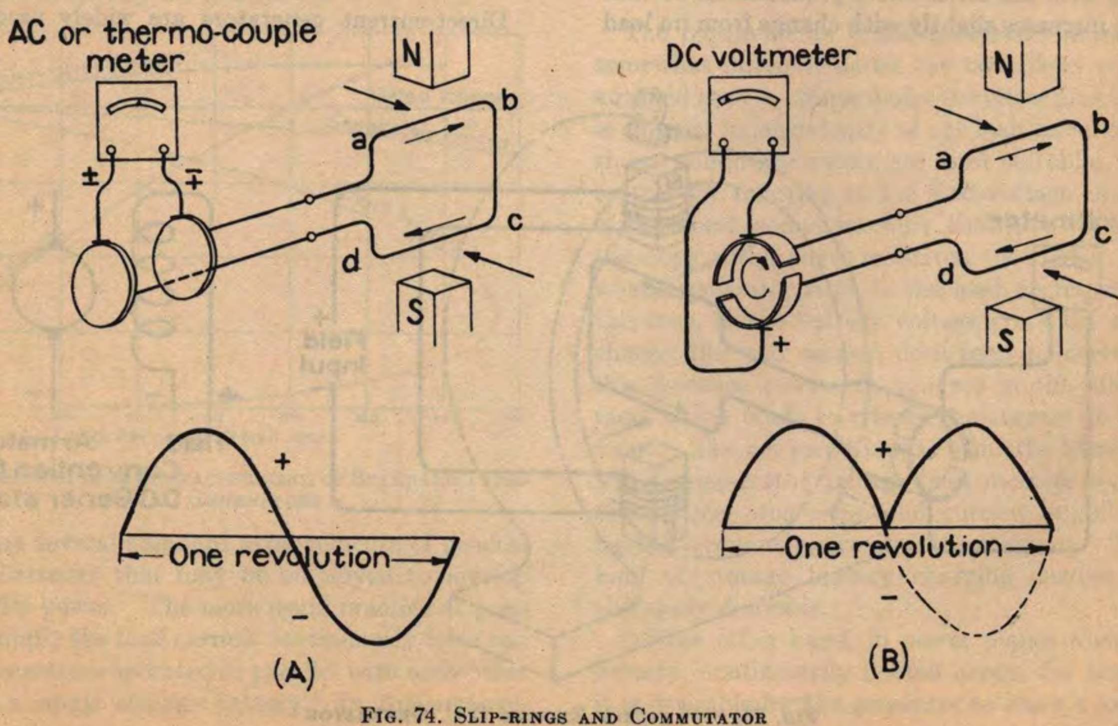


Fig. 74. Slip-rings and Commutator

shown schematically in Figure 77. As the E.M.F. induced in the armature is proportional to the magnetic flux, which in turn is proportional to the current in the field windings, a variation in the field current will cause a change in the armature E.M.F. With the shunt wound generator, an increase in load current causes a decrease in field current, as may be seen from a study of parallel resistances, and consequently causes a decrease in armature E.M.F. On the other hand, in the case of the series wound generator, the armature E.M.F. increases with the load current. The compound wound generator is designed to neutralize this change in armature E.M.F. by balancing the series effect against the shunt effect. An **over-compounded** generator is constructed with the series effect predominant so that the voltage increases slightly with change from no load to full load. Figure 78 shows curves representing armature voltage plotted against load for these various types of generators.

Generators may be further classified by the number of poles, a four-pole machine being represented by Figure 79. In every case there are the same number of brushes as poles, alternate brushes being connected together, as shown, to form the armature terminals. The voltage with four poles will be double that with two poles if the same armature winding and the same machine speed are used.

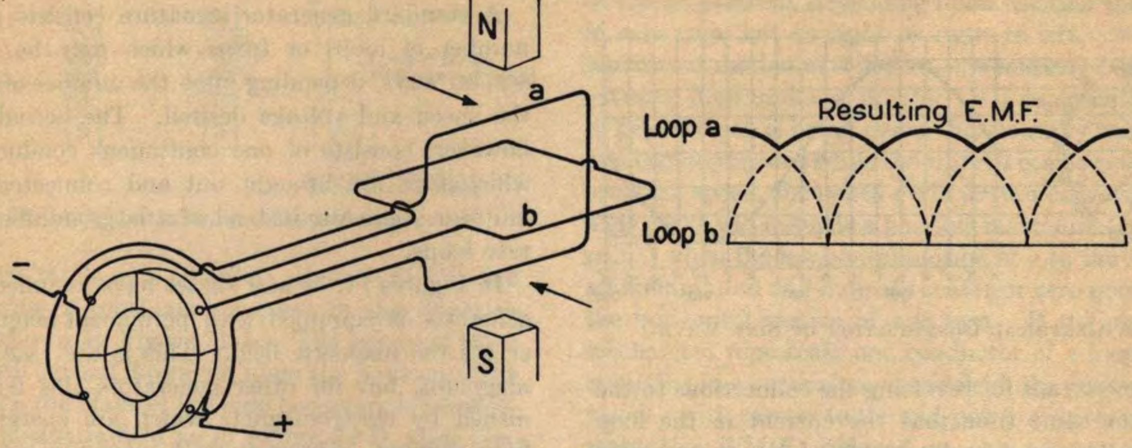


Fig. 75. Effect of Additional Loops

45. D.C. Generators for Supplying Central Office Power

Direct-current generators are widely used in tele-

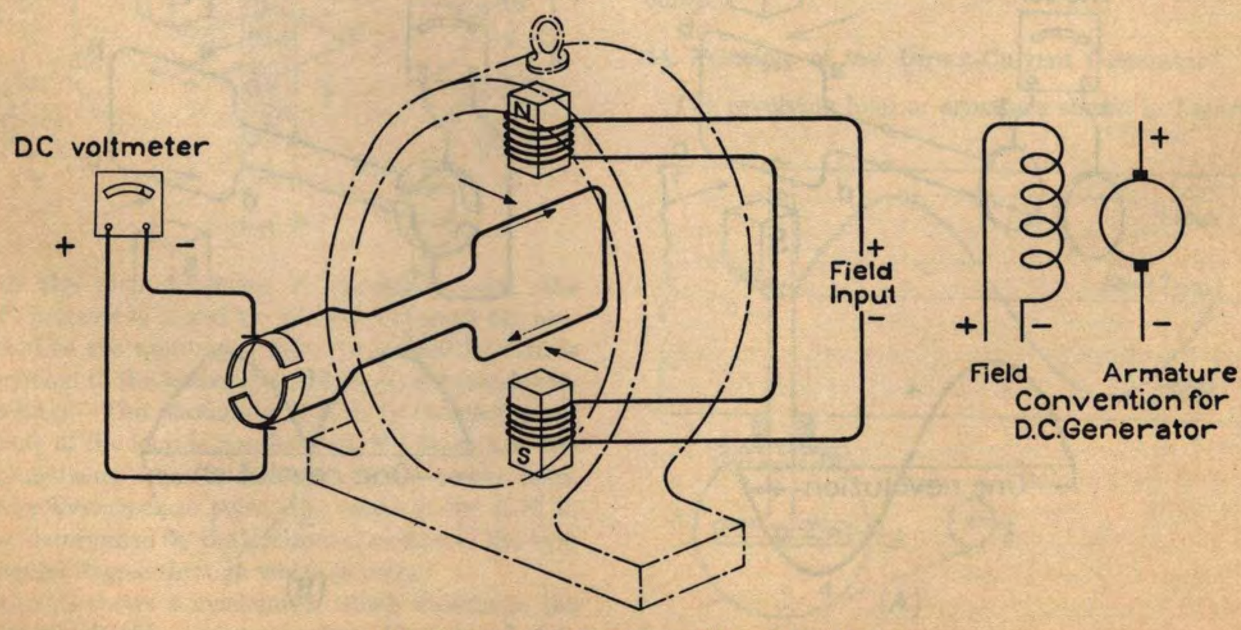


Fig. 76. Separately Excited D.C. Generator

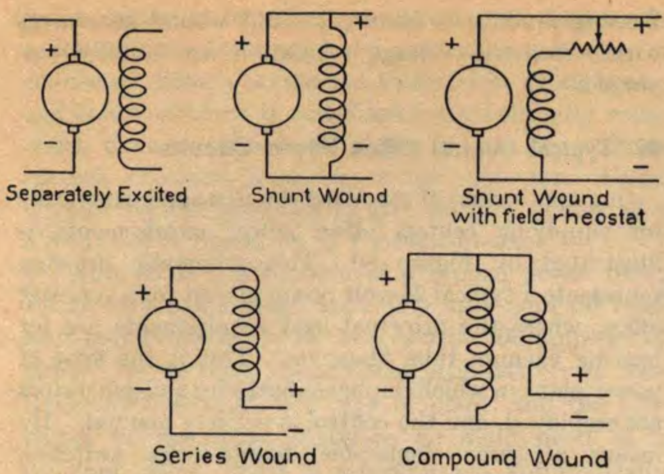


FIG. 77. CONVENTIONS FOR STANDARD TYPES OF D.C. GENERATORS

phone and telegraph work for supplying the several voltages required to operate the central offices. These include 24 and 48 volts used for "talking battery" and for operating vacuum tube filaments, relays, etc.; and also several higher voltages, ranging up to a maximum of 152 volts, for vacuum tube plate supply and operation of telegraph circuits. The motors which drive the generators are ordinarily supplied with power from commercial power lines and to guard against the possible failure of this supply, storage batteries are always provided in central offices. These batteries are kept charged by the central office generators so that they can take over the load temporarily in case of failure of the primary power supply. Being always connected to the load, the storage batteries also act as filters to reduce noise caused by the generators.

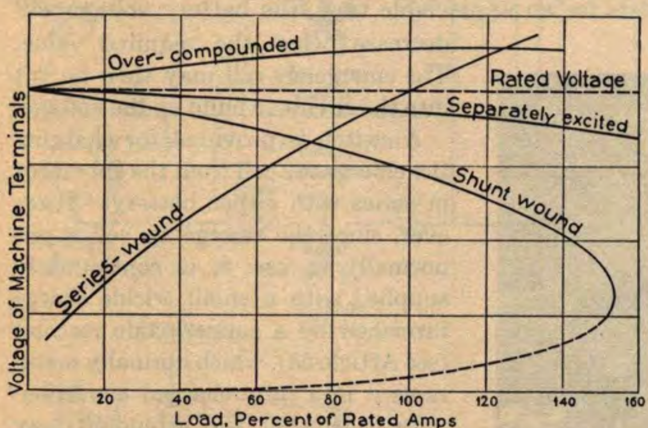


FIG. 78. LOAD-VOLTAGE CHARACTERISTICS OF STANDARD TYPES OF D.C. GENERATORS

There are several standard arrangements of generators and batteries that may be employed to develop central office power. The more usual practice at present is to supply the load current continuously from one or more generators operated in parallel with each other and with a single storage battery. In this arrangement, the storage battery is "floated", or connected continuously across the main bus-bars. The normal generator voltage is then maintained at a value sufficiently high to take care of the load requirements and to supply a small "trickle" charge to the battery, thus keeping it fully charged. Another standard practice is to employ duplicate storage batteries and two or more generators. Under this plan, one battery may be charged while the other battery is either supplying the load by itself, or is floated across another generator which is supplying the load. Both of these methods are considered more fully in the next article.

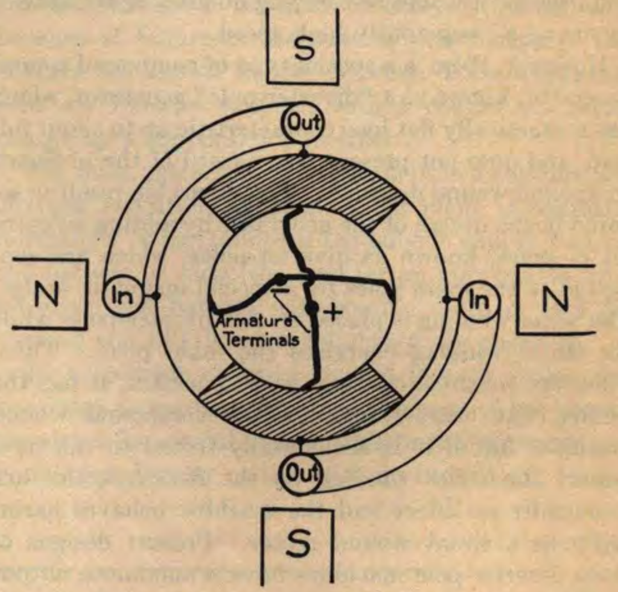


FIG. 79. FOUR-POLE GENERATOR

The requirements which generators must meet are somewhat different under the two plans of operation outlined above. Since under the latter plan the battery is charged independently of the load circuits, standard shunt wound generators are most suitable. This may be seen by referring to the load-voltage characteristic of the shunt wound machine, shown in Figure 78. As the slope of this curve indicates, the E.M.F. of a shunt wound generator rises as the load decreases. And in this case, as the battery voltage gradually rises under charge, the load current does tend to decrease. But this decrease causes an increase in the machine voltage, which tends to restore the current to its former value. The net result is that while the battery voltage and the generator voltage both increase somewhat as the charge progresses, the current supplied to the battery remains approximately constant. This is the kind of storage battery charging current which is ordinarily desirable.

On the other hand, in power plants where a single battery, continuously floated across the line, is used, it is desirable for the generator to have a load-voltage characteristic as nearly flat as possible. Because of its drooping characteristic, the shunt wound generator is therefore not suitable unless its voltage is constantly controlled by manual or automatic means.

The required flat characteristic could be obtained from an ordinary compound wound generator of proper design; but such a machine is not safe to use because in case of failure of the outside power, the floating battery would run the generator as a motor if the reverse current circuit-breaker failed to operate. In this situation, the reversed current in the series winding would cause the generator to operate like a series motor and because it would be carrying no load, it would tend to run at a dangerously high speed.

However, there is a special type of compound wound generator, known as a "diverter-pole" generator, which has a practically flat load characteristic up to about full load, and does not present this hazard of the ordinary compound wound machine. This desirable result is secured in the design of the generator by adding an extra set of poles, known as diverter-poles, which are connected to the main poles by a special magnetic bridge. The series winding is placed on the diverter-poles while the shunt winding energizes the main poles. Then, when the machine operates as a generator, it has the desired flat characteristic of the compound wound machine; but if it is accidentally forced to run as a motor, the series winding on the diverter-poles has practically no effect and the machine behaves harmlessly as a shunt wound motor. Present designs of these diverter-pole machines have a maximum output of 200 amperes, and their use is accordingly limited to offices where the total current requirements are relatively small. At larger offices where the continuous floating practice is followed, shunt wound generators with automatic voltage regulation are usually employed.

46. Typical Central Office Power Circuits

One application of standard shunt wound generators for supplying central office power requirements is illustrated by Figure 80. This schematic drawing represents a typical 24-volt power circuit for a repeater office, where the principal load requirements are for heating vacuum tube filaments. This is the type of power plant in which duplicate batteries and generators are employed, and the control is entirely manual. By means of three single-pole double-throw switches, either generator may be used to charge either battery while the other battery is carrying the load; or either machine may be connected to the load with one battery floated, while the other battery is being charged by the other machine. A filter, consisting of a "choke coil" and three "electrolytic condensers" connected in parallel to ground, is inserted in the output circuit to the filament supply panel. This smooths out small variations in current which might have an adverse affect on the sensitive apparatus supplied from this panel.

The circuit also includes an emergency storage battery cell which is connected to two single-pole double-throw switches in such a way that it can be connected in series with either of the main batteries. This cell is provided to take care of emergency conditions where the outside power supply fails and the generators are therefore inoperable. In such a case, the load must be carried by the batteries alone and if the failure persists for an appreciable time, the battery voltage will decrease below the required value. The emergency cell may then be cut into the circuit to build up the voltage.

A switch is provided for charging the emergency cell from the generator in series with either battery. However, since the emergency cell is not normally in use, it is continuously supplied with a small trickle charge furnished by a copper-oxide rectifier (see Article 53), which normally maintains it in a fully charged condition.

In order that the attendant may observe the operation of the circuit at all times, an ammeter is inserted in each generator lead, and another is inserted in the main output lead to the fuse panels. On the power panel also, is a voltmeter connected to a circular switch so arranged that it can read the voltage of either gen-

Typical Repeater Station Power Plant

erator, either main battery, or either main battery in series with the emergency cell. Another voltmeter is connected directly across the load circuit at all times, and this voltmeter is paralleled by a voltmeter relay which gives an automatic alarm in case the voltage exceeds specified limits in either direction. Alarm circuits are also provided to give warning in case of a blown fuse or operation of a circuit-breaker.

Figure 81 is a schematic of a typical power plant where only one storage battery is used, and the total load requirements are in the order of 500 amperes or more. As we have already seen, the battery is continuously floated in this type of plant, and the generator voltage must therefore be maintained at a constant value. This is accomplished automatically by means of motor driven field rheostats associated with the shunt wound generators, as indicated in the figure. A voltage relay (designated Gen. Reg. Voltage Relay in the drawing) is bridged directly across the main battery. As long as the battery voltage remains at its proper value, this relay is not operated; but if the battery voltage becomes too high, or too low, one or the other of the two relay contacts is closed. This causes either relay L or relay R to operate, as the case may be, and the operation of either of these relays causes the motor driven field rheostat to move in the direction which will restore the generator voltage to its normal value. To avoid the possibility of overloading the generator, an ammeter relay is inserted in series with the line. When the generator is fully loaded, a contact of this relay closes causing relay A to operate and open the voltage relay circuit. This prevents any further attempt on the part of the voltage relay to increase the generator output.

Like the circuit previously discussed, this circuit is equipped with emergency cells which are automatically cut into the circuit in series with the main battery by means of another voltage relay bridged across the line. As before, these cells come into operation only when the storage battery is required to carry the load alone because of failure of the outside power supply. Although not shown in the drawing, these emergency cells are likewise supplied with a continuous trickle charge by a copper-oxide rectifier.

The main battery is, of course, kept in a continuously charged condition as long as the plant is operating normally. When failure of the outside supply requires the battery to carry the load for an appreciable time, however, the battery will become more or less discharged and will therefore require special charging. In order to provide charging current in such a case, it is necessary to increase the output voltage of the generator above its normal value. But since the generator is connected directly to the load, an increase in its out

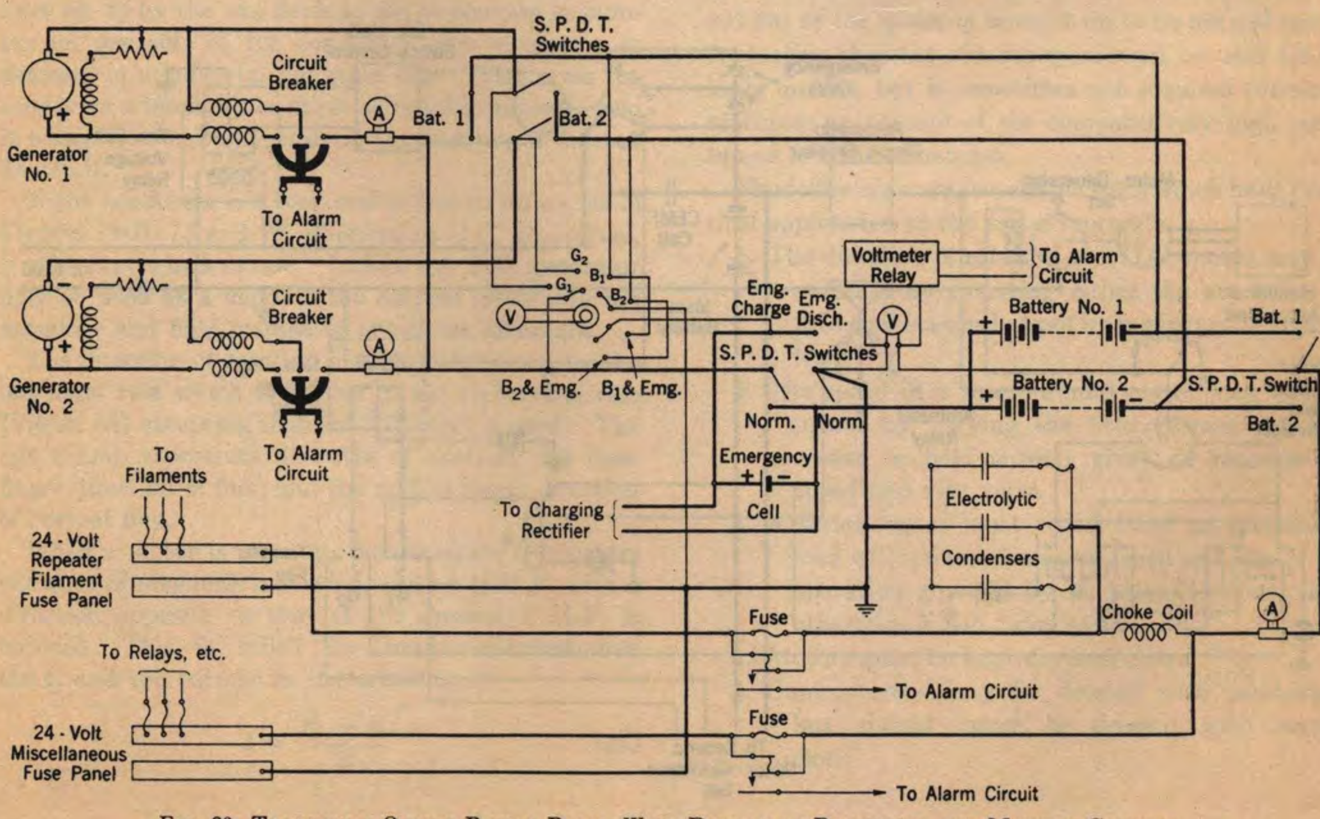
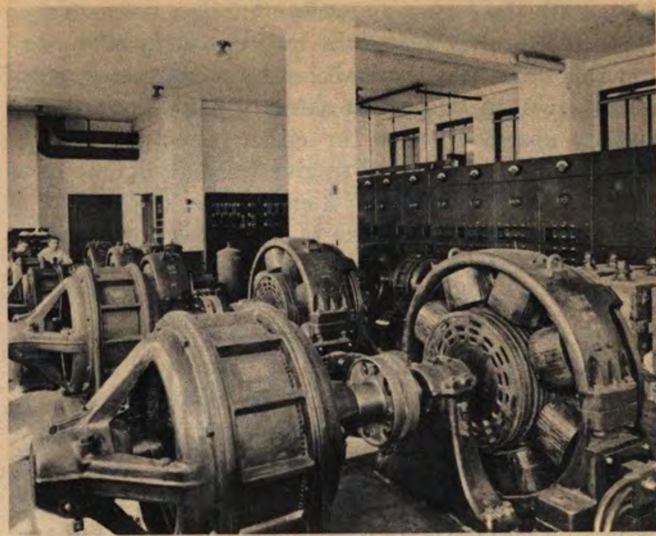


Fig. 80. Telephone Office Power Plant With Duplicate Batteries and Manual Control

put voltage would also increase the load voltage. To avoid this, the circuit includes a "counter-E.M.F. cell" which is automatically inserted in series with the load circuit when the output voltage of the generator is increased above its normal value.

The C.E.M.F. cell has the property, when current flows through it, of setting up a voltage opposing the voltage which is driving the current. The counter voltage is approximately 2 volts per cell and is developed without any appreciable loss of energy. Physically, the C.E.M.F. cell consists of two plates of pure nickel immersed in a caustic soda solution. As in the case of storage batteries, which are discussed in Article 50 following, the size of the nickel plates depends upon the amount of current which the cell is required to handle. The cells are usually mounted along with the storage battery cells.

Figure 81 shows only one generator but, as indicated at the bottom of the drawing, additional generators may be included. To insure continuity of operation, a practical power plant always includes at least two generators, and as many more may be added as are necessary to handle the maximum load. The second generator is equipped with a motor driven field rheostat like the first generator. When the first generator becomes fully loaded, this is put into operation by throwing the transfer key shown on the drawing.



Dial Central Office Power Plant

Additional generators, when required, are connected across the main leads to the battery, but are manually controlled.

This power plant may be arranged so that the motor-generator sets will start automatically upon restoration of the outside power supply after failure. By including additional relay circuits, this general type of plant may also be arranged so that needed generators will be

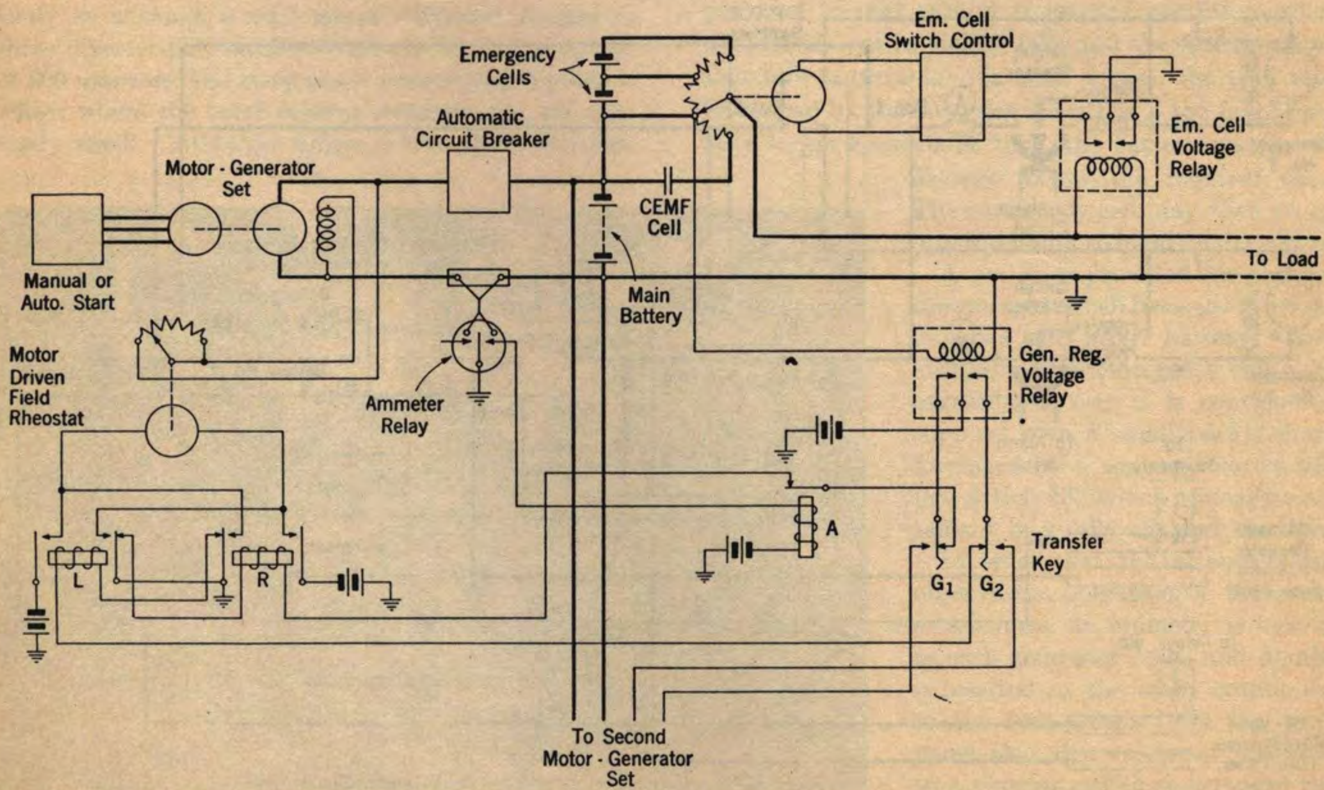
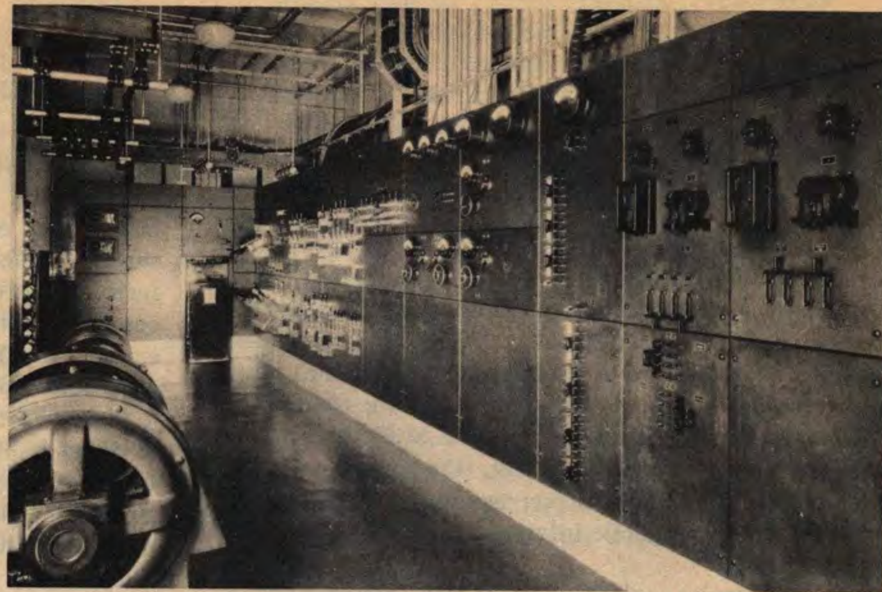


Fig. 81. Telephone Office Power Plant With Single Floating Battery and Automatic Control

automatically started and connected to the line as the load increases, and automatically disconnected and stopped as the load decreases.



POWER SWITCHBOARD OF LARGE TELEPHONE CENTRAL OFFICE

47. Direct-Current Motors

In the telephone plant, most of the electric motors used to drive charging generators, ringing machines and other minor units such as polishing machines and fans, operate on alternating current as supplied by the regular power distribution systems. However, direct-current motors are occasionally used. There are many classes of such motors (series, shunt, compound, multipolar) and each has different characteristics of power and speed. It is impracticable to discuss particular types in this book, and only a brief explanation of the fundamental working principle will be given.

When a conductor carrying an electric current is placed in a magnetic field at right angles to the lines of magnetic induction, there is a reaction between the circular field about the conductor and the field in which it has been placed. This reaction causes the lines set up by the two fields to aid or increase in number on one side of the conductor and to oppose or decrease in number on the other side. This gives the conductor a tendency to move across the magnetic field in a direction which depends on the direction of current flow in it.

If the conductor is a loop and is free to rotate, as in Figures 74-B, 75 and 76, illustrating D.C. generators, it will revolve as a motor. In fact **any D.C. generator may be used as a motor if the current flows into the armature and field instead of out of the armature.**

The direction of rotation may be determined by the **left-hand rule** which is similar to the right-hand rule (Figure 68) excepting that the left hand is used. The left thumb represents direction of motion; the forefinger direction of flux; and the middle finger direction of current flow.

When a motor is running, the armature conductors cut lines of magnetic induction, and an E.M.F. with a direction opposite to that of the applied E.M.F. is induced. This is called the **counter-electromotive force,** and the current in the armature is—

$$I = \frac{E_i - E_c}{R} \qquad (27)$$

where E_i is the impressed E.M.F., E_c is the counter-E.M.F., and R is the armature resistance.

Since there is low counter-E.M.F. until the motor has reached about its normal speed, it will draw a very large current at starting unless this is prevented by a **starting rheostat.** This is a variable resistance placed in series with the motor's armature which is gradually cut out as the motor is brought up to its normal speed. A starting rheostat of some type must be used for all large motors, but is sometimes not required for small machines on account of the comparatively high resistances of their armatures.

The following are a few simple rules which have practical application to the use of motors:

1. The direction of rotation of a D.C. motor may be reversed by reversing either the armature or field connections but **not** by reversing the supply leads.
2. The speed of a shunt wound motor may be adjusted by varying the field current. A **decrease** in field current gives an **increase** in speed and vice versa.
3. A series motor must either have an increasing load with increase in speed, such as a fan, or its operation guarded by an attended controller; otherwise it will "run away".
4. Motors must be kept dry and clean.
5. Commutators may be dressed with sandpaper but should never be dressed with emery cloth.

CHAPTER VII

OTHER SOURCES OF DIRECT ELECTROMOTIVE FORCE

48. Types of D.C. Energy Sources

For an electrical circuit to become energized, some source of electromotive force must be connected to it either by direct connection or through inductive relations. In the case of a direct current, the circuit must be energized by the actual connection of the conductors to the terminals of the source of E.M.F.; but in the case of an alternating current, the circuit may be energized either by such connection or by inductive effects due to magnetic interlinkages or capacity relations.

If any device maintains an E.M.F. and sustains a current of electricity in a circuit, energy is supplied to the circuit. But the law for conservation of energy states that energy cannot be created or destroyed. Any source of E.M.F. may then be defined as a device for supplying electrical energy by converting it from some other form. The battery converts chemical energy into electrical energy, the generator converts mechanical energy into electrical energy, and the thermocouple changes heat to electrical energy. A rectifier is in one sense a source of direct E.M.F. but it converts alternating-current energy into direct-current energy, changing it from one electrical form to the other rather than from some other form to the electrical.

Figure 82 shows the circuit conventions for sources of direct E.M.F. that are common in electrical work. In the operation of the telephone plant we are interested principally in the battery, the generator and the rectifier. The theory of the generator is covered in Chapter VI. We shall at this time consider the various types of batteries and the general battery requirements of telephone service, and make some mention of rectifiers and other interesting though perhaps less important sources of direct E.M.F.

49. Primary Batteries

Chemical batteries are divided into two classes,

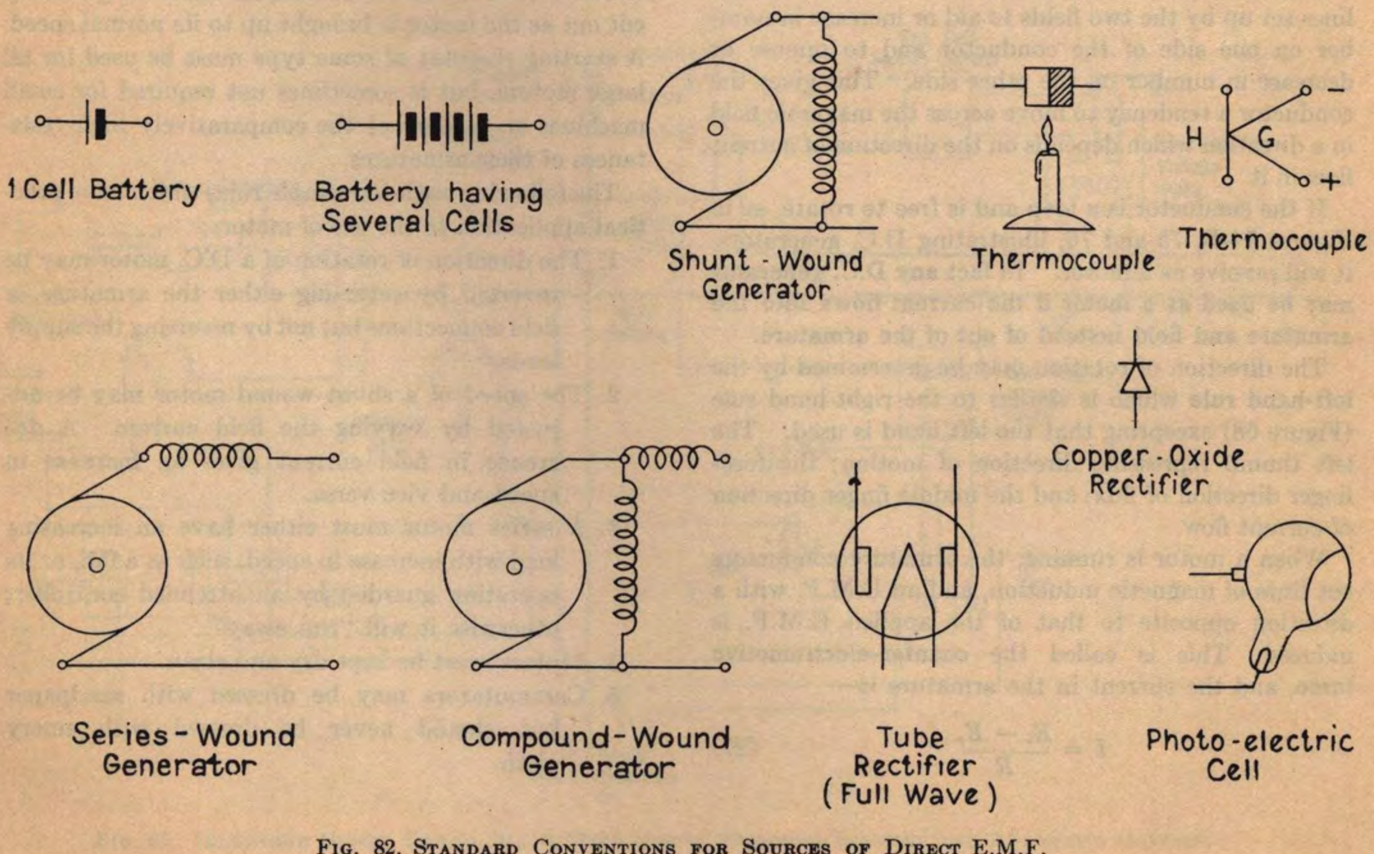


Fig. 82. Standard Conventions for Sources of Direct E.M.F.

primary and secondary. A primary battery is one that generates an E.M.F. by virtue of certain chemicals coming in contact with submerged metals or other substances which constitute the positive and negative terminals. A secondary battery stores electrical energy but does not directly generate an E.M.F. unless a current is first passed through the battery in a direction opposite to that in which it will flow when supplying energy to an external circuit.

The unit of a battery is the cell, consisting of a single couple of submerged positive and negative poles or plates. As illustrated in Figure 83, cells may be connected in parallel or in series, depending upon the value of the E.M.F. desired and the value of current to be sustained. If they are connected in series, the E.M.F.'s are added, making the total E.M.F. of the battery the sum of the E.M.F.'s of the individual cells. If they are connected in parallel, the E.M.F. of the battery is that of a single cell, but the current supplied to the circuit is divided between the several cells.

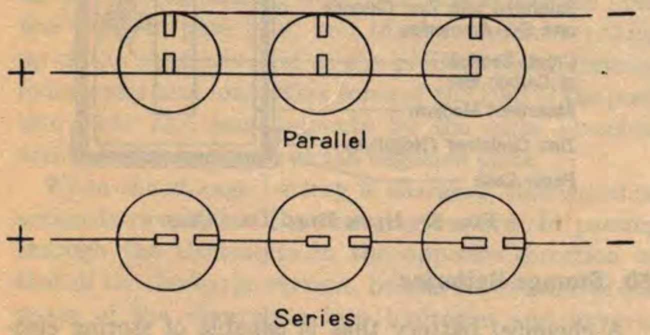


FIG. 83. CELLS IN SERIES AND PARALLEL

A battery may consist of groups of cells connected in parallel and these in turn connected in series, or vice versa. Figure 84 shows two methods of connecting six cells where the E.M.F. desired is that of only three cells and a single string is not sufficient to sustain the current required. Theoretically, the two methods give the same results, but in the case of dry cells, method "B" has some advantage from the standpoint of deterioration of the battery due to the uneven electrical characteristics of the individual cells.

The various types of primary batteries may be divided into two general groups called "wet" cells and "dry" cells. There are numerous kinds of wet batteries, the more common of which are as follows:

a. The **Daniell cell** which consists of a zinc plate in a solution of zinc sulphate and a copper plate in a solution of copper sulphate (blue vitriol). In earlier forms of this cell the two solutions were separated by a porous cup which contained one solution and was submerged in the other. One of the later and more commonly used types, called the "gravity cell", dispenses with the porous cup on account of the two solutions having different specific gravities. The copper sulphate, being heavier than the zinc sulphate, is placed in the bottom of the jar with a copper plate, and the zinc sulphate in the top of the jar with a zinc plate (or "crow's foot").

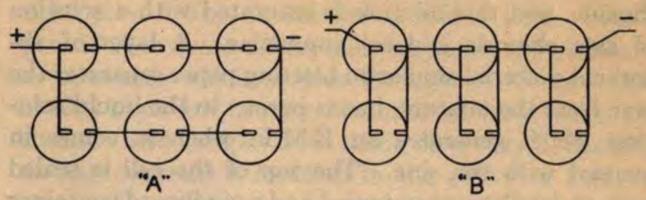


FIGURE 84

b. The **sal ammoniac cell,** which consists of a zinc negative rod or plate and a carbon positive plate in a solution of sal ammoniac (ammonium chloride).

c. The **Lalande cell,** which consists of a zinc negative cylinder in a solution of caustic soda and a perforated sheet iron cylinder in which is embodied black copper oxide. The two are separated by cylindrical insulators.

About the only **wet cell** used to any considerable extent in the telephone plant is a special form of Lalande cell, known as the "air-cell battery". It consists essentially of a negative zinc plate and a porous carbon rod immersed in a caustic soda solution. The chemical action of the cell is such that hydrogen is liberated at the positive carbon electrode, which combines with oxygen from the air breathed in by the porous carbon to form water. It is necessary to keep the top of the carbon electrode clean and to locate the battery in a well ventilated cabinet or room, as each cell must absorb some 45 cubic inches of air per hour for proper operation at full load. The air cell has a nominal voltage of 1.25 and a capacity of 600 ampere-hours with a maximum current drain of .66 ampere. Its principal use in practice is for supplying current for operators' transmitters in magneto offices and for operating certain types of interrupters.

The **dry cell** is a special form of chemical battery, so constructed that the chemicals used in its action are sealed. It is most convenient for shipping and general use. There are two important and general classes of service for which dry cells are designed. They may be constructed for heavy current duty, such as for flashlights, at a sacrifice of life; or they may be intended for connection to a high resistance and a correspondingly low current output. In the latter case, the batteries do not require replacement for a much longer period, particularly if the service is required only at intervals and the battery is allowed to "rest" on open circuit.

The Blue Bell dry cell is representative of the "long-life" type. Its construction is illustrated in Figure 85. The negative terminal consists of a zinc container in which is centered a bar of carbon as a positive terminal. The carbon is surrounded by a porous medium consisting principally of ground carbon and manganese dioxide, and this mixture is saturated with a solution of zinc chloride and sal ammoniac. A layer of absorbent material similar to blotting paper separates the zinc from the mixture, but is porous to the liquid solution which generates an E.M.F. when it comes in contact with the zinc. The top of the cell is sealed with an insulating compound and a cardboard container acts as an insulating cover for the zinc. Spring connectors, which are securely fastened to the zinc and carbon electrodes, form suitable terminals. When new, this dry cell gives a voltage of about 1½ volts, which decreases with age, and has an internal resistance of .2 to .3 ohm, which increases with age. For average use, its capacity may be roughly estimated at 20 to 30 ampere-hours but this will vary considerably depending upon conditions. For example, the capacity when connected to a high resistance circuit may be several times the capacity when connected to a low resistance circuit. Intermittent use is also an important factor.

Dry cells are commonly used in the telephone plant for service where connections to central office storage batteries are not feasible, or cannot be used because the storage battery is grounded. In addition to transmitter batteries for magneto subscribers' stations, such uses may include battery supply for telegraph sounders on subscribers' premises, Wheatstone bridge testing battery for toll testboards, plate and grid batteries for vacuum tube circuits, and testing batteries for portable testing sets. For many of the above purposes, the current requirements are comparatively low while the voltage needed may be considerable. To meet these conditions, it is the usual practice to employ small or miniature cells which are connected in series and assembled in sealed "battery blocks" in the manufacturing process. Standard battery blocks of this type are available having nominal maximum voltages of 3, 4½, 22½, and 45 volts. The higher voltage blocks usually have intermediate taps giving various voltage values below the maximum.

As noted above, the life of a dry cell or a battery block depends upon the kind of service in which it is used. Standard maintenance provisions call for replacement if batteries fail to meet the following tests:

a. When the current drain is negligible, as in vacuum tube grid batteries, a voltage of at least 1.33 volts per cell should be obtained when the battery is tested with a voltmeter having a resistance of 1000 ohms per volt of full scale deflection.

b. When the current drain is steady and fairly large, as for vacuum tube plate supply, the voltage should be at least 1.13 volts per cell when tested under load with a voltmeter having a resistance of 60 to 100 ohms per volt.

c. When the current drain is fairly large but variable, the voltage should measure at least .9 volt per cell when tested 10 seconds after the application of an artificial load of 5 ohms per cell for large cells and 10 ohms per cell for small cells, with a voltmeter having a resistance as in (b) above.

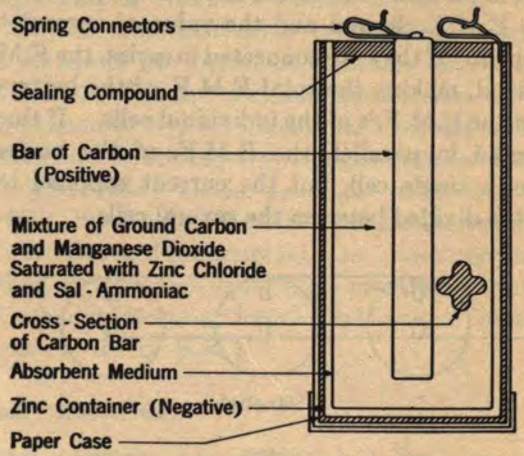


Fig. 85. Blue Bell Dry Cell

50. Storage Batteries

A chemical battery that is capable of storing electrical energy delivered to it from some other source, and delivering this energy to an electrical circuit at some later time, is called a storage battery. Other names commonly used are "accumulator" and "secondary battery". The two general types of storage batteries are the lead-acid and the Edison (iron-potassium hydroxide-nickel). On account of its low internal resistance and more constant terminal voltage, the lead-acid type more nearly meets the exacting requirements of the telephone central office. (These requirements are discussed in Article 51 following.)

When the lead-acid cell is in a fully charged condition, the active constituents are a positive plate of lead peroxide (PbO_2) and a negative plate of spongy lead (Pb) in a dilute solution of sulphuric acid (H_2SO_4 + H_2O). When the battery is discharging, the current, passing from the positive to the negative plate through the external circuit, must return from the negative to the positive plate through the dilute acid (electrolyte). In doing so, it breaks the electrolyte into its component parts resulting in first, the spongy lead of the negative plate combining with the positively charged component (SO_4) of the electrolyte, forming lead sulphate ($PbSO_4$) and losing its negative charge; second,

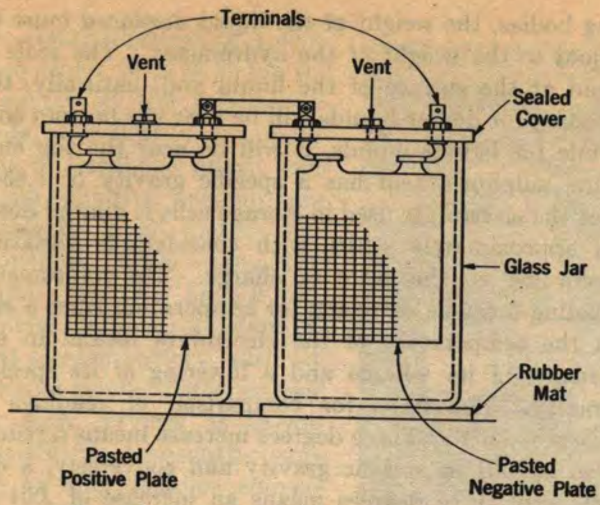


FIG. 86. ENCLOSED TYPE STORAGE BATTERY

the oxygen of the lead peroxide of the positive plate combining with a part of the hydrogen liberated from the electrolyte, forming water, and converting the positive plate to pure lead; and third, a similar breaking up of the sulphuric acid at the positive plate, forming more water and converting some of the lead of the positive plate into lead sulphate by the same chemical action that takes place at the negative plate.

When the storage battery is charging, this chemical action is reversed. The charging current, in passing through the electrolyte in the opposite direction to that of the discharge current, breaks down some of the water of the electrolyte into hydrogen and oxygen. The oxygen travels against the current to the positive plate where it combines with the lead sulphate of that plate to form lead peroxide. The sulphate (SO_4) released by this action combines with hydrogen to form sulphuric acid. At the same time, hydrogen, travelling with the current to the negative plate, combines with the lead sulphate of that plate to form sulphuric acid. This leaves pure metallic or sponge lead on the negative plate, and the two plates and the electrolyte are thus gradually restored to their original charged condition.

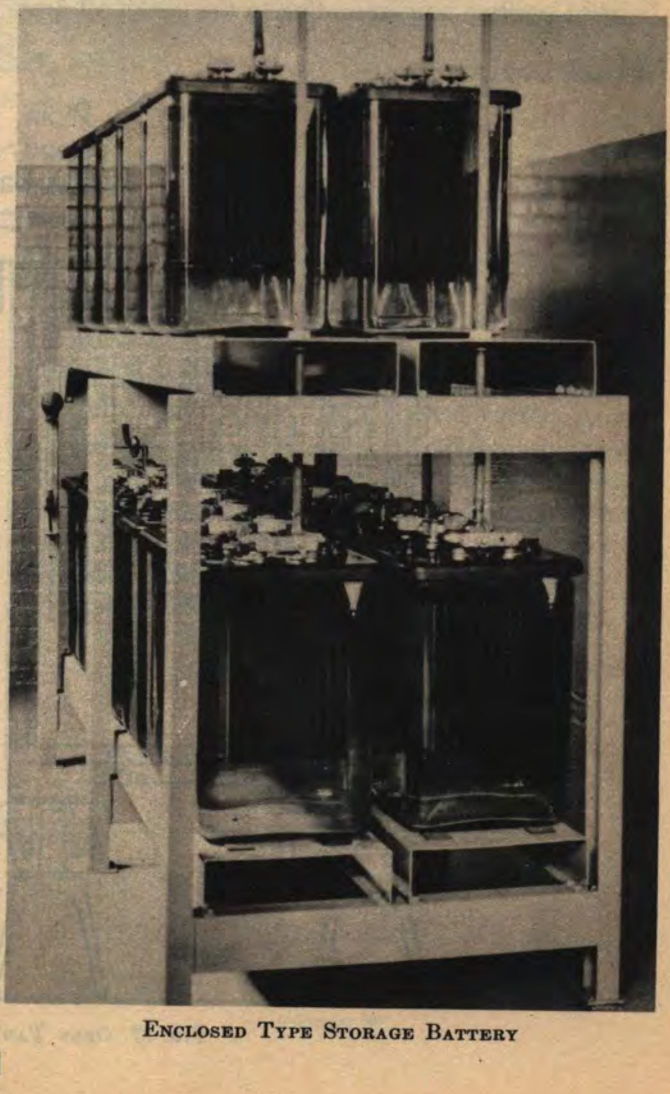
The following chemical equation may be used to explain the action of discharge when reading from left to right and the action of charge when reading from right to left:

$$PbO_2 + Pb + 2H_2SO_4 \rightleftarrows 2PbSO_4 + 2H_2O \quad (28)$$

Storage batteries are built in a wide variety of sizes to meet the various load requirements. The capacity of a cell naturally depends on the total area of plate surface which is exposed to the electrolyte. The smallest cell consists of a single pair of plates having a total area of only a few square inches, while the largest cells may have more than 100 plates, each with an area of more than three square feet.

In modern central office practice, the plates of the smaller storage cells (up to a maximum ampere-hour capacity at an 8-hour discharge rate in the order of 1000) are mounted in glass jars. These are enclosed at the top with a cover of glass or acid-resisting compound which is sealed to the glass jar to form an acid-tight joint. The covers are provided with vents to permit the release of gas and the plates are connected to terminals which project through the cover. Cells of this type are usually mounted on iron racks, and since no acid escapes from them, they may be installed at any convenient location in the power or terminal room. Larger cells are assembled in lead-lined wooden tanks. These cells are open at the top and must therefore be installed in a separate well-ventilated room.

The general arrangement of the two main types of cells is illustrated in Figures 86 and 87. As these drawings indicate, different methods are employed for impressing the active material (lead peroxide and sponge lead) on the plates. In all cases, the basic structure of the plate is a cast antimony-lead grid.



ENCLOSED TYPE STORAGE BATTERY

The method of applying the active material is purely a mechanical problem and has no effect on the electrical behavior of the cells. The plates are kept separated by wooden or rubber separators, and are supported in the manner indicated in the drawings.

In the practical operation of a storage battery, we must be able to determine the state of charge or discharge at any time. It is not convenient to do this by chemical analysis, but in the foregoing explanation of the cycle of charge and discharge, there are two changes taking place that may be easily determined. One is the change in the electrical charge held by the plates, resulting in a change in the E.M.F. of each cell. The other is the increase on discharge, and the decrease on charge, of the amount of water contained in the electrolyte, which increase or decrease, as the case may be, changes the specific gravity of the electrolyte. This latter condition gives the better index to the cell's operation and is the one ordinarily used.

Figure 88 shows a hydrometer designed for determining the specific gravity of the electrolyte. A weight at the bottom makes the hydrometer float in an upright position and, according to the law of all floating bodies, the weight of the liquid displaced must be equal to the weight of the hydrometer. The scale is read at the surface of the liquid and, naturally, the reading for denser liquids will be near the bottom end, while for lighter liquids, it will be near the top end. Pure sulphuric acid has a specific gravity of 1.8342 but the electrolyte used in storage cells is diluted down to approximately 1.210, with considerable variation according to the state of charge. The hydrometer reading must be corrected for temperature since a rise in the temperature of the electrolyte means an expansion of its volume and a lowering of its specific gravity. The basis for comparison of readings is taken at 70°F. Three degrees increase means a reduction of .001 in specific gravity and conversely, a decrease of three degrees means an increase of .001 in specific gravity.

51. Power Plant Requirements in Telephone Offices

The telephone central office power plant must be not only absolutely reliable at all times but must meet other exacting requirements. Modern practice has led to

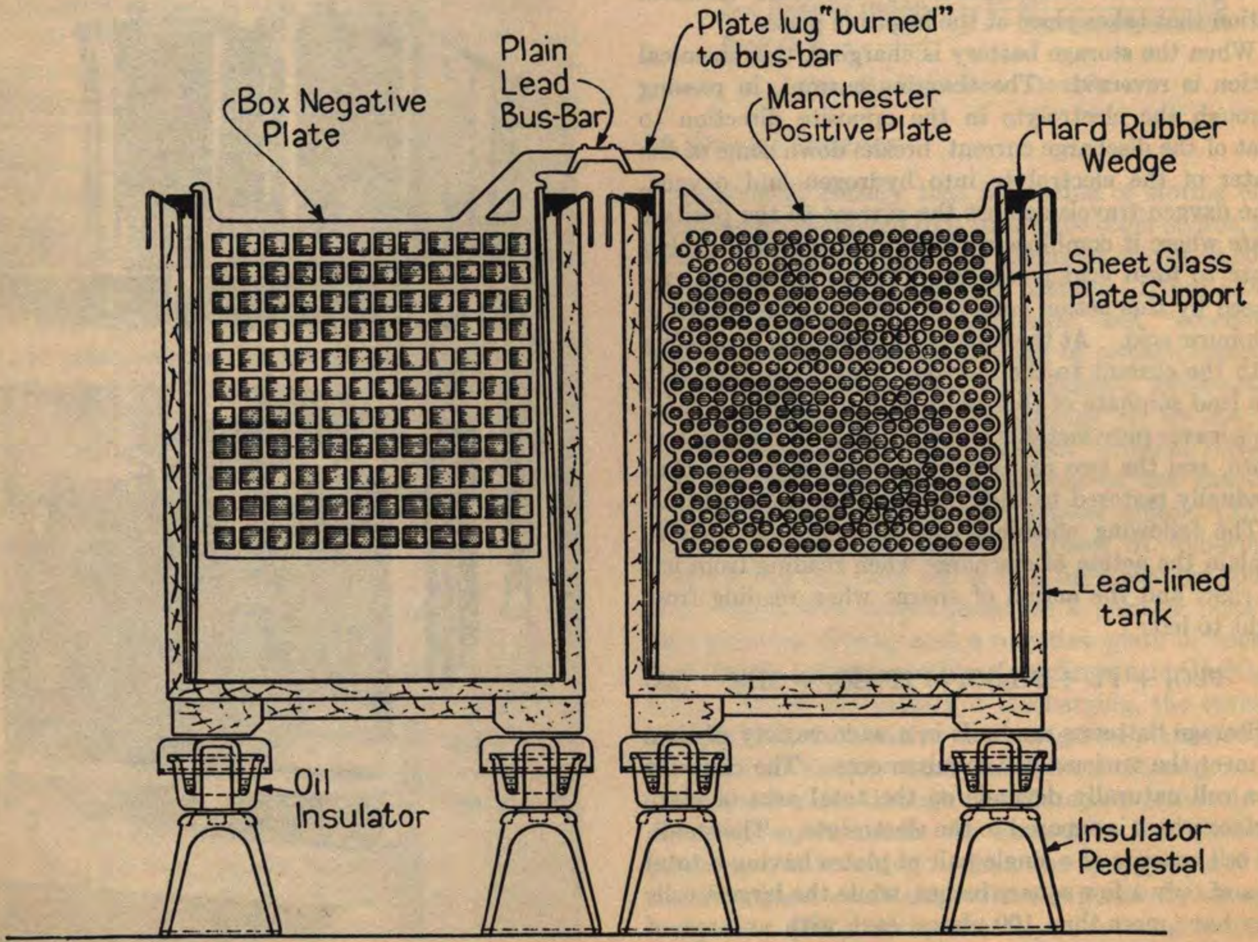
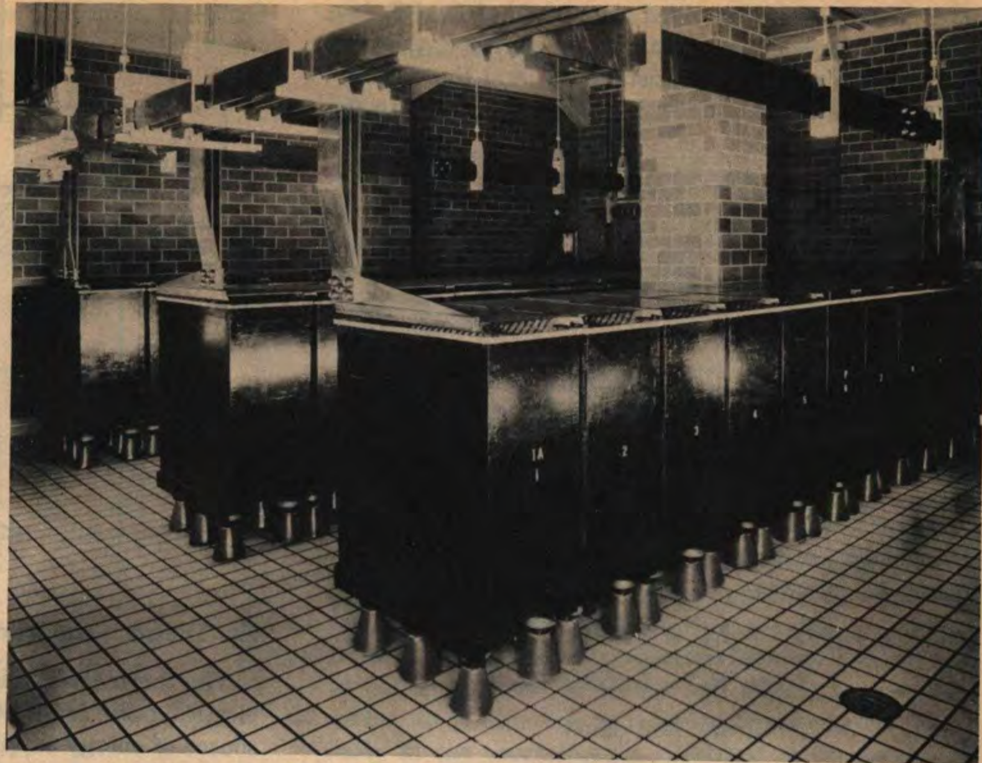


Fig. 87. Open Tank Type Storage Battery

the standardization of a common source of E.M.F. for the majority of the talking circuits, as well as for the operation of telephone relays, telegraph sounders, vacuum tube filaments, small motors, and numerous other apparatus units. We thus have a very general use of the standard 24-volt storage battery, with additional smaller batteries used for such services as 48-volt subscriber's transmitter supply on long distance connections, 120-volt supply, both positive and negative, for telegraph repeater operation, and other voltages, for telephone repeater operation. The common battery results in a number of plant economies, but, on the other hand, imposes certain exacting electrical requirements. Probably the most essential of these requirements is **low internal resistance.**



Open Tank Type Storage Battery

In our study of simple electrical circuits, we have considered a single source of E.M.F. for each individual circuit. But we have learned that any number of resistances may be connected in parallel, as shown by Figure 89, and that the current in any single resistance is independent of that in any other resistance provided all resistance branches are connected directly to the terminals of the battery as indicated. This follows naturally from the application of Ohm's Law to a single resistance branch, since the E.M.F. impressed on any single branch is the E.M.F. of the source and, theoretically, is independent of current flowing through other branches. This assumes, however, that the battery is a perfect source of E.M.F. without internal resistance.

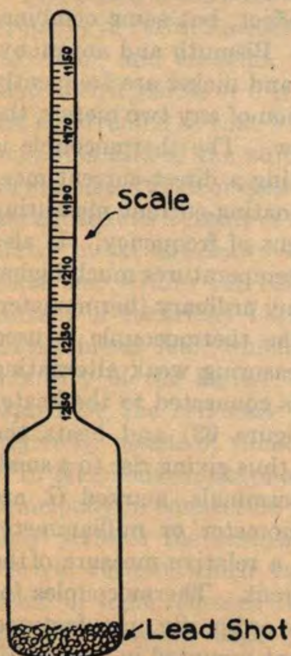


Fig. 88. Hydrometer

Figure 90 represents the central office storage battery connected to bus-bars at the fuse panel. The central office circuits are cabled to this fuse panel and receive their battery supply through taps to the small panel busses. Thus hundreds of circuits of varying resistance are connected in parallel to a common battery, and we have in practice a circuit arrangement identical to that shown in theory by Figure 89, excepting that as indicated in Figure 90, fuses for protection against excessive currents due to short-circuit or overload are used, and the positive terminal of the battery is connected to ground. This ground connection stabilizes the potential of all circuits in the central office by short-circuiting their capacities to ground. It also simplifies the central office wiring and affords circuit protection, but it cannot in any way affect the total current supplied by the battery or the current in any individual circuit that may be connected to the bus-bars.

Returning to Figure 89, in which the current in any

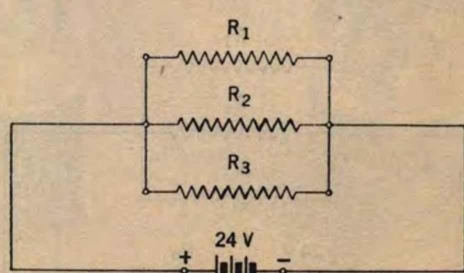


Figure 89

one resistance branch was seen to be independent of that in any other (provided the source of E.M.F. is a perfect one), let us assume, on the contrary, that the battery has an internal resistance R_0 and that the circuit is actually that shown by Figure 91. Due to the

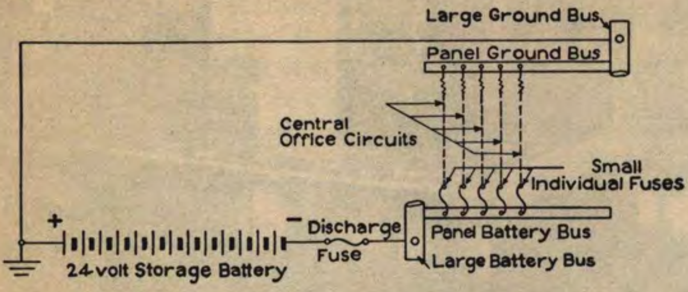


Fig. 90. Storage Battery Load Connections

resistance R_0, the current in one branch is no longer independent of that in other branches. Let us assign values as follows:

$$R_0 = 2 \text{ ohms}$$

$$R_1 = 5 \text{ ohms}$$

$$R_2 = 4 \text{ ohms}$$

$$R_3 = 3 \text{ ohms}$$

$$V = 24 \text{ volts}$$

If we solve this network, we shall find that the current through R_1 is 1.87 amperes. If we should suddenly open resistances R_2 and R_3, however, it would immediately change to 3.43 amperes. Applying the same principle to Figure 90, **unless the central office source of E.M.F. has negligible resistance, including both the internal resistance of the battery and that of the supply leads from the battery to the bus-bars where individual circuit leads are connected, there will be ever-changing current values in the individual circuits. This will result in noise and crosstalk in all talking circuits and unreliable operation of various other telephone apparatus.** From this it follows that common battery operation for any number of circuits may be substituted for local or individual batteries only when the common source of E.M.F. has negligible internal resistance.

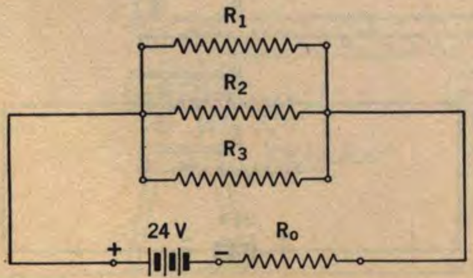


Figure 91

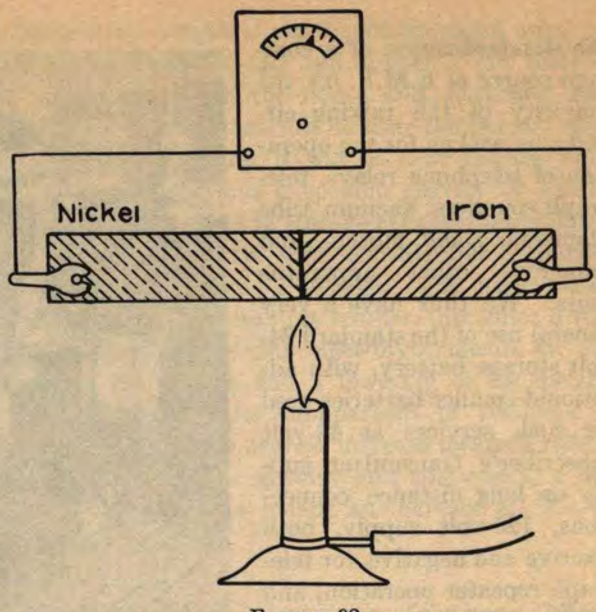


Figure 92

52. Thermo- and Photo-Electric Effects

The thermocouple is probably the simplest source of electromotive force but it has little practical use as a source of energy supply. Nevertheless, because it is a direct means of establishing an electrical current from heat, its principle of operation is important. It consists of two dissimilar metals in contact, with heat applied to their junction. The different characteristics of the two metals result in a difference of potential between them when heated and if a galvanometer is connected as shown in Figure 92, current can be detected. Almost every combination of dissimilar metals will give the thermocouple effect, but some combinations are better than others. Bismuth and antimony, iron and constantin, copper and nickel are frequently used. For a single combination of any two metals, the E.M.F. generated is very low. The thermocouple is used extensively for converting a direct-current measuring instrument to an alternating-current measuring instrument that is independent of frequency. It also permits the measurement of temperatures much higher than can be measured with any ordinary thermometer.

In the telephone plant, the thermocouple is used primarily as a device for measuring weak alternating currents. The A.C. circuit is connected to the heater terminals (marked H in Figure 93) and heats the junction of dissimilar metals, thus giving rise to a small direct current. The D.C. terminals, marked G, are connected to a D.C. galvanometer or milliammeter which gives by its deflection a relative measure of the input A.C. or "heating" current. Thermocouples for laboratory and general use are ordinarily manufactured with the glass-enclosed element mounted in a cylindrical metal container equipped with either ordinary terminals or a vacuum tube base for use in standard sockets. For general use, there are three standard resistance values for the heater element, viz. 5 ohms, 40 ohms, and 600 ohms.

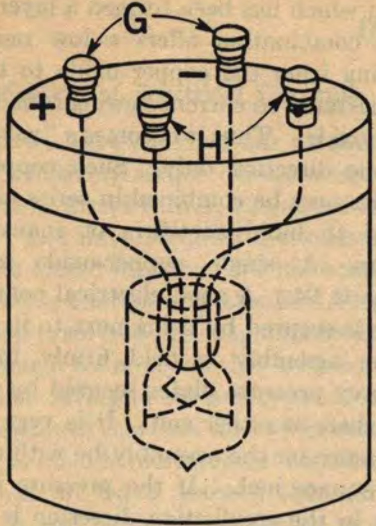


FIG. 93. STANDARD THERMO-COUPLE

Like the thermocouple, the photo-electric cell is not important as a source of electrical energy. Its ability to convert the energy of light into electrical energy is extremely useful, however, in connection with telephotography, television, and various systems for controlling relays or other electro-magnetic devices by means of light.

The action of the photo-electric cell depends upon the property of certain metals, notably sodium, potassium, rubidium, and caesium, to emit electrons when irradiated with visible or ultra-violet light. Practical cells are constructed so that a beam of light may fall upon a very thin film of the pure metal which is contained in an evacuated glass or quartz bulb. The electrons which are emitted as a result of the light falling on the photoactive metal are drawn away to a positively charged plate also within the tube, thus establishing a small current of electricity between this plate and the irradiated metal film, which varies in strength with the intensity of the light. The very weak current generated by the cell may be amplified to any desired degree by means of vacuum tube amplifiers.

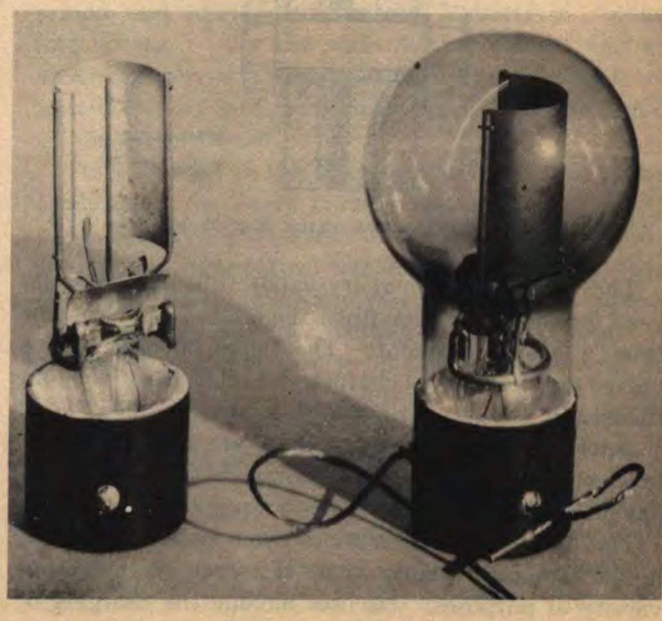
In telephone work, photo-electric cells are now used principally in connection with telephotography. Here they perform the essential function of producing an electric current which varies with the intensity of a beam of light reflected from the picture being transmitted. A much larger field of use for these cells is in sound motion picture projection where they translate the "sound-track" on the edge of the film into sound. They are also employed extensively for operating relays under light control in various industrial applications.

53. Rectifiers

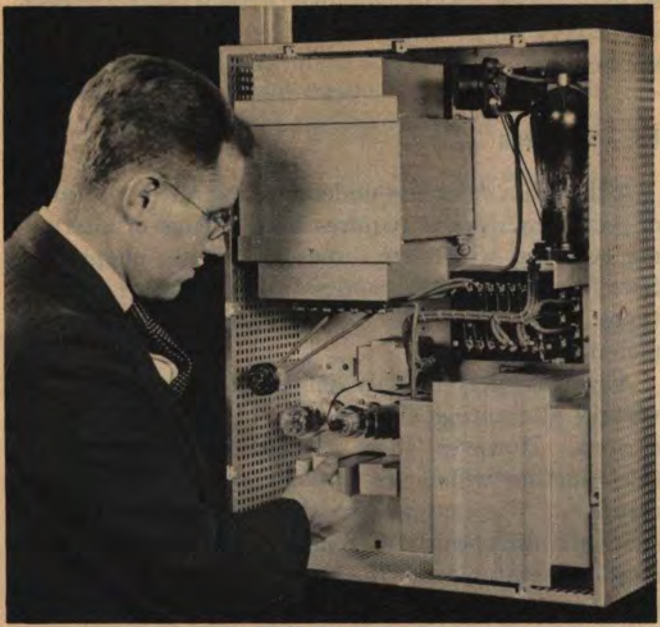
Although a complete understanding of the operation of rectifying devices requires a knowledge of alternating currents, which are discussed in later chapters, a few of the essential characteristics of the more common types of rectifiers may be mentioned here for completeness.

A rectifier is commonly defined as a device for converting alternating electric current to direct electric current. However, there are devices which perform this function which are not normally referred to as rectifiers. For example, an alternating-current motor driving a direct-current generator is referred to as a motor-generator set. If the motor and generator of such a set are combined in one housing with a single rotor, the machine is referred to as a rotary converter. In either case, however, electrical energy is first converted to mechanical energy and this in turn is converted to a different type of electrical energy. Similarly in the thermocouple described in the preceding section, A.C. electrical energy is converted to heat and this in turn generates D.C. energy. Accordingly, rectifiers may be somewhat more precisely defined as devices for converting A.C. energy to D.C. energy directly or without an intervening step.

All rectifying devices depend for their operation upon the characteristic of permitting electric current to flow through them freely in one direction only. They include a variety of vacuum and gas filled tubes such



PHOTOELECTRIC CELL



Mercury-Vapor Tube Rectifier

as the older mercury arc tube, the newer mercury-vapor tube, and the Tungar tubes, as well as nearly all other types of vacuum tubes when properly connected. In addition, there is the copper-oxide type of rectifier.

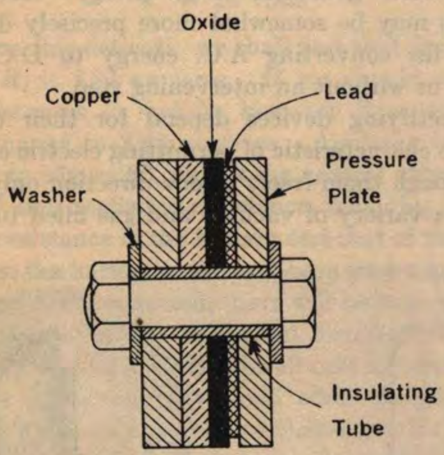
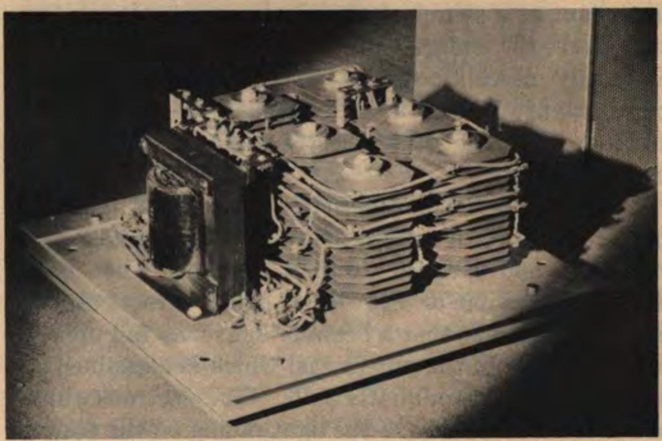


Fig. 94. Copper-oxide Rectifier

The Tungar and mercury-vapor tube rectifiers depend for their operation upon the emission of electrons from a heated filament. The basic phenomena involved are the same as characterize all types of vacuum tubes and these are discussed more fully in a later chapter. Both types of rectifiers are used quite extensively for charging small storage batteries and similar purposes.

The copper-oxide rectifier is even more widely used in the telephone plant, where it serves an increasing variety of purposes. Its uses include the charging of small storage batteries and emergency cells, the furnishing of a direct source of power for small repeater installations, etc. In addition, as we shall see in a later chapter, it has extensive applications in many carrier circuits.

The copper-oxide rectifier element consists of a copper disc upon which has been formed a layer of copper oxide. This combination offers a low resistance to current flowing from the copper oxide to the copper but a high resistance to current flowing from the copper to the copper oxide. Thus it becomes a "valve" to pass current in one direction only. Such copper copper-oxide elements may be combined in series and parallel arrangements to build rectifiers of many different characteristics. A single copper-oxide element is shown in Figure 94. A good electrical contact to the copper oxide is secured by using next to it a soft lead washer. The assembly is held firmly together by means of heavy pressure plates secured by a bolt and nut with washers at either end. It is very important that the pressure on the assembly be within 500–2000 pounds per square inch. If the pressure is low, the current flow in the conducting direction is not maximum and if the pressure is high, the resistance in the reverse direction breaks down, permanently injuring the element.



Copper-Oxide Rectifier

Copper-oxide rectifiers suffer a rapid decrease in the amount of current they will pass in the first three or four months of use. This deterioration is called aging. The decrease in initial output does not usually amount to more than 25%, however, and this is anticipated in initial design. Plating the lead washers with tin and graphiting the oxide surfaces reduces the aging somewhat. The amount of aging a rectifier undergoes is also a function of temperature. For this reason current drains in excess of the rated output should be avoided as excess current will raise the temperature, increase the aging and may destroy the rectifying action of the discs.

CHAPTER VIII

INDUCTANCE AND CAPACITY

54. Classification of Electrical Currents

Thus far we have confined our attention largely to circuits of relatively simple characteristics. We have had a source of direct E.M.F. connected to one or more resistances, and have assumed a resultant steady current in each closed branch. We have noted, however, the alternating character of the E.M.F. generated by a closed loop revolving in a magnetic field; but we have not attempted to analyze the behavior of such an E.M.F. when acting in various types of circuits.

It is desirable at this time that we broaden our studies somewhat to include more general conditions and while nothing that we have learned thus far will be invalidated, it will be necessary for us to study certain additional properties of electrical circuits and their effect on the current set up in them by impressed E.M.F.'s.

Broadly speaking all electrical currents may be classified into five groups as follows:

a. The current that results from a constant direct source of E.M.F. connected to a resistance network (i.e., the condition assumed in the earlier chapters for the calculation of direct-current networks through the application of Ohm's and Kirchoff's Laws).

b. The current immediately after opening or closing a circuit, varying its resistance, or in some way interrupting the steady direct current for a short period of time during which the current values readjust themselves before again becoming fixed or steady.

c. Current where the source of E.M.F. is an alternating one, having the simplest, most common and most convenient wave form, viz. the sine wave.

d. Current where the source of E.M.F. is an alternating one having a definite wave shape other than the sine wave.

e. Alternating current immediately after opening or closing the circuit, or immediately after effecting some other change in circuit conditions.

We can further classify the currents in the foregoing: *a*, *c*, and *d* are those relating to **steady state** currents, while *b* and *e* refer to temporary currents, sometimes called **transients.** In practice we are mostly interested in steady state currents in so far as the actual determination of current values is concerned, but under certain conditions the effects of transients are important. Certainly, in a telephone connection, we are concerned with any "clicks" or "scratches" that may be heard in a telephone receiver due to the opening or closing of circuits which are electrically connected to the telephone system. For example, when sending telegraph signals over a telegraph circuit superposed on a telephone circuit, there should be no appreciable "telegraph thump" in the telephone circuit. The successful operation of both telephone and telegraph circuits introduces certain important considerations having to do with changes in current values.

In fact, we deal with all five of the circuit conditions mentioned above in the telephone plant. Let us consider a long distance line wire not only composited for telegraph service but having a carrier current telegraph channel superposed as well. The resulting current in the wire can best be studied by scrutinizing the behavior of its separate components. When analyzed, the current due to the composited telegraph connection alone is an illustration of two of the classifications, namely *a* and *b*. At the instant of "make" or "break" of the key, conditions are as described by *b*. When the key is closed, i.e., when signals are not being sent, conditions are as described by *a*. For the carrier channel, we likewise have condition *c* for a part of the closed key period and condition *e* for the instants of "make" and "break". For the main talking circuit, we have an application of *d* when a vowel sound is being transmitted, and an application of *e* when a consonant sound is being transmitted.

Thus we find in the telephone plant no scarcity of applications for every current classification. It happens, however, that some of these are by no means simple and for practical telephone work we may limit our study to a thorough analysis of steady state currents only, and to concepts, rather than calculations, of transients in either direct- or alternating-current circuits.

55. Changes in Direct-Current Values

We may analyze classification *b* (changes in direct-current values) since this will lead us to certain of the new circuit properties that we wish to examine. In Figure 95, with the switch open we have a circuit with infinite resistance and zero current; with the switch closed we have, by Ohm's Law, a current—

$$I = \frac{E}{R} = \frac{10}{5} = 2 \text{ amperes.}$$

In spite of the apparent promptness with which electricity responds to the operation of any controlling device, we cannot conceive of the current changing from zero to two amperes without going through the range of every intermediate value between zero and two amperes; neither can we conceive of the current building up in the circuit in zero time to the value given by the application of Ohm's Law. If such were the case, the current would have every value from 0 to 2 amperes at the instant of closing the circuit. Reverting to our water analogy with the circulating mechanism in Figure 2, when a valve is shut we know there is no flow of water in a long pipe line and when the valve is opened we know that, due to the inertia of the water, a definite time is required for the flow to become a maximum. A current in an electrical circuit cannot be established instantaneously any more than the water flow can be established instantaneously.

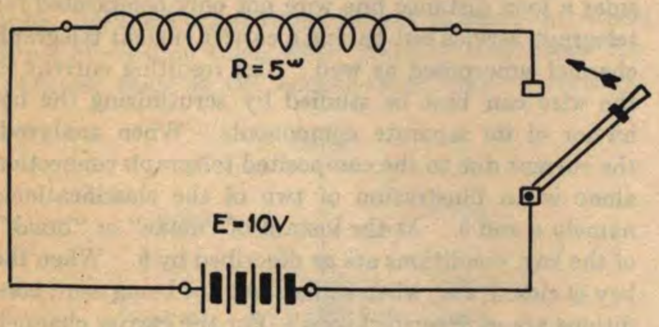


Fig. 95. Simple Inductive Circuit

Again, if in Figure 95 we suddenly open the switch in a dark room while there is a current of two amperes in the circuit, we shall observe a spark at the contacts of the switch. Though the electrical current is reducing in value, it continues to flow for an instant after the switch points are no longer in contact, forcing itself through the air, and thereby forming an "electrical arc" which gives the illumination.

We thus have two conditions where the current in a brief interval of time assumes all intermediate values between two amperes and zero, and we may compare these with other less abrupt changes in a circuit, such as those due to a varying transmitter resistance. It may be said that an electrical circuit "reacts" to such current changes. But this reaction cannot be explained by our previous understanding of either resistance or E.M.F. The circuit has other properties which are latent when the current is a steady unidirectional one but which are immediately brought into play when the current attempts to change its value. There are two such additional properties, namely, **"inductance"** and **"capacity". Inductance tends to give the circuit something that is analogous to inertia in a mechanical device, and capacity something analogous to elasticity.**

56. Inductance

When an E.M.F. is connected to a circuit, the conditions are somewhat analogous to those obtaining when a locomotive starts a train. The locomotive exerts considerable force which, in the circuit, corresponds to the impressed E.M.F. A part of this force is used in overcoming resisting forces such as the friction of the moving wheels, the grade of the track, and others that apply to the train as a definite resistance to its motion at all times. The second part of the force is used in setting the train in motion, i.e., accelerating the heavy inert body. As soon as the train is accelerated to full speed, the entire force applied is available for overcoming the resistance alone. Likewise in the electrical circuit, for any given E.M.F., the current does not instantaneously establish itself to that value which represents the effect of the full voltage overcoming the resistance.

We have learned that there is a magnetic field about every current-carrying conductor, and when a conductor is wound into a coil or is in the presence of iron, the magnetic field is intensified. The magnetic field cannot be established instantaneously any more than the train can be instantly changed from its state of rest to that of full speed. What actually happens in the case of the electrical circuit is that the E.M.F. endeavors to start a current; the current in turn must establish a magnetic field; this field reacts upon the circuit in a manner similar to that in which the counter-E.M.F. generated by a motor opposes the applied voltage, and for an instant a part of the E.M.F. that is connected to the circuit must be used in overcoming these reactions. The current, therefore, increases gradually and as it does so, the magnetic field becomes more nearly established and the reaction becomes less pronounced, until finally the entire E.M.F. is applied to overcoming the resistance of the circuit alone, thereby sustaining the established current at a value determined by Ohm's Law.

This may be more clearly understood by referring to the circuit shown in Figure 96 and following the change in current that is taking place immediately after the switch has been closed. When the switch S is closed, the E.M.F. E endeavors to establish a current in the circuit equal in value to E/R, or two amperes. But the current, as has been stated, must go through every intermediate value from zero to two amperes. By directing our attention to only one turn of the coil, for example, T_1, we can imagine the current building up and in consequence establishing lines of magnetic induction around this single turn which will, however, cut every other turn of the coil. This action will set up in the other turns an induced E.M.F. tending to establish a current in the opposite direction in much the same way as we learned a back E.M.F. was set up in the electric motor. And as in the case of the motor, the two currents are in one and the same circuit and the induced current is opposed to the current established by virtue of the battery E.M.F.

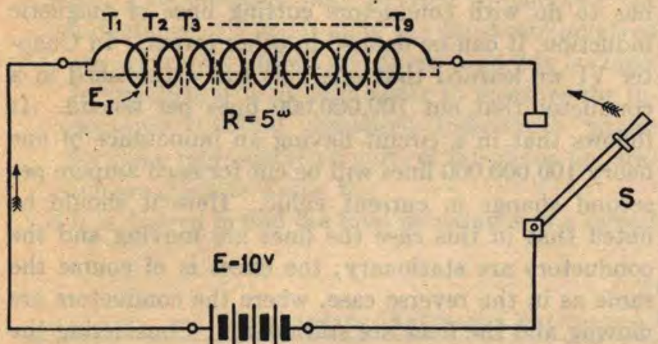


FIGURE 96

Figure 97 represents graphically the current in this circuit. With the switch open, the current is zero. When it is closed (or when sufficiently near the contacts for the E.M.F. to break down the insulation of the narrow separation of air, since the current starts to flow before actual contact is made), the 10-volt battery will attempt to establish a current of two amperes in accordance with Ohm's Law. But the current cannot be completely established until after an interval of time represented by t_2; and at the start, it cannot be increasing at a rate greater than that which would induce a back E.M.F. of 10 volts, because if it did so, the induced E.M.F. would be equal to the applied E.M.F. and since they oppose each other, there would be no current whatsoever. As would be expected, however, the maximum rate of increase of the current occurs at the instant the switch is closed.

Now let us consider the conditions at some intermediate time between the closing of the switch and t_2. If, from the value represented by point P, the current increased at a rate that continued without changing, the line PM would represent the trend of current values that would follow. But with the current increasing at this rate, the lines of magnetic induction are cutting other turns of wire and inducing an E.M.F. which we might represent in Figure 96 as a second battery E_1, and which must be of the value necessary to establish a current equal to two amperes minus the current which has been already established at the point P. This follows from the earlier explanation regarding the directional property of an induced E.M.F. If the battery voltage E acted alone, the current value would be E/R or two amperes. Since the actual current flowing is less than two amperes, the difference between the actual current and two amperes may be regarded as due to a current flowing in a direction opposite to that of the two amperes set up by the battery. This current is established by the induced E.M.F. and we may designate it as an **induced current** to distinguish it from the two-ampere current which the supply voltage tends to set up. The actual current in the circuit at any instant, then, is the numerical difference between the two-ampere battery current and the induced current.

If we now assume for the sake of reasoning that the induced voltage E_1 remains unchanged, the resulting induced current will oppose the battery current, and the net amount of current flow will remain at the value P. We know, however, that the current which will eventually flow is two amperes, and furthermore, if the current becomes constant at a value P, no lines of magnetic induction are in motion; hence, there is no induced voltage and consequently no induced current. But with no induced current, the battery will set up two amperes; therefore our assumption that the induced

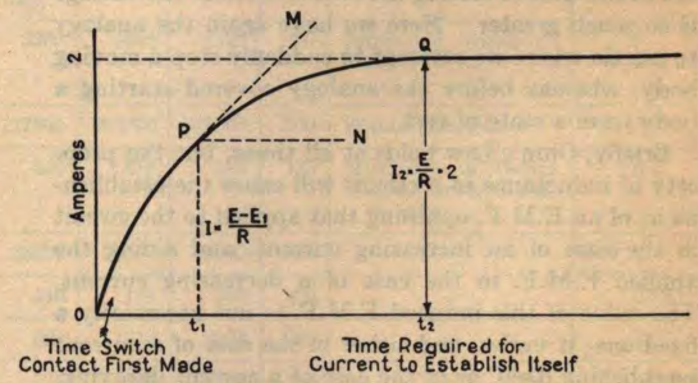


FIG. 97. BUILD-UP OF CURRENT IN INDUCTIVE CIRCUIT

voltage E_1 remains constant, keeping the current down to a value such as that represented by the line PN is false. On the other hand, it is clear that the induced voltage E_1 cannot become zero until the current becomes two amperes, though it does continue to decrease in value, since we know that a current is always accompanied by a magnetic field which must change if the current changes, and the result of such a change is an induced voltage. From this we conclude that there must be a compromise trend for the curve of current as it establishes itself, somewhere between the two extremes. This compromise is that shown by the curve PQ which is tangent to but bending away from PM. The current is neither maintaining the same rate of change as it approaches the value fixed by Ohm's Law nor does it cease entirely its increase in value before it reaches two amperes. This is true because although the induced E.M.F. that would stop the change in current is gradually becoming less in value, the IR drop is becoming greater, and the sum of these two must always equal the impressed voltage in accordance with Kirchoff's second law. Thus we see from the curve in Figure 97 the "choking" effect of an inductively wound coil to increases in current value.

The case of a decreasing current value, and the E.M.F. induced at the time of opening a circuit, is of course another application of the same theory, but the effects are different in their practical aspects. Because this E.M.F. is induced as a result of a decreasing current instead of an increasing one, it aids rather than opposes the existing E.M.F. Moreover, the current change is a very rapid one because the opening of the switch tends to change the resistance of the circuit from a definite value to infinity with great suddenness. As a result, the induced E.M.F. may become much greater than the applied E.M.F. besides being additive to it, whereas in the closed circuit it can never be greater than the applied E.M.F. This total E.M.F. of the opening circuit tends to force an arc across the switch contacts, which is much more evident than the arc at the time of closing the switch because the voltage is so much greater. Here we have again the analogy to inertia where we attempt to suddenly stop a moving body, whereas before the analogy covered starting a body from a state of rest.

Briefly, Ohm's Law holds at all times, but the property of inductance in a circuit will cause the establishment of an E.M.F. opposing that applied to the circuit in the case of an increasing current, and aiding the applied E.M.F. in the case of a decreasing current. The value of this induced E.M.F. is not necessarily a fixed one; it varies, and either in the case of a current establishing itself, or in the case of a current decaying, eventually becomes zero. The magnitude or influence of the induced E.M.F. as a reactive effect is determined by two factors:

a. The first is a property of the circuit having to do with the number of inductive turns, whether or not each coil has a magnetic core and if magnetic, the permeability of the iron, etc.

b. The second is the rate of change of current. This employs the property of the circuit as a tool or facility for creating the induced E.M.F.

The **property of the circuit** which we have called **inductance** is represented by the symbol L and is measured in a unit called the **henry.** The rate of change in current, though a varying quantity, would naturally be measured in amperes per second or I/t. (In this case I/t represents the rate of change of current which if maintained constant, would permit the current to rise from a value 0 to a value I in a time interval t.)

The unit value of the henry is defined as the inductance of a circuit that will cause an induced E.M.F. of one volt to be set up in the circuit when the current is changing at the rate of one ampere per second. From this we may write—

$$E_1 = \frac{LI}{t} \tag{29}$$

where E_1 is the chosen symbol for induced E.M.F. and L represents inductance in henrys.

Since L depends upon a property of the circuit which has to do with conductors cutting lines of magnetic induction, it can be defined in other terms. In Chapter VI we learned that one volt was established in a conductor that cut 100,000,000 lines per second. It follows that in a circuit having an inductance of one henry 100,000,000 lines will be cut for each ampere per second change in current value. Here it should be noted that in this case the lines are moving and the conductors are stationary; the effect is of course the same as in the reverse case, where the conductors are moving and the lines are stationary. Considering the inductance of any given coil, the lines threading through the coil, as they build up or decrease, will cut each turn, or we may write—

$$E_1 = \frac{\phi N}{100{,}000{,}000 \times t} \tag{30}$$

where ϕ is the flux through the coil, N is the number of turns and t is the number of seconds required for the flux to cut the turns. But, from Equation (29), $E_1 = \frac{LI}{t}$; therefore, $\frac{LI}{t}$ can be substituted for E_1 in Equation (30) and we have—

$$\frac{LI}{t} = \frac{\phi N}{100{,}000{,}000 \times t} \tag{31}$$

or, with the t cancelled on both sides of the equation—

$$LI = \frac{\phi N}{100{,}000{,}000} \tag{32}$$

But in Equation (14) we learned that $\phi = M/R$ where M is magnetomotive force and R is reluctance. Also, in Equation (16) we found that for a solenoid, $M = 1.26NI$. Therefore—

$$\phi = \frac{M}{R} = \frac{1.26NI}{R} \tag{33}$$

which may be substituted in Equation (32) giving—

$$LI = \frac{1.26NI}{R} \times \frac{N}{100{,}000{,}000}$$

$$L = \frac{1.26N^2}{R \times 100{,}000{,}000} \tag{34}$$

The reluctance for any entire coil is determined by the dimensions of the coil and the permeability of the iron core. We may substitute in Equation (34) an expression for reluctance that may be derived from Equations (14), (16) and (17); namely, $R = \frac{l}{\mu A}$ where μ is the permeability, l is the length of the core in centimeters and A is the area of the core in square centimeters. Thus, we have finally—

$$L = \frac{1.26N^2\mu A}{100{,}000{,}000 \times l}\,. \tag{35}$$

Note: This equation may be used to calculate the inductance of a coil if all of the constants involved are accurately known and there is no flux leakage. In practice, it is usually easier to measure the inductance. Actual measured inductance values for some representative units of telephone apparatus are given in Table V.

If it is desired **to find the total inductance of a circuit having several coils in series, the inductances should be added in the same way that resistances in series are added. Similarly, parallel inductances are calculated by the same formulas as are parallel resistances.** For example, see Equation (4) and substitute L, L_1, and L_2 for R, R_1 and R_2, respectively, etc.

This property of a circuit which creates an E.M.F. from a change of current values when the reaction effects are wholly within the circuit itself is called self-inductance to distinguish it from the relation permitting electromagnetic induction between coils or conductors of separate circuits. This latter property of the two circuits taken jointly is called mutual inductance. It is discussed in a later chapter.

57. Capacity

There remains that property of the circuit that we have called "capacity", which gives it something

TABLE V

APPROXIMATE INDUCTANCE VALUES FOR WINDINGS OF VARIOUS ELECTRICAL APPARATUS

APPARATUS		NO. WINDINGS	DC RESISTANCE	IMPEDANCE					INDUCTANCE (HENRYS) (SEE NOTE 2)
Name	Code No.			AC Resistance (See Note 1)	Reactance (See Note 1)	Impedance	Freq.	Connections	
Relays, AC Type	172-B	2	Inductive 540 Non-inductive 2000 Comb. 406	3260	5650	6520	900	Inductive winding only	1.0
Relays, AC Type	196-A	2	1600; 1600	117500	203000	235000	900	Windings in Series	36.0
Relays, AC Type	218-B	2	117; 117	450	678	810	135	Windings in Series Armature held stationary	.8
Relays, AC Type	J-1	1	1090	38000	39400	55000	900	—	7.0
Receivers	144	—	83	140	164	250	800	—	.0325
Receivers	525	—	276	620	1910	2000	800	—	.38
Receivers	528	—	56	106	237	260	800	—	.0475
Receivers	557-B	—	30	46	110	120	800	—	.022
Retardation Coils	5-U	2	(1-2) 500	—	—	—	16	Single Winding	3.0 to 4.0
			(2-3) 500				16	Series Aiding	12.0 to 15.0
Retardation Coils	5-AA 77-A	2	74 ea.	—	270	—	16	Two windings as connected for 1 wire	2.7
Retardation Coils	12-A	1	165	—	5024	—	800	—	1.0
Retardation Coils	44-D	2	83 ea.	2480	39565	39580	900	Windings in Series	7.0
Retardation Coils	47-B	1	150	160	1700	1710	900	—	.3
Retardation Coils	57-B	2	175 ea.	1620	22610	22700	900	Windings in Parallel	4.0
Retardation Coils	82-H	1 (tapped)	(1-8) 25.2	47	1640	1640	1800	Entire winding	.145
Retardation Çoils	182-B	1 (tapped)	(1-8) (30.0)	56	1640	1640	1800	Entire winding	.145

Note: (1) Impedance and impedance components are discussed in Chapter XVI. A.C. Resistance or the resistance component of impedance is often widely different from the resistance to direct-current flow.

(2) Inductance values vary greatly depending upon conditions under which apparatus is operated, age of iron, degree of saturation, etc. This table gives only representative and approximate values.

analogous to elasticity. While a storage battery stores electricity as another form of energy, in a smaller way a **condenser** stores electricity in its natural state.

As a container, a condenser is hardly analogous to a vessel that may be filled with water, but more nearly to a closed tank filled with compressed air. The quantity of air, since air is elastic, depends upon the pressure as well as the size or capacity of the tank. If a condenser is connected to a direct source of E.M.F. through a switch as shown by Figure 98, and the switch is suddenly closed, there will be a rush of current in the circuit. This will charge the condenser to a potential equal to that of the battery, but the current will decrease rapidly and become zero when the condenser is fully charged.

Fig. 98. Simple Capacitive Circuit

The insulated conductors of every circuit have to a greater or less degree this property of capacity. A certain quantity of electricity, representing a certain quantity of energy, is accordingly delivered to a circuit before the actual transfer or transmission of energy from a sending device to a receiving device takes place. The capacity of two parallel open wires, or a pair of cable conductors of any considerable length, is appreciable in practice.

The quantity of electricity stored by a condenser depends upon the condenser's capacity and the electromotive force impressed across its terminals. The following equation expresses the exact relation:

$$Q = EC \tag{36}$$

where Q is the quantity of electricity in coulombs, E is the impressed E.M.F. in volts, and C is the capacity of the condenser in **farads.** The farad is a very large unit and is seldom used in practice. The microfarad (from "micro", meaning one one-millionth) is the practical unit more commonly used; and with C expressed in these units, Equation (36) becomes—

$$Q = \frac{EC}{1{,}000{,}000} \tag{37}$$

Figure 99 illustrates a condenser in its simplest form together with one convention used for a condenser connected to a battery. Two wires are connected to two parallel metal plates having a definite separation as shown. This is called an "air condenser" because air is the "dielectric" medium between the plates. The capacity of such a condenser is directly proportional to the area of the plates, and inversely proportional to their separation. At the instant a battery is connected to its terminals, there is a rush of electricity which charges the plates to the potential of the battery, but as the plates become fully charged, the current in the connecting conductors becomes zero. Were we to insert a sensitive high resistance galvanometer in series with the battery, we would observe an instantaneous "kick" of the needle when the connection is made, but the needle would return and come to rest at zero. If the capacity of the condenser were increased, the kick would become more noticeable. If now the battery were disconnected and the condenser short-circuited through the galvanometer, there would be a kick of the needle in the opposite direction. This would result from the quantity of electricity, which had been stored in the condenser, establishing an instantaneous current in the opposite direction and discharging the condenser through the winding of the galvanometer.

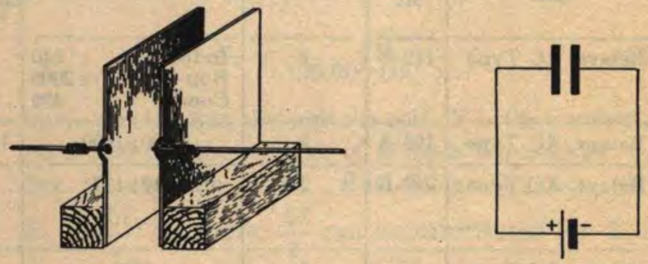


Fig. 99. Elementary Condenser and Convention

In addition to the size of its plates and their separation, the capacity of a condenser depends upon the insulating medium between the plates. For example, if mica is inserted between the plates of an air condenser its capacity is increased about five times. The insulators, in addition to being classified in the order of their insulating properties as given in Table III, may be classified in the order of their "dielectric powers", or "specific inductive capacities", i.e., their ability to increase the capacity of a condenser over that of an air condenser. Such a classification is given in brief in Table VI.

The equation for the capacity value of a two-plate condenser is—

$$C = K\frac{A}{d} \tag{38}$$

where C is capacity in microfarads, K is the constant taken from Table VI, A is inside area of one plate in square centimeters, and d is separation of plates in centimeters. There are similar equations for calculating the capacity per unit length of parallel open wire conductors or cable conductors. These may be found in various electrical handbooks, but for telephone and telegraph work, tables giving measured values, which vary for each class of open wire or cable pairs, are preferable and are usually available.

TABLE VI

DIELECTRIC POWER OF VARIOUS INSULATING MATERIALS

Values are only approximate and are given for value of K in Equation (38) rather than compared to air as unity.

SUBSTANCE	K IN EQUATION (38)
Glass—Very dense flint...........	$.9 \div 10^6$ approx.
Mica.............................	$(.3 \text{ to } .7) \div 10^6$ approx.
Glass, ordinary..................	$.3 \div 10^6$ approx.
Shellac..........................	$.3 \div 10^6$ approx.
Gutta-Percha.....................	$(.2 \text{ to } .4) \div 10^6$ approx.
India Rubber.....................	$.2 \div 10^6$ approx.
Paraffin paper...................	$(.2 \text{ to } .3) \div 10^6$ approx.
Air (at atmospheric pressure).....	$.0885 \div 10^6$ standard

An inspection of Equation (38) will show that if two identical condensers are connected in parallel as shown by Figure 100, the effect is that of doubling the plate area of a single condenser, and therefore doubling the capacity. On the other hand, if two identical condensers are connected in series as shown by Figure 101, the middle or common plates have a neutral potential and the effect is that of doubling the thickness of the dielectric of a single condenser, which cuts the capacity in half. It follows that capacities in parallel and series act inversely to resistances or inductances in parallel and series. This may be stated in a single rule covering all conditions—

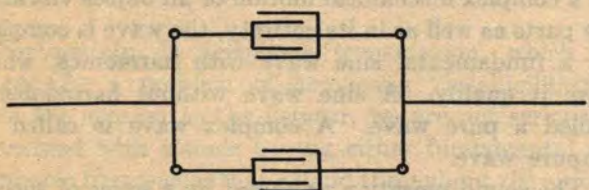


FIGURE 100

Capacities in parallel should be added to find the total capacity in the same way that resistances in series should be added to find the total resistance; and the reciprocal of the sum of the reciprocals must be taken to find the total capacity of capacities in series in the same way that the reciprocal of the sum of the reciprocals must be taken to find the total resistance of resistances in parallel.

This rule may be expressed by two simple equations: For several parallel capacities—

$$C = C_1 + C_2 + C_3 \text{ etc.} \tag{39}$$

For several series capacities—

$$\frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} \text{ etc.} \tag{40}$$

Or for only **two** series capacities, a third equation may be written as follows:

$$C = \frac{C_1 \times C_2}{C_1 + C_2}. \tag{41}$$

Note: Equation (39) may be compared with Equation (4) and Equation (41) may be compared with Equation (8).

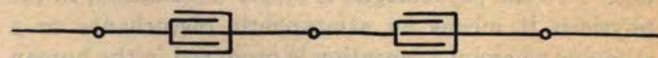


FIGURE 101

58. Effects of Inductance and Capacity in Direct-Current Circuits

The circuit reactions coming from the presence of inductance and capacity offer their most common applications in alternating-current circuits where we deal with their effects singly or jointly as "reactance", a quantity measured in ohms just as resistance is measured in ohms. Direct-current applications in telephone and telegraph work are nevertheless common. Figure 102 shows one way to apply the property of capacity to neutralize the detrimental effects of the self-inductance that is always present where there is a relay winding. Here the key contacts are bridged with a condenser which prevents excessive arcing when the circuit is opened because the sustained current is charging the condenser instead of forcing an arc. In practice the condenser usually has a non-inductive resistance in series, its purpose being to avoid oscillatory effects which are discussed in a later chapter.

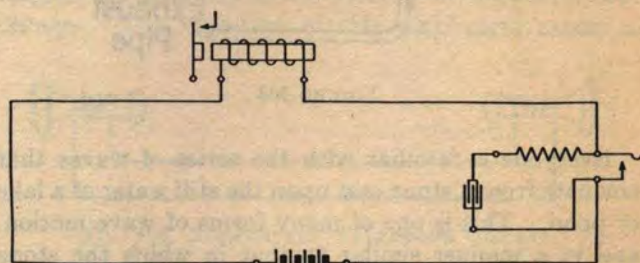


FIG. 102. CONDENSER AS SPARK-KILLER

CHAPTER IX

PRINCIPLE OF THE TELEPHONE

59. Sound

The telephone accomplishes the electrical transmission of speech by employing the mechanical energy of the speaker's voice to produce electrical energy having similar characteristics, and in turn converting this electrical energy into sound waves having similar characteristics at the listener's station. To understand its principle of operation we may well consider the nature of "sound".

Sound in the scientific sense has two distinct meanings. To the psychologist it means a sensation, to the physicist it means an atmospheric disturbance or a stimulus whereby a sensation is produced in the human ear. In other words, it is a form of wave motion produced by some vibrating body such as a bell, tuning fork, the human vocal cords, or similar objects capable of producing rapid to-and-fro or vibratory motion.

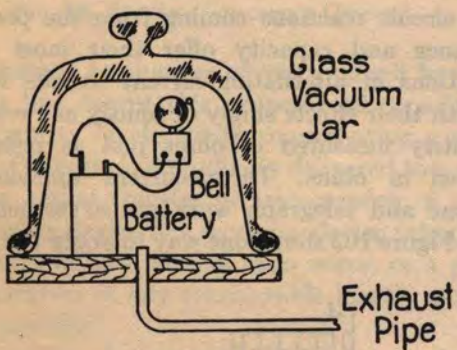


FIGURE 103

Everyone is familiar with the series of waves that emanate from a stone cast upon the still water of a lake or pond. This is one of many forms of wave motion, and in a manner similar to that in which the stone coming in contact with the water establishes radiating rings formed by circular wave crests alternating with wave troughs, there emanates from a source of sound alternate condensations and rarefactions of the air. Instead of being rings on a single plane or surface, however, they are a series of concentric spheres expanding at a definite rate of travel. This rate of travel (or the velocity of the sound wave) is approximately 1,075 feet per second but varies to some extent with altitude and atmospheric conditions. The velocity of sound is very low as compared with the velocity of light, heat or wireless waves, which are also a form of wave motion. We thus see a flash of lightning before we hear a clap of thunder or see the smoke dispelled from the muzzle of a gun before we hear the gun's report.

Unlike light, heat or electrical wave transmission, sound is an atmospheric disturbance. If as shown in Figure 103, a vibrating bell is placed under an inverted glass bowl resting on a brass plate that has an outlet through its center to which an exhaust pump is connected, it may be heard almost as distinctly as though there were no glass container. But if the air is exhausted until there is a vacuum about the bell, no sound can be heard; yet the bell may be seen vibrating as clearly as before the glass container was exhausted. We thus learn that there must be an atmospheric medium for the transmission of sound.

If the sound's source is a vibrating mechanism in simple form, such as a simple to-and-fro motion of the prong of a tuning fork, and is sustained without decay for a definite interval of time, the wave motion is said to be "simple harmonic". (A simple harmonic wave may be represented by the sine curve already discussed in Article 43.) On the other hand, if the source consists of a complex mechanical motion or an object vibrating by parts as well as in its entirety, the wave is **complex,** or a fundamental sine wave with **harmonics,** which give it **quality.** A sine wave without harmonics is called a pure wave. A complex wave is called an impure wave.

The sound sensation produced by a series of successive waves identical in form is called a **tone,** and if each wave is complex, it is a tone having **timbre or quality,** but if simple or a sine wave, it is a **pure tone.**

A vibrating mechanism giving a pure tone is said to establish a tone of low pitch if it is vibrating slowly, but if vibrating rapidly, it establishes a tone of high pitch. The lowest pitch which is audible to the average ear lies somewhere in the octave between 16 and 32 vibrations per second. The ear does not respond to a slower vibration. On the other hand, the average ear has an upper limit of audibility lying somewhere in the octave between 16,000 and 32,000 vibrations per second. These two octaves are the extreme limits of the scale of audibility.

Audible sound is thus defined as a disturbance in the atmosphere whereby a form of wave motion is propagated from some source at a velocity of 1,075 feet per second, the transmission being accomplished by alter-

nate condensations and rarefactions of the atmosphere in cycles having a fundamental frequency ranging somewhere between 16 per second and 32,000 per second.

The superposed waves on the fundamental, which we have called harmonics, are present in most distinctive sounds, and particularly in the human voice. They permit us to distinguish notes of different musical instruments when sounded at the same pitch. They also establish subtle differences in the voice which may indicate anger or joy, or permit us to distinguish the voice of one person from that of another. Figure 104 illustrates wave forms for different kinds of sound and, similarly, Figure 105 shows the predominating wave shapes of certain spoken vowels.

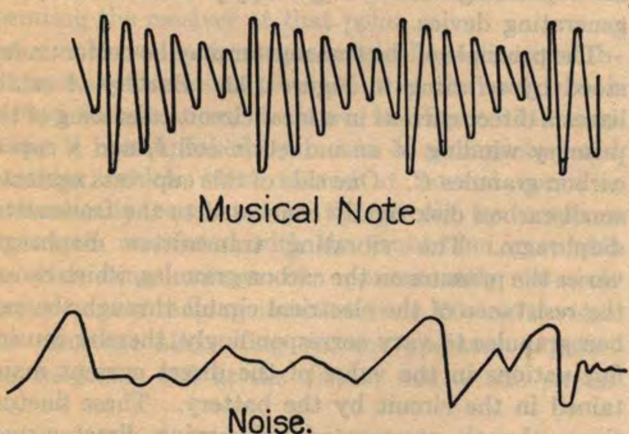


FIGURE 104

Fortunately, in telephone transmission, which is essentially a problem of conveying "intelligibility" from the speaker to the listener, we are not seriously concerned with sounds having either fundamental or harmonic frequencies that extend throughout the entire scale of audibility. The sound frequencies which play an important part in rendering the spoken words of ordinary conversation intelligible are the band of frequencies within the audible scale ranging from approximately 200 to 2,500 cycles per second. Within this band the frequencies between 700 and 1,100 cycles per second are perhaps of greatest importance.

60. The Simple Telephone Circuit

The original telephone, as invented by Bell in 1876, consisted of a ruggedly constructed telephone receiver. It was used both as a transmitter and a receiver at that time. The telephone circuit in its simplest form consisted of two wires terminated at each end by such an instrument but without transmitter or battery and without signaling features. Figure 106 shows such a circuit.

At the speaker's station, the sound waves of the voice

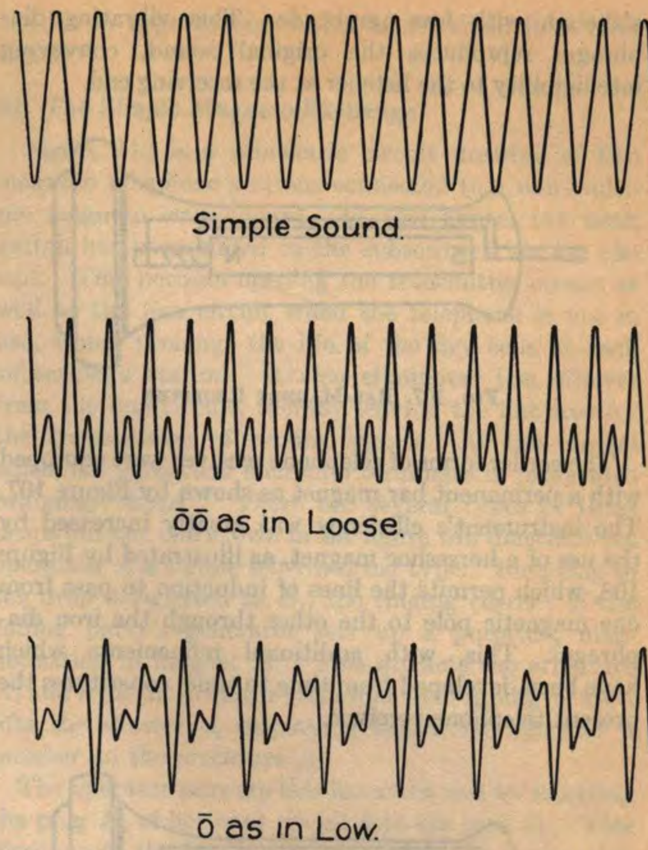


FIGURE 105

strike the metal diaphragm of the telephone receiver, and the alternate condensations and rarefactions of the air on the side of the diaphragm establish in it a sympathetic vibration. Behind the diaphragm is a permanent bar magnet and the lines of induction leaving the magnet are crowded in the vicinity of the metal diaphragm. The vibration of this diaphragm causes a

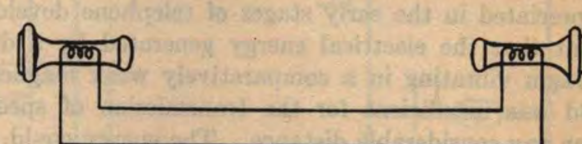


FIG. 106. ELEMENTARY TELEPHONE CIRCUIT

corresponding change in the number of lines that thread through the receiver winding, resulting in the turns of the winding being cut by the building up and decaying lines. This establishes a varying electrical current in the winding of the telephone receiver, having wave characteristics similar to the characteristics of the sound wave. This current, in passing through the receiver winding at the distant end, alternately strengthens and weakens the magnetic field of the permanent magnet, thereby lessening and increasing the pull upon the receiving diaphragm, which causes it to vibrate in unison with the diaphragm at the transmitting end, although with less amplitude. This vibrating diaphragm reproduces the original sound, conveying intelligibility to the listener at the receiving end.

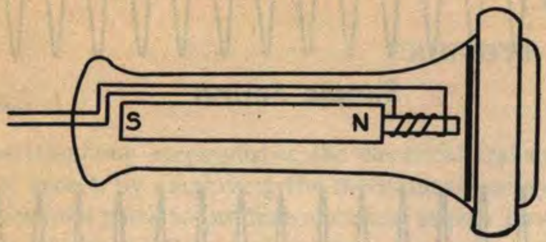


FIG. 107. BAR-MAGNET RECEIVER

The earlier forms of telephone receiver were equipped with a permanent bar magnet as shown by Figure 107. The instrument's efficiency was greatly increased by the use of a horseshoe magnet, as illustrated by Figure 108, which permits the lines of induction to pass from one magnetic pole to the other through the iron diaphragm. This, with additional refinements which have been developed from time to time, constitutes the present telephone receiver.

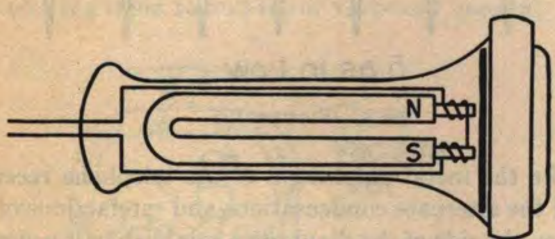


FIG. 108. HORSESHOE-MAGNET RECEIVER

Although the principle of Bell's original telephone applies to the present day telephone receiver, it was appreciated in the early stages of telephone development that the electrical energy generated by a diaphragm vibrating in a comparatively weak magnetic field was insufficient for the transmission of speech over any considerable distance. The energy could, of course, be increased by using stronger magnets, louder sounds, and the best possible diaphragms, but even with any ideal telephone receiver that might be perfected, voice transmission would be limited to comparatively short distances. One year after the invention of the original telephone, the Blake transmitter was introduced. It worked on the principle of a diaphragm varying the strength of an already established electrical current, instead of generating electrical energy by means of electromagnetic induction. By this means it was possible to establish an electrical current with an energy value much greater than that conveyed to the instrument by a feeble sound wave. The battery in this case was the chief source of energy

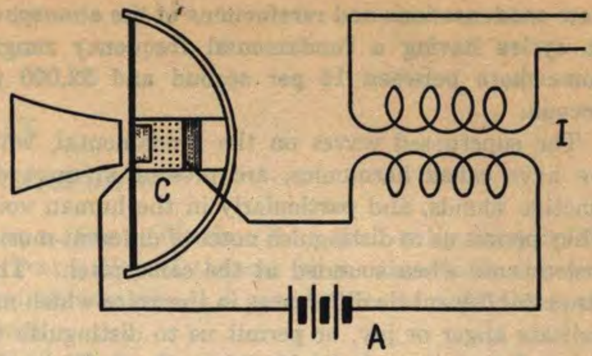


FIG. 109. PRINCIPLE OF THE TELEPHONE TRANSMITTER

and the vibration of the diaphragm acted as a means for regulating this energy supply rather than as a generating device.

The principle of the transmitter may be better understood by referring to Figure 109. Battery *A* establishes a direct current in a local circuit consisting of the primary winding of an induction coil *I*, and a cup of carbon granules *C*. One side of this cup rests against a small carbon disk rigidly connected to the transmitter diaphragm. The vibrating transmitter diaphragm varies the pressure on the carbon granules, which causes the resistance of the electrical circuit through the carbon granules to vary correspondingly, thereby causing fluctuations in the value of the direct current maintained in the circuit by the battery. These fluctuations, though represented by varying direct-current values instead of by an alternating current, as in the case of the telephone circuit in Figure 106, establish an alternating E.M.F. in the secondary winding of the induction coil. This, in turn, sets up an alternating current through the local receiver, over the line, and through the distant receiver. The operation of the distant receiver is no different than that explained in connection with Figure 106.

Figure 110 shows transmitters used at the ends of a simple telephone circuit. When the magnetic field is established by the fluctuating current through the primary of the induction coil, an alternating current is induced in the secondary of the coil. This current flows through the receiver at the same end of the cir-

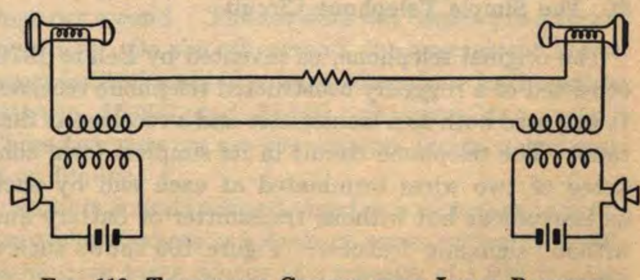


FIG. 110. TELEPHONE CIRCUIT WITH LOCAL BATTERY TRANSMITTERS

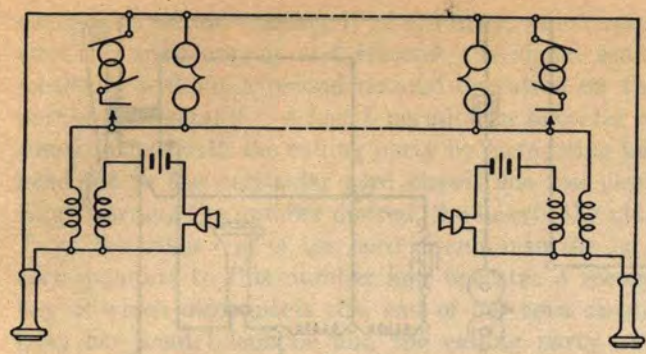


FIG. 111. TELEPHONE CIRCUIT WITH SIGNALING EQUIPMENT

cuit, giving "side-tone" to the receiver at the home station. It is also transmitted to the distant station, operating the receiver at that point.

A simple two-party magneto telephone circuit without central office connections and with the hook switch omitted for clearness, is shown in Figure 111. Signaling is accomplished by means of a magneto hand generator, which when turned at normal speed is automatically connected in the circuit by a spring mechanism associated with the crank and generates an alternating voltage of approximately 20-cycle frequency and ranging in value from 50 to 75 volts. The resultant alternating current operates a polarized telephone bell at the distant end of the circuit similar in type to one which is described in a later article.

61. The Simple Magneto Exchange

Figure 112 is a schematic circuit drawing of two magneto telephone stations connected to a non-multiple magneto switchboard. In this figure, the hook switch has been added to the subscriber's station circuit. This permits opening the transmitter circuit as well as the line circuit when the telephone is not in use, which prolongs the life of the dry cells at each subscriber's station. It also eliminates the receiver from the line circuit, thereby leaving the line free for the transmission of ringing signals. At the central office, the subscriber's circuits terminate in jacks with bridging "drops". There are several types of these drops but the one shown in the figure will illustrate the operation of a simple self-restoring type. Referring to the drop designated as D_1, the ringing current of the calling party's generator sets up a pulsating magnetization in magnet M. This attracts the armature A_1 and trips the armature A_2 which in dropping forward lifts the shutter S, displaying before the operator a number on the armature A_2.

The operator answers this incoming call by inserting the plug P_1 of her cord circuit into the jack J_1. This disconnects the bridged drop winding M, preventing it from affecting the talking current transmission, and

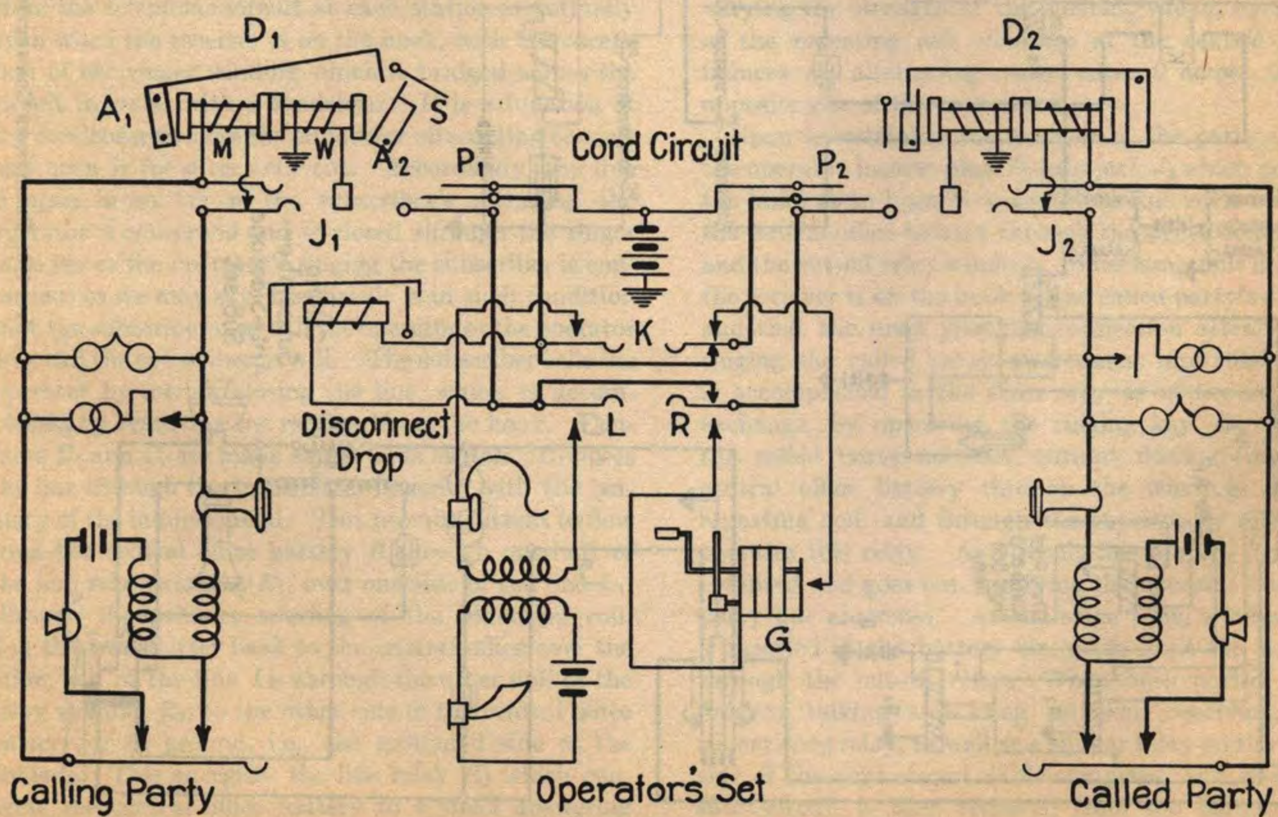


FIG. 112. TELEPHONE CONNECTION THROUGH MAGNETO EXCHANGE

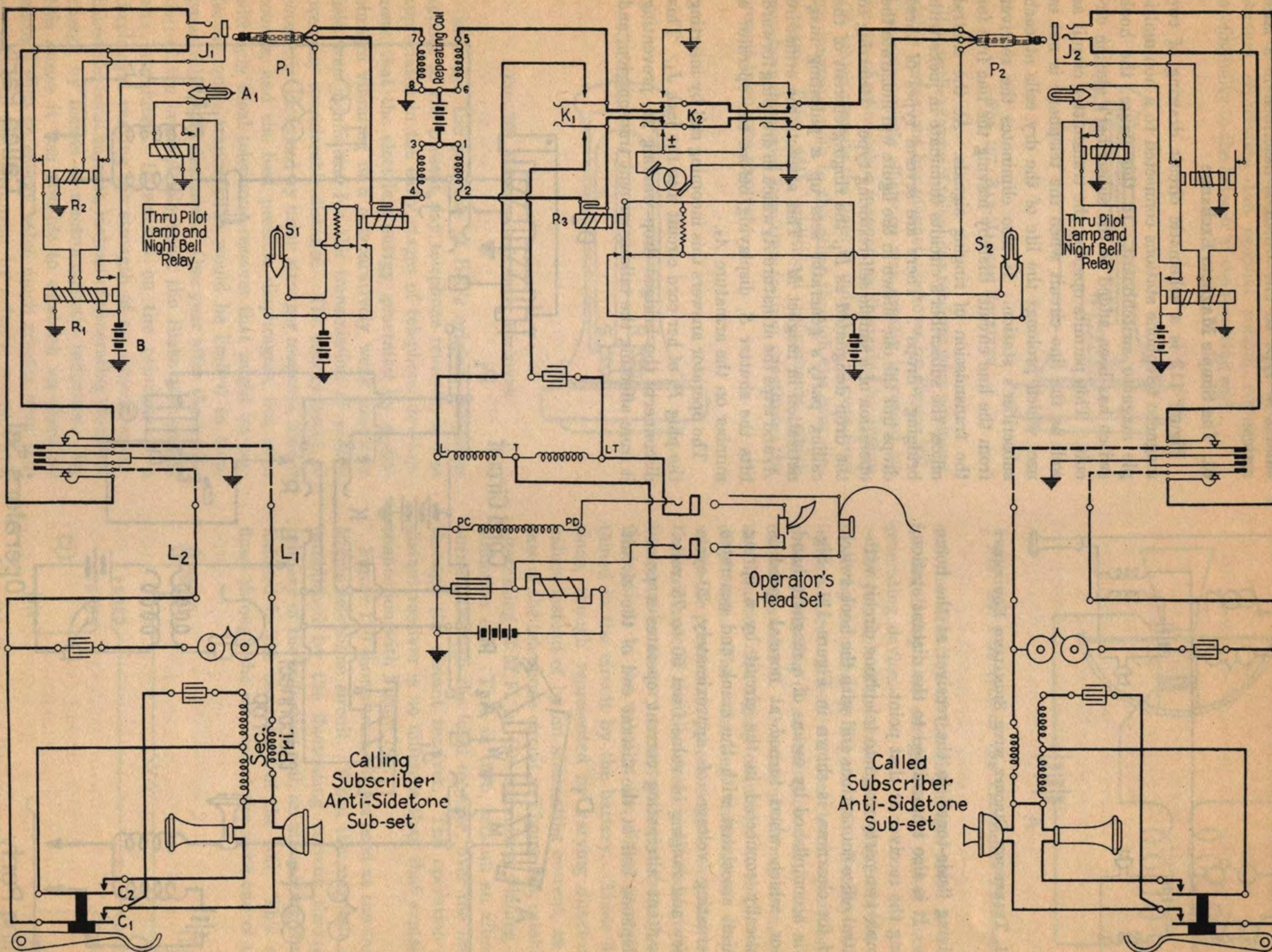


FIG. 113. TELEPHONE CONNECTION THROUGH COMMON BATTERY EXCHANGE

energizes a second winding W of the drop, which operates the armature A_2 and restores the signal automatically without a second manual operation on the part of the operator. A key L permits the operator to communicate with the calling party by connecting her head set to the particular cord circuit she has used. After learning the number desired, she inserts the plug P_2 of the other end of the cord circuit into the jack corresponding to this number and operates a second key R which disconnects this end of her cord circuit from her head telephone and the calling party and connects it to the generator G. She is now able to ring the called party by turning the generator crank. In the larger magneto offices, which are equipped with ringing machines, the operation of key R connects the cord circuit with ringing leads that are energized at all times and the necessity of turning a ringing generator associated with the switchboard position is eliminated.

62. The Common Battery Telephone Exchange

In Chapter VII we learned that it is possible for a number of circuits to be energized from a single battery, and that if the battery has a very low internal resistance, the operation of any one of these circuits does not interfere with the operation of any other. Figure 113 shows a telephone connection between two common battery stations terminating at the same central office. Here the telephone circuit at each station is normally open when the receiver is on the hook, with the exception of the ringer winding which is bridged across the circuit in series with a condenser. It is a function of the condenser to close the circuit for alternating current and open it for direct current. Accordingly, the line is open in so far as the subscriber's signaling the operator is concerned and is closed through the ringer in so far as the operator's ringing the subscriber is concerned; or we may say, the circuit is in such condition that the subscriber may call the operator or the operator may call the subscriber at will. The subscriber calls the operator by merely closing the line, which is accomplished by removing the receiver from the hook. Contacts C_1 and C_2 are made at the hook switch. C_1 closes the line through the transmitter in series with the primary of the induction coil. This permits current to flow from the central office battery B through one-half of the line relay winding R_1, over one side of the line L_1, through the primary winding of the induction coil, and the transmitter back to the central office over the other half of the line L_2, through the other half of the relay winding R_1, to the other side of the central office battery or to ground, i.e., the grounded side of the battery. This energizes the line relay R_1 which connects the central office battery to a small answering lamp A_1 in the face of the switchboard in front of the operator. This lamp lighting, indicates to the operator that this particular line is calling. She answers the call by inserting plug P_1 into the jack associated with the lighted lamp and to which the line of the calling party is connected. A third battery connection to the sleeve of the plug closes a circuit through the winding of a second relay R_2, known as a "cut-off" relay, which disconnects the line relay from the circuit, putting out the burning answering (or line) lamp A_1. The operator learns the calling subscriber's wishes by connecting her telephone set to the cord circuit by means of the listening key K_1. She talks over the two heavy conductors of the cord circuit through the windings of the repeating coil, which, by means of transformer action, induces current into the other windings of the same coil; this flows back over the calling subscriber's line and induces a current in the secondary of the induction coil, which flows through the telephone receiver.

Not only does the operator's voice current flow from the central office cord circuit to the subscriber's receiver, but there is a direct current furnished by the central office battery through two of the four windings of the repeating coil of the cord circuit, over the line, and through the subscriber's transmitter. This corresponds to the transmitter current furnished by a local battery in the magneto set. It permits the subscriber to talk by virtue of the transmitter carbon resistance varying the strength of the current, which, by means of the repeating coil windings at the central office, induces an alternating voice current across to the opposite side of the cord circuit.

Upon ascertaining the number of the party called, the operator inserts plug P_2 into jack J_2 which permits the lamp S_2 to burn because the circuit is closed from the central office battery through the sleeve connection and the cut-off relay winding. This lamp tells her that the receiver is on the hook at the called party's station and that she must give this connection attention by ringing the called party at frequent intervals. This is accomplished in the same way as in the magneto exchange, by operating the ringing key K_2. When the called party answers, current flowing from the central office battery through the windings of the repeating coil, and through the supervisory relay R_3, operates this relay. As a result the lamp S_2 is short-circuited and goes out, notifying the operator that the party has answered. At the same time, a resistance is inserted in the battery circuit to limit the current through the cut-off relay. When both parties have finished talking and hang up their receivers, this supervisory relay, as well as a similar relay on the other side of the cord circuit, is de-energized, and since the short-circuit is then removed from the lamps, they light. This notifies the operator that both parties are through talking and that both cords are to be taken down. When the operator pulls down both cords, the sleeve circuit of the cord is opened at the jack and the lamps go out.

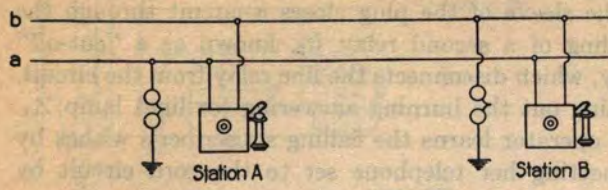


Fig. 114. Signaling on Two-Party Line

It is seen that the operator depends upon burning lamps for each operation excepting that of connecting the calling cord to the jack of the called station. In all common battery operating, **a burning lamp means attention.** Thus a burning lamp in the face of a switchboard signifies "line to be answered"; one burning lamp on a cord signifies "continue ringing on the corresponding cord"; two burning lamps signify "disconnect both cords as both parties have 'hung up' ". A flashing lamp means one party is not hanging up but wishes to place another call or desires the operator to answer in on the connection.

63. Party Lines and Selective Ringing

In order to put more than one subscriber on the same line in common battery operation, it is necessary that some means be provided for signaling each party independently. One method of doing this is represented by Figure 114, where the same 2-wire circuit is used for two subscribers, but one subscriber is rung over wire *a* to ground and the other over wire *b* to ground. A more complex system makes use of the "biased" ringer, which is shown in Figure 115. In this ringer, the magnetic circuit through the cores of the two windings is completed through a permanent steel magnet which gives what is known as a "polarized magnetic circuit". To give the bias effect, a small spring is provided to keep the soft iron armature normally in one position. Without tension on the biasing spring, a current flowing through the windings in one direction will increase the pull on one end of the armature and decrease the pull on the other. This permits the tapper to strike one gong. Likewise, if the current flows in the opposite direction, it will permit the tapper to strike the other gong. An alternating current will, therefore, ring the bell. But if two such ringers are placed in the same circuit and they are biased in opposite directions, a pulsating direct current in one direction will operate the first ringer, while a similar current in the other direction will operate the second.

These two systems may be combined by placing two biased ringers between each wire and ground, thus making a four-party system.

Another system that is used to some extent is known as the "harmonic system". Each ringer is constructed with a special spring armature having a weighted tapper to give it a natural period of vibration. The period of vibration is different for each ringer on a single line and the alternating ringing current must have a corresponding frequency to select a particular ringer. This system requires ringing current taps at the operator's cord circuit of various frequencies, instead of the several arrangements of a single frequency required for the systems described above.

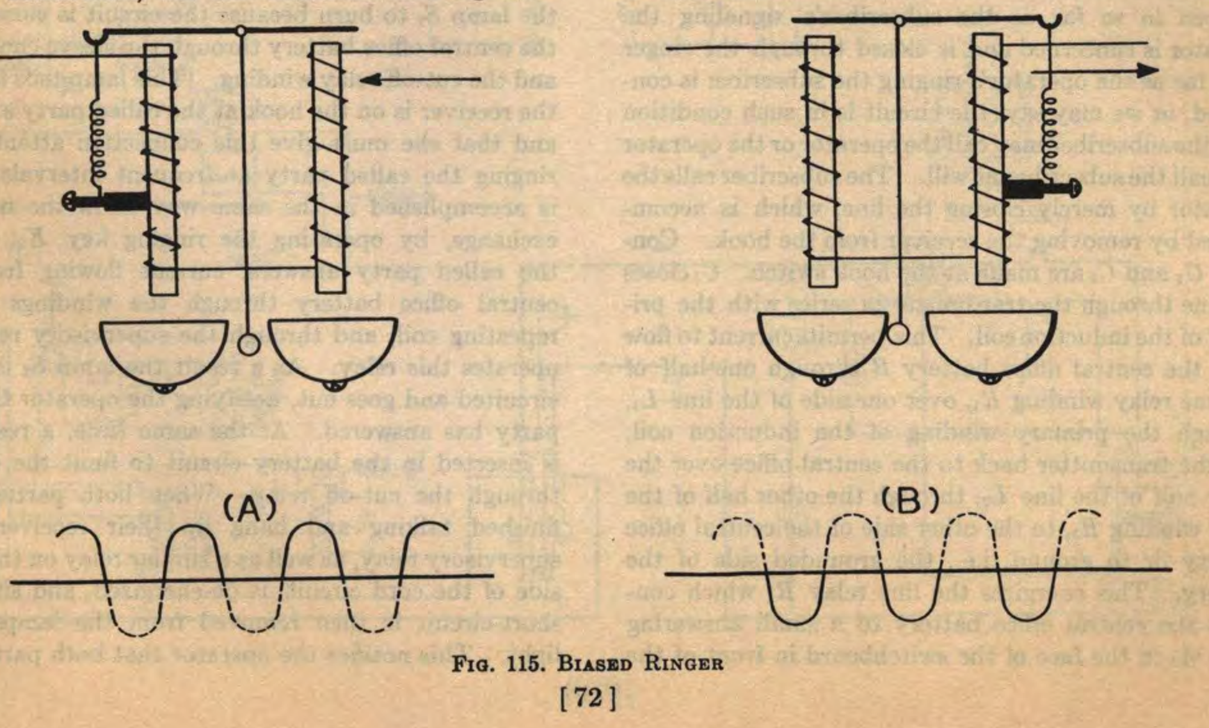


Fig. 115. Biased Ringer

"A" Switchboard

64. The Multiple Switchboard

In a small exchange where a single operator can handle all of the subscribers, it is possible to connect any two subscribers together when each subscriber's line terminates in a single jack only. As described in the foregoing, if subscriber "A" signals the operator, she will plug into his answering jack, which is next to the signal by which he calls her, and upon ascertaining the number he is calling, for example subscriber "B", she will connect him by plugging into the answering jack of subscriber "B" with the other plug of the same cord circuit. However, when there are more than a few hundred subscribers, all of whose lines terminate at the same switchboard, it is obviously impossible for one operator to answer all of these lines. To apportion the work and to make it possible to mount the switchboard apparatus in such a way that it will permit interconnection for a large number of lines, the **multiple switchboard** was developed.

The principle of the multiple switchboard is that the answering jacks and signals are divided up among the various operators, each operator handling on the average about two hundred lines and being responsible for answering any signals from these subscribers. In addition to these answering jacks, there may be as many as 3,300 **calling jacks** in the position in front of each operator. These calling jacks do not have any signals mounted with them, as they are for calling only. The calling jacks are each multipled, that is, connected in parallel with a similarly located jack in the third position to the left and right, and with the answering jack. Any operator can reach any one of about 10,000 calling jacks, either directly in front of her or in the adjacent positions on her left or right. A multiple switchboard is shown diagrammatically in Figure 116. In this figure should subscriber Number 109 call subscriber Number 567, the signal would come in at position "1" where the answering jack for subscriber 109 is located and the operator would connect him by plugging into calling jack Number 567 in the multiple to her right (Position 2). On the other hand, if subscriber 567 called subscriber 109, the operator at position 3 would answer his call and connect him to subscriber 109 by means of the calling jack in the multiple to her right (Position 4). Each operator is warned against plugging into a busy line by means of a "click" which is heard in her head receiver when she starts to

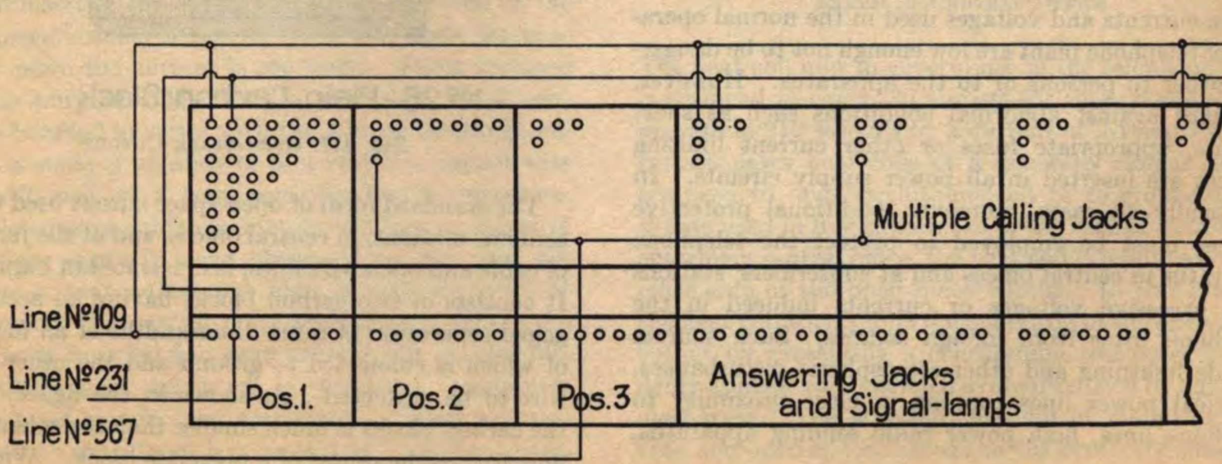
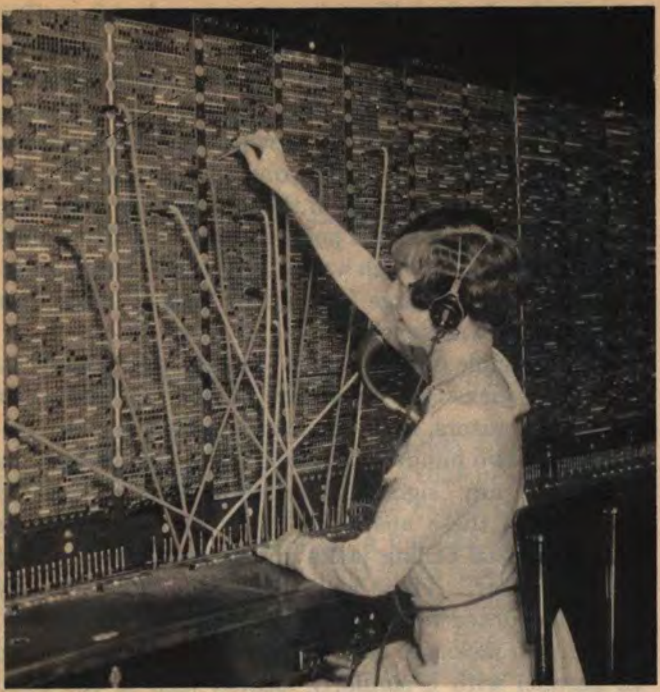


Fig. 116. Multiple Switchboard



"B" Switchboard

plug into a calling jack already in use somewhere else in the multiple.

Note: It is not practicable to take up here any comprehensive study of the apparatus and switching systems used for providing local telephone service. Some branches of the subject, such for instance as machine or dial switching systems, are much too specialized to lend themselves to brief analysis. Trunked connections between subscribers where lines terminate at different central offices, however, involve apparatus features essentially similar to those of the toll switching trunk described later in connection with toll switchboard circuits.

65. Protection Against Foreign Currents and Voltages

The currents and voltages used in the normal operation of telephone plant are low enough not to be dangerous either to persons or to the apparatus. However, to guard against abnormal conditions such as short circuits, appropriate fuses or other current limiting devices are inserted in all power supply circuits. In practically all cases, moreover, additional protective devices must be employed to protect the telephone apparatus in central offices and at subscribers' stations from excessive voltages or currents induced in the telephone wires from foreign sources. Such sources include lightning and other atmospheric disturbances, electrical power lines running in close proximity to telephone lines, high power radio sending apparatus, etc.

Practically all outside telephone plant, except such conductors as are completely underground from terminal to terminal, may be exposed from time to time to one or more of these foreign hazards. Accordingly, whenever exposed wires are led into a central office or a subscriber's station, they are connected first through certain protective apparatus. The particular protective units employed and the manner in which they are connected into the telephone circuits vary somewhat with particular situations, but in general protective devices are of three principal types—namely, open-space cutouts, fuses, and heat coils.

The first and last of these devices ordinarily operate to ground the protected wire, while the fuse opens the wire in which it is inserted. Each of the protective units is designed so that, for the particular situation in which it is used, it will be sufficiently sensitive to operate before the plant which it is protecting is damaged, but on the other hand, not so sensitive as to cause an unnecessary number of service interruptions.

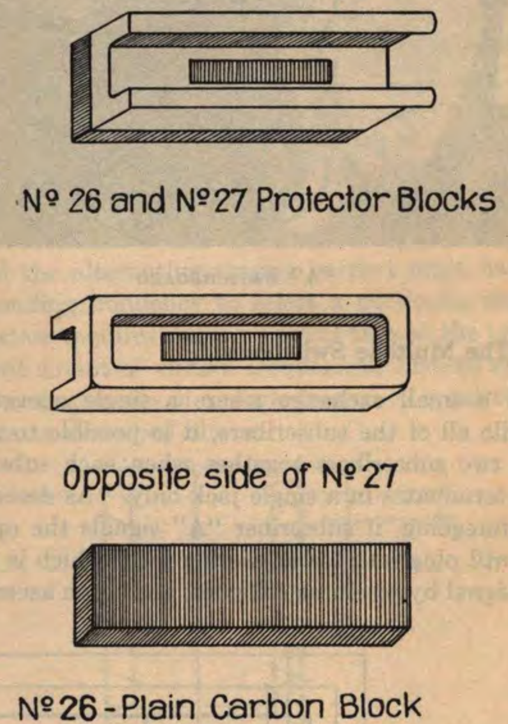


Fig. 117. Open-Space Cutout

The standard form of open-space cutout used at subscribers' stations, in central offices, and at the junctions of cable and open wire lines, is illustrated in Figure 117. It consists of two carbon blocks having an accurately gaged separation of a few thousandths of an inch, one of which is connected to ground and the other to the wire to be protected. As shown in the figure, one of the carbon blocks is much smaller than the other and is mounted in the center of a porcelain block. When the voltage of the telephone wire becomes too high, the wire will be grounded by arcing across the small air-gap between the carbon blocks. If a considerable current flows across the gap in this way, enough carbon may be pulled from the blocks by the arc to partially fill in the gap and cause permanent grounding. Or, in the extreme case, when the discharge is prolonged and sufficiently high, the glass cement with which the small carbon insert is held in the porcelain block may be melted, with the result that the blocks are forced into direct contact by the mounting springs in which they are held. However, in the majority of protector operations the blocks do not become permanently grounded.

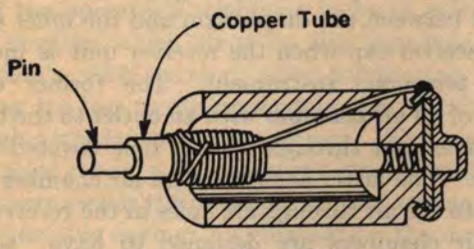


Fig. 118. Heat Coil

The air-gap space between the blocks is designed so that the operating voltage of the protector will be less than the breakdown voltage of the weakest point of the circuit which it is designed to protect and greater than the maximum working voltage of the circuit. The average operating voltage of the open-space cutouts used at subscribers' stations and in central offices is about 350 volts. For the cutouts used at junctions between open wire and cable lines, an average operating voltage of about 710 volts is standard.

When a telephone conductor is grounded by the operation of an open-space cutout, current will continue to flow through the telephone conductor to ground so long as the exposure continues. This current may be large enough to damage the telephone conductor or the protective apparatus itself. Accordingly, it is necessary to insert in the conductor, on the line side of the open-space cutout, a device which will open the conductor when the current is too large. Fuses are used for this purpose. The fuse is simply a metal conductor inserted in series with the wire to be protected, which is made of an alloy or of a very fine copper wire that will melt at a comparatively low temperature. Short lengths of cable conductors (six feet or more) of 24 or finer gage will serve effectively as fuses and will fuse on current values not high enough to overheat dangerously the central office protectors. Where the use of such inserted fine gage cable is not practicable, lead alloy fuses, mounted in fire-proof containers or fire-proof panels, are employed. These are also designed to operate with a current of 7 to 10 amperes.

Finally, it is frequently necessary to protect telephone apparatus against external effects in which the voltage is not high enough to operate the open-space cutout, nor the current high enough to operate fuses, but still high enough to damage apparatus if allowed to flow over a long period. Such currents are usually called "sneak" currents and are guarded against by the use of heat coils. As illustrated in Figure 118, the heat coil consists of a small coil of wire wound around a copper tube which is connected in series with the wire to be protected. Inserted within the copper tube and held in place by an easily melting solder is a metal pin which is connected to the line side of the coil. If sufficient current flows through the coil to melt the solder, this pin will move under the pressure of its mounting spring and thus connect the line to ground.

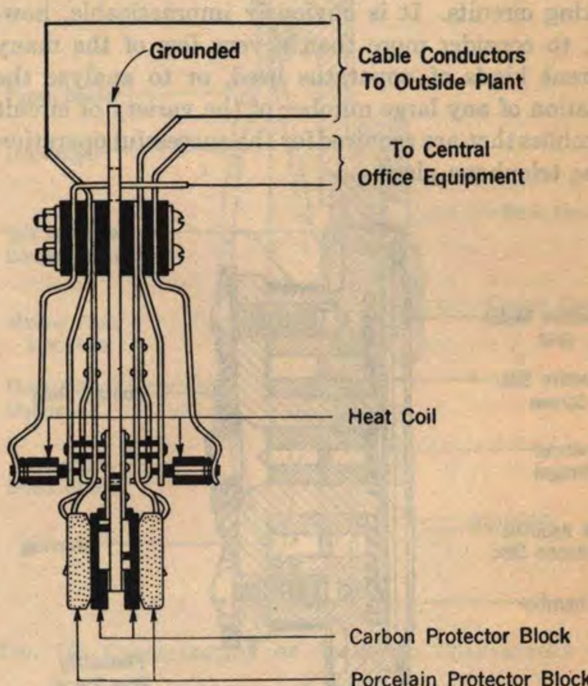


Fig. 119. Heat Coils and Cutouts Mounted on Protector Frame in Central Office

The heat coil now in general use in the telephone plant is designed to carry .35 ampere for three hours and to operate in 210 seconds on a current of .54 ampere. In certain cases heat coils of a generally similar nature are used to open circuits instead of to ground them. Where used in line circuits, as in the case of conductors entering a central office, the heat coil is mounted on the office side of the open-space cutout. In this position the heat coil wiring aids the operation of the open-space cutout by presenting a considerable resistance to suddenly applied voltages such as are produced by lightning discharges. The standard method of mounting heat coils and open-space cutouts on the protector frames in central offices is illustrated in Figure 119.

CHAPTER X

TELEPHONE APPARATUS AND CIRCUITS

66. The Telephone Receiver

The next subject for consideration in our study of electricity would logically be that of the theory of alternating currents. But before, taking up that subject, it may be profitable to examine in more detail certain of the more common and relatively simple types of apparatus used in telephone work, and to observe how apparatus units are connected together to form working circuits. It is obviously impracticable, however, to consider more than a very few of the many different kinds of apparatus used, or to analyze the operation of any large number of the variety of circuit assemblies that are required for the successful operation of the telephone plant.

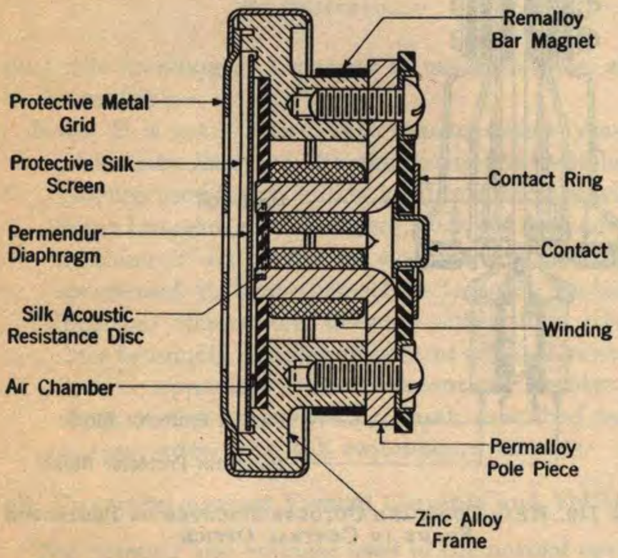


FIG. 120. CROSS-SECTION OF STANDARD RECEIVER UNIT

As the earliest and perhaps most fundamental of telephone apparatus units we may consider first the telephone receiver. Many models of this device have been designed and employed in service since Bell's original invention, and numerous different models are still in use. All operate on the same general principle as was outlined in Article 60, but the details of design show substantial variations.

Figure 120 is a cross-sectional view of the receiver unit which is the present standard in the Bell System. This receiver which is of the bipolar permanent magnet type employs in its construction no less than three of the comparatively new magnetic alloys that were mentioned in Article 23. It is substantially more efficient than any previous design. It also differs notably from earlier types in the extent to which the motion of the diaphragm, which is made of vanadium-permendur, is affected by "acoustic controls". One acoustic control is directly behind the diaphragm, and the other is enclosed between the diaphragm and the inner surface of the receiver cap when the receiver unit is mounted in the telephone instrument. The former control consists of an air chamber with an outlet to the back of the receiver unit through a small hole covered with a silk disc. The latter consists of an air chamber which opens into the air through six holes in the receiver cap. These air chambers are designed to have "acoustic impedances" which match the "electrical impedances" of the receiver and improve its overall efficiency very appreciably. The diaphragm, which is unclamped, rests on a ring-shaped ridge and is held in place by the pull of the magnet. In this way variations in receiver efficiency at different frequencies are practically eliminated. The two permalloy pole pieces are welded to a pair of very strong remalloy or cobalt-steel bar magnets, and the assembly is fastened to a zinc alloy frame. The whole unit is held together by a brass ferrule on the back of which are mounted two silver plated contacts for the electrical connections.

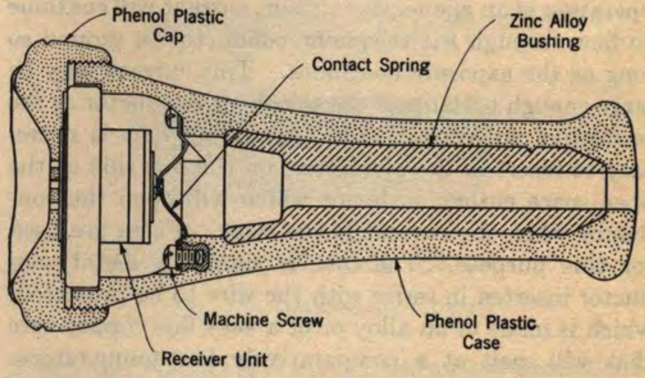


FIG. 121. RECEIVER UNIT AS MOUNTED IN DESK STAND

The receiver unit is designed so that it can be mounted in the standard hand set or in the standard receiver casing used with desk stands, as shown in Figure 121. The receiver cap and other external casings are made of a phenol plastic material and suitable terminals and contact springs are included for connecting the receiver unit contacts to the external wiring. The receiver shell of Figure 121 also encloses a zinc alloy bushing which is included merely to give the instrument sufficient weight to operate the hook switch properly.

The small metal-cased receiver used in standard operators' head sets is coded #528. Formerly monitoring arrangements for operating supervisors, chief operators, and testboardmen involved the use of a special high resistance receiver, coded #525, which could be bridged across a talking connection without causing appreciable transmission loss. However, the #528 receiver with vacuum tube monitoring is now used with most testboards. Similarly, a repeating coil is now installed in the receiver leads of the left-hand pair of the operator's telephone jacks of each switchboard position which steps up the impedance of the supervisor's #528 receiver when monitoring, thereby reducing the loss to approximately that obtained with the high impedance #525 receiver. The use of but one type of receiver for operators, chief operators, and supervisors avoids the changing of head sets and results in considerable savings by eliminating the duplication of sets required under the former plan of monitoring.

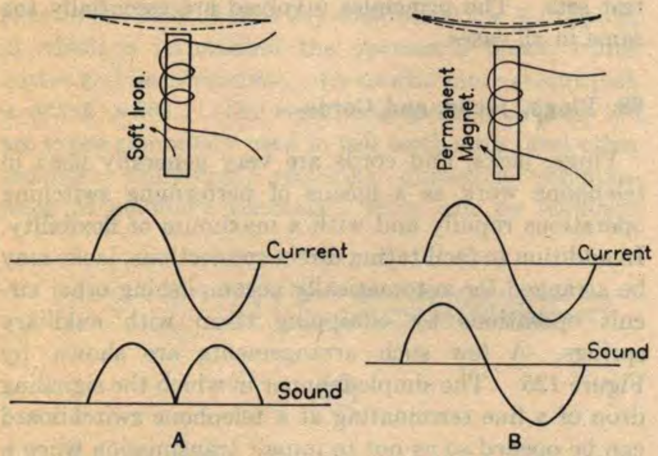


Figure 122

The telephone receivers discussed above are equipped with permanent magnets, and it is of course important that the magnetism should not be impaired by jarring or other abuse. A permanent magnet not only increases the amplitude of vibration of the diaphragm when the voice current is flowing through the windings but prevents the diaphragm vibrating at twice the voice frequency. This principle is illustrated in Figure 122. When a piece of soft iron is held near an electromagnet, it is attracted by the magnet regardless of the direction of the current in the windings. Thus, an alternating current in a winding on a soft iron core will assert an attraction during each half cycle, which in the case of the receiver diaphragm will establish a vibration with a frequency twice that of the current. If, on the other hand, a permanent magnet is used, the alternating current establishes a vibration of the same frequency as the current by merely increasing or lessening the pull already exerted on the diaphragm.

67. The Telephone Transmitter

As we have already learned (Article 60), the operation of the telephone transmitter depends upon the variation in resistance of carbon granules with changes in pressure. Figure 123 shows in cross-section the transmitter unit which is the present standard for subscribers' telephone sets. This transmitter is of the "direct action" type; that is, the movable element attached to the diaphragm which actuates the granular

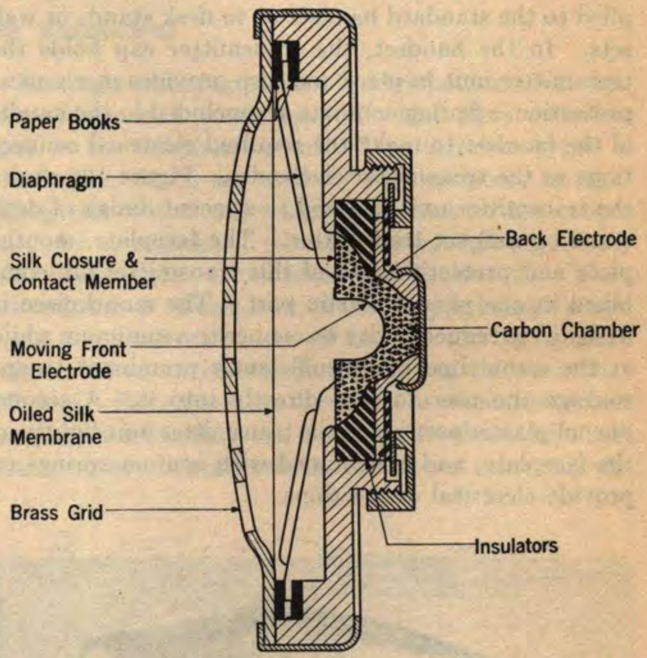


Fig. 123. Cross-section of Standard Transmitter Unit

carbon is an electrode, and serves the dual purpose of contact and pressure surface. As the drawing shows, this dome-shaped electrode is attached to the center of a conical diaphragm, and forms the front center surface of the bell-shaped carbon chamber.

The diaphragm is made of aluminum alloy .003 inch thick with radial ridges to increase stiffness. Paper books, which consist of a number of thin impregnated paper rings, support the diaphragm at its edge on both sides. The carbon chamber is closed on the front side by a silk membrane clamped under the flange of the diaphragm electrode. A light spoked copper contact member, clamped under the diaphragm electrode, is the means of providing a flexible connection between this front electrode and the supporting metal frame. The fixed back electrode is held in place in the frame by a threaded ring and is insulated by a phenol fibre washer and a ceramic insulator which also forms one of the surfaces of the carbon chamber. The active surfaces of both electrodes are gold plated. A brass plate which is perforated with large holes protects the vibrating parts against mechanical injury. Moisture is kept out of the working parts by an oiled silk moisture-resisting membrane placed between the brass plate and the diaphragm.

The shape of the electrodes and the carbon chamber provides sufficient contact force between the diaphragm electrode and the granular carbon in the zone of maximum current density so that this transmitter operates satisfactorily in any position. When new, it has a resistance of around 30 to 40 ohms.

This transmitter unit is designed to be readily applied to the standard handset or to desk stands or wall sets. In the handset, the transmitter cap holds the transmitter unit in place and also provides mechanical protection. Spring contacts are included in the handle of the handset to make the required electrical connections to the transmitter electrodes. Figure 124 shows the transmitter unit adapted to a recent design of desk stand or wall set transmitter. The faceplate, mouthpiece and protective grid of this transmitter are combined in one phenol plastic part. The mouthpiece is designed to reduce cavity resonance to a minimum while at the same time being sufficiently prominent to encourage the user to talk directly into it. A second phenol plastic part holds the transmitter unit tightly in the faceplate, and is equipped with contact springs to provide electrical connections.

Standard Handset

The above description is confined to recent transmitter designs only. Many older types of subscribers' transmitters are of course in use in the telephone plant. Different designs are also used for the standard operator's breast-set transmitter, and in linemen's

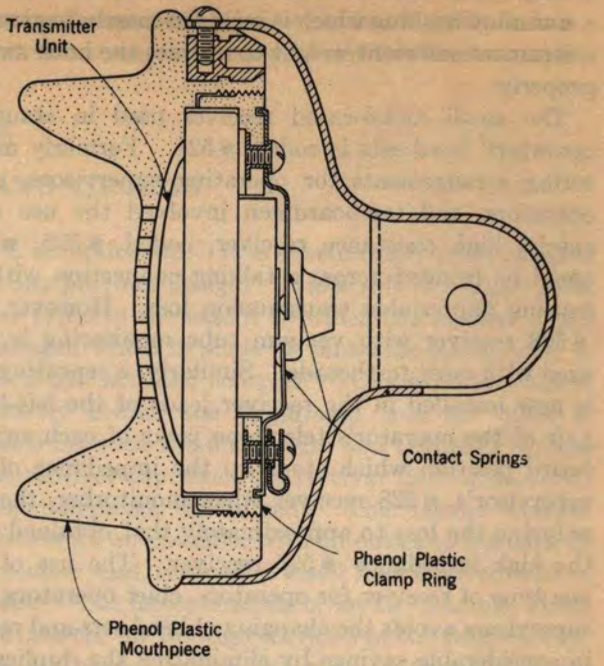


Fig. 124. Transmitter Unit as Mounted In Desk Stand

test sets. The principles involved are essentially the same in all cases.

68. Plugs, Jacks, and Cords

Plugs, jacks, and cords are very generally used in telephone work as a means of performing switching operations rapidly and with a maximum of flexibility. In addition to facilitating direct connections, jacks may be arranged for automatically accomplishing other circuit operations by equipping them with auxiliary springs. A few such arrangements are shown by Figure 125. The simple manner in which the signaling drop of a line terminating at a telephone switchboard can be opened so as not to impair transmission when a plug is inserted in the jack is illustrated by Figure 125-A. Figure 125-B shows another use of the same auxiliary contact. Here a telegraph set terminated with a two-conductor plug may be looped with the wire at a single operation, or an ammeter connected to a cord may be inserted for measuring the current in the wire. Figure 125-C illustrates the use of normals for two springs of a three-conductor jack such as is used in connection with the #10 local switchboard to perform a function similar to that of the cut-off relay in the #1 switchboard. Figure 125-D illustrates a commonly used two-conductor jack which in this case is wired to operate a self-restoring drop in the same way as the three-conductor jack shown in Figure 112.

The mechanical construction of a few types of jacks widely used in connection with long distance service, is illustrated in Figure 126. The #49 jack is mounted

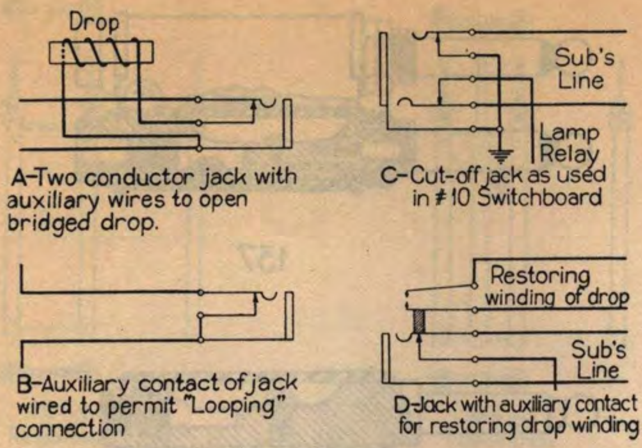


FIG. 125. TYPICAL USES OF AUXILIARY JACK CONTACTS

in strips of 5, 10, or 20 for use in the face of both local and toll switchboards. It takes the #110 plug. A smaller jack, coded as the #92, takes the #109 plug, and is used in the face of larger switchboards where the multiple must accommodate a greater number of lines than the toll or small local switchboard multiple. The #99 jack, illustrated in the same figure, is mounted in pairs in the switchboard key shelf to take a #137 plug, in which is terminated the operator's breast transmitter and head receiver. (A similar more recent jack is coded #364.) The remaining jacks in Figure 126 are types commonly used in toll testboards, and other testroom equipment, requiring numerous combinations of auxiliary contacts. They can be mounted either singly or in pairs to accommodate single one- or two-conductor plugs such as the #116 and #47 or 2-, 3- or 4-conductor plugs such as the #209, #241, and #289 types, respectively. Jacks of this type are made with a sherardized metal frame having a brass sleeve mechanically fastened to its front face. The channel shape readily permits the mounting of german silver contact or auxiliary springs properly insulated from each other by bushings and washers.

Figure 127 illustrates both the mechanical and electrical features of various plugs and Figure 128 shows the construction of a commonly used type of switchboard cord. While this is only one of many cords in use, it represents the standard features and gives an insight into cord manufacturing processes.

69. Resistances

No single unit of apparatus is more fundamental than the resistance, several types of which have countless uses in the telephone plant. Two common types of resistances, which are used for such purposes as regulating the central office supply current to the proper value for operating and releasing relays, lighting switchboard lamps, etc., are illustrated in Figure 129. They are coded as #18 and #19, the #18-type being a single plain resistance, and the #19-type having a third connection to an intermediate point of the resistance winding. Both types are furnished in resistance values ranging from less than one ohm to a few thousand ohms. The accuracy is ordinarily within 5 per cent, and the

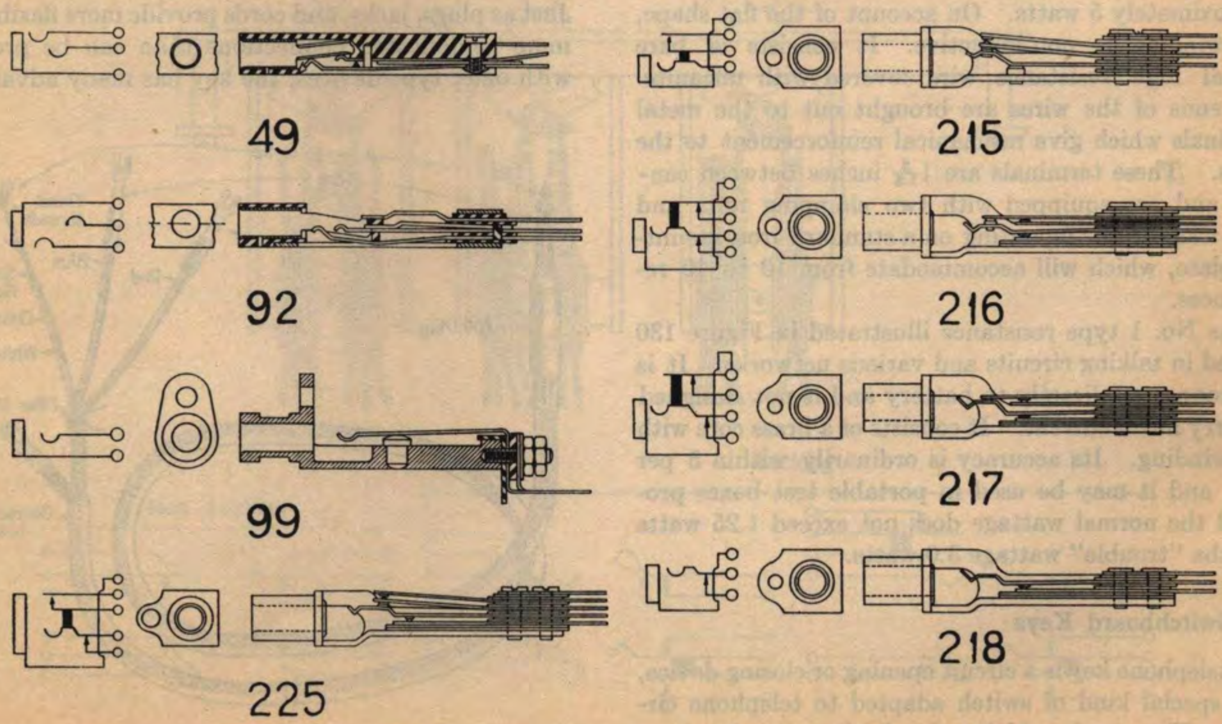


FIG. 126. MECHANICAL CONSTRUCTION OF STANDARD JACKS

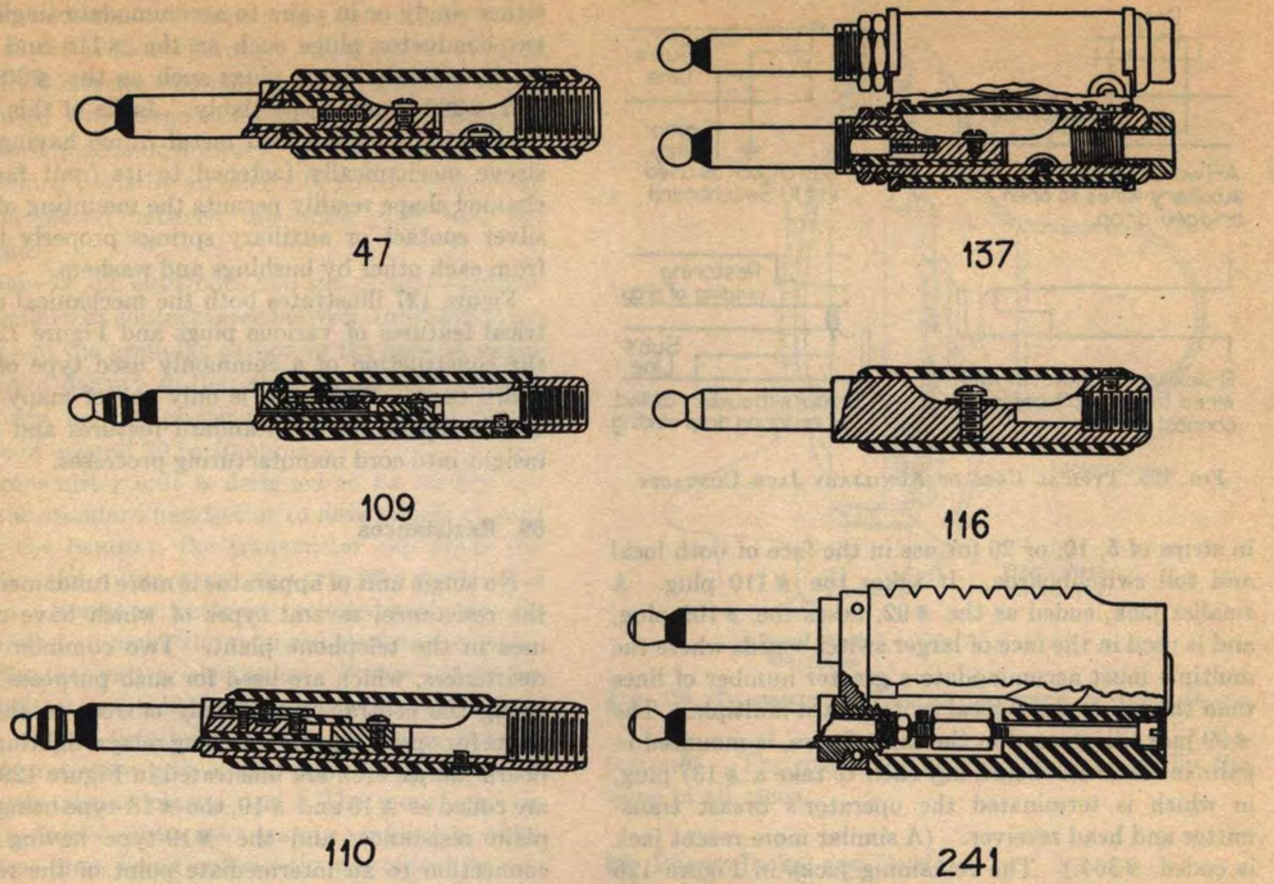


Fig. 127. Mechanical Construction of Standard Plugs

safe radiating capacity, which depends upon the mechanical design rather than the resistance value, is approximately 5 watts. On account of the flat shape, the winding is non-inductive. It consists of bare special high resistance wire covered with micanite. The ends of the wires are brought out to the metal terminals which give mechanical reinforcement to the edges. These terminals are $1\frac{5}{16}$ inches between centers, and are equipped with two clamping nuts and fibre washers for mounting on a standard iron mounting plate, which will accommodate from 10 to 40 resistances.

The No. 1 type resistance illustrated in Figure 130 is used in talking circuits and various networks. It is not connected directly to battery and is not designed to carry much current. It consists of a brass core with one winding. Its accuracy is ordinarily within 5 per cent, and it may be used in portable test boxes provided the normal wattage does not exceed 1.25 watts and the "trouble" wattage 3.0 watts.

70. Switchboard Keys

A telephone key is a circuit opening or closing device, or a special kind of switch adapted to telephone circuits. The way in which a simple six-spring key may perform the same circuit functions as a double-pole, double-throw switch, was illustrated in Figure 6. Just as plugs, jacks, and cords provide more flexible and more complicated connections than can be provided with older type devices, the key has many advantages

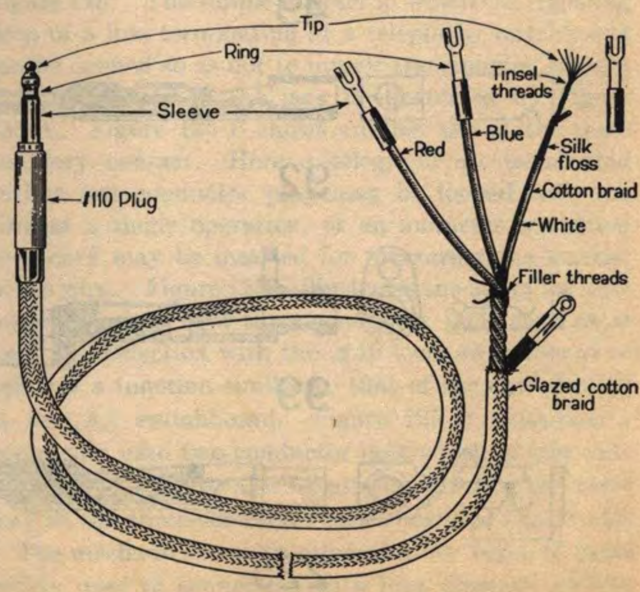


Fig. 128. Switchboard Cord

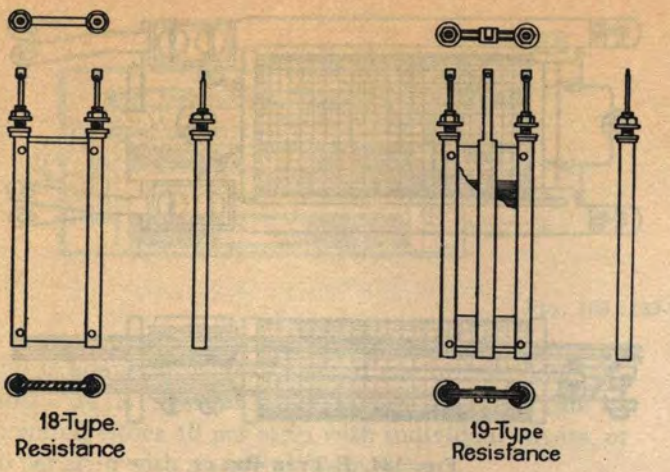


Fig. 129. Flat Resistances

over the knife switch, and facilitates additional features essential to telephone operation. Contacts to be made or broken may be delicately adjusted through the use of german silver springs. These contacts, which are adequate for carrying the current values ordinarily used in telephone circuits, are made through special contact metal welded to the springs, thereby preventing excessive resistances from being inserted in the sensitive telephone circuits. Auxiliary contact springs permit the operation of additional or more complicated circuit features, which could not be easily provided on any other form of switch.

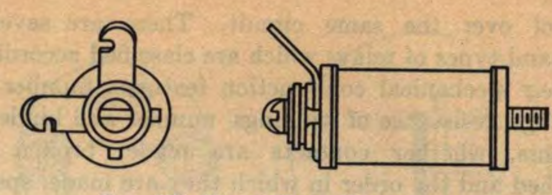


Fig. 130. No. 1 Type Resistance

Figure 131 illustrates a key used in the operation of switchboard circuits, which is especially important. This key was designed for use in connection with the universal switchboard key shelf and has the so-called "unit" construction. This permits one or more key spring units to be mounted, as illustrated, on a standard metal base which is equipped with a hard rubber top. Two types of spring units are provided, the lever type (Figure 131) and the push-button type. The convenient manner in which individual units can be removed, and in which any key combination can be had by selecting various units for one standard base, has certain obvious maintenance advantages.

71. Relays

A relay may be defined as an electrically operated switch or key. It gives one electrical circuit control over one or more other electrical circuits; or as in the case of the locking type relay, it may give certain

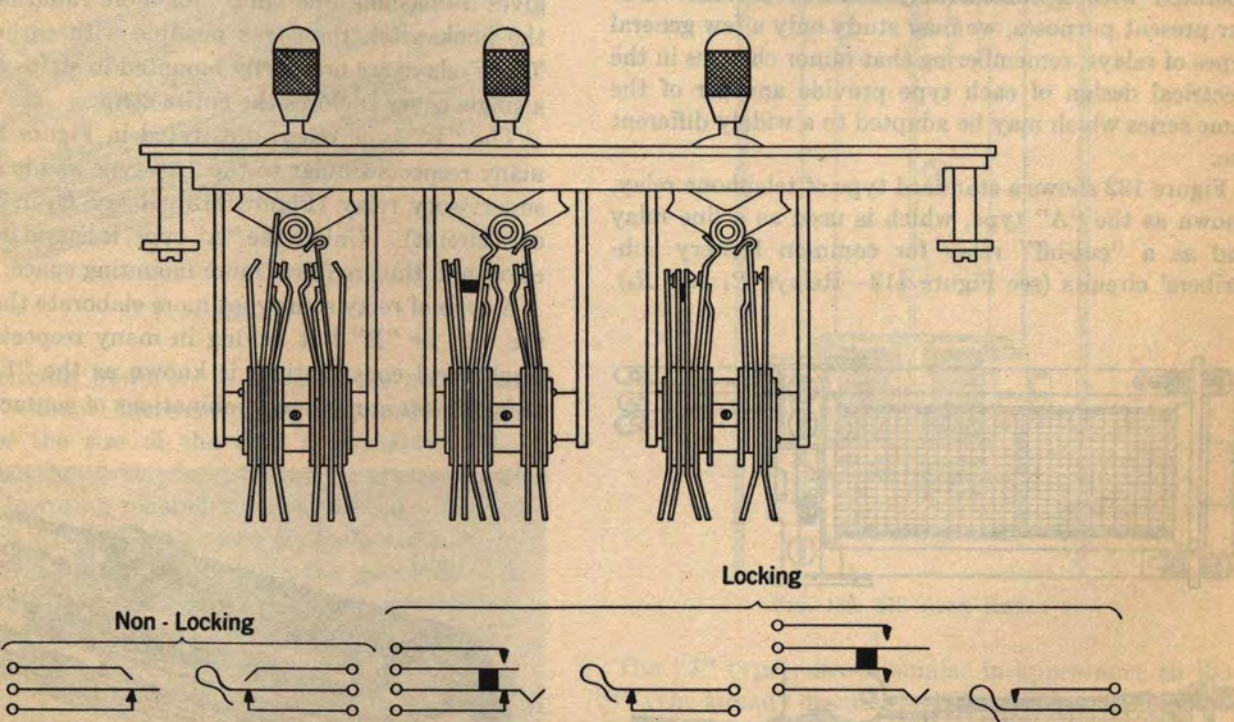


Fig. 131. Standard Switchboard Key Assembly

control over the same circuit. There are several thousand types of relays which are classified according to their mechanical construction features, number of windings, resistance of windings, number and kinds of contacts, whether contacts are made, broken or switched and the order in which they are made, speed of operation, and current values required for operation. In connection with speed of operation, relays may be designed to have a time-delay in operating or releasing.

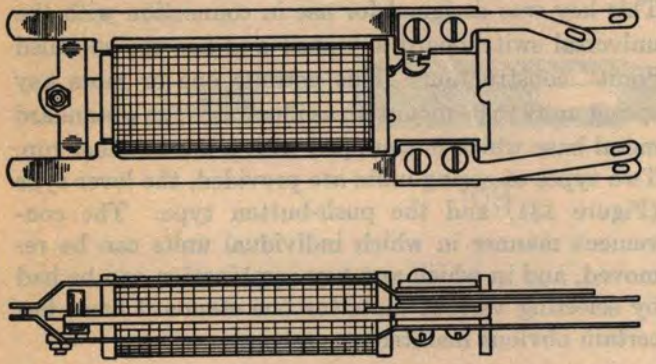


Fig. 132. A-Type Relay

Such relays are classified as "slow operate" and "slow release" respectively. Relays which are termed "marginal" are also used. These are usually designed to operate only on a fairly high current value but to stay operated with a considerably reduced current. For our present purposes, we may study only a few general types of relays, remembering that minor changes in the electrical design of each type provide another of the same series which may be adapted to a widely different use.

Figure 132 shows a standard type of telephone relay, known as the "A" type, which is used as a line relay and as a "cut-off" relay for common battery subscribers' circuits (see Figure 113—Relays R_1 and R_2).

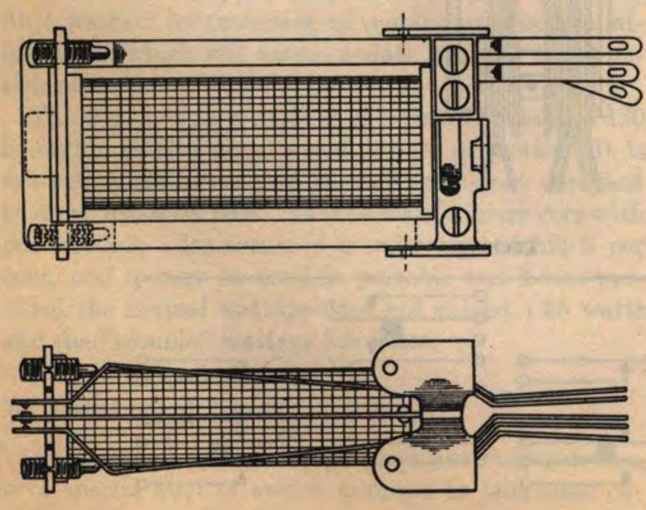


Fig. 133. B-Type Relay

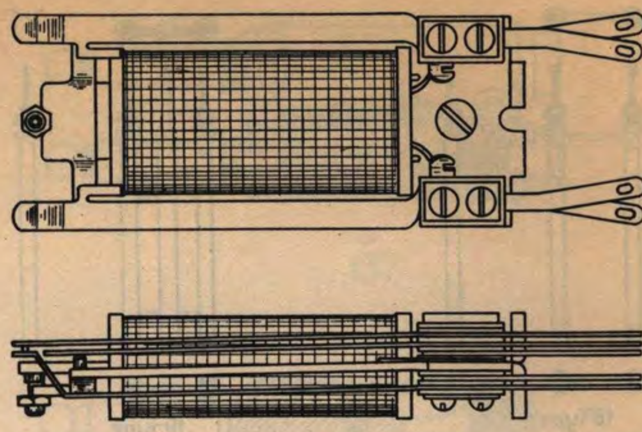
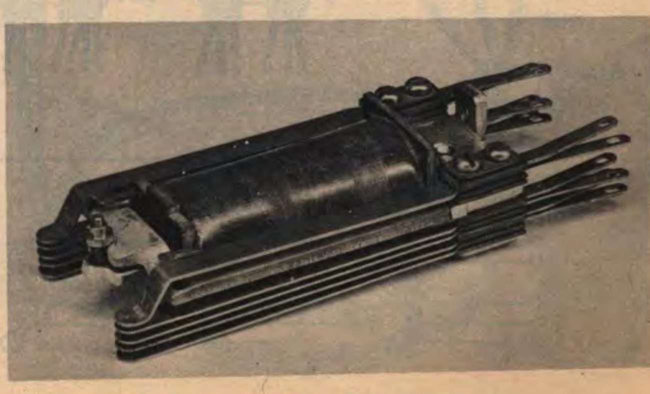


Fig. 134. E-Type Relay

The mechanical construction of the "A" type is typical of the great majority of modern relays. In dimensions it is both small and narrow, thereby permitting a large number to be mounted in a comparatively small space, which results in a notable saving of relay rack space in local central offices where many line and cut-off relays are in use. The soft iron armature forms a loop which completes the magnetic circuit from the core through the two halves of the loop, and mechanically operates the contact springs. The winding is of enamel insulated wire, which also aids in reducing the size of the relay. The "A" type is very quick in operation, and gives a "flashing line lamp" for more rapid moving of the hookswitch than was possible with earlier types. These relays are ordinarily mounted in strips of 20 and a single cover encloses the entire strip.

The "B" type relay, illustrated in Figure 133, is in many respects similar to the "A" type and is used as a supervisory relay (Figure 113—Relay R_3 in the local cord circuit). Unlike the "A" type, it has an individual cover and thus requires more mounting space.

A type of relay somewhat more elaborate than either the "A" or "B" but having in many respects similar mechanical construction, is known as the "E" series. It facilitates numerous combinations of contact springs



E-Type Relay

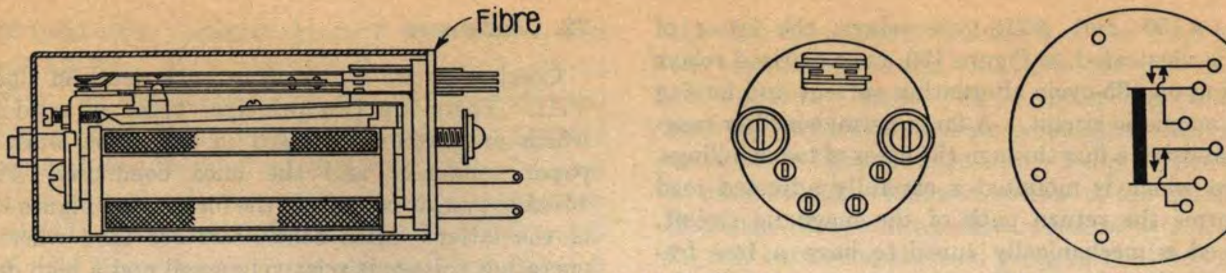
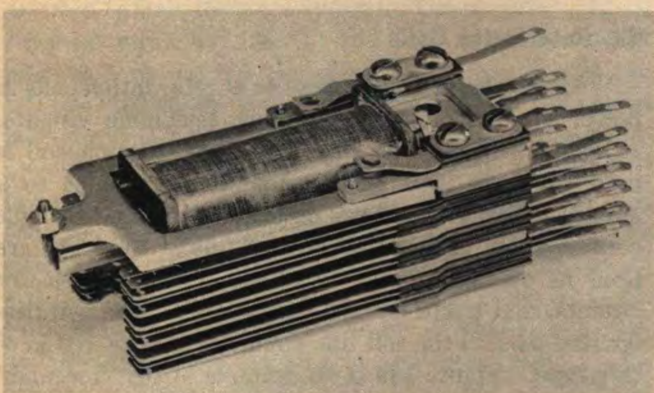


Fig. 135. 122-Type Relay

and is adapted for very general use in telephone circuits. It is illustrated in Figure 134. It can be mounted either 10 per strip with individual covers, or 20 per strip with an overall strip cover.

The "R" type relay is a flat relay which is similar to the "E" type, except that the core, although having the same cross-sectional area, is of a semi-elliptical shape. This affords a greater winding space and permits of a shorter length of turn than is possible on the "E" type core.

The "U" type relay is an all-purpose relay of the same general design as those already mentioned. Certain improvements have been incorporated in its



U-Type Relay

magnetic circuit, however, to increase the pull of its armature without increasing the required current strength or the size of the core appreciably. This permits this type of relay to be built with as many as 24 springs. Operating reliability has also been increased above older types by using two separate contacts on each spring. This greatly reduces the possibilities of faulty operation due to dirt particles between contact points. The "Y" type relay is similar to the "U" type except that it is especially designed for accurately timed slow-release operation.

A type of relay still standard for certain telephone circuit uses, though of older design than the types discussed above is coded as the #122-type. It is shown in Figure 135. Its mechanical appearance is similar to the #178-type and with the exception of the number of springs, it is similar to the #125 (3 groups), #149 (one group) and #162 (one group). It has a spool winding and a laminated iron core with the magnetic circuit completed through the armature and the soft iron framework at each end of the core. Relays of this design ordinarily have round caps which are fastened by means of a nut as shown in the figure.

The relays thus far described are intended to operate only on direct current. There are a few other types that may be classified as alternating-current relays, the more common of which are the "J" type, the #87-type, the #172-type, the #196-type, the #150-type, and the #218-type.

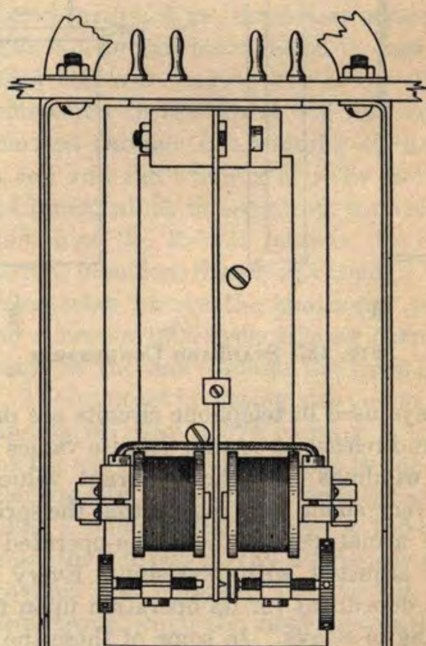
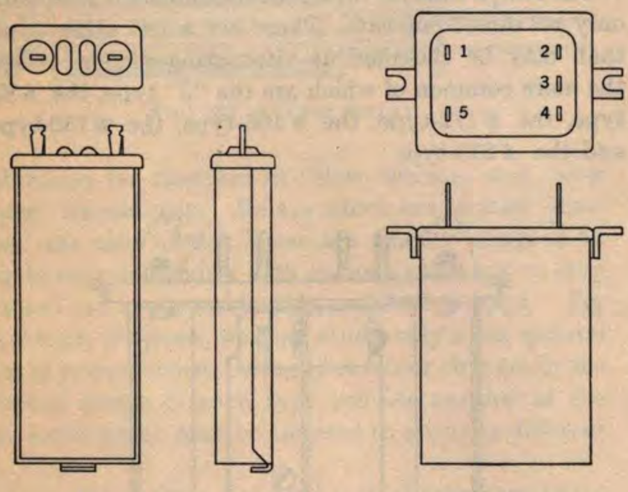


Fig. 136. 218-Type Relay

The "J" type relay is similar in appearance to the "B" type already discussed. The #172 and the #196 relays are not illustrated, but these more nearly resemble direct-current types. The #172 is a toll line ring-down relay and the #196 is used for operating the toll line supervisory lamp of toll cord circuits and in telephone repeater circuits designed for 20-cycle signaling.

The #150- and #218-type relays, the latter of which is illustrated in Figure 136, are polarized relays operating on 135-cycle alternating current and having a split magnetic circuit. A large permanent bar magnet establishes a flux through the cores of two windings, between which is mounted a carefully adjusted reed that forms the return path of the magnetic circuit. This reed is mechanically tuned to have a free frequency of vibration of 135 cycles per second, with the result that a very weak alternating current of that frequency in the relay windings will set it into active motion. The vibration of the reed controls the operation of an associated direct-current relay by opening and closing a pair of contacts connected in series with its windings. The #150- and #218-type relays are used in composite ringer circuits and in connection with telephone repeaters designed for 135-cycle signaling.



A B

Fig. 137. Standard Condensers

All relays used in telephone circuits are designed to operate and release at certain definite values of current in their windings. As these current values are frequently very small, this means that the springs which hold the armatures in their non-operated positions must be adjusted with precision. Every telephone circuit is dependent for its operation upon the proper functioning of relays. In some of these the operating limits are sufficiently liberal to allow considerable margin in adjustments. But in others—and these are frequently the more important ones—the difference between an adjustment giving satisfactory operation and one under which the relay will fail to function properly, may be very small. In practice, specific instructions, giving the exact operating and release current values for which each type of relay should be adjusted for each kind of circuit in which it may be used, are provided.

72. Condensers

Condensers are discussed in some detail in Chapter VIII. There are two principal types of condensers which are extensively used in telephone work—the paper condenser and the mica condenser. Figure 137-A shows an example of the former and Figure 137-B of the latter. Mica condensers are used where the operating voltage is relatively small and a high degree of stability with respect to temperature changes and time is required. The mica condenser illustrated here should not be subjected to more than 200 volts. Paper condensers are designed to withstand much higher voltages with safety. They are ordinarily less stable than mica condensers and their capacity values are naturally less precise. On the other hand, they are more economical.

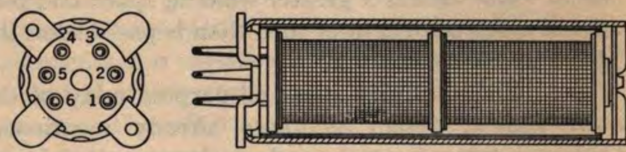


Fig. 138. Straight Core Retardation Coil

73. Retardation Coils

We have already noted the use of retardation coils in power circuits and certain other telephone circuits. In general, it may be said that a retardation coil is used wherever it is desired to add inductance to a circuit. Figure 138 shows an example of the "straight core" type of retardation coil. Coils of this type are used in operators' telephone sets, incoming selector circuits, dial trunk circuits, and so on. The coil designated 54-L in the toll line circuit of Figure 140 is of this type. Figure 139 is an example of the "toroidal" type of retardation coil. This particular type is designed for use in the balancing network of composite ringers where the network is mounted on the relay rack. It has two windings, the rated resistance of each being approximately 202 ohms.

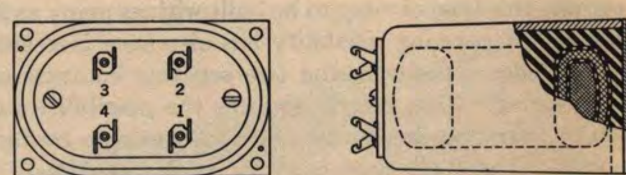


Fig. 139. Toroidal Retardation Coil

74. Typical Long Distance Central Office Circuit

Telephone apparatus such as we have been discussing, is wired together in practice to form a very large number of different types of circuits, each designed to perform some particular function necessary in providing service. As a representative example, let us consider the circuit—or series of circuits—illustrated in Figure 140. This is intended to show schematically the typical circuits required at one terminal of a long distance toll circuit for establishing the connection between a subscriber's telephone and the toll circuit itself. In this example it is assumed that the toll switchboard is of the No. 3 type, and that the toll line circuit is a 4-wire cable circuit. Other types of switchboard, as well as other types of line facilities are, of course, also in use in the telephone plant.

It may be noted that the drawing is separated by heavy dividing lines into three principal parts in which are shown (1) the subscriber's station circuit, (2) the local central office circuit, and (3) the long distance central office circuit. The latter two parts of the drawing are again divided by the open dashed lines into a number of parts, each of which can be considered more or less separately from the other parts in analyzing the circuit. Thus, the circuit units associated directly with the toll switchboard are shown in several blocks in the left part of the large section of the drawing devoted to long distance central office circuits. These separate units include (1) the long distance central office end of the toll switching trunk circuit, (2) the toll line circuit, (3) the toll cord circuit, (4) the operator's position circuit, and (5) the operator's telephone set circuit. The various apparatus units installed in the toll terminal room are shown in a similar way in the group of blocks at the right of the drawing. The drawing includes a large amount of circuit detail, although numerous minor signaling and auxiliary functions are omitted. It should be particularly noted, however, that the drawing is not necessarily a true picture of actual circuit connections at any existing central office. It is intended to be merely representative in a general way and may differ in various respects from the layout in any particular office, which is naturally designed to fit the specific operating requirements applying.

In analyzing the entire circuit drawing, it will be convenient to follow through the completion of a long distance telephone connection, beginning at a point where the outward long distance operator has received an order for placing a call from a telephone subscriber in her city, and is ready to pick up the toll line to the called city. We shall also assume that this is a delayed call and that the calling subscriber is not "on the circuit" when our analysis starts.

The operator's first step is to connect the rear plug of a toll cord circuit to the outward jack of an idle toll line circuit to the called city. This establishes a connection from battery through the supervisory lamp of the Toll Cord Circuit, contacts of the Talk key, and the cord sleeve, to the winding of the B-1025 relay in the Toll Line Circuit. This relay operates but, due to its relatively high resistance (1800 ohms), the current is too weak to light the supervisory lamp. The operation of relay B-1025 is followed by that of the R-1500 relay. This relay in operating closes a contact which connects 24-volt battery to the lead going to the idle indicating lamp circuit. Relays on this circuit (which are not shown in the drawing) then operate to give visual indications that the circuit is busy at all points where it appears in the switchboard multiple. Another contact of the R-1500 relay is opened to lift off from across the line a bridge of 600 ohms in series with a 1 mf. condenser, the purpose of which is to provide the proper termination of the toll line when it is idle. The purpose of the remaining contacts of the R-1500 relay will appear later in the discussion.

The operator now signals on the circuit by operating the ringing key (Toll Cord Circuit) toward the left. The lower contacts of this key break the connection of the tip wire of the cord to the R-857 relay and connect it instead to 24-volt battery through 60 ohms. At the same time, the upper contacts bridge a 600-ohm resistance in series with a 1 mf. condenser across the front cord to give it a proper termination in case it is connected when the rear ringing key is operated. The battery connected to the tip of the rear cord establishes a current through one winding of the 54-L retardation coil and the winding of relay R-1649 in the Toll Line Circuit, which in operating, connects battery to the winding of the R-1721 relay in the 1000-Cycle Direct-Current Terminal Signaling Circuit. The operation of this relay breaks the continuity of the line circuit and connects 1000-cycle ringing current to the outgoing side of the line through the leads designated T and R. One contact in closing also connects ground to the winding of relay E-374 causing it to operate. This opens the signal receiving circuit to the amplifier, thus preventing any possibility of ringing "kick-back". It also closes a contact that bridges a 750-ohm resistance across the output of the 1000-cycle generator, which provides the outgoing line circuit with the proper termination while the ringer is operating.

Now, going to the distant end of the circuit and assuming that the terminal circuit arrangements there are identical with those at the near end, we may follow the effect of the incoming 1000-cycle signaling current. The repeating coil circuit connected across the line in the ringer circuit is designed to offer an attractive path to the incoming 1000-cycle current. It accordingly passes through the windings of this coil, the contacts of the 267-A testing jack, and contacts of the non-operated E-374 relay, to the amplifier. After amplification, the signaling current is sufficiently strong to operate the B-1042 relay. The operation of this relay shorts out the winding of relay 149-CD, which, in releasing, connects ground to the winding of relay E-6035. This relay, in turn, grounds the lead to the winding of relay R-233 and the latter, in operating, connects battery to the signal lead. As a result, a current is set up through the 320-ohm winding of the R-1330 relay in the Toll Line Circuit, causing its operation. One armature of this relay, in closing, operates the busy signals; another closes a circuit from the inward line lamps through contacts of relays R-1648, R-1330, R-994, and R-6015, to the 600-ohm winding of the latter relay, and thence to battery through a contact of the R-1500 relay. This causes the inward lamp signals to light and also operates the relay R-6015, which locks up under control of relay R-1500. Relay R-6015, when operated, connects another battery to the lead to the inward signal lamps and also closes a contact by means of which the R-1330 relay is locked up through its 410-ohm winding under control of relay R-1500.

The inward operator recognizes the lamp signal by connecting one end of a cord circuit to the corresponding inward jack. This permits current to flow from the sleeve, through contacts of the non-operated R-1648 relay, to the winding of the B-1025 relay. The operation of this relay is followed by that of the R-1500 relay, which disconnects the battery from the locked up R-6015 and R-1330 relays, permitting them to release and extinguish the lamp signals. At the same time, operation of the busy signals, now relinquished by the R-1330 relay, is continued by the closing of an auxiliary contact of the R-1500 relay. Another pair of contacts of the same relay opens to lift the 600-ohm terminating bridge from the incoming toll line.

The inward operator answers on the connection by throwing her talk key which, through the resultant operation of the R-857 relay in the toll cord circuit, and the two R-1084 relays, as well as certain other relays in the operator's position circuit to be discussed later, connects her telephone set across the line.

The operator has a standard operator's telephone circuit, which consists of a 528 receiver and a 396A transmitter. The receiver is connected through contacts of the non-operated E-106 relay to the terminals of the #65 induction coil designated T and LT. The transmitter is connected to the primary winding of the same coil, in series with a 24-volt battery and a 54-R retardation coil. The retardation coil tends to steady the transmitter current, thereby reducing noise and compelling the fluctuations (which are in effect a superimposed alternating current) to flow through the 4 mf. bridged condenser instead of through the 24-volt battery. This tends to give the effect of a lower voltage battery, which is preferable for the operator's transmitter in that it prevents both loud clicks and other disagreeable loudness being heard in the receiver as side-tone. The talking current is induced from the primary of the #65 induction coil to the secondary winding, and flows over the connections to the contacts of the closed R-857 relay and thence to the tip and ring conductors of the cord. The 20-G repeating coil and the associated 239-FL relay, shown in the telephone set circuit, are required for the ordinary busy test. Thus, if a trunk circuit is busy when the operator tests it with the tip of her cord, the 239-FL relay will operate. This will cause a rush of current through the 40-ohm winding of the 20-G coil, which will produce a sharp click in the operator's receiver connected across the other winding of the coil.

We now have the outward operator at the near end of the circuit in communication with the inward operator at the distant end of the circuit (assuming, of course, that the outward operator has also operated her Talk key). When the distant operator gets the called party on the line, she throws her talking key to normal. This leaves the procedure in connection with handling the call entirely up to the outward operator at this end, with the exception of taking down the connection at the distant end when the conversation is finished. This is merely a matter of disconnecting the plugs when that operator finds the supervisory lamps associated with her cord circuit burning.

In the meantime, the outward operator at the near end has connected the other end of the cord circuit into the proper jack of the toll switching trunk multiple. This operation closes a sleeve connection through the 1800-ohm winding of the B-199 relay in the outgoing end of the trunk. The B-199 relay then operates to close a circuit through the winding of the B-1009 relay, over the trunk and through the two windings of the 124-F relay in the incoming end of the trunk, operating both relays.

At the same time, the B-operator in the local central office, having been passed the number, connects the plug of the incoming trunk to the proper subscriber's jack in the B-board multiple. When this connection is made, the E-122 relay (300 ohms) of the incoming toll switching trunk circuit is operated by a 24-volt battery circuit through its winding, through the sleeve of the B-board multiple jack, through the cut-off relay (A-26) of the subscriber's line circuit, to ground. The operation of the E-122 relay, in addition to disconnecting the B-operator's busy test, connects a guard and disconnect lamp, associated with the cord of the incoming trunk, to the contact of the 124-F relay, and disconnects it from the contact of the B-15 relay and the 30-ohm winding of the E-126 relay. If the B-board operator finishes her connection to the subscriber's line before the outward long distance operator finishes her connection with the trunk, this lamp will burn but as soon as the outward operator finishes her connection, it will go out. This assures the B-board operator that the long distance operator has plugged into the right trunk.

The reason the guard and disconnect signal goes out is the operation of the 124-F relay, which breaks the lamp connection to ground, and in turn, closes a ground connection to the 30-ohm winding of the E-126 relay but does not operate it because the other side of this winding is open. As we have seen, the 124-F relay is operated because the operation of the B-199 relay in the toll office end of the trunk closed a path from the 24-volt battery at the 124-F relay, through a 500-ohm winding of this relay to the contact of the E-126 relay, through one winding of the 25-S repeating coil, over one conductor of the trunk to the long distance office, through a contact of the E-65 relay and one winding of the repeating coil, to the winding of the B-1009 relay; and thence back by a like path to the other 500-ohm winding of the 124-F relay. The resistances shown in series with the 500-ohm windings of the 124-F relay and designated as "X" are adjusted in value to compensate for different lengths of trunk circuits.

We now have a connection established from the long distance operator to the subscriber's station. The 124-F relay in the local office end of the switching trunk is operated and the B-operator's guard and disconnect lamp is not burning, telling her that the trunk is in use, needs no attention, and must not be disconnected. It will be remembered that the same current that operated the 124-F relay also operated the B-1009 relay in the outgoing end of the trunk. The operation of this relay connects the 85-ohm winding of the B-199 relay, in parallel with its 1800-ohm winding, to the sleeve wire. The resultant reduction in series resistance permits sufficient current to flow to light the cord circuit supervisory lamp, thereby signifying to the long distance operator that she must ring the subscriber. She then operates her ringing key to the front position which connects battery to the tip wire of the cord and so operates the E-65 relay in the outgoing trunk. This relay operates to connect 20-cycle ringing current across the trunk, which causes the operation of the 87-A relay during the interval that the ringing current flows. This is followed by the operation of the E-122 (220-ohm) relay, which connects ringing current to the subscriber's line. Incidentally, this ringing current flows through the contacts of a special key so wired that it can be set to reverse the ringing connection and permit party line ringing from the toll cord circuit. This key is set at the same time the trunk is plugged into the B-board multiple, in case the called number is for a party line.

OUTWARD LONG DISTANCE SWITCHBOARD

We thus have the ringing current properly relayed at the local office. When the subscriber answers, a 48-volt battery current flows through the winding of the B-15 relay, through a 40-ohm non-inductive resistance and one winding of the 25-S repeating coil, over the subscriber's line; and back to ground through the other winding of the 25-S repeating coil and a second 40-ohm non-inductive resistance. This 48-volt circuit is the subscriber's battery supply and is used in connection with toll switching trunks to improve the subscriber's transmission over long subscriber's loops. The 40-ohm resistance in series with the repeating coil prevents the current being too great on very short loops. Neither this resistance nor the winding of the B-15 relay can appreciably weaken the voice current on account of a condenser being bridged between terminals 3 and 8 of the 25-S repeating coil; likewise, the winding of the 87-A relay between terminals 1 and 6 does not weaken the voice current on account of the bridged condenser on the other side of the repeating coil.

When the B-15 relay operates, due to the subscriber taking his receiver off the hook, its armature contact closes the 200-ohm winding of the E-126 relay through a 600-ohm resistance, through the 30-ohm winding of the same relay, to the ground connected to the armature of the 124-F relay. This operates the E-126 relay, which disconnects the 124-F relay from its bridged position across the trunk, but connects one of its windings to ground, thus holding it operated and keeping the guard and disconnect signal from burning. The interruption of the current through the trunk bridge releases the B-1009 relay and thus puts out the cord circuit supervisory signal, thereby notifying the long distance operator to cease ringing because the subscriber has answered. She then, having her talking key depressed, notifies the subscriber that she is "ready on his long distance call". After this she throws the key lever of her talking key in the other direction to the monitoring position. This disconnects her telephone circuit but connects the 94-G repeating coil across the cord and also operates the E-106 relay. The E-106 relay connects the telephone receiver of the head set circuit directly to terminals 1 and 2 of the 94-G repeating coil, thereby permitting the operator to monitor on the circuit.

As soon as the subscriber starts talking, the operator stamps the ticket and dissociates her head set from the connection altogether, unless for some reason it is desirable to continue monitoring.

When the E-106 relay is operated it establishes contacts other than those for connecting the operator's telephone receiver to the 94-G repeating coil. These additional contacts are associated with monitoring taps which connect the operator's circuit with the service observing board, thereby permitting the service observer to listen in on the circuit either when the operator is talking or monitoring.

When the subscriber has finished talking and hangs up his receiver, the B-15 relay of the toll switching trunk circuit is released. This, in turn, releases the E-126 relay and again connects the 124-F relay across the trunk. The B-1009 relay then operates, lighting the supervisory signal, thus notifying the long distance operator that the subscriber has finished. After stamping the ticket, she pulls down the connection to the toll switching trunk, releasing the 124-F relay and in doing so, lights the guard and disconnect signal in front of the B-operator. This time the B-operator knows that the burning lamp means "disconnect" and pulls down the cord. This releases the E-122 relay which lets the guard and disconnect signal again go out, telling the B-operator that the trunk is not in use and no further attention is needed.

At the same time that she takes down the trunk connection, the outward operator rings on the toll line. Since the circuit is now connected at the distant end, the effect of the incoming ringing signal is somewhat different from the former case. The terminal signaling circuit operates in exactly the same way to connect ground to the 320-ohm winding of the R-1330 relay in the toll line circuit. However, because a cord is now connected to the circuit, the R-1500 relay is operated. Accordingly, the operation of the R-1330 relay closes a circuit from ground through the winding of relay B-1020, the non-operated contacts of relay E-661, and the operated contacts of relay R-1500, to the energized sleeve wire, operating the B-1020 relay. The low resistance of this latter relay, now connected in parallel with the 1800-ohm winding of the B-1025 relay, reduces the series resistance of the sleeve wire to about 80 ohms, and so permits the cord circuit supervisory lamp to flash. The operation of relay B-1020 also connects interrupted ground to the winding of the E-661 relay, which then operates intermittently to open and close the connection of relay B-1020 to the sleeve wire. The latter relay is held operated, however, by battery supplied through a closed contact of the E-661 relay to one end of its winding, the other end being grounded through a pair of its own operated contacts. The net result is to cause the cord circuit signal lamp to flash intermittently, and this will continue even after ringing stops until the inward operator either pulls down the cord or answers on the circuit.

The inward operator may answer by throwing the talking key in her cord, which disconnects the sleeve wire from the lamp circuit and connects it to battery through the windings of the B-1022 and B-1023 relays in the operator's position circuit, in series. The combined resistance of these two relays is about 600 ohms, a value sufficiently high to so reduce the current flowing in the sleeve wire that the B-1020 relay, which is marginal, is released. This breaks the connection to the E-661 relay, stopping its action, and also increases the sleeve resistance to 1800 ohms. The B-1022 relay in the operator's position circuit is also marginal and does not operate but the B-1023, and following it, the R-1084, are operated. The operation of the talking key also establishes a circuit through a non-operated contact of the R-1586 relay in the position circuit and the 175-ohm winding of the R-857 relay in the cord circuit, operating the latter. This breaks the direct connection between the two ends of the cord and connects them to the splitting key from which they are connected to the telephone set circuit through closed contacts of the two R-1084 relays, which are both operated if both ends of the cord are connected to jacks.

The operator's position circuit is, as its name implies, common to all the cord circuits at a position. This means that each wire shown in Figure 140 as connecting from this circuit to the cord circuit, is also connected in the same way to every other cord circuit in the position. It may be noted that with the monitoring and talking keys of the cord normal, every one of these wires is open at one end or the other. When the position circuit is connected to a cord by operation of the talking key, the cord may be split for talking in either direction by operation of the splitting key in the position circuit.

Overseas Toll Switchboard

If when the talking key is operated, a ringing signal is received over the line, it will operate the relays in the signaling circuit and the R-1330 relay of the toll line circuit. The latter connects ground through the winding of the B-1020 relay, the non-operated contacts of relay E-661, and operated contacts of relay R-1500 to the sleeve, and thence through contacts of the talking key and the windings of the B-1022 and B-1023 relays, to battery. The current set up is not great enough to operate the B-1020 relay but, due to the connection of the B-1020 relay in parallel with the B-1025 relay in the sleeve circuit, the current is of sufficient value to operate the marginal B-1022 relay in the position circuit. Its operation reduces the resistance of the circuit of the cord circuit supervisory lamp from 1800 ohms to 80 ohms, by connecting an 85-ohm resistance in parallel with the 1800-ohm resistance already connected to the lamp and grounded through a closed contact of the R-1084 relay, and permits the lamp to light.

Similarly, a switch-hook signal from the subscriber, when the talking key is operated, will operate the B-1009 relay in the trunk circuit. This will connect the low resistance winding of the B-199 relay into the sleeve, reducing its net resistance, and so allowing the B-1022 relay associated with that end of the cord to operate and light the other signal lamp.

The position circuit is so designed that if the talking keys of two cords are operated, only the cord whose key was thrown first will be connected to it. This is effected by means of the R-1586 relay which, it will be noted, operates immediately after the operation of the R-857 relay in the cord circuit whose key is thrown first, because a circuit is closed from its winding through a closed contact of the R-857 relay and a contact of the Talk key to ground. This operation of the R-1586 relay opens the ground connection to the 175-ohm winding of the R-857 relay and replaces it with a ground connection through auxiliary contacts of the R-857 relay to its own 700-ohm winding. This holds the R-857 relay in the first cord operated but makes it impossible for the R-857 relay in any other cord to operate even though its talking key is operated, because both windings will be opened, one by its own non-operated contacts and the other by the operated armature of the R-1586 relay.

It is possible, on the other hand, to monitor on two or more cords at the same time by operating the monitoring keys. It is possible also to talk and listen on two cords simultaneously by operating the talking key of one and the monitoring key of the other. In this case, only the cord whose talking key is thrown is connected through the position circuit for splitting or transferring, but the operation of the R-506 relay in the operator's position circuit (which follows the operation of the Talk key) connects the monitoring leads through two 2 mf. condensers to the leads running to the telephone set circuit. Operation of the monitoring key of another cord will accordingly connect the telephone set across the cord.

The position circuit of Figure 140 is arranged for transferring an incoming call from an inward to an outward position. The circuit may also be arranged for transferring from inward to "through" or, by adding several more relays in the toll line circuit, for transferring from inward to both through and outward. The transfer from inward to outward is accomplished by operating the transfer key shown in the drawing. This connects battery through 1000 ohms to the ring wire of the cord, establishing a current over this wire through the 1-2 winding of the 54-L retardation coil and the 370-ohm winding of relay R-1648, to the grounded lead to the switching pad circuit. The resultant operation of the R-1648 relay opens the sleeve connection from the inward jack, thus releasing the B-1025 and R-1500 relays. The simultaneous establishment of a ground connection to the 340-ohm winding of the R-1330 relay, however, causes it to pull up and hold the busy signals operated. At the same time a circuit is completed from battery at an open contact of the R-1500 relay through the 600-ohm winding of relay R-6015 and a pair of its contacts, through open contacts of relay R-994, and through closed contacts of relays R-1330 and R-1648 to the outward lamp signals. Relay R-6015 operates and locks up, connecting another battery to the outward lamps. Its operation also closes a contact which causes the R-1330 and R-1648 relays to lock up through their 410- and 457-ohm windings respectively, under control of relay R-1500. The lamps at the outward position therefore remain lighted until an outward operator plugs into a jack, even though the inward operator restores the transfer key to normal and takes down her cord. When the outward operator connects to the circuit, the resultant operation of the R-1500 relay breaks the battery connection to relay R-6015, which releases, followed by relays R-1330 and R-1648, and extinguishes the lamp signals.

From the switchboard, the line circuit is led through two pairs of jacks at the "patching jack board", the purpose of which is to facilitate quick changes in the inter-connections between line and drop facilities. It is to be noted that while the line circuits at this board may include several different types of layouts, the drop circuits are all identical and may therefore be interchanged in any desired manner. Before it reaches the drop jacks at the patching jack board, the test and control board multiple is bridged to the circuit. The connections to the jacks of this multiple, which is provided solely for testing purposes, are the same as those to the jacks of the toll circuit multiple at the outward switchboard. The make-busy key associated with each jack of the test and control board multiple, is operated by the testboardman whenever he picks up a circuit at this point for testing. Its operation connects ground to the winding of the R-994 relay in the toll line circuit. This relay operates the busy signals, connects busy tone to the sleeve wire, and opens the circuit to the line signaling lamps at the switchboards. It also operates signal lamps at the Plant and Traffic "control boards" through contacts not shown in the drawing. Its purpose is, of course, to prevent any attempts by operators to use the circuit while it is being tested.

Leaving the line jacks in the patching jack board, the circuit is next connected to the "switching pad circuit". This is an artificial line causing a loss of 2, 3 or 4 decibels, which is normally connected in the circuit but is automatically removed through the operation of the R-1124 relay when the circuit is used for a switched or through connection. When two such circuits are connected together, therefore, the net loss of the overall connection is reduced by the sum of the loss values of their respective switching pads. The effect is the same as if a "cord circuit repeater" were used.

As we have already noted, the signaling circuit is bridged across the line between the switching pad circuit and the 4-wire terminating set circuit. This latter is a device for breaking the circuit into two parts, a transmitting and a receiving circuit, each requiring a pair of wires. It consists of two 124-A repeating coils connected in the hybrid coil arrangement to be discussed in a later chapter.

From the terminating set, the transmitting and receiving circuits pass through four-jack circuits in and

No. 8 Test and Control Board

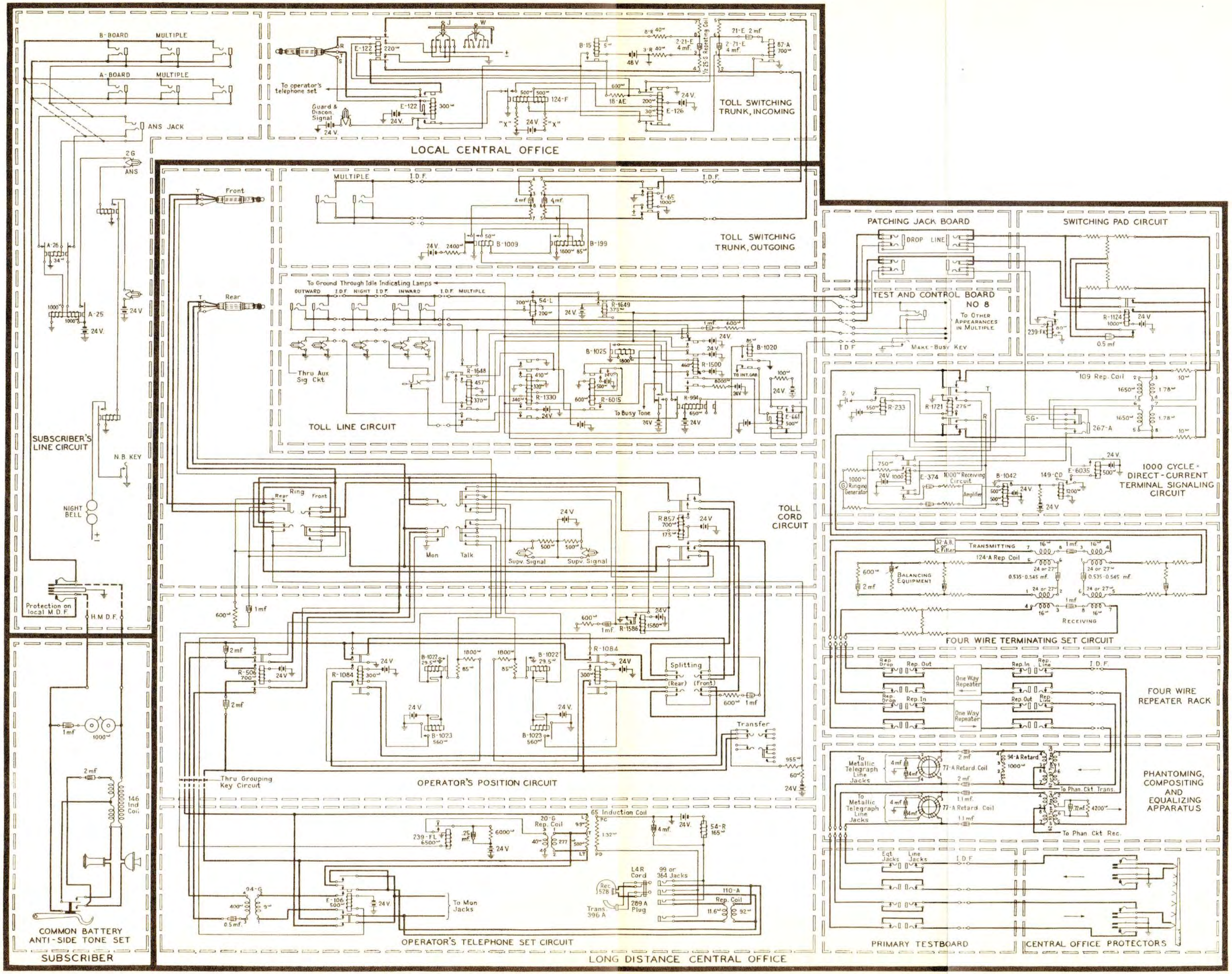


Fig. 140. Typical Long Distance Terminal Circuits—Four-Wire Cable Circuit—No. 3 Switchboard

out of the terminating amplifiers or repeaters, indicated by blocks in the drawing, and from thence to the line equipment. This consists of composite sets, equalizing equipment, and phantom sets. The composite sets shown in this drawing are arranged for metallic telegraph circuits. The equalizing apparatus consists of arrangements of resistances, inductances and capacities connected across the line on the line side of the phantom repeating coil in the transmitting circuit and in series with the drop windings of the repeating coil in the receiving circuit. The purpose of this is to broaden the band of frequencies through which transmission over the circuit will be practically uniform.

For line testing purposes and for patching the equipment, the line circuits are next connected through four-jack circuits at another testboard position called the primary testboard, where a large number of cable pairs may be terminated conveniently since each pair occupies only a relatively small space in the testboard jack panels. From the line jacks here, the circuits are connected to the distributing frame again, and thence through protectors to the toll cable itself.

CHAPTER XI

TELEGRAPH CIRCUITS

75. Means of Obtaining Telegraph Circuits

As a medium of communication, telegraphy depends upon the transmission of electrical signals which are arranged according to some definite code that can be readily translated into a language form by an operator or a machine. In the usual practice, telegraph signals are formed by interrupting or reversing the direction of a continuous current between the sending and receiving stations according to some standard code pattern.

Many telegraph circuits are derived from wire facilities created and used primarily for telephone service, through the aid of apparatus and circuit designs that make possible the simultaneous use of the same facilities for both telephone and telegraph transmission. There are three principal methods commonly employed for accomplishing this result, as follows:

1. "Simplexing" or "compositing" open wire or cable facilities to obtain direct current grounded telegraph circuits.
2. Superposing on open wire facilities high-frequency alternating-current circuits known as "carrier telegraph channels".
3. Superposing on two wires of a cable telephone circuit a very low current "metallic" telegraph circuit.

In addition to the circuits obtained from facilities which are also used for telephone service, telegraph circuits are often secured by the use of a carrier system employing much lower channel frequencies than the open wire system mentioned above. This is known as the "voice-frequency carrier system". Facilities employed for this purpose, however, cannot be used at the same time for telephone service.

It is not practicable here to describe fully all of these methods of obtaining telegraph circuits but we may take up those electrical principles that are fundamental to each and more or less common to all. In general, these are applications of theory already discussed.

It is obvious that where the same wire facilities are used for both telephone and telegraph circuits, some means of separating the telephone and telegraph currents at the line terminals must be employed. The oldest device for this purpose is the simplex set by means of which one grounded telegraph circuit is obtained from the two wires of a "non-phantomed" telephone circuit, or from the four wires of a "phantom group" by applying the simplex to the "phantom" circuit. The simplex principle is illustrated by Figure 141. The telegraph currents cannot interfere with the telephone currents because they divide equally at the mid-point of the line winding of the simplex or "repeating" coil to which each telegraph set is connected. Any change in current value at the "make" or "break" of the telegraph key is not induced into the telephone circuit because the magnetic field established by half of the telegraph current in one-half of the repeating coil winding is exactly neutralized by the field produced by the other half of the telegraph current in the opposite direction in the other half of the same winding.

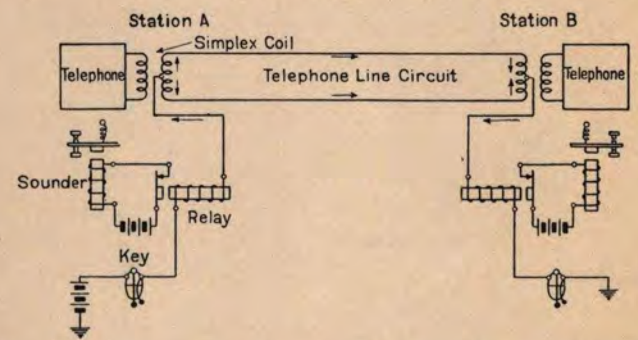


Fig. 141. Telegraph Circuit on Simplexed Telephone Circuit

Referring to Figure 141, the arrows represent the telegraph currents and the total current is shown dividing at the mid-point of the simplex coil line winding at Station A. The two halves join again at the mid-point of the line winding of the coil at Station B. It is imperative that the two line conductors have identical electrical characteristics, including not only equal or "balanced" series resistances but equal capacities and leakages to other conductors and to ground. If the two line conductors are not so balanced, the telegraph current will not divide into equal parts at the mid-points of the simplex coil windings and the larger part will induce a current in the "drop winding" of the coil, which will not be neutralized by the current induced by the lesser part.

Differing radically in principle from the simplex set, the composite set, illustrated by Figure 142, permits a grounded telegraph circuit to be derived from each of the two wires of a telephone circuit, and this without interfering with the use of the telephone circuit as one side of a phantom circuit. Referring to the figure, it may be seen that there are two reasons why the telegraph currents do not interfere with the telephone circuit. In the first place, the inductance of the retardation coil in series with the telegraph "leg", together with the capacitance of three 2 mf. condensers connected in parallel to ground, prevents sudden changes in the telegraph current values, which would tend to be audible as "clicks" in the telephone circuit. The inductance serves here as a "choke coil"; that is, it opposes the sudden building up of the current at the make of the key and retards the rate of decay of the current when the key is opened. The condensers assist the inductance by storing up a small quantity of electricity while the key is closed and discharging this through the inductance when the key is opened. The net result is that the current reaching the line wire changes in value less abruptly than the current at the telegraph key; also, the voltage induced at the break is kept at a relatively low value, thereby preventing a voltage substantially greater than the operating voltage being impressed on the line at that instant.

The second feature of the set that is necessary for keeping the telephone and telegraph signals separated is the 2 mf. condenser in series with the telephone drop, which prevents the direct telegraph currents from reaching the telephone equipment. The bridge across the telephone circuit on the drop side of the series condensers is provided to prevent "crossfire", a condition where telegraph signals sent on one wire of a telephone circuit induce voltages sufficient to interfere with telegraph signals on the other wire, or to operate the signaling relays of the telephone circuit. The bridged arrangement of two 2 mf. condensers in series with the windings of a second retardation coil, which is connected to ground at its mid-point, tends to stabilize the potential of the two line wires by providing a path for unbalance currents to "leak" to ground.

76. Principle of Neutral Telegraph Operation

A telegraph circuit in its simplest form consists of a single wire between two points, equipped at each end with a manual telegraph set consisting of relay, sounder and key. These are so arranged that one set is connected to ground and the other to grounded battery, or both sets are connected to grounded batteries of opposite polarities. Because the operation of the relays is independent of the direction of the current through them, such a system is called a "neutral" system to distinguish it from "polar" systems in which the direction of operation of the relays is determined by the polarity or direction of the current through their windings.

Neutral operation makes use of a flow of current on the line for the operated or "marking" position of the relay armature and zero current for the open or "spacing" position. The line current furnishes the power to operate the receiving relay to marking position while either a spring on the armature or a "biasing" current in another winding of the relay furnishes the energy to operate the relay to the spacing position.

Figure 143 illustrates such a simple neutral telegraph circuit. To analyze its operation, let us assume that the West Station key is closed and the East Station key is open ready for sending. If now the East operator closes his key for only an instant, current flows through

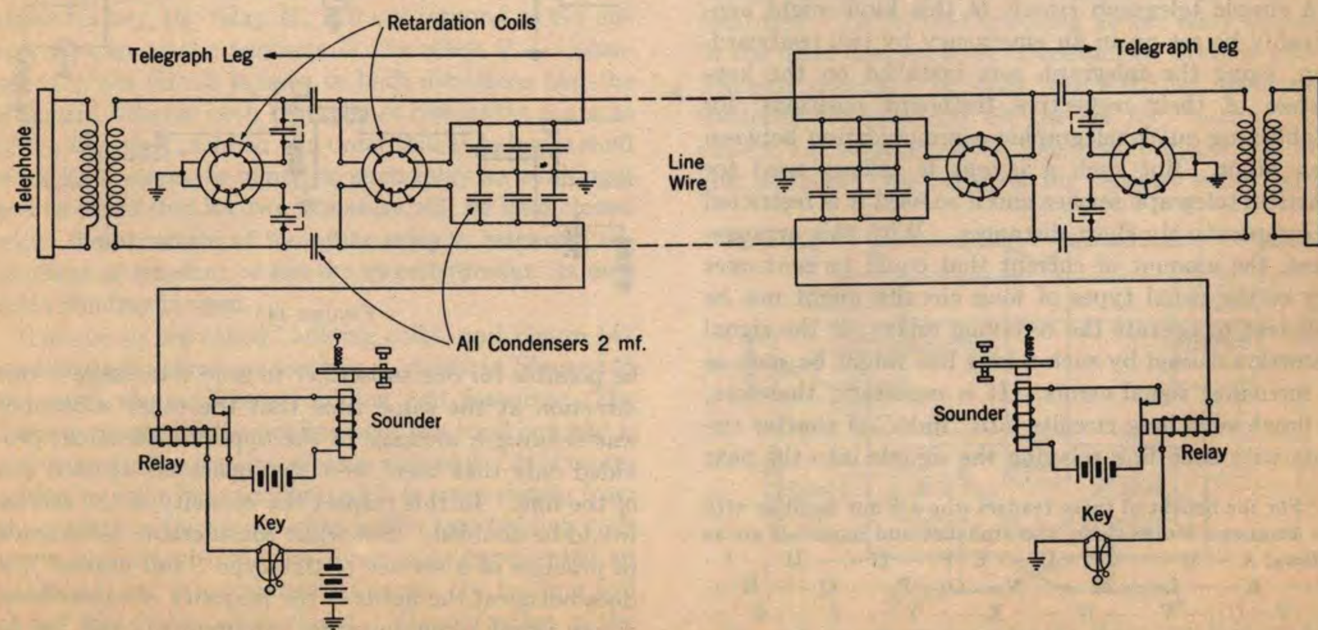


Fig. 142. Telegraph Circuit on Composited Telephone Circuit

both the East and West relay windings in series and both relays operate. This in turn closes the local sounder circuits causing a quick, complete stroke of the sounder lever corresponding to a "dot". If the key lever is held closed for a little more than one-tenth of a second, a longer signal is transmitted giving a greater interval between the up and down strokes of the sounder lever corresponding to a "dash".* If the West Station operator desires to stop the East Station from sending, he "breaks", i.e., he opens his key, thereby opening the circuit and the operator at the East end, noting the failure of his own relay to respond to his signals, knows that the West operator wishes to send to him. He accordingly closes his key by means of the "locking" lever, which short-circuits the contacts of the sending or "non-locking" lever, and the operator at the West Station can then send.

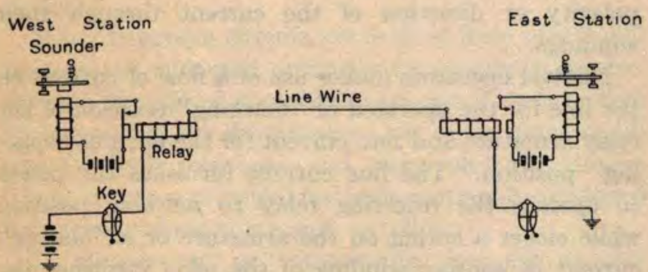


Fig. 143. Elementary Neutral Telegraph Circuit

From Figure 143 it may be noted that **neutral transmission is that form of transmission in which the sending end "impedance" (resistance) changes from some finite value under the marking condition, to infinity in the spacing condition.**

A simple telegraph circuit of this kind might conceivably be set up in an emergency by two testboardmen, using the telegraph sets installed on the keyshelves of their respective testboard positions, for establishing quick telegraphic communication between each other. But such a layout is seldom used for practical telegraph service and if so used it is restricted to comparatively short distances. With this arrangement, the amount of current that could be sent over any of the usual types of long circuits might not be sufficient to operate the receiving relays; or the signal distortion caused by such a long line might be such as to introduce signal errors. It is necessary, therefore, to break such long circuits into "links" of shorter circuits with each link relaying the signals into the next

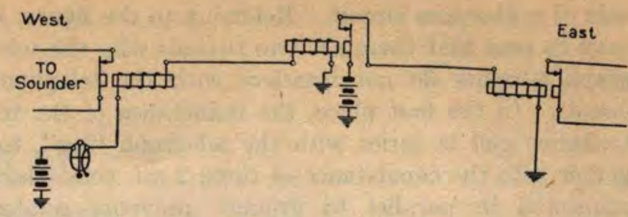


Figure 144

adjacent one. Figure 144 represents a telegraph circuit similar to that shown in Figure 143 but with an intermediate relay at a central point in the circuit. This will permit the West Station to send to a much more distant East Station because the signal is re-energized at the intermediate station by a new battery connected to the contacts of the relay at that point. However, the East Station cannot send to or "break" the West Station; the circuit will work in one direction only. Under these conditions, two separate circuits between the two stations would be required to provide communication in both directions. The two-circuit arrangement is indicated by Figure 145. It involves twice the number of wire facilities and is therefore economically less desirable than a system by which equally satisfactory service could be furnished using only one wire. From the telegraph subscriber's standpoint, the class of service which it would provide would be entirely different from that given by the circuit of Figure 143 and might or might not be preferable to it. With the two-circuit arrangement, it would

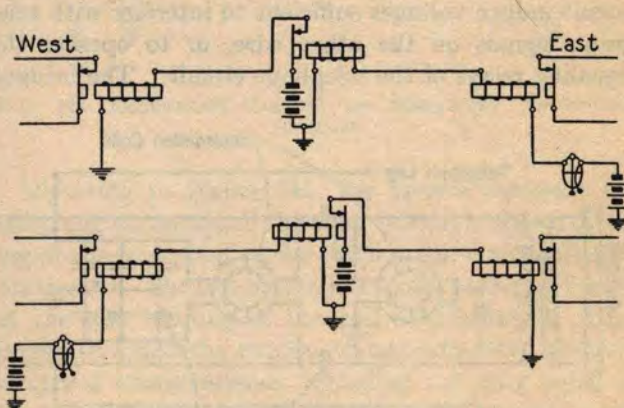


Figure 145

be possible for one subscriber to send a message in one direction at the same time that the other subscriber was sending a message in the opposite direction, provided only that there were two operators at each end of the line. In this respect the capacity of the service would be doubled. But while considerable use is made in practice of a service of this type ("full duplex"), it does not meet the needs of the majority of subscribers, whose usual communication requirements call for an interchange of messages in the manner of a conversa-

*For the benefit of those readers who are not familiar with the American Morse Code, the alphabet and numerals are as follows: A·— B—··· C·· · D—·· E· F·—· G——· H···· I·· J—·—· K—·— L—— M —— N—· O· · P····· Q··—· R· ·· S··· T— U··— V···— W·—— X·—·· Y·· ·· Z··· · &· ··· 1·——· 2··—·· 3···—· 4····— 5 ——— 6······ 7 ——·· 8—···· 9—··— 0——.

tion rather than for continuous simultaneous transmission in both directions.

77. The Single Line Repeater

One solution of the problem of providing two-way service over a long single circuit is the use of the "single-line repeater". The theory of this ingenious device can be best understood by studying its operating features step by step. As has been implied, it is expected to relay energy just as the relays in Figure 145 do, but its operation is restricted to a single circuit and it must permit one operator to break the other. First, let us suppose that two intermediate relays are connected

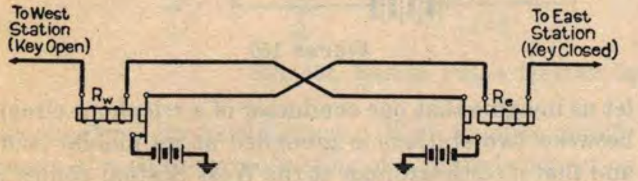


FIGURE 146

into a single circuit as shown in Figure 146, with the winding of one relay in series with the contacts of the other and vice versa. Although a step in the right direction, such an arrangement is not by itself sufficient to effect the desired result. For let us assume as a test that the West operator starts to send a message to the East operator. He opens his key, which at the intermediate point lets the armature of the relay designated as R_w fall back and open the circuit east. This will result in the armature of the relay designated as R_e falling back and again opening the circuit west, which is already open at the key. If now the West operator closes his key, the relay R_w will not respond as the circuit is open at the contacts of the relay R_e. Consequently, the circuit is open in both directions and the closing of either or both keys cannot restore the contacts of the R_e and R_w relays. In order that this device shall work **it is necessary to add to each relay an additional coil so wired that its own armature will be held closed while the armature of the other relay is released, regardless of whether or not the circuit through its own main winding is open.**

These coils are called "holding coils" and Figure 147 represents the same connections as shown in Figure 146 but with the additional holding coil features. The battery circuit for the holding coils is a local one and is not connected to the line wires in any way. It is represented by light lines to distinguish it more clearly from the main line telegraph wires. The two holding coils are in series and each line relay is equipped with an additional set of contacts that shunt the holding coil of the other relay when closed. The operation of the repeater is now as follows: As before, let us assume that

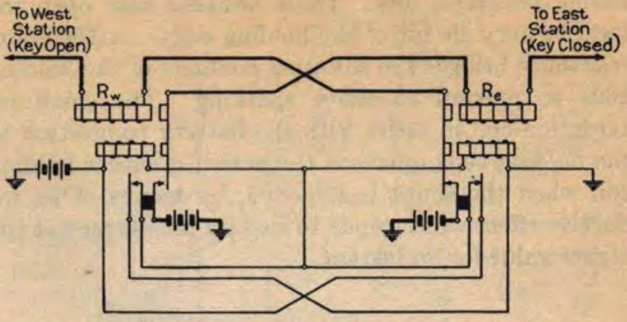


FIG. 147. SIMPLIFIED SINGLE LINE REPEATER CIRCUIT

the key at the west end of the line is open and that the main line contacts of the corresponding relay R_w of the repeater are open. The key at the distant end of the east line is assumed to be closed, i.e., we are assuming for the time being that a signal is being transmitted from west to east. This can now be accomplished because the holding coil of the R_e relay is not shunted and will not permit its armature to fall back and open the west line when the signal is repeated from the west line into the east line by means of the R_w relay's armature. If the East operator desires to break while the West operator is sending, he merely opens his key. As the West operator continues to send, the next signal that closes his circuit and so shunts the holding coil of the R_e relay, will render this holding coil inoperative and permit the R_e armature to fall back. Likewise if the East operator is sending to the West operator and the latter desires to break, the circuit will operate in exactly the same manner in the reverse direction.

A standard telegraph repeater set of this type is shown by Figure 148. Here a few other features of the circuit are shown that were omitted in Figure 147 for clearness. To prevent sparking at the relay contacts in the main line, each set of contacts is bridged with a 300-ohm resistance in series with a 0.5 mf. condenser. Switches are provided on the set for "cutting" the circuit, i.e., separating the line east from the line west and using the two halves of the repeater set as termi-

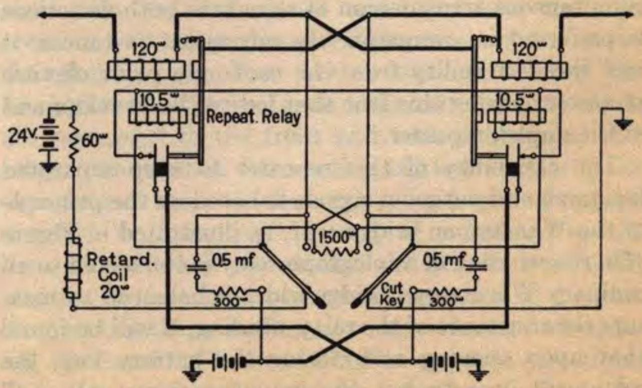


FIG. 148. STANDARD SINGLE LINE REPEATER EMPLOYING NEUTRAL RELAYS

nating telegraph sets. These switches also open the local battery circuit of the holding coils. A 1500-ohm resistance bridges the shunting contacts of the holding coils to prevent excessive sparking. The small retardation coil in series with the battery connection to the holding coils quickens the action of either holding coil when the shunt is removed, by means of its inductive effect which tends to sustain the current at the higher value for an instant.

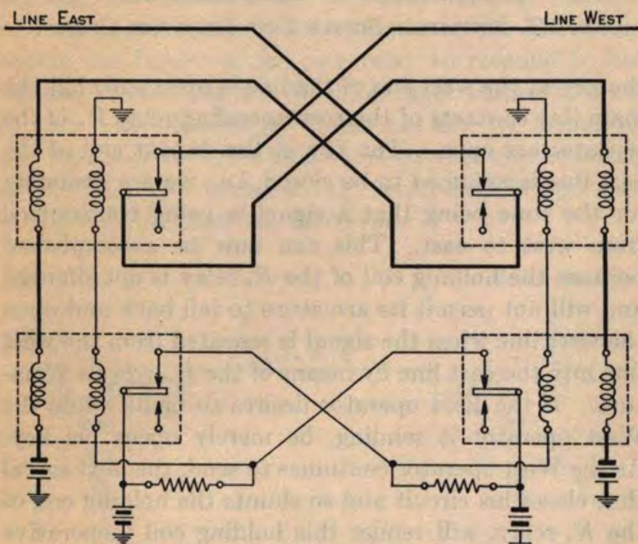


Fig. 149. Single Line Repeater With Polar Relays

Another design of single-line telegraph repeater is illustrated by Figure 149. In general principle it is not radically different from the set that we have been discussing but it employs "polar relays" instead of neutral relays. Its operation will be better understood after studying the discussion of how relays of this type are used in differential duplex systems as given in Article 81.

78. Principle of Polar Duplex Operation

While a type of telegraph service that permits the simultaneous transmission of signals in both directions is preferred in comparatively infrequent instances, it was the desirability from the economic point of view of accomplishing this feat that led to the development of the duplex repeater.

The capability of this repeater to keep separated incoming and outgoing signals is based on the principle of the Wheatstone bridge. If, as illustrated in Figure 150, the winding of a telegraph relay is connected to an ordinary Wheatstone bridge which is balanced to measure the resistance of the relay winding, it will be found that upon opening and closing the battery key, the relay will operate but the galvanometer needle will remain stationary because the bridge is balanced. Now

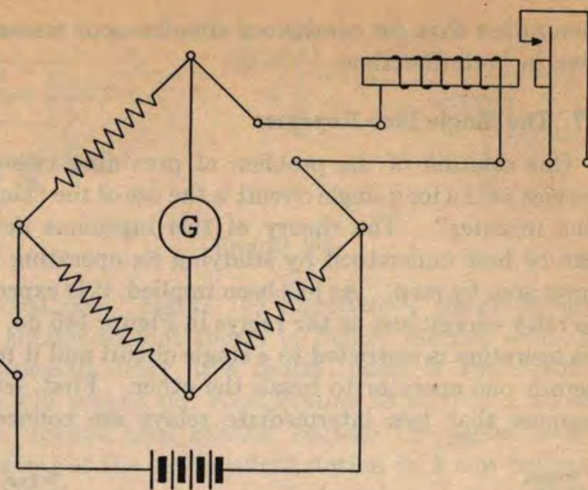


Figure 150

let us imagine that one conductor of a telephone circuit between two stations is grounded at its middle point and that a testboardman at the West Station connects his Wheatstone bridge to the circuit to locate the ground by the Varley method. Further, let us suppose that the testboardman at the East Station instead of crossing the circuit for this test, connects his Wheatstone bridge to the circuit at the same time in order to locate the ground from his end. The connections will then be those shown in Figure 151. When the testboardman at the West Station adjusts the value of the variable resistance in his bridge to equal that of the line and the East Station bridge considered together as a complex network, the bridge will be balanced and no current will flow through the west galvanometer due to the battery at that station.

It is furthermore conceivable that the East Station testboardman might at the same time balance his bridge. Then, with the battery keys closed at each station, a current will flow in the East Station galvanometer due to the West Station battery and an equal current will flow through the West Station galvanometer due to the East Station battery, but no current will flow in either galvanometer from the battery at the same station. If either bridge battery key is opened, however, this condition will be upset because opening the key at one station will destroy the balance at the other and result in current flowing through its galva-

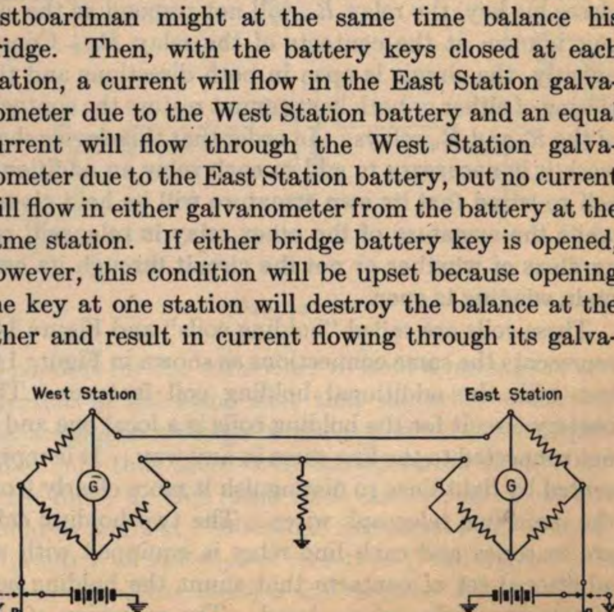


Figure 151

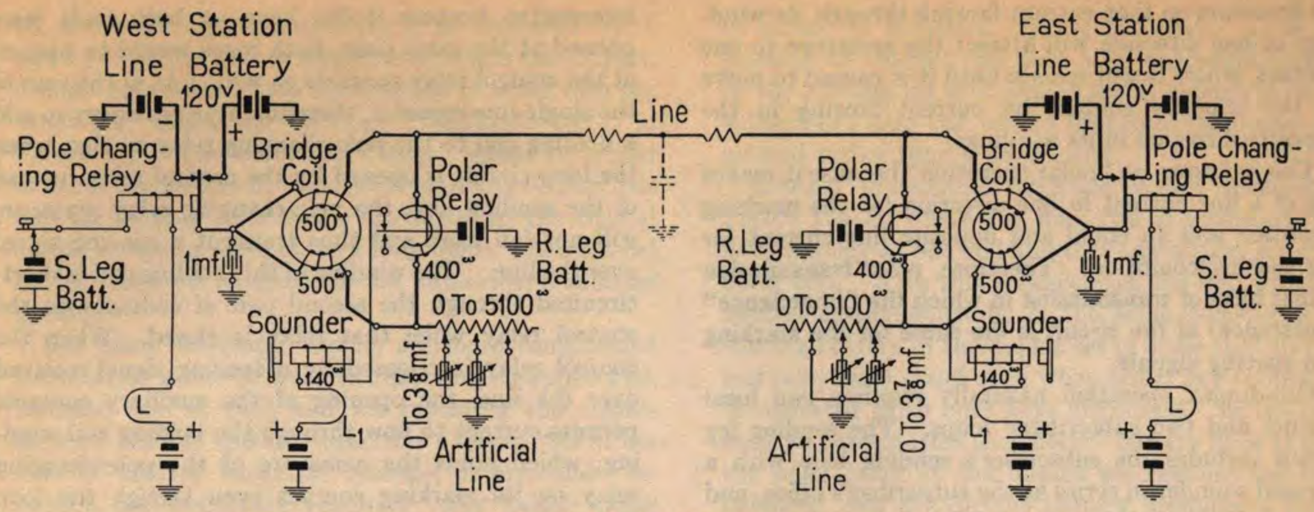


FIG. 152. BRIDGE POLAR DUPLEX SETS ARRANGED FOR FULL DUPLEX SERVICE

nometer from the local battery. With a simple modification of each bridge consisting of a ground connection to the back contact of the battery key (which will not interfere with its use as a bridge), balance will be maintained regardless of the position of the battery keys and the current in either galvanometer will be the same when the associated battery key is open as when it is closed. But opening the key at either station will cause the galvanometer current at the other station to fall from some finite value to zero.

Thus with the two bridges balanced, we have a condition where each station has control over the galvanometer at the other station but has no control over its own galvanometer. Consequently, the battery keys at both stations may be opened and closed at the same time and the operation in one direction will not interfere in any way with the operation in the other direction. It follows that if the galvanometers are now replaced with sensitive relays, and telegraph keys are substituted for the battery keys, the two stations can transmit telegraph messages to each other simultaneously. As in the arrangement pictured in Figure 145, however, one station can break the other only by sending a special break message over the channel transmitting in the opposite direction. The service provided by such a circuit arrangement is called "full duplex". Figure 151 illustrates the principle of the duplex system of operation; but in practice, instead of employing two line wires with a ground at the midpoint of one, the second or defective wire is replaced with an artificial line to ground at each station. These artificial lines take the place of the variable arms of the Wheatstone bridges as well as the resistance of the grounded wire.

79. Bridge Polar Duplex Systems

Figure 152 is a simplified schematic drawing of a telegraph circuit equipped at each end with a "bridge polar" duplex set arranged for full-duplex service. Here are shown the more fundamental modifications that adapt the Wheatstone bridge principle to practical telegraph service. Instead of the bridge arms consisting of simple non-inductive resistances, the two balanced windings of a retardation coil are employed. Each winding has a resistance of 500 ohms and an inductance of three to four henrys, but the two windings in series, i.e., considered as a shunt around the bridged receiving relay, have a very much larger inductance. The artificial line consists of a 0–5100-ohm rheostat variable in steps of 10 ohms which can be grounded, first, directly at any point, thereby simulating the resistance of the line grounded at the distant end or through any intermediate shunt leakages; and second, through two variable capacities at two other points, thus simulating the capacity of the line wire to ground. Together the adjustments are such that the artificial line may be made almost identical electrically to the actual line with another set at its distant end.

The telegraph key, instead of being inserted in the main battery connection to the bridge, is placed in the local "sending leg" circuit which includes the subscriber's transmitting loop and the winding of a "pole-changing" relay. The current transmitted over the line is furnished by batteries of opposite polarity which are connected to the front and back contacts of this relay. Thus when the sending leg key is closed, the armature of the pole-changing relay is pulled up and negative or "marking" battery is connected to the line; with the key opened and the armature released, positive or "spacing" battery is connected to the line. Accordingly there is a flow of current during both spacing and marking intervals, but in opposite directions. This requires that the receiving relay be polarized, i.e., that it have a split magnetic circuit completed through its armature so that current flowing through its windings in one direction will attract the armature to one contact, where it will remain until it is caused to move to the opposite contact by current flowing in the opposite direction in its windings.

This is known as "polar operation" because it makes use of a line current in one direction for the marking condition and an equal and opposite line current for the spacing condition. Therefore, **polar transmission is that form of transmission in which the "impedance" (resistance) of the circuit is the same for the marking and spacing signals.**

Full-duplex operation naturally requires two local circuits and two subscribers' loops. The sending leg circuit includes the subscriber's sending loop with a key and sounder in series at the subscriber's office, and a key and the winding of the pole-changing relay at the duplex set. Two batteries poled to aid each other are connected to the two ends of this circuit. Similarly the "receiving leg" circuit, which is connected to the operating contacts of the receiving polar relay, contains a sounder at the duplex set and a sounder and key at the subscriber's end of the receiving loop. As in the case of the sending leg circuit, grounded batteries aiding are connected to the two ends of the circuit.

The bridge polar duplex set may also be used as a "half-duplex" repeater on an "ordinary" telegraph circuit where the simultaneous transmission of messages in both directions is not desired, providing that a feature whereby one operator may break the other is incorporated. In practice any standard set may be quickly converted from a full-duplex to a half-duplex repeater, or vice versa, by the operation of certain switches. When arranged for half-duplex service, the essentials of the circuit are as shown in Figure 153. Here there is naturally only one loop to the subscriber's station since he will never be sending and receiving at the same time. The winding of the sending relay is connected in series with this loop as it was in the sending loop of the set when arranged for full-duplex service, but the receiving leg circuit instead of being connected into the loop, is connected in series with the winding of an additional relay known as the "control relay". One of the two sets of contacts with which this relay is equipped is in series with the loop. Received signals cause the operation of the receiving polar relay, followed by that of the control relay, which in turn opens and closes the loop in accordance with the incoming signals. For sending, the subscriber opens and closes the loop with his key which, assuming the control relay to be operated as a result of the key at the distant end being closed, operates the pole-changing relay and connects positive or negative battery to the line.

Under these conditions, however, and without any additional features, the circuit would be practically inoperative because if the keys at both ends were opened at the same time, both loops would be opened at the control relay contacts as well. As in the case of the single-line repeater, therefore, it is necessary to add a holding coil to the pole-changing relay so that when the loop circuit is opened by the control relay instead of the sending key, the pole-changing relay armature will not fall back and thus transmit a spacing signal over the line. The winding of this holding coil is short-circuited through the second pair of contacts on the control relay when that relay is closed. When the control relay is released by a spacing signal received over the line, the opening of the auxiliary contacts permits current to flow through the holding coil winding, which holds the armature of the pole-changing relay on its marking contact even though the loop circuit through its main winding is open.

Thus when a subscriber is receiving and has his key closed, the pole-changing relay will be held steadily on its marking contacts, while the loop is opened and closed at the control relay in accordance with the incoming signals. Now if this subscriber wishes to break, he opens his key. If at that instant a spacing signal is being received, this will have no effect because the loop is already opened at the control relay contacts and the pole-changing relay is held operated by the holding coil. But as soon as the distant station sends a marking signal, the holding coil windings will be shunted out and the pole-changing relay will release, thus transmitting a spacing signal back over the line to stop the distant station from sending. If the distant station happens to be sending a series of rapid dots when the near station breaks, the action of the control relay armature may be so rapid that the armature of the pole-changing relay will not have time to fall back during the short intervals that the holding coil winding is shorted out. To insure positive breaking action in such a case, an auxiliary pair of contacts on the pole-changing relay are provided and connected in series with the winding of a "repeating sounder". Then even though the control relay is not closed long enough to allow the armature of the pole-changing relay to fall back to the spacing contact, it will at least leave the marking contact, which will open the circuit through the repeating sounder, causing it to release. The release of this sounder short-circuits the contacts of the polar receiving relay thus locking up the control relay, which results in the pole-changing relay armature moving to the spacing contact and so transmitting the break signal.

Thus far we have considered the duplex set circuit only as a terminal set for repeating signals between a line wire and the subscriber's loop. At the terminal points, as we have seen, the subscriber's loops for full-duplex service are connected in series with the sending leg and receiving leg respectively, as indicated by L and L_1 of Figure 152; or when the service is half duplex, a single subscriber's loop is connected in series with the single leg, called the "dummy", indicated by L in Figure 153. The subscriber's loop consists of a pair of cable conductors or other local subscriber's facilities and all battery connections are made in the central office (except in cases of very long loops, when a local battery may be required for the subscriber's sounder circuit).

But many telegraph circuits are more complicated than merely a wire between two stations, equipped with terminal sets at each end for repeating the signals between the line and the subscriber's loops. As in the case of neutral telegraph systems, intermediate repeaters are frequently required. Any bridge polar duplex set can be used as half of an intermediate repeater. When so used for full-duplex service, it is only necessary to connect the sending leg of one set to the receiving leg of the other set and vice versa as illustrated by Figure 154-A. The half-duplex repeater, on the other hand, is made up by connecting the sending legs of the two sets in series as shown in Figure 154-B. In half-duplex service it frequently happens that the layout consists of a number of branches radiating from certain repeater points, instead of a single direct circuit between two stations. In this case it is necessary for all of the sets connected to branch lines at any one station to have their sending legs connected in series, as indicated by Figure 154-C. It is also possible to connect one side of a single-line repeater to the sending leg of a duplex set as though it were a subscriber's loop and work a branch of the layout on a neutral rather than a polar basis.

Another practicable telegraph layout is shown in Figure 154-D, where the line wire is connected to a neutral relay at one end and to a half-duplex set at the other end. With this arrangement the battery at the neutral station must be of the same polarity as the spacing battery of the duplex set in order that the neutral relay shall stand open when either key is open and be operated when both keys are closed. In the latter case, no current from the "home" battery flows through the polar relay at the duplex set because the set is balanced, but the distant battery produces a flow of current in the polar relay in such a direction as to hold its armature to the marking contact. When the key at the neutral station is opened, the duplex set balance is "upset" and current from the home battery flows through the polar relay in the opposite direction, causing the armature to move to the spacing contact.

It is not only possible to operate a neutral station in conjunction with a half-duplex set in the manner just described, but also at an intermediate point on a circuit equipped at both ends with a half-duplex set, as illustrated in Figure 154-E. It is necessary for such operation, however, that the marking and spacing batteries at one of the duplex sets be reversed in order that there

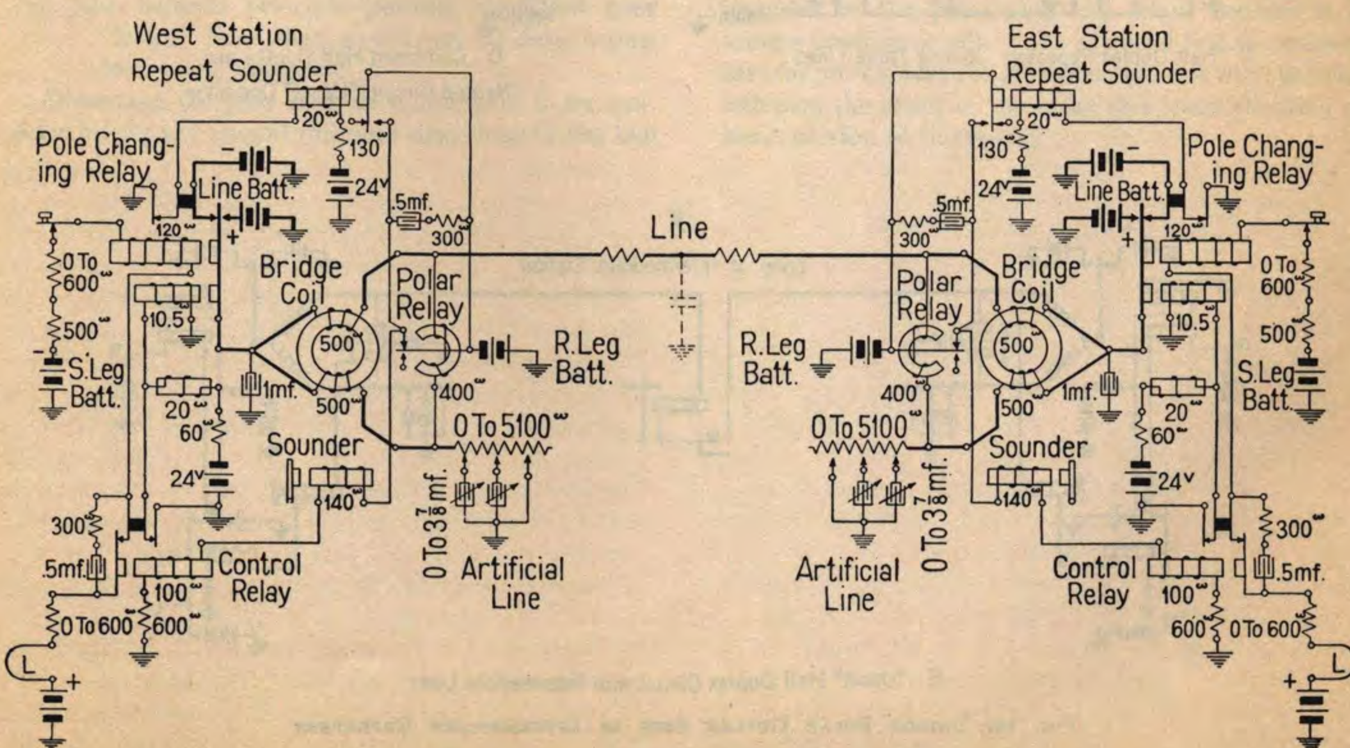


Fig. 153. Bridge Polar Duplex Sets Arranged for Half Duplex Service

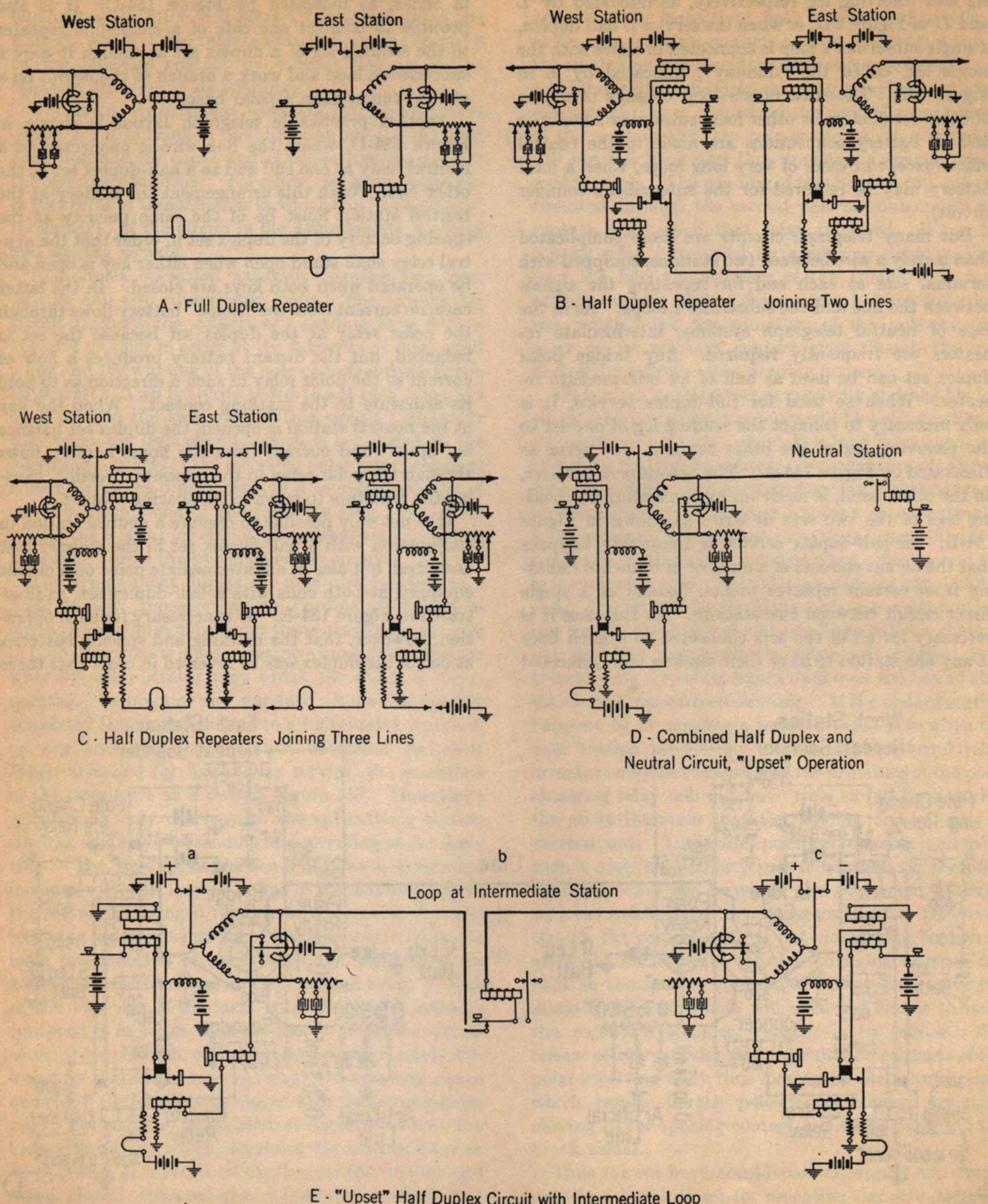


Fig. 154. Bridge Polar Duplex Sets as Intermediate Repeaters

shall be current in the line to hold the intermediate neutral relay closed when the keys are closed at both duplex stations. When either duplex key is opened, the current in the line is reduced to zero because batteries of like polarity are then connected to the line at each duplex set. The neutral relay is accordingly released. When sending from the neutral station, opening the line circuit upsets the balance of both duplex sets, thereby causing both receiving relays to operate to the spacing position.

80. Advantages of Half Duplex Over Neutral System

Since in half-duplex service messages are sent in only one direction at any one time, and this same type of operation is provided by neutral systems, it is natural to inquire why the latter, requiring less equipment, should not be used exclusively. Upon analysis it will be found that half-duplex operation has several marked advantages over neutral operation, the more important of which are briefly as follows:

a. The same set can be used for either full- or half-duplex service, thereby making for maximum flexibility in office layout and keeping the total number of sets at a minimum.

b. The transmission of current for spacing gives the effect of increased voltage without increasing the current values in any part of the apparatus or subjecting the circuit to high working voltages that might be unsafe.

c. The balance principle permits operation over "leaky" lines that would not be satisfactory for neutral operation.

Advantage (b) may best be understood if we consider briefly the magnitudes and directions of the line currents for the two methods of operation. In neutral operation the nominal line current for the marking condition is in the order of 60 milliamperes, and of course, zero current for the spacing condition. As the half duplex operates on a polar basis, the line currents for the marking and spacing conditions are in the order of 30 milliamperes, but flow in opposite directions depending upon whether a mark or space is being transmitted. Therefore, in effect the magnitude of the line currents in the half duplex are only half that in the neutral circuit. Incidentally, it will be noted after studying Articles 91 and 93 of Chapter XIII, covering both neutral and polar "wave shapes", that certain circuit conditions change the signal length during transmission on neutral circuits but have no effect on polar circuits.

For an explanation of (c) above we may refer to Figure 179 of Article 91. If here the East Station is sending to the West Station, there will be a definite current through the shunt, S, and the West Station relay while the key at the East Station is open. This tends to keep the relay energized all of the time and requires that it be so adjusted that its normal release current is appreciably greater than the current that flows as a result of the shunt. In other words, each relay when receiving must work as a "marginal" relay, which makes it very difficult to keep in adjustment when the value of the leakage resistance to ground is varying with changing weather conditions, etc. In half-duplex operation there are no such limitations imposed on the polar receiving relays because it is always possible to adjust the artificial line to compensate for normal leaks to ground on the line wire, thereby reducing the effect of the leaks to a mere shunting of some portion of the energy.

CHAPTER XII

TELEGRAPH CIRCUITS—(*Continued*)

81. Differential Duplex Systems

Although the bridge polar duplex is still in use on many grounded telegraph circuits it is gradually being replaced by a similar device known as the "differential duplex". This has certain advantages that will appear as the discussion proceeds. Its major departure from the older type of set lies in the use of differential polar relays in both the sending and receiving circuits.

The principle of the differential relay depends upon winding a magnetic core with two equal but opposing windings so that if equal currents flow in the same direction through both windings, the magnetic field produced by one winding will be exactly neutralized by that set up by the other winding. Furthermore, by the use of a permanent magnet and a split magnetic circuit, such a relay may be polarized like the receiving relay of the bridge polar duplex set. The magnetic circuit of a typical relay of this type is illustrated in Figure 155.

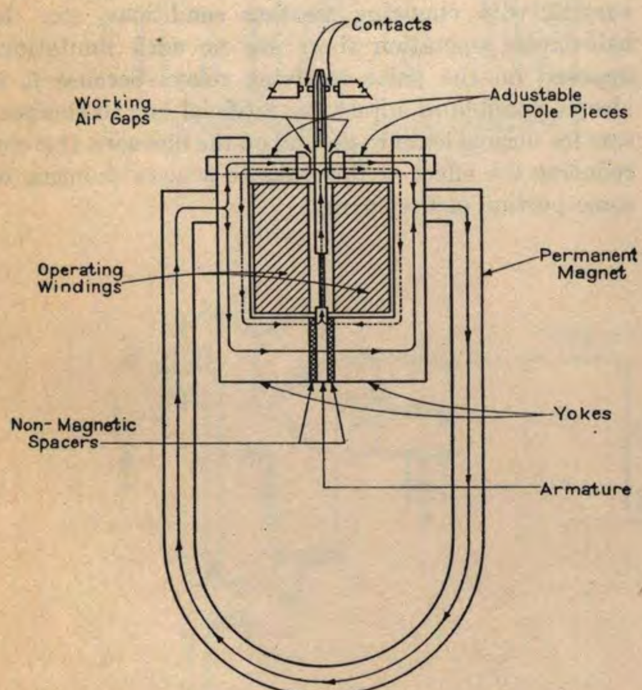


Fig. 155. Magnetic Circuit of the Differential Polar Relay

The manner in which the differential polar relay is employed to repeat signals in duplex telegraph operation may be best understood by referring to Figure 156, which is a schematic drawing of a terminal differential duplex set arranged for full-duplex service. It will be observed that the familiar bridge arrangement of the bridge polar set is here replaced by a differential polar receiving relay, the two windings of which are connected at one end to the real and artificial lines respectively, and have their other ends connected together to the armature of the sending relay. When the artificial

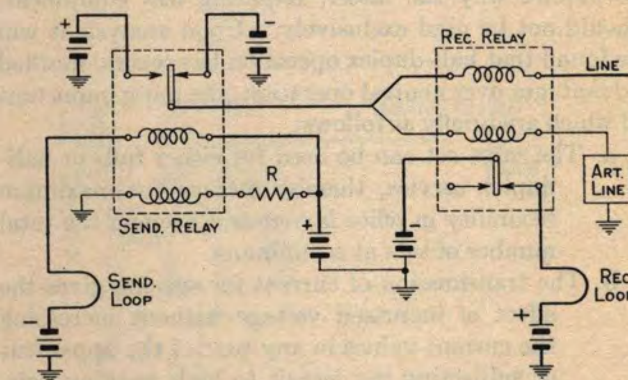


Fig. 156. Terminal Differential Duplex Set Arranged for Full Duplex Service

line is adjusted to exactly balance the real line, currents coming from the sending battery divide equally between the two windings in parallel and, since these are connected differentially, the resultant magnetic flux is zero and the relay is not operated. Current coming from the line, on the other hand, flows through the two windings in series, which produces aiding magnetic fields and causes the relay armature to move to one or the other of its contacts depending upon the polarity of the incoming current. Thus it is evident that there is no interference between the sending and receiving circuits and the two can be operated quite independently of one another—in other words, full-duplex. It would be possible to operate this circuit with a neutral sending relay as in the bridge polar set but in practice a differential polar relay is used for this purpose also.

Referring again to Figure 156, the upper winding of the sending relay is known as the **operating winding** and the lower as the **biasing winding.** With the sending loop key closed, the magnetic fields produced by the two windings are in opposition because current is flowing in the same direction through each. The resistance, *R*, however, is of such value that the current

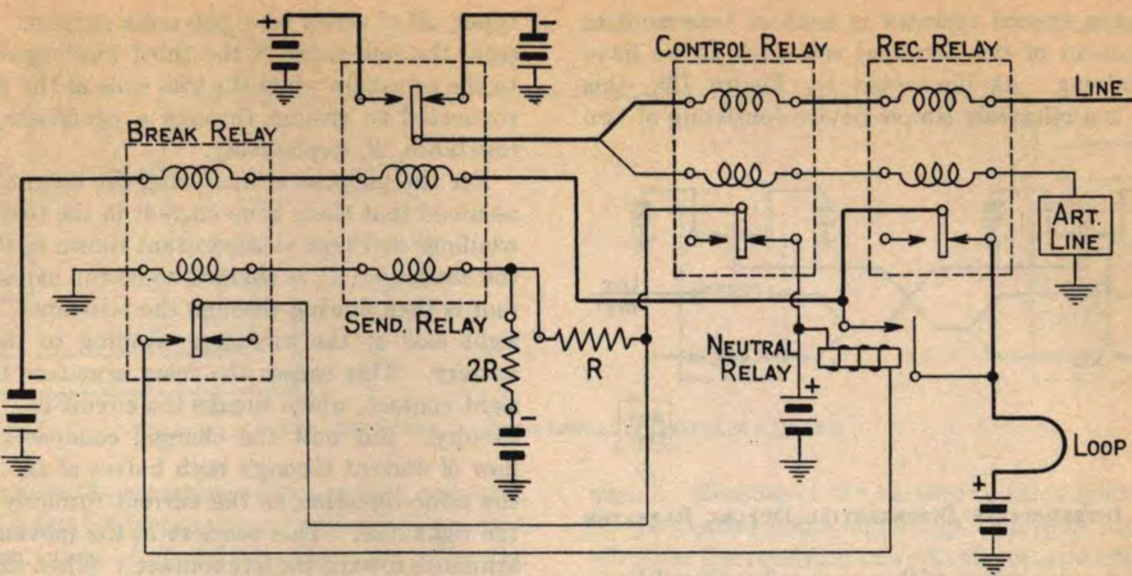


Fig. 157. Terminal Differential Duplex Set Arranged for Half Duplex Service

in the biasing winding is limited to a substantially smaller value (usually half) than that in the operating winding and the armature is accordingly held to the marking contact. But when the sending loop key is opened, only the biasing winding is effective and the armature is drawn over to the spacing contact.

Half-duplex operation necessitates the inclusion of several additional relays in the differential duplex set circuit. The essentials of the circuit under these conditions are shown by Figure 157. The control relay, wired in series with the receiving relay, is provided to prevent the operation of the sending relay when the loop is opened and closed by the receiving relay. Two batteries of opposite polarity are connected in parallel to the biasing winding of the sending relay but the resistances, R and $2R$, are so adjusted in value that when the control relay contact is closed, the positive or spacing battery will be in control and the operation of the sending relay will be identical with its operation in the full-duplex circuit. When signals are being received, on the other hand, a spacing signal opens the loop at the contacts of the receiving relay and no current can flow through the operating winding of the sending relay. Then if it were not for the simultaneous operation of the control relay, the sending relay would be operated to spacing; but the opening of the control relay contacts breaks the circuit to the positive battery and allows current to flow in the opposite direction through the biasing winding and the resistance, $2R$, to the negative battery. This holds the sending relay armature on its marking contact.

An additional polar relay, known as the break relay, is connected in series with the sending relay. Its purpose will be understood if we analyze a condition where signals are being received from the distant station and the local operator wishes to break. To do this he opens the loop circuit with his key. If at that instant a spacing signal is being received, this will have no effect because the loop circuit is already opened at the contacts of the receiving relay. But as soon as a marking signal is received and the control relay armature closes, spacing battery is connected to the biasing windings of the sending and break relays and since there is no current in their operating windings, both relays operate to spacing. The operation of the sending relay of course results in the transmission of the desired spacing or break signal to the line. The operation of the break relay insures that the break signal, once begun, will not be interrupted in case a spacing signal is received from the line. Such a received signal causes the control relay contacts to open and so would permit the sending relay to operate to marking were it not for the second connection between the spacing battery and the biasing windings of the sending and break relays, established through the spacing contacts of the latter relay. The neutral relay in series with this circuit is provided to take care of the possible contingency of the break and sending relays at both ends of a circuit becoming simultaneously operated to spacing. In such a case it would be impossible for either operator to regain control of the circuit because the loop circuits at both ends would be opened at the receiving relays. However, the operation of the neutral relay, which occurs whenever the control relay is opened after the break relay is operated to spacing, short-circuits the contacts of the receiving relay so that the sending and break relays will be operated to their marking contacts when the loop key is closed.

One of the important practical differences between bridge polar and differential duplex systems is that in the latter a special repeater is used at intermediate points instead of two terminal sets such as we have been studying. As illustrated by Figure 158, this repeater is a relatively simple device consisting of two

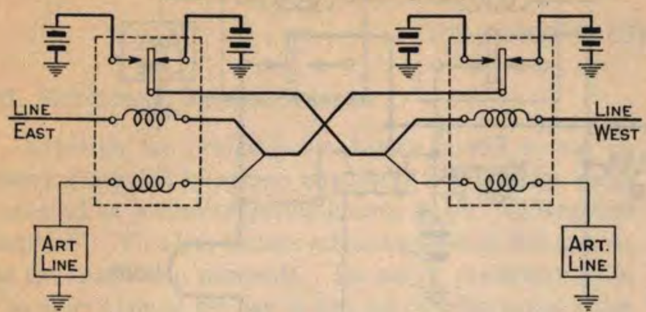


FIG. 158. INTERMEDIATE DIFFERENTIAL DUPLEX REPEATER

differential polar relays with associated artificial lines, by means of which signals are repeated directly from the Line East to the Line West and vice versa. Where a subscriber or a branch line is connected at an intermediate point, it is of course necessary to use terminal sets in much the same way as illustrated in Figure 154 for bridge polar duplex operation. Differential type intermediate repeaters can be used on circuits equipped at their terminals with bridge polar sets and bridge polar intermediate repeaters may be used on circuits equipped at one or both terminals with terminal differential sets.

From the foregoing it will be seen that the differential duplex system offers about the same general possibilities of operation as the bridge polar system. Several of its design features, however, make it more suitable for use on grounded lines than the bridge polar system, particularly where high-speed teletypewriter circuits are involved. The principal advantages are the use of a break relay operating simultaneously with the sending relay, which establishes the break circuit when the latter relay operates instead of after a sounder is released by the sending relay, and the employment of polar relays for transmitting as well as for receiving.

82. Principle of the Vibrating Circuit

Undoubtedly the most important advantage of the differential system lies in the rapid response to signals of the polar relays used. In addition to being inherently much more sensitive than the types of relays ordinarily used in bridge polar sets because of their more delicate design, these relays are equipped with a special third winding which forms a part of a "vibrating circuit" that adds further to their sensitivity and rapidity of response. The principle of the vibrating circuit may be understood by referring to Figure 159, which is a schematic drawing of one of several possible types, all of which have the same purpose. As will be seen, the mid-point of the third winding is connected to the armature while the two ends of the winding are connected to ground through a condenser, C, and a resistance, R, respectively.

For the purpose of analyzing the circuit, it may be assumed that there is no current in the two main relay windings and that at the instant shown in the diagram the condenser, C, is charged to its full capacity. Current is then flowing through the resistance, R, and the right side of the vibrating winding to the negative battery. This causes the relay armature to leave its right contact, which breaks the circuit to the negative battery. But now the charged condenser sets up a flow of current through both halves of the winding in the same direction as the current formerly flowing in the right half. This accelerates the movement of the armature toward the left contact. When the armature reaches the left contact, there is a strong initial current from the positive battery through the left side of the winding to charge the condenser in the opposite direction. This current flows in the same direction in this

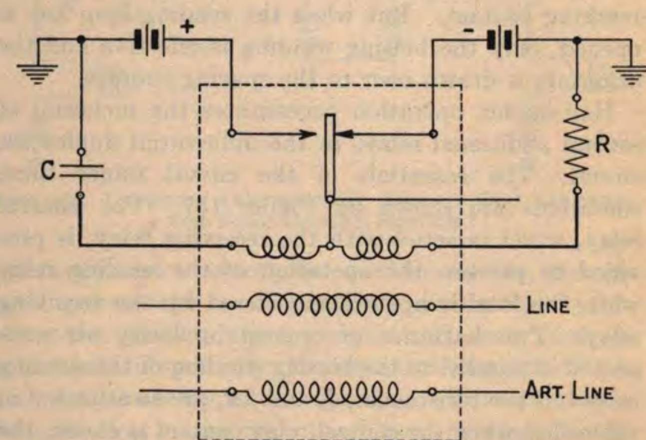


FIG. 159. VIBRATING CIRCUIT OF DIFFERENTIAL RELAY

half of the winding as the current flowing before the contact was made and its effect, therefore, is to hold the armature firmly against the contact without rebound or "chatter". As the condenser becomes charged, the current in the left side of the winding falls off until it becomes smaller than the current flowing in the opposite direction through the right side of the winding and the resistance. The armature then pulls away from the left contact; the current flowing from the condenser, which is now charged in the opposite direction, hastens its travel to the right contact, through which current then flows from the negative battery to again charge the condenser, thus holding the armature solidly against the contact. When the condenser is charged, the cycle is completed and it continues to repeat itself indefinitely.

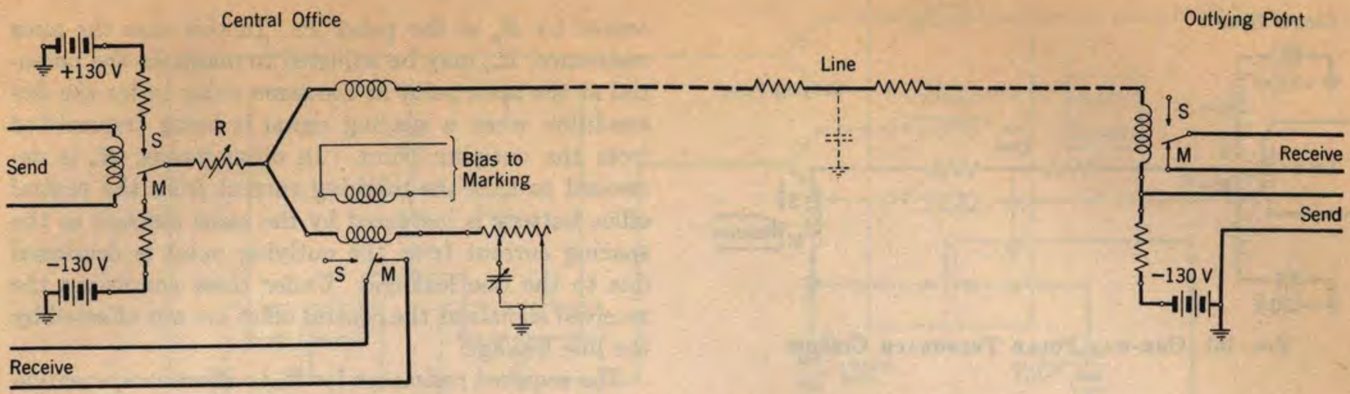


Fig. 160. Type A Polarential Telegraph Circuit

Thus we have the armature vibrating back and forth between the two contacts at a rate which is dependent only on the values of R and C. However, when the main windings of the relay are connected into the repeater circuit in normal fashion, the marking or spacing currents flowing in them will prevent the armature from vibrating freely under the influence of the vibrating circuit. But the tendency to vibrate is nevertheless present and whenever the current in a main winding is reversed due to the transmission of a signal, the vibrating circuit causes the armature to move from one contact to the other a little in advance of the time that it otherwise would. It also causes the movement to be more rapid and the contact to close more positively than would be the case if it were not operative.

83. Polarential and One-Way Polar Systems

For furnishing service to subscribers at outlying points, two special types of grounded telegraph systems known respectively as "Polarential" and "One-Way Polar" are frequently used. The polarential system permits true polar operation from the central office out and a modified polar operation from the outlying point into the central office. Thus the advantages of polar transmission are secured and at the same time the equipment arrangements in the subscriber's office are relatively simple. A schematic diagram of such a circuit is shown in Figure 160, with only the essential elements included for the sake of simplicity.

By inspection of the above diagram it may be seen that with the central office duplex repeater balanced while the outlying sending loop is closed, the transmission from the central office out is true polar. The relay at the outlying point receives signal combinations of equal marking and spacing currents of opposite polarity applied to the line by the sending relay at the central office.

In transmitting to the central office, ground is applied to the line at the outlying point for the marking signal. Because of the balance of the duplex repeater at the central office under this condition, there is no effect on the receiving relay at the central office and it is held in the marking position by the current through the biasing winding. For the spacing signal, negative battery is applied to the line at the outlying point, which produces an effective spacing current in the receiving relay at the central office. This current comes from two sources, the first being due to the current flowing in the relay windings from the negative battery at the outlying point, and the second coming from the home battery as a result of the duplex unbalance caused by the resistance in series with the battery at the outlying point.

The variable resistance R at the central office is adjusted to such a value that the spacing line battery at the outlying point is higher than the potential applied to the apex of the repeater circuit at the central office. This assures that the line current will reverse when a spacing signal is transmitted from the outlying point. This is necessary in the case of teletypewriter operation to assure getting "home copy" at the outlying point.

Service conditions are sometimes such that only one-way transmission from the central office to an outlying point is required; as for example, in the transmission of news copy in certain cases. In such situations a somewhat simpler arrangement, known as the "One-Way Polar" system, is used. This also gives the advantage of polar operation but does not include the duplex feature at the central office. The essentials of this circuit arrangement are shown in Figure 161.

84. Leakage Compensation System

A somewhat novel method of operation employing polar transmission is shown schematically in Figure 162. The important advantage of this method of operation is that the circuit is self-compensating to a considerable extent for line leakage. This system is also sometimes called "Polarential (Type B)" in which

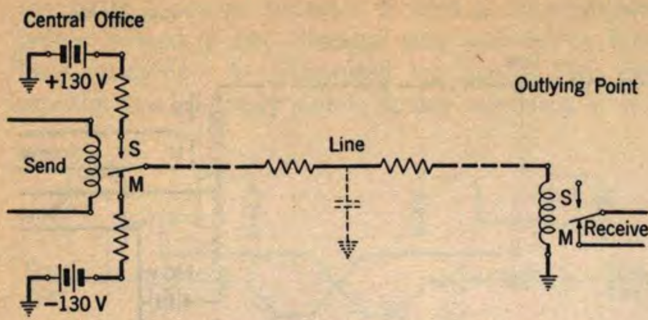


FIG. 161. ONE-WAY POLAR TELEGRAPH CIRCUIT

case the polar system discussed in the preceding article is designated "Polarential (Type A)".

It may be noted that true polar transmission is employed in sending from the central office to the outlying point. When sending at the central office, the receiving relay is held on its marking contact by the biasing current as shown. Transmission in this direction is therefore the same as in the polarential method of operation.

When transmitting from the outlying point to the central office on a **dry line**, the marking line current has no effect on the receiving relay at the central office due to the balancing network precisely balancing the line. For the spacing signal, aiding battery is applied to the line at the outlying point which causes an effective spacing current, $\frac{E}{R_L}$, to flow in the receiving relay, where E is the potential of the outlying battery and R_L is the resistance of the line (and the artificial line) from apex to ground. The biasing current is adjusted to a value of $\frac{E}{2R_L}$. With this set-up the signals sent at the outlying point will be satisfactorily received at the central office. When the line is **wet** there will be a comparatively large leakage to ground which may be represented by R_g at the point P. In this case the apex resistance, R_a, may be adjusted to maintain the potential at the apex point at the same value as for the dry condition when a spacing signal is being transmitted from the outlying point. In other words, R_a is decreased so that the marking current from the central office battery is increased by the same amount as the spacing current from the outlying point is decreased due to the line leakage. Under these conditions the received signals at the central office are not affected by the line leakage.

The required resistance for R_a to effect compensation for the line leakage may be determined from the following equation:

$$R_a = \frac{R_L(2R_2 - R_1)}{2R_1 + R_2} \tag{42}$$

where R_a and R_L are as previously defined, R_1 is the resistance in the line from the apex to the point of leak, and $R_2 = R_L - R_1$.

From inspection of this equation, it is evident that if R_1 is greater than $2R_2$ complete leakage compensation cannot be effected unless the transmitting voltages at the central office are made higher than those at the outlying point. However, since R_2 includes the resistance, R_b, in the outlying point, this latter resistance may readily be made great enough so that this condition will not ordinarily occur.

In sending a spacing signal from the outlying point, the batteries applied to the two ends of the line are aiding and the line is charged to a high positive potential near the outlying point. When the sending relay at the outlying point goes from space to mark, the high charge on the line causes a surge of current to flow to ground through the receiving relay in a direction opposite to that of the normal marking current. This would tend to cause the receiving relay at the outlying

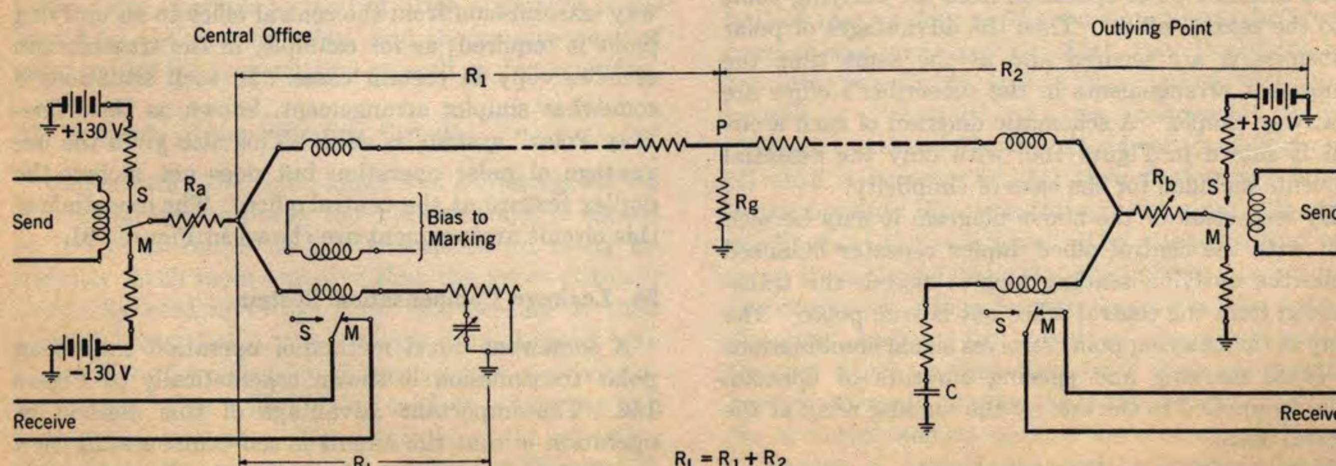


FIG. 162. TYPE B POLARENTIAL TELEGRAPH CIRCUIT

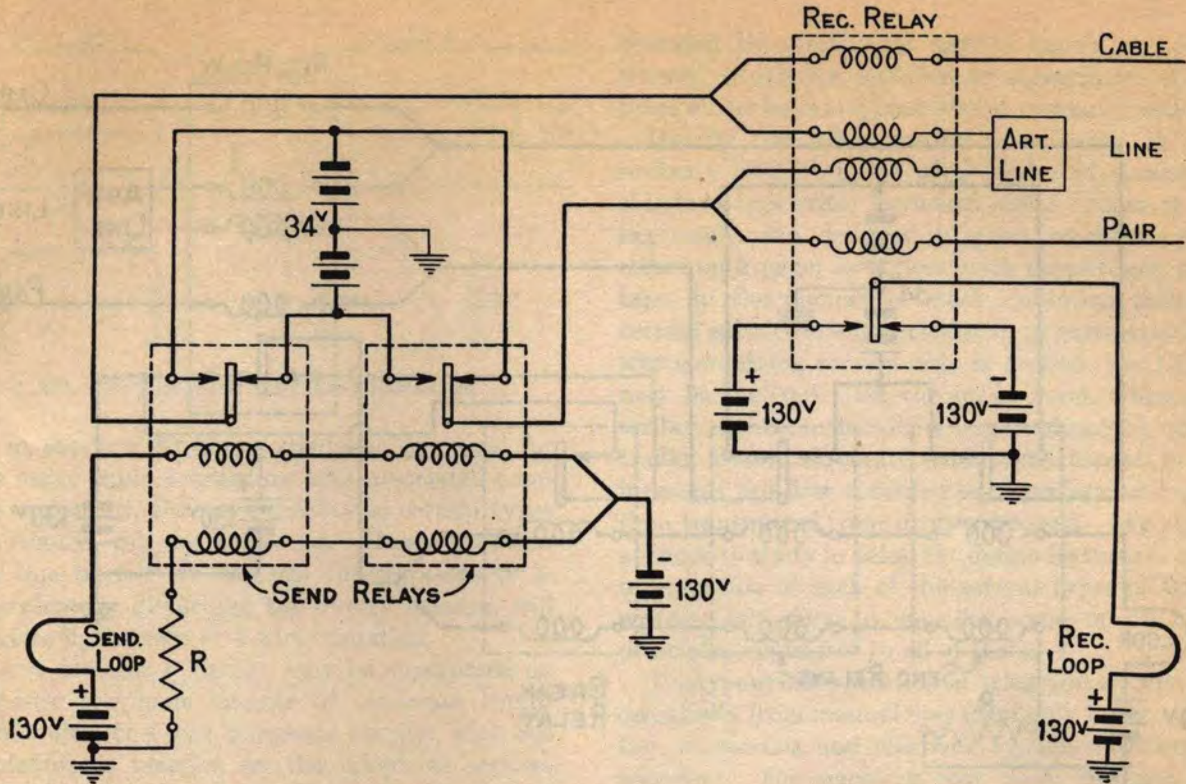


FIG. 163. TERMINAL METALLIC DUPLEX SET ARRANGED FOR FULL DUPLEX SERVICE

point to "kick off" and mutilate the home copy or produce false breaks. To neutralize the effect of this line "kick", the bridge arm containing the condenser C is added. This produces a "kick" affecting the relay oppositely to the line "kick" and of such magnitude that the receiving relay remains steadily on its marking contact when the sending relay is operating.

85. Metallic Telegraph Systems

The operation of telegraph circuits on composited cable conductors usually imposes certain requirements differing from open wire operation. In order to avoid interference with the telephone circuits, it is necessary in the first place that the telegraph currents be limited to values of the same order of magnitude as the telephone currents. Furthermore, in order to eliminate interference from ground potentials and crossfire and also from power circuits, it is preferable to use a second metallic conductor instead of an earth return, as is done in open wire operation. This means that at least two line wires are employed for each telegraph circuit. However, metallic telegraph systems are operated in practice both 2-wire and 4-wire.

The 2-wire metallic cable telegraph system is quite similar to the differential duplex system for grounded lines and, like the latter, employs different types of repeaters at through and terminal points. Figure 163 is a schematic drawing of a terminal metallic set arranged for full-duplex service, with the monitoring connections and all other auxiliary circuit details omitted for clearness. Here it will be noted that polar differential relays and balanced circuits are employed in both the sending and receiving circuits. The receiving relay has four balanced operating windings connected differentially so that when the artificial line is adjusted to balance the real line, there is no interference between incoming and outgoing signals. Polar transmission over the line is accomplished by means of a 34-volt battery which is reversed by the two sending relays to produce the marking and spacing signals. The line current, when the system is operating on 19-gage conductors, is approximately five milliamperes in each conductor of the pair. The sending loop is balanced by the resistance, R, and when the key is closed, the current in the upper or operating windings of the sending relays is exactly twice that in the lower or biasing windings because there are two batteries aiding in the loop circuit as against a single battery in the balancing circuit. Although the magnetic fields set up by the currents in the two sets of windings are in opposition, the preponderance of current in the operating windings holds the relay armatures on their marking contacts. When the loop key is opened, however, only the spacing windings are energized and the armatures are accordingly operated to spacing.

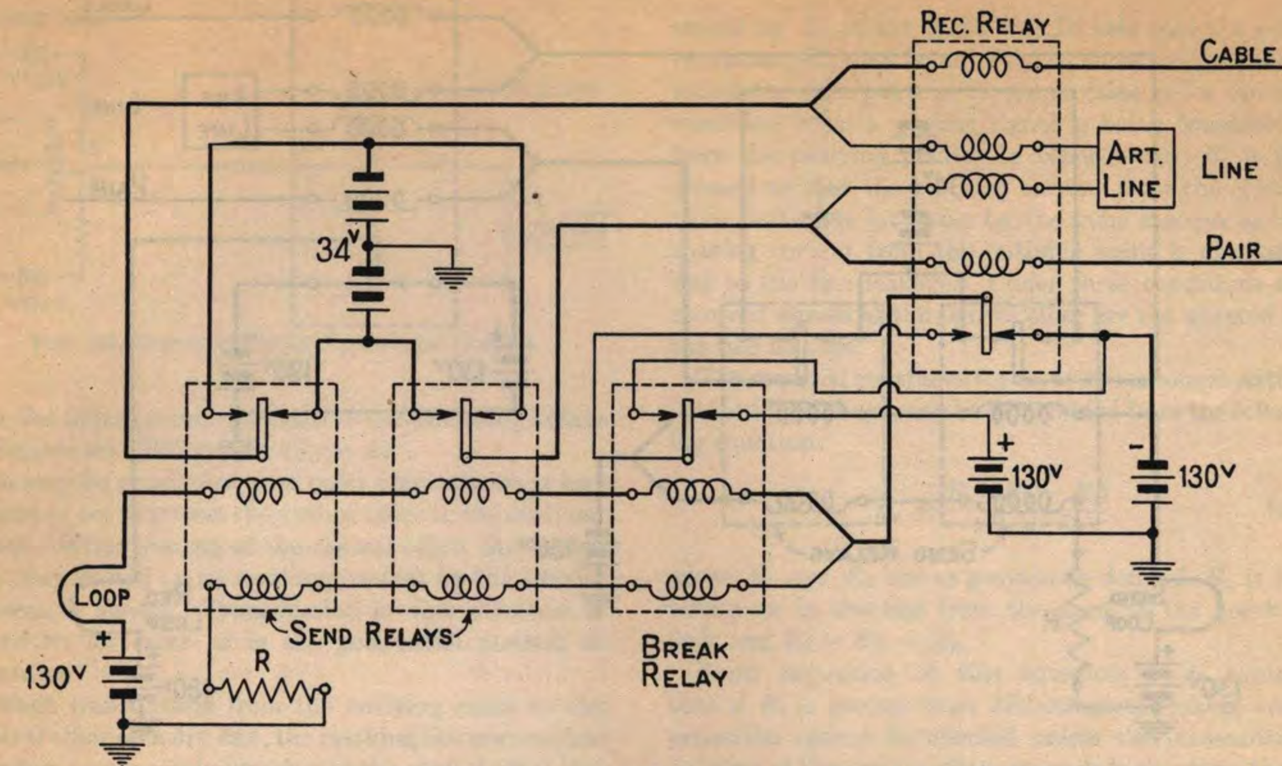


FIG. 164. TERMINAL METALLIC DUPLEX SET ARRANGED FOR HALF DUPLEX SERVICE

Half-duplex operation requires the addition of a break relay to the circuit, as shown in Figure 164. This permits the transmission of a clean-cut break by insuring that the armatures of the sending relays, once shifted to their spacing contacts by the opening of the key, remain so shifted. If the sending and receiving circuits were connected together to the loop without the break relay, opening the key in the loop would cause the sending relays to be controlled by the current flowing from the receiving relay contacts through their biasing windings and the balancing resistance to ground. Then if signals were being received at the time the key was opened for a break, the sending relays would be operated in accordance with the received signals, and these would be transmitted back over the line inverted, instead of the clean-cut spacing signal desired. But the break relay, connected in series with the sending relays as shown, operates simultaneously with the sending relays when the key is opened for a break. The shifting of its armature to the spacing contact connects negative battery to both contacts of the receiving relay so that the sending relay armatures are held on their spacing contacts as long as the loop key is open, regardless of the operation of the receiving relay by signals coming in from the line.

It is of course possible to use two terminal sets for through repeating at an intermediate point where no subscriber's loops are involved. In practice, however, a special type of repeater is used for through operation which is much simpler in design. As shown schematically in Figure 165, this consists merely of two relays with artificial lines associated. In order to avoid the use of four relays instead of two, separate positive and negative 34-volt batteries are used. This is known as "single-commutation", as distinguished from the method of reversing the connections to a single battery in the manner described in connection with the terminal sets, which is known as "double-commutation". The polar relays in both terminal and through sets are equipped with vibrating windings which operate in substantially the same manner as described in Article 82.

In the 4-wire metallic telegraph system different paths are employed for transmission in the two directions, thus avoiding the necessity for networks or artificial lines to balance the line circuits. As indicated in Figure 166, which is a schematic of a 4-wire metallic telegraph circuit between two terminal type repeaters, the sending and receiving paths are separated from each other as regards transmission over the line. The local circuit arrangements of the repeaters are the same as for 2-wire operation.

Telegraph transmission with this 4-wire arrangement will, in general, be better than that obtained with 2-wire operation because of the improvement in stability produced by eliminating the duplex balance requirements. As the use of different paths for trans-

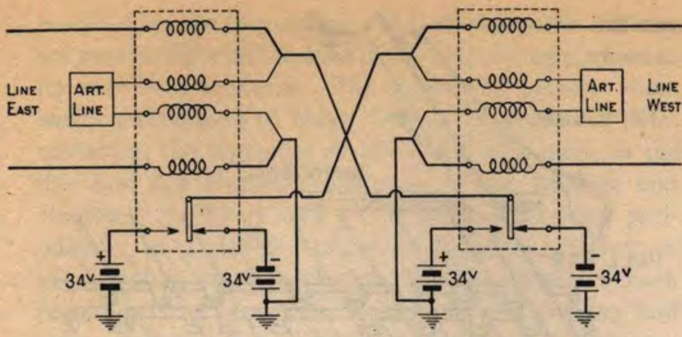


FIG. 165. INTERMEDIATE METALLIC REPEATER

mission in the two directions results in the need for twice as many cable conductors with associated compositing equipment, the susceptibility to certain types of line trouble will be increased. However, other types of line trouble which cause interruptions to 2-wire operation by disturbing the duplex balance, will not interfere in the case of 4-wire operation.

Four-wire metallic telegraph may be superposed on either 2-wire telephone circuits of moderate length (500–1000 miles) or 4-wire telephone circuits, with but little unfavorable reaction on the telephone service. However, very long 4-wire telephone circuits are not composited throughout their entire length because of low-frequency "delay distortion" introduced by the composite sets.

86. Principles of the Teletypewriter

In discussing telegraph circuits in the preceding articles, we have tacitly assumed that the signals to be transmitted were produced manually by the hand operation of an ordinary telegraph key. As a matter of fact, a large percentage of the telegraph circuits are operated by mechanical devices known as teletypewriters, which are installed in subscribers' offices in place of the keys and sounders of manual practice.

Usually the teletypewriter installation at a subscriber's office consists of a keyboard similar to a standard typewriter keyboard and a typing or printing mechanism designed to print received messages either on a page, as is done with typewriters, or on a tape, in the manner of stock quotation tickers. At certain subscriber's stations, such as newspaper offices, where receiving service only is desired, the keyboard may be omitted. On the other hand, when a subscriber wishes to handle a large volume of outgoing traffic, a more elaborate sending mechanism in which messages are first recorded on a perforated tape and then transmitted from it, may be used. We shall not attempt to study in detail the design features or method of operation of each of the several types of teletypewriters, but only to consider some of the general principles applicable to all of them.

Teletypewriter operation of telegraph circuits differs essentially from manual operation only in the substitution of sending and receiving machines for keys and sounders. The signaling code used, however, is not the Morse code of manual operation but a special one in which each letter or signal is made up of five units or elements of equal length. As illustrated by Figure 167, this code provides for the letters of the alphabet, the numerals, and several miscellaneous symbols of common use, as well as for the special operations or "stunts" that the machines must perform, such as line feed, carriage return, and miscellaneous switching and signaling features. The machines must then be so designed that when a certain letter key is operated at the sending machine, the marking and spacing signals

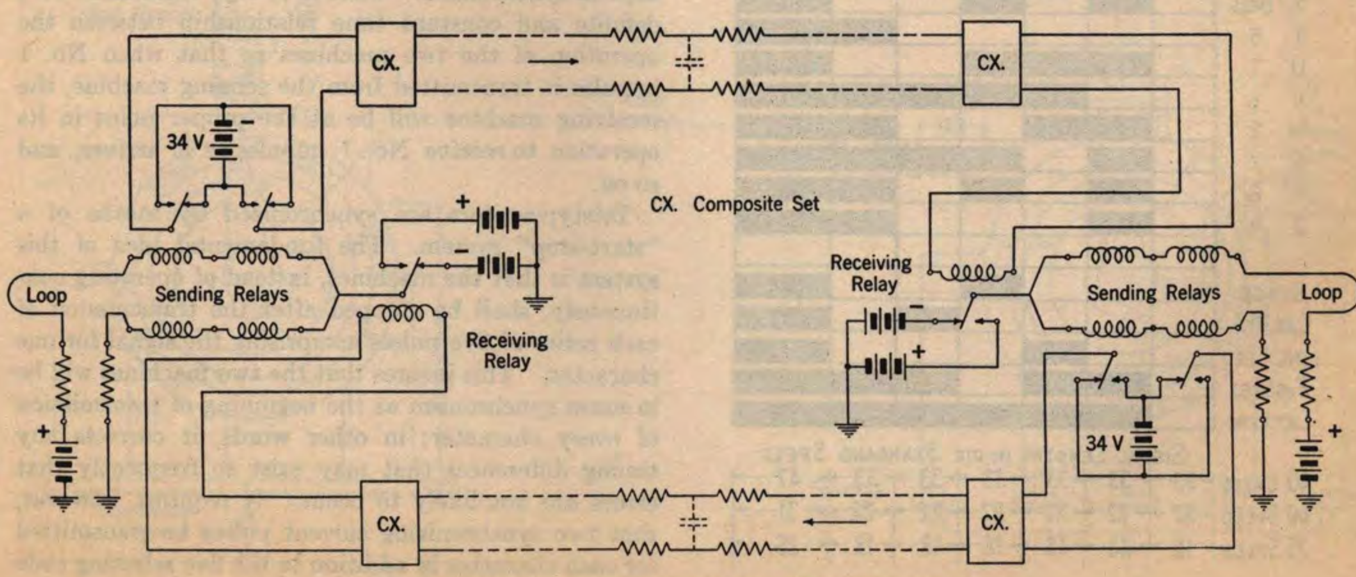


FIG. 166. 4-WIRE METALLIC TELEGRAPH CIRCUIT

corresponding to the code for that letter are sent out on the line; and when this signal combination comes in at the receiving machine, the corresponding type bar is selected and operated to print the letter.

The principle of selection may be understood by referring to Figure 168. The five "code bars" shown are under the control of the five signal units or pulses making up the code for each letter. If the first pulse is a marking signal, code bar No. 1 will be moved endwise a slight amount. Similarly code bar No. 2 will be moved or left in position accordingly as the second pulse is a marking or spacing signal, and so on through the five pulses of the code. When all five pulses have been received, the code bars are so arranged that the slots under the "pull bar", corresponding to the particular code combination received, are in line and all other groups are out of line. This one pull bar is then allowed to drop down a small distance where it engages the "main bail" which pushes it forward and so causes the corresponding type bar to print the character.

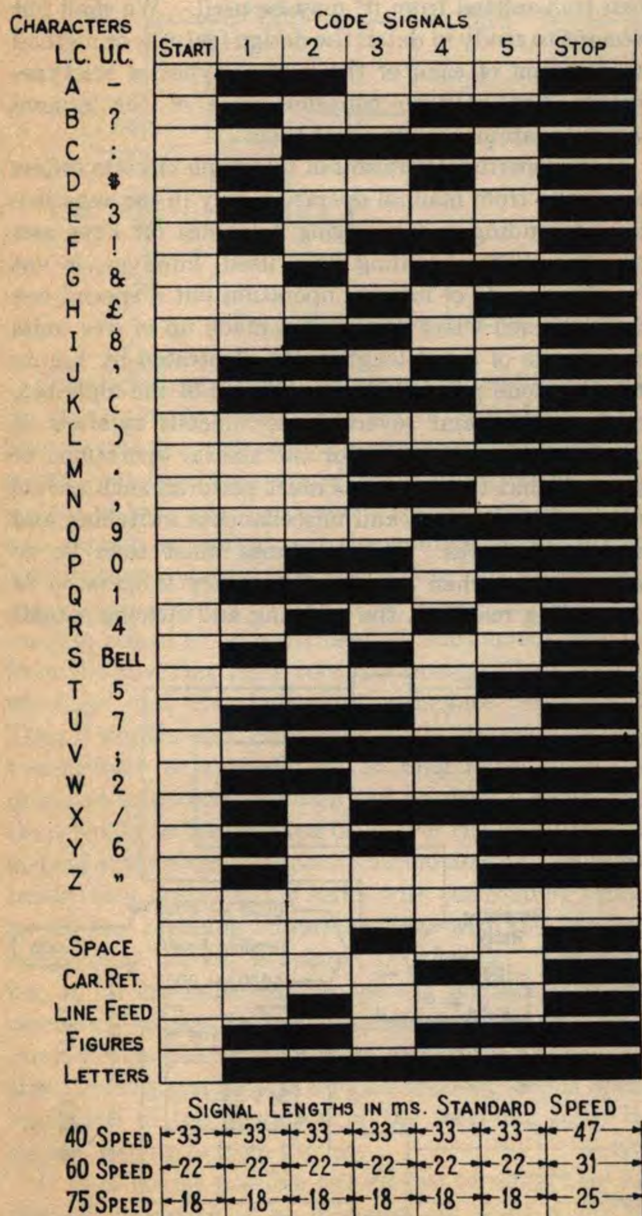


Fig. 167. Teletypewriter Code

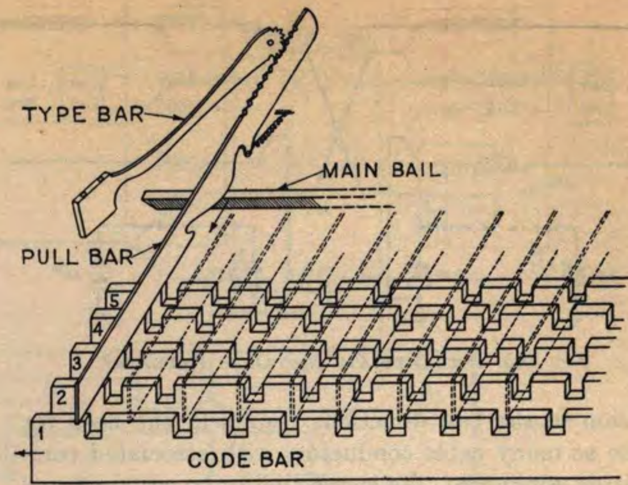


Fig. 168. Selecting Mechanism of Receiving Machine

Assuming some such selection method as has just been outlined and some analogous mechanical arrangement for producing the proper series of current pulses when a key is depressed at the sending machine, there remain two additional essential features that must be provided for. First, there must be a means of positioning the code bars in accordance with the incoming current pulses; this may be effected by electromagnets or by purely mechanical means. Second, and of vital importance, the sending and receiving machines must be kept in synchronism. That is to say, there must be a definite and constant time relationship between the operation of the two machines so that when No. 1 impulse is transmitted from the sending machine, the receiving machine will be at the proper point in its operation to receive No. 1 impulse as it arrives, and so on.

Teletypewriters are synchronized by means of a "start-stop" system. The fundamental idea of this system is that the machines, instead of operating continuously, shall be stopped after the transmission of each series of five pulses comprising the signal for one character. This insures that the two machines will be in exact synchronism at the beginning of transmission of every character; in other words, it corrects any timing differences that may exist so frequently that errors are not likely to occur. It requires, however, that two synchronizing current pulses be transmitted for each character in addition to the five selecting code pulses, a feature which necessarily extends the time required for the transmission of each character. There are several different designs of the controlling apparatus for start-stop systems. The oldest and perhaps most easily understood of these consists of a pair of commutators, the segments of which are connected to the line and the electrical elements of the sending and receiving machines and are connected together periodically in a definite order by brushes rotating in synchronism and stopping at the completion of each revolution. A simplified diagram of the sending and receiving faces of a pair of these commutator devices, known as "distributors", is given in Figure 169.

To follow the operation, let us assume that the letter, D, is to be transmitted. By referring to Figure 167, we find that the code signal for this letter consists of a mark, two spaces, a mark, and a space. We must also remember that the start-stop system requires the transmission of two additional pulses, one to start the brushes revolving and one to complete the operation. The brush or "distributor arms" are coupled to the driving shafts of motors by friction clutches and are normally held stationary by the latches of the sending and receiving start magnets. The motors at the sending and receiving ends are governed to rotate at approximately the same speed. Now when a keyboard key (that for D in our example) is operated, the first effect is to close the circuit through the sending start magnet windings, which pulls up the latch and allows the sending distributor arm to start to rotate. As the inner pair of brushes passes over the start segment in the outer ring of the sending face, the line circuit is opened and a spacing signal is transmitted. This, known as the "start-pulse", releases the receiving line relay, which connects battery to the receiving start magnet and permits the receiving distributor arm to start to rotate.

The operation of the key for D in the keyboard will also have connected battery to segments 1 and 4 of the sending distributor face in accordance with the code for that letter, so that when the sending distributor arm passes off from the start segment on to segment 1, the line will be closed to battery and the receiving line relay will operate. This will connect battery to the large inner segment of the receiving face with the result that when the receiving distributor arm passes over segment 1 (which it will do while the sending distributor arm is still on segment 1 of the sending face), selecting magnet No. 1 will be energized and will move its associated code bar in the printer mechanism. As the sending distributor arm passes over segments 2 and 3, "opens" will be transmitted to the line and accordingly no battery will be connected to the corresponding segments of the receiving face and the associated selecting magnets will not be operated. Continuing, selecting magnet 4 will operate while 5 will not. At this point the code bars in the receiving machine are properly placed for printing the letter, D, and as the receiving distributor arm passes off from segment 5 and on to the stop segment, battery is connected to the "printing magnet" which actuates the printing mechanism and causes the letter to be printed. In the meantime, the sending distributor arm has passed on to its stop segment, thereby transmitting a marking signal to the line, and is stopped by the start

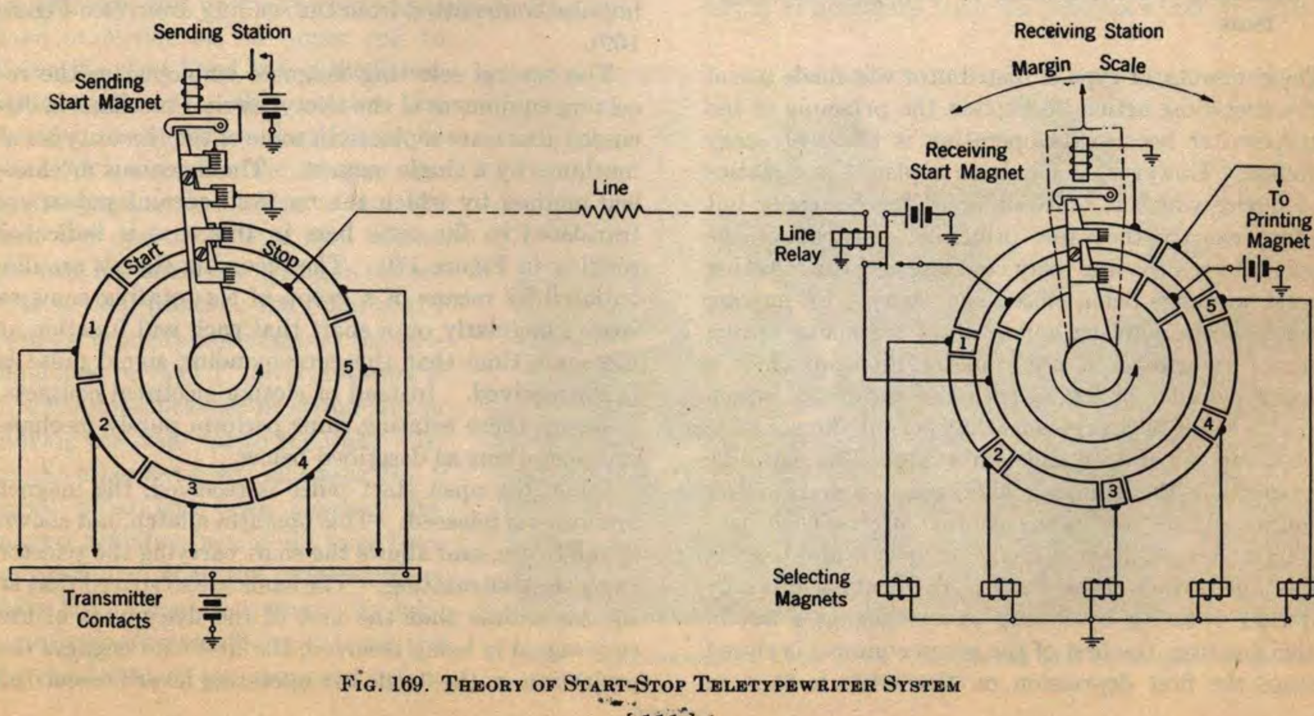


Fig. 169. Theory of Start-Stop Teletypewriter System

No. 3A TWX Switchboard

magnet latch which was released as soon as the outer pair of brushes opened the circuit through the winding at the beginning of the operation. The received marking signal holds the receiving line relay closed so that the receiving start magnet is not operated and the receiving distributor arm is also stopped by its latch as it completes the revolution. Both distributors are then in position to handle the next character.

87. Operating Characteristics of Teletypewriter Systems

The commutator type of distributor was made use of in the preceding article to explain the principle of the teletypewriter because its operation is relatively easy to follow. However, it has been replaced in practice by a device which is quite different mechanically but employs exactly the same principle. Instead of the circuit between the line relay contacts and the selecting magnet windings being closed successively by moving brushes short-circuiting commutator segments, spring contacts are closed in order under the control of a rotating cylinder or drum, into the surface of which are cut a series of depressions that permit the contacts to close at the proper time intervals. The cam-like depressions in the drum are arranged in a spiral order about its surface and seven contact levers, each controlling a contact, are mounted side by side and bearing against the surface of the drum. The drum is normally held from rotating by a stop arm engaging a notch. In this position, the first of the seven contacts is closed because the first depression on the drum is then in position to allow the contact lever to move forward. The circuit to the start magnet is connected through this closed contact. When a start pulse is received, the start magnet releases the drum and it starts to rotate, which immediately opens the first contact. Following this, the second depression on the drum comes under the second contact lever which then moves forward closing the contact to the first selecting magnet, and this magnet will be operated or not, depending upon whether the line relay is at that instant on its marking or its spacing contact. As the drum continues to rotate, the remaining four selecting contacts operate in order, which results in the remaining four selecting magnets being operated in accordance with the incoming signals; and finally, the seventh contact is closed to operate the printing magnet causing the character set up by the selecting magnets to be printed.

In order to increase the maximum overall speed of operation, the system may be so arranged that when sending at a maximum speed, the sending distributor rotates continuously instead of stopping after the transmission of each character. To preserve synchronism, however, it is still necessary that the receiving distributor come to a full stop after each complete revolution. This is effected by arranging the receiving distributor to rotate at some fourteen per cent greater speed than the sending distributor, thus providing a brief time interval during each revolution for it to stop. This requires, of course, that the depressions on the receiving cylinder be spaced fourteen per cent farther apart angularly than those on the sending drum in order that the receiving drum will close the receiving contacts during the exact middle portion of each signal impulse transmitted from the sending drum (see Figure 169).

The several selecting magnets employed in the receiving equipment of the teletypewriter mechanism discussed above are replaced in some of the recent types of machines by a single magnet. The ingenious mechanical method by which the received current pulses are translated to the code bars in this case is indicated roughly in Figure 170. The incoming signals are distributed by means of a group of six rotating cams so spaced angularly on a shaft that each will function at the same time that the corresponding signal pulse is being received. Instead of closing electrical contacts, however, these rotating cams perform purely mechanical operations as described below.

When the open start pulse is received, the magnet armature is released. This operates a latch, not shown in the figure, and allows the shaft carrying the selector cams to start rotating. The cams are so spaced that at the same time that the first of the five pulses of the code signal is being received, the first cam engages the projection on the "code bar operating lever" associated

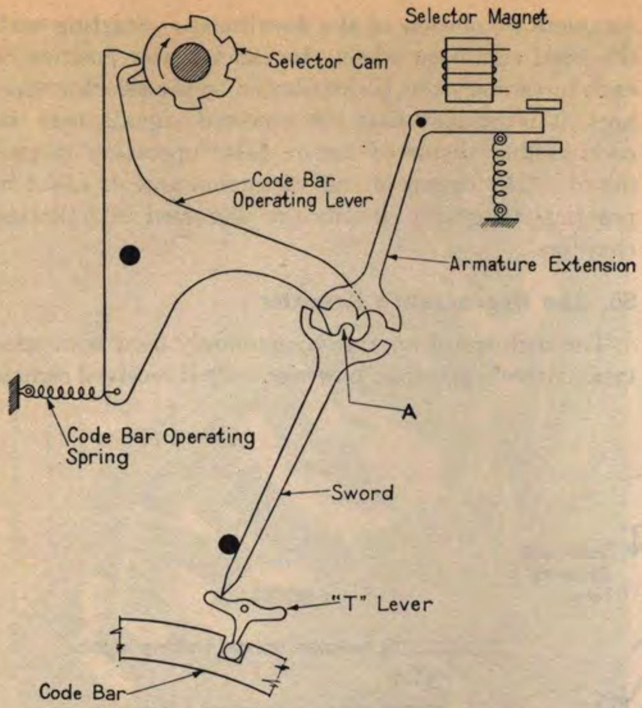


Fig. 170. Receiving Mechanism of Single Magnet Teletypewriter

with the first code bar and rotates it slightly in a counterclockwise direction. The effect of this movement depends upon whether or not the magnet armature is operated. If the received No. 1 pulse is a marking signal, the armature will be operated as shown in the figure, whereas if this pulse is spacing, it will not be operated. But assuming that it is operated, the movement of the code bar operating lever by the first selector cam lifts up the "sword" and causes the right-hand projection on its upper end to strike the right-hand end of the "armature extension". This rotates the sword in a clockwise direction in its pivot "A", and when the selector cam in its continued rotation clears the code bar operating lever and allows the code bar operating spring to restore it to normal position, the point of the sword is brought down against the left-hand side of the "T" lever, rotating it in a counter-clockwise direction and so moving the code bar to the right. If, on the other hand, the incoming No. 1 signal pulse had been spacing, the magnet armature would not have been operated and when the code bar operating lever raised the sword, its left-hand projection would have struck the left-hand side of the armature extension causing the sword point to move to the right and the code bar to the left.

In exactly the same manner, when No. 2 pulse is received, the second selector cam will have arrived at the proper position in its rotation to operate the code bar operating lever associated with code bar No. 2 and it will be positioned according to the position of the magnet armature at the time. After all five signal pulses have been received and the code bars properly positioned, the sixth cam releases a clutch allowing the printing mechanism to operate.

One of the advantages of the teletypewriter over manual telegraph service is the high speed of operation that can be consistently maintained. There are three standard operating speeds, namely, 40, 60 and 75 words per minute. The lowest one of 240 operations or 40 words per minute corresponds to very fast manual operation. A speed of 360 operations or 60 words per minute is used for a majority of the teletypewriter services. The highest speed consists of 450 operations or 75 words per minute. Naturally these high signaling speeds require not only that the transmitting and receiving machines be sturdy and dependable but also that the connecting lines and telegraph repeating apparatus be of the highest grade.

Even though synchronism between sending and receiving machines may be satisfactorily maintained by means of the start-stop method of operation, it is clear that any distortion of the transmitted signals, due to unsatisfactory line conditions or other reasons, will tend to cause receiving errors. Therefore since a certain amount of distortion is practically inevitable in long telegraph circuits as we shall see in the next chapter, it is necessary that the teletypewriter systems be

Page Teletypewriter with Tape Perforator and Automatic Sending Equipment designed with the maximum possible operating margin. By referring to Figure 169, it will be noted that the segments on the receiving face of the distributor are considerably shorter than those on the sending face. This means that only a relatively small portion of each transmitted signal pulse is used for operation of the receiving machine and that these pulses may therefore vary appreciably before causing false operation of the receiving machine. This is illustrated by Figure 171 in which the narrow spaces bounded by the vertical lines indicate the time intervals during which the five selecting magnets are connected to the line during one complete revolution of the distributor. Starting with the ideal condition where the exact center position of each incoming pulse is distributed to the selector magnets, it is evident that the received signals may be considerably distorted before false operation is produced. The causes of this distortion and its effect in practical telegraph circuits are discussed in following chapters.

88. The Regenerative Repeater

The high speed operation commonly used with teletypewriters is possible, however, only if received signals

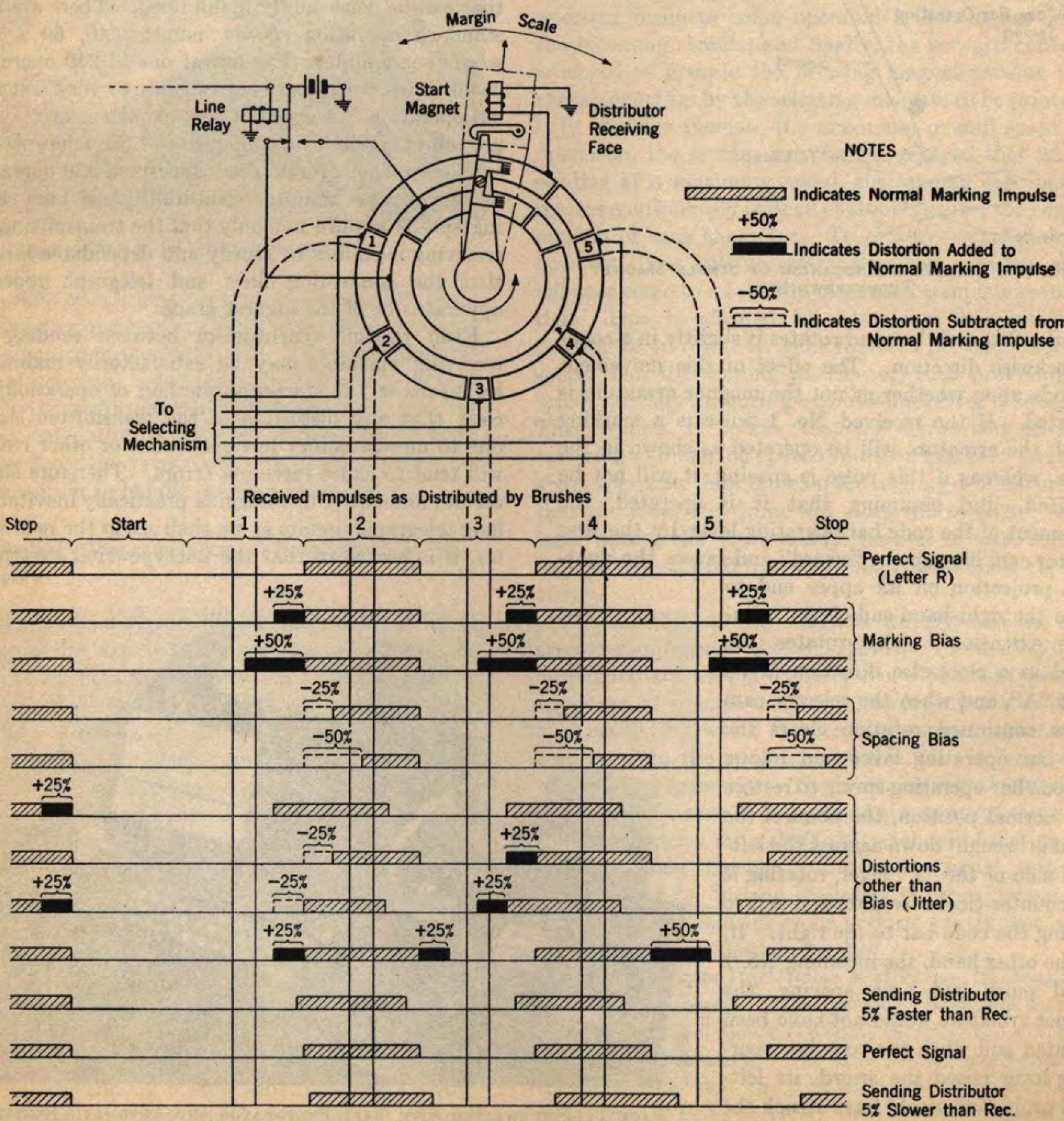


FIG. 171. TIME RELATIONSHIPS OF RECEIVED SIGNALS

Large TWX Switchboard

are free from any large amount of distortion. Since telegraph signals are invariably distorted to a greater or less extent in the process of transmission and since the ordinary telegraph repeaters repeat the greater part of such distortion so that it increases cumulatively with the length of the overall circuit, the maximum distance over which a teletypewriter circuit can be operated tends to be limited by this factor. Fortunately, the fact that the signals are of standard length and are transmitted with mechanical uniformity permits the use in long circuits of a special type of telegraph repeater which is capable of eliminating distortion from the signals.

This is known as the start-stop "regenerative" repeater. Consisting essentially of a sending and receiving drum or distributor·similar to those used in the teletypewriter mechanism, it is capable of receiving without error any set of signals that would be satisfactorily received by an ordinary teletypewriter set, and of sending these same signals out as free from distortion as the signals formed by the sending teletypewriter. Therefore by spacing regenerative repeaters at intervals that would be sufficiently short for satisfactory operation of a standard teletypewriter circuit, it is theoretically practicable to operate a circuit of any length whatever. -

Figure 172 shows schematically the simplified arrangement of a type of regenerative repeater employing a flat distributor face and rotating brush arm. Here the two outer rings of segments represent the receiving commutator face and are shorted together by a pair of rotating brushes at the same time that the two inner rings, comprising the sending face, are shorted together by another pair of brushes mounted on the same rotating brush arm. To follow its operation, let us assume that the letter, R, is to be transmitted. Referring to Figure 167, we find that the incoming signals will consist of the starting spacing pulse, a space, a mark, a space, a mark, a space, and the final marking pulse. As the spacing start impulse is received, the brush arm will be releasedtthrough a mechanism not shown in the drawing) and he two sets (of brushes will start to rotate. The receiving brush passes first over a blank segment and then connects the short No. 1 receiving segment to the receiving relay armature. This occurs at the instant that the first spacing signal of the five-impulse code is being received, and the receiving relay is therefore operated to its spacing contact. The right storing condenser will accordingly be charged positively. In the meantime, the brush of the sending face has moved over segment 7, connecting spacing battery to the sending relay and so repeating the start signal to the line in the other direction. Just after the receiving brush moves off from No. 1 segment on to a blank, the

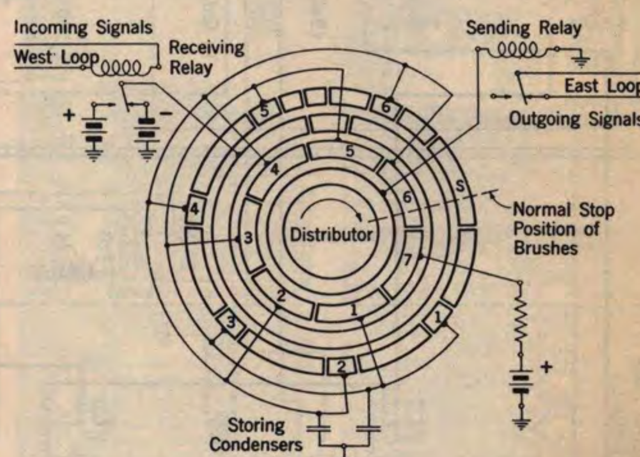


Fig. 172. Principle of the Regenerative Repeater

sending brush moves on to No. 1 segment of the sending face and the positively charged right storing condenser discharges through it to the sending relay, thus repeating the first code spacing signal. During this operation, the receiving face brush moves on to segment No. 2 of the receiving face and charges the left storing condenser negatively in accordance with the incoming marking signal. This condenser is in turn discharged through No. 2 segment of the sending face

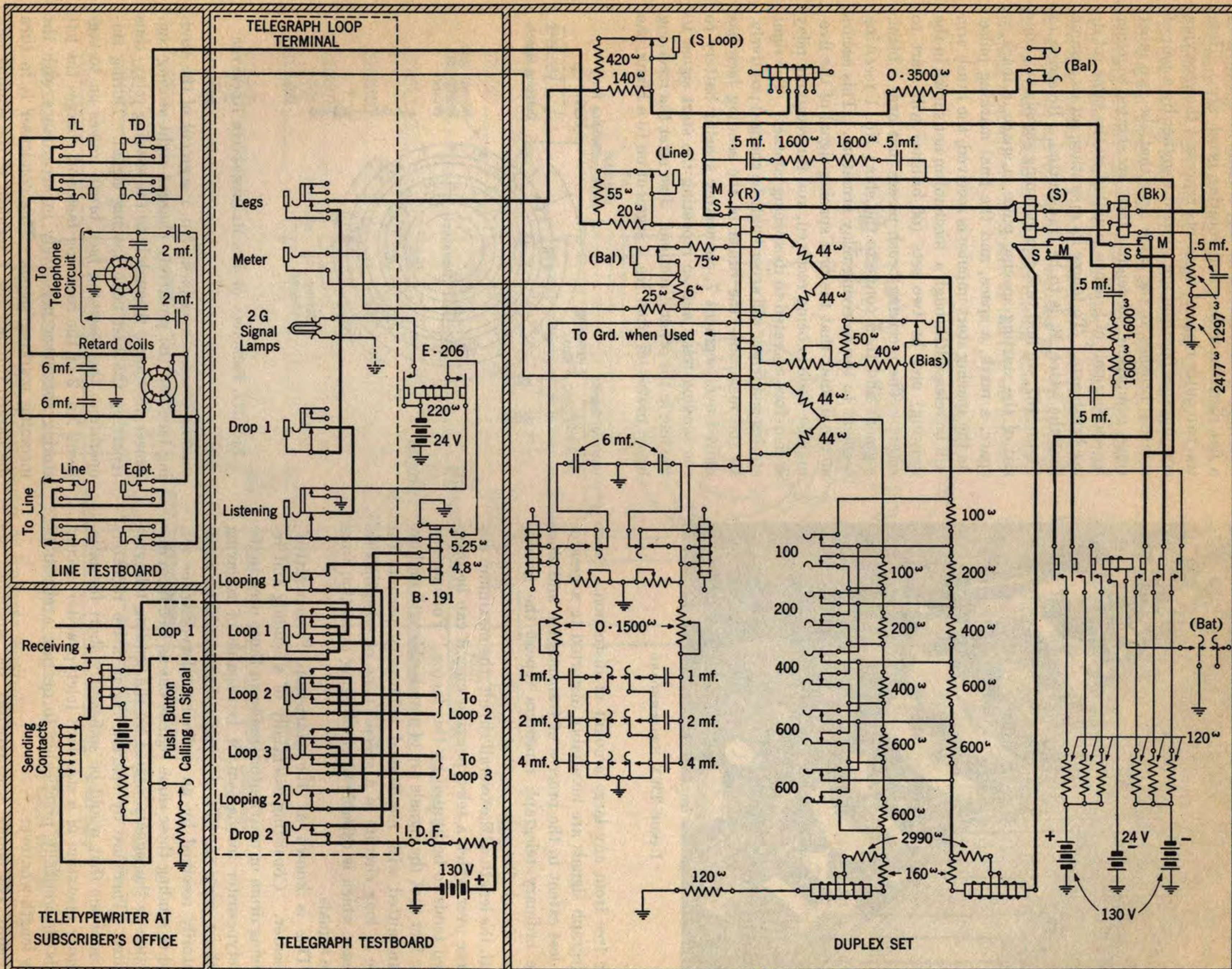
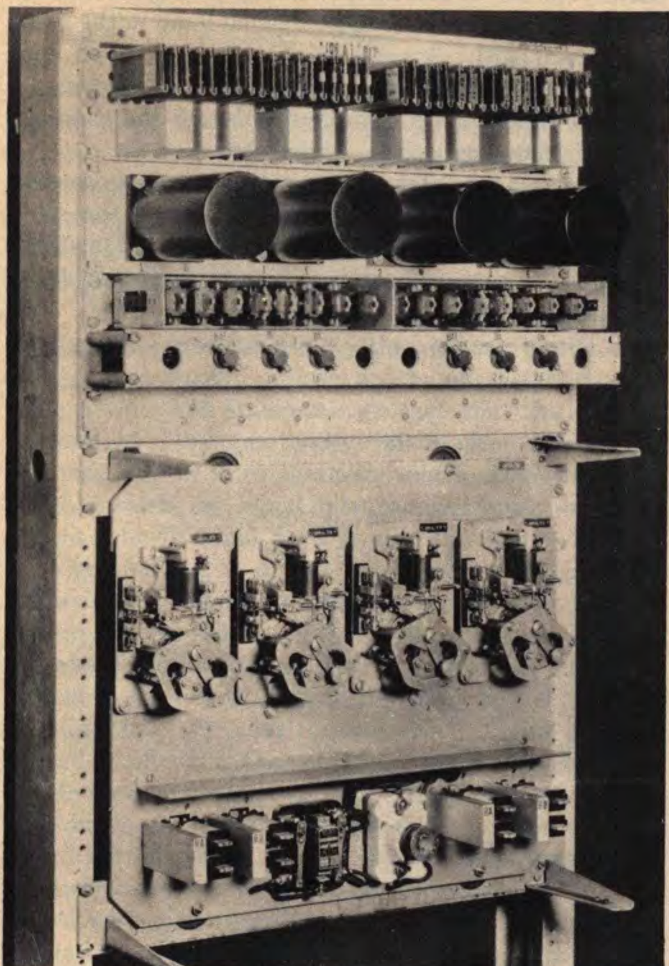


Fig. 173. Typical Terminating Telegraph Circuit Equipped with Modern Differential Duplex Set.

while the right condenser is being charged through No. 3 segment of the receiving face. This alternate operation continues until all five of the received code impulses and the final stopping mark impulse have been repeated to the outgoing line. As it completes its revolution, the brush arm is stopped until the next starting impulse is received.



CAM-TYPE REGENERATIVE TELEGRAPH REPEATERS

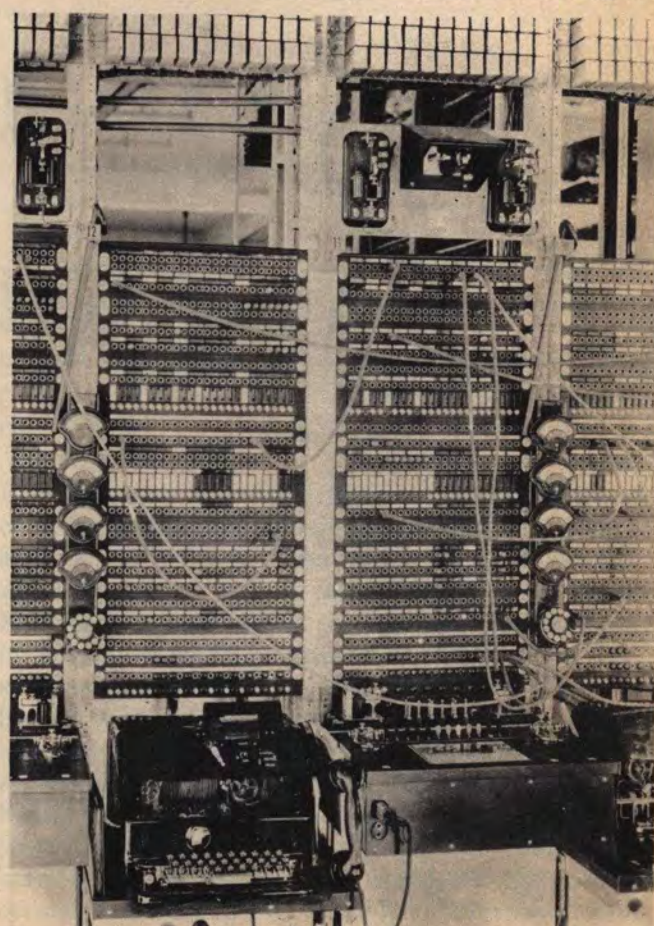
The fundamental value of the repeater lies in the fact that the short receiving segments pick up only the mid-portion of the incoming signal thereby allowing for considerable distortion, while the signals sent out are uniform and of equal length because the sending face segments are of equal length and spaced in exactly the same way as those of the regular sending distributor at the sending teletypewriter. It is apparent, accordingly, that the repeater will receive and convert to perfect signals any signals that are good enough to cause satisfactory operation of an ordinary receiving teletypewriter.

There is also a type of regenerative repeater which uses a cam type distributor. Its principle of operation is, however, the same as discussed above.

89. Typical Terminating Telegraph Circuit

In this and the preceding chapter we have studied the general principles of some of the more important equipment units that are used with long distance telegraph circuits, as well as the apparatus at the subscriber's station. In order that we may get a clear picture of how these several parts are coordinated in practice, Figure 173 shows in some detail the wiring arrangement of a representative terminating telegraph circuit from the subscriber's office to the long distance line wires. This includes in addition to the subscriber's station equipment and the terminal duplex set, the connections at the telegraph testboard and at the line testboard.

The repeater shown is a recent design of a terminal differential duplex set arranged for half-duplex operation and using a 2-wire line circuit. The usual ground return is employed for transmission and no signals are transmitted over the second line wire which is known as a "neutralizing wire". Its purpose is to balance out interference in the receiving relay that may be produced by the line section. This is accomplished by connecting the receiving relay windings to the regular



No. 9 TELEGRAPH TESTBOARD

line and to the neutralizing line so that they magnetically oppose each other. Any interference present on these line wires is then neutralized and has no effect on the receiving relay.

At the subscriber's station is a teletypewriter arranged for both sending and receiving. The telegraph testboard apparatus and wiring, which we have not hitherto considered, provide a flexible patching arrangement and also permit rapid testing of the telegraph facilities in both directions. The line circuits, coming from the regular telephone line testboard are connected direct to the differential duplex set, while the subscriber's loop is connected into "telegraph loop terminal" circuits (TLT) which are in turn connected to the duplex set.

The relays associated with the telegraph loop terminal circuit are a part of an auxiliary circuit by means of which a subscriber can signal an attendant at the telegraph testboard. Normally the same current flows through the two opposing windings of the B-191 relay so that its armature is not attracted. But if the subscriber grounds one side of the loop by operating the push-button calling-in signal, the current in one winding of the B-191 relay becomes much larger than that in the other and the relay is operated. This connects battery to the E-206 relay winding through contacts of the Listening jacks to light a signal lamp in the telegraph testboard. An attendant at the telegraph line terminal answers the signal by plugging into the Listening jack which opens the circuit through the winding of the E-206 relay and the signal lamps.

Connections at intermediate repeater points on long telegraph circuits do not, of course, include the subscriber and telegraph line terminal circuits but in other respects the arrangement is generally similar to that shown by Figure 173.

CHAPTER XIII

TELEGRAPH TRANSMISSION PRINCIPLES

90. Nature of Telegraph Signals

In telegraph transmission we are concerned with the reproduction of the sent telegraph message at the receiving end at a satisfactorily rapid rate, without error, and without interference to other services. Telegraph transmission differs from telephone transmission in that intelligence is conveyed from some sending point to one or more receiving points by means of a signal code. In the preceding chapters we assumed this was satisfactorily accomplished by the various circuits and apparatus discussed. However, the characteristics of these circuits and apparatus are such that the signals transmitted sometimes tend to fail to reproduce at the receiving end the same character that was transmitted. In other words, the signal in transmission may undergo certain changes which tend to alter its characteristics.

As pointed out in Article 76, there are two general methods of transmitting telegraph signals—(1) neutral transmission in which current is sent over the line to operate the relays to the marking position, and the current is stopped to operate the relays to the spacing position; and, (2) polar transmission which is accomplished by changing the polarity of the sending battery for the mark and space signals. Thus, telegraph signal transmission is accomplished on what may be termed a two current basis, that is, by transmitting spurts of steady current interspersed by intervals of no current in the case of neutral operation, or by transmitting spurts of current in one direction interspersed by reversals of current in the case of polar operation. In neutral operation, the closed circuit signal is referred to as a "mark" or "marking signal" and the open circuit signal is known as a "space" or "spacing signal". In polar operation, the marking and spacing nomenclature is retained but here it refers to the direction of current flow rather than to the open and close condition as in neutral operation. In either type of operation the change from one current condition to the other, that is, from mark to space or space to mark, is known as a **transition.**

The change of the current from the marking to the spacing condition, or vice versa, can be plotted with respect to time. A drawing showing this change of the current from the one condition to the other is called a "wave shape diagram" or, more commonly, simply a **wave shape.** Wave shapes, since they depict the change of current in a telegraph circuit are an important aid in the study of telegraph transmission.

91. Wave Shapes in Neutral Telegraph Systems

In any neutral telegraph circuit, if we could ignore the times required for the direct current to establish itself and to decay, the wave shape of a telegraph signal for the letter A in the Morse code would be as illustrated in Figure 174-A. As a practical matter, however, every telegraph circuit has some series inductance.

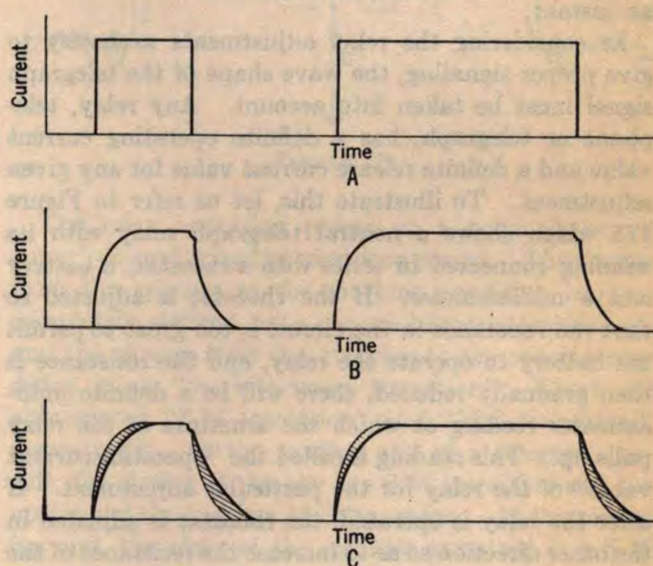


FIGURE 174

Each line relay adds some inductance and, in the case of the composited circuit, each retardation coil winding adds several henrys. The signal wave shape with series inductance is more nearly that represented by Figure 174-B, each current pulse having a sloping curve from zero to maximum value at the make of the key, and from maximum value to a point where the arc is broken at the break of the key. If in addition to the inductance we consider the condensers of the composite set, we have a further sloping of the pulse as shown by Figure 174-C. Here the shaded portion represents the effect of the condensers over and above the effect of the inductance. When the key is closed, the first rush of current flows only in part to the line; the inductance of the retardation coil in the composite set opposes any sudden change and diverts the current to the con-

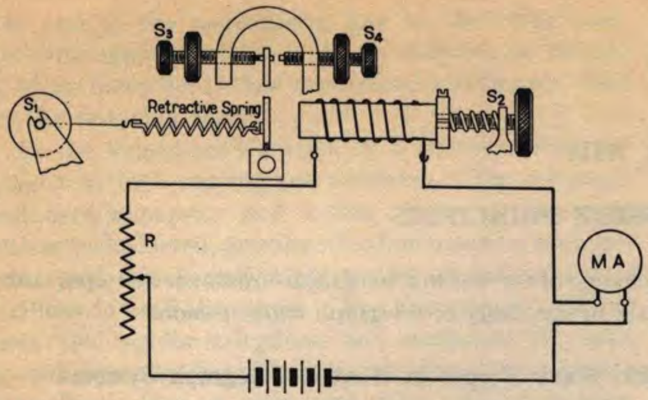


FIGURE 175

densers until they are charged to about the potential of the battery. When the key is opened, on the other hand, the current does not stop at the breaking of the arc because the discharging condensers sustain it for an instant.

In considering the relay adjustments necessary to give proper signaling, the wave shape of the telegraph signal must be taken into account. Any relay, telephone or telegraph, has a definite operating current value and a definite release current value for any given adjustment. To illustrate this, let us refer to Figure 175 which shows a neutral telegraph relay with its winding connected in series with a rheostat, a battery and a milliammeter. If the rheostat is adjusted so that the resistance in the circuit is too great to permit the battery to operate the relay, and the resistance is then gradually reduced, there will be a definite milliammeter reading at which the armature of the relay pulls up. This reading is called the "operating current value" of the relay for the particular adjustment. If after the relay is operated, the rheostat is adjusted in the other direction so as to increase the resistance of the circuit, the relay armature will fall back at a definite milliammeter reading. This is called the "release current" for the relay at the particular adjustment. The release current is smaller in value than the operating current for two reasons—(1) the magnetic circuit is much stronger when the armature is closer to the pole pieces so that the magnetic pull which holds the armature is greater than the pull which advances the armature; and (2) there is some residual magnetism in the iron core at the time the circuit is broken that did not exist at the time the circuit was made.

We might represent by the points O and R in Figure 176-A the operating and release current values respectively for a relay like that illustrated in Figure 175. With this particular adjustment, the length of the signal repeated by the relay will be the time indicated by T. If we should now make certain adjustments of the relay either by weakening the tension of the retractive spring with the screw S_1, lessening the air gap between the pole pieces and the armature with the screw S_2, or decreasing the stroke of the armature by adjustments of the contact and back stop screws S_3 and S_4, we may greatly decrease the operating and release current values, say to those represented by O_1 and R_1 of Figure 176-B. The effect would be to increase the length of the signal repeated by the relay from that represented by T to that represented by T_1. These adjustments would have changed the signal from "light" to "heavy". For the sake of contrast, let us imagine that the wave shape of the signal was that shown by Figure 174-A. Here it is evident that we could neither increase nor decrease the length of the signal by relay adjustments.

To a degree this explains the frequent adjustments that are necessary on telegraph apparatus in practice. If additional inductance is added to a circuit by inserting a relay winding in series, the slope of the make and break of the signal is increased and a new adjustment may be required. The adjustment might be to lengthen the signal in one case and to shorten it in another. It would depend upon the original positions of points O and R on the curve.

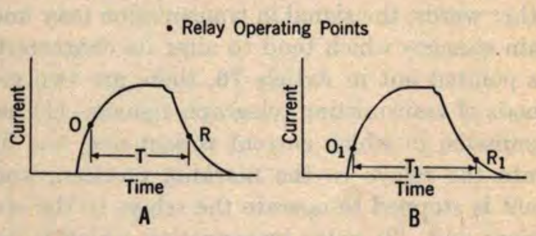


FIG. 176. EFFECT OF RELAY ADJUSTMENT ON TELEGRAPH SIGNAL LENGTHS

Another factor that will change the length of the signal is a change in the current value, resulting from a change in the voltage or in the series resistance. Let us consider the case of increasing the current by using higher voltage or taking series resistance out of the circuit. Naturally the operating and release current values of the relay before and after the change are the same, but they are more nearly the maximum current values before the change is made than after. Since the increase in current with constant inductance steepens the sides of the curve, the net result is an increase in the length of the signal. It is to be noted, however, that current values are limited in practice by considerations of crossfire and interference with telephone circuits, so that this is not ordinarily a practicable method of increasing signal length in actual operation.

In practice there is a third changing condition that affects adjustments. This is fluctuation in current values due to leakage along the line. To a degree it can be compensated for by using grounded battery connections at both ends of the circuit, one end being positive and the other end being negative, instead of

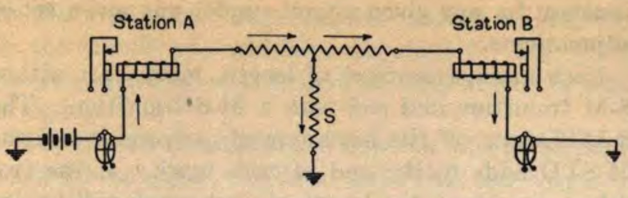


Figure 177

using a single battery at one end and only a ground at the other end. To understand this let us assume the condition shown in Figure 177 where S represents a leak to ground along the line, either distributed or otherwise. First, let us suppose that Figure 178-A represents the current curve when there is no leak. The leak will increase the current in the station A relay on account of the additional path through S. Since this path has less inductance we may represent the leakage current alone by the curve shown in Figure 178-B, which is smaller in value. Now, the curve B is going to influence the current through the A station relay by making it more nearly that represented by curve C, which is steeper and will therefore give a heavier signal. On the other hand, the leak S has a shunting effect on the current through the B station relay, and will not only tend to decrease the current but to flatten the curve as shown by Figure 178-D. This wide variation in the current through the two relays could not take place if the battery at Station A had about half its voltage, and an equal battery of opposite polarity was used at Station B, instead of a direct connection to ground. The current under these conditions will be more nearly constant through the relays at the two ends because we can assume that each battery is furnishing current to ground through the leak, and these currents as illustrated by Figure 179, tend to neutralize each other because they are flowing in opposite directions.

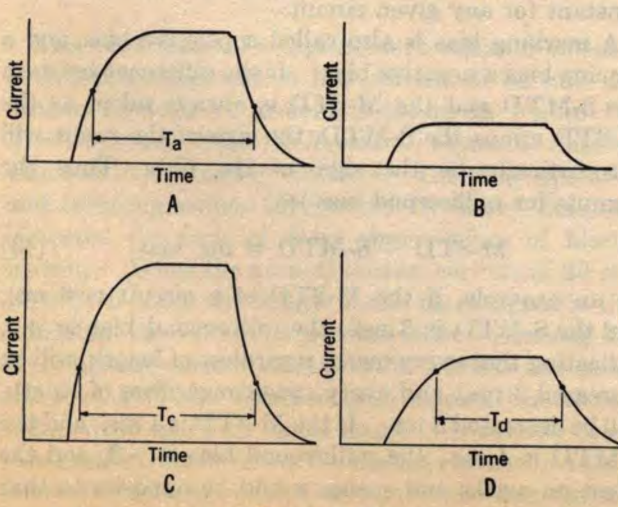


Figure 178

The shunting effect to ground of a leak, such as is shown by Figure 177, is a special case. On every telegraph wire, regardless of insulation conditions, we have in effect a leak to ground through the capacity between the wire and ground, or a condition that might be illustrated by substituting a condenser for the resistance S in Figure 177. Since the telegraph current wave shapes are somewhat similar to alternating-current cycles, the condenser may properly be considered as shunting the current to some extent. Furthermore, this capacity not only decreases the current value that reaches the distant station but tends to further distort the wave shape, thereby limiting the distance over which satisfactory signals can be sent without additional repeaters.

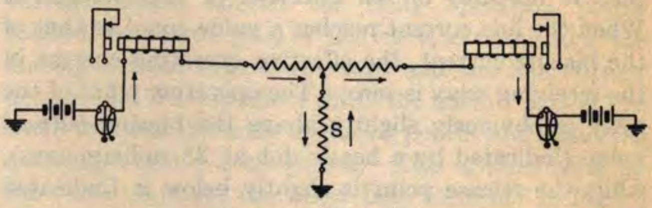


Figure 179

The relay operating points on a wave shape, of course, occur during the transition period. The change from the spacing to the marking condition is more completely defined as a "space-to-mark transition", and the change from the marking to the spacing condition as a "mark-to-space transition". These are abbreviated "S-M transition" and "M-S transition", respectively. At the sending end of a telegraph circuit, the closing of the key is a S-M transition and the opening of the key is a M-S transition. At the receiving end, the close of the sounder armature is a S-M transition while the release of the sounder armature is a M-S transition.

Let us consider the wave shapes of the telegraph signals in the neutral circuit, schematically illustrated in Figure 180, using an electrically biased receiving relay as discussed in Article 81. It will be seen from an inspection of this circuit that when the sending key is closed, the capacity between line and ground is charged by the current flowing from the battery at the sending end. The inductance in the circuit retards the current from building up to its full value instantaneously. The wave shape takes the form shown in Figure 181 (letter A in Morse code). As the biasing current in the receiving relay tends to hold the armature to the spacing contact, the line current, which magnetically opposes the biasing current, does not operate the receiving relay to the marking contact until it reaches a value slightly in excess of the biasing current. While the operating points of the relay are determined by its

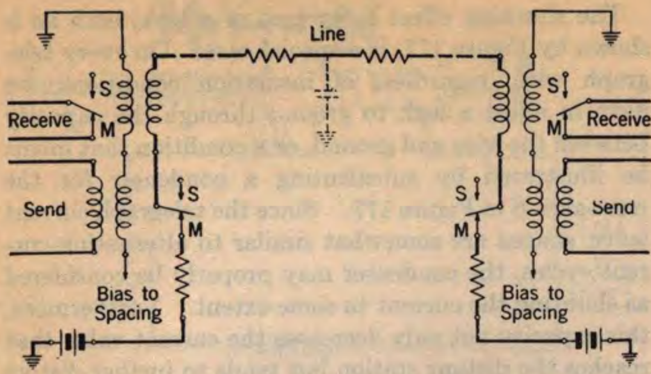


FIG. 180. NEUTRAL TELEGRAPH CIRCUIT EMPLOYING BIASED RELAYS

design and adjustments, we will consider for simplicity that it operates on an effective ± 3 milliamperes. When the line current reaches a value equal to that of the biasing current, the effective operating current in the receiving relay is zero. The operating point of the relay is obviously slightly above the biasing current value (indicated by a heavy dot at 33 milliamperes), while the release point is slightly below it (indicated by a heavy dot at 27 milliamperes).

At the instant the key at the sending end is closed, the line current starts to rise in the receiving relay as indicated by the wave shape but does not reach the relay operating point until a few milliseconds later. This means there is a delay between the closing of the sending key on a S-M transition and the operation of the receiving relay. This may be called a "space-to-mark transition delay" and abbreviated as S-MTD.

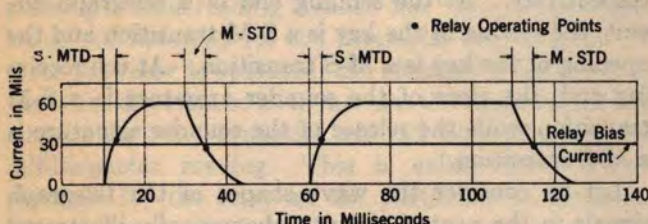


FIG. 181. SIGNAL WAVE SHAPES IN NEUTRAL TELEGRAPH CIRCUIT OF FIG. 180

In a similar manner when the sending key is opened, the line current in the receiving relay does not become zero instantaneously. The receiving relay will be held on its marking contact for an interval of time after the circuit is opened at the sending end. This time delay is a "mark-to-space transition delay" and is abbreviated as M-STD.

The magnitudes of these delays range from a fraction of a millisecond to several milliseconds. The S-MTD and M-STD are determined entirely by the characteristics of the circuit, and, though the two delays may not be equal, **each transition delay will always be a constant for any given circuit, under any given set of adjustments.**

Each mark, regardless of length, must start with a S-M transition and end with a M-S transition. The S-MTD cuts off the beginning of each mark and the M-STD adds to the end of each mark. If the two delays are equal, the length of each mark will be unchanged by transmission over the circuit. Each space, regardless of length, starts with a M-S transition and ends with a S-M transition. The M-STD cuts off the beginning of each space and the S-MTD adds to the end of each space. Each delay thus has the opposite effect on a space that it has on a mark. If the two delays are equal, the length of each space will be unchanged by transmission over the circuit. The transmission is considered perfect if the received marks and spaces are exactly the same length as the sent marks and spaces.

92. Bias Distortion

The requirement for perfect transmission then is that the S-MTD equal the M-STD. If the two delays are not equal, as for instance if the M-STD is greater than the S-MTD, all marks will be lengthened, and all spaces will be shortened. This is a common condition on circuits and is called "marking bias" because the circuit lengthens the marks. If the S-MTD is greater than the M-STD, all spaces will be lengthened and all marks shortened. This is another common condition and is called "spacing bias".

Since the lengths of the marks and spaces may be indicated in milliseconds (ms.), the amount that is added to or subtracted from each mark or space due to a bias condition may also be indicated in milliseconds. It is equal to the difference between the S-MTD and the M-STD expressed in milliseconds. This is referred to as the "millisecond bias" of a circuit, and is a constant for any given circuit.

A marking bias is also called a positive bias, and a spacing bias a negative bias. If the difference between the S-MTD and the M-STD is always taken as the M-STD minus the S-MTD, the sign of the result will automatically be the sign of the bias. Thus the formula for millisecond bias is:

$$\text{M-STD} - \text{S-MTD} = \text{ms. bias} \qquad (43)$$

As an example, if the M-STD of a circuit is 6 ms., and the S-MTD is 3 ms., the millisecond bias is +3, indicating that every mark, regardless of length, will be increased 3 ms., and every space, regardless of length, will be decreased 3 ms. If the M-STD is 1 ms., and the S-MTD is 4 ms., the millisecond bias is −3, and the effect on marks and spaces would be opposite to that in the first example. It is important to keep in mind that a millisecond bias condition is determined entirely by the equipment, line facilities, overall length, etc. of the circuit and will be a constant for any given circuit, regardless of the speed of transmission or kind of signals.

The effect on transmission, however, of a given millisecond bias condition, does vary with the length of marks and spaces transmitted even though the millisecond bias condition itself is constant. As an example of this, let us consider a manual telegraph circuit where the dashes (long marks) are normally about two and a half to three times the length of the dots (short marks). In manual telegraph the lengths of the dots and dashes decrease as the speed of transmission increases. Assume first a slow speed of transmission where the dots are 30 ms. long and the dashes are 90 ms. long. A millisecond bias condition of +10 will make the dots 40 ms. long and the dashes 100 ms. long. The signals will be quite readable since the three to one ratio has been changed very little. Next, assume a much faster speed where the dots are 5 ms. long and the dashes 15 ms. long. The same +10 bias will make these dots 15 ms. long and the dashes 25 ms. long. Greater difficulty will be experienced in reading these signals since the dashes now are not even twice the length of the dots.

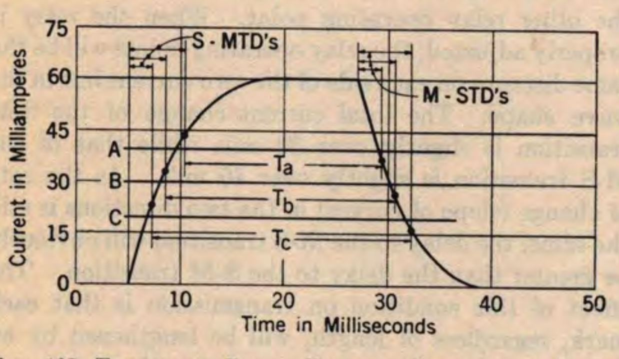


FIG. 182. EFFECT OF RELAY BIASING CURRENT ON SIGNAL LENGTH

The wave shape of a typical mark signal in a neutral circuit operating with a line current of 60 mils and having capacity to ground is shown in Figure 182. The horizontal lines A, B, and C, represent different values of relay biasing currents. The relay operating and releasing points (designated by heavy dots) are indicated for each of these three values of biasing current. When the normal biasing current of 30 mils (line B) is used, the length of the mark signal is that indicated by T_b. It is obvious that increasing the relay biasing current increases the S-MTD and shortens the M-STD. This produces spacing bias since it reduces the marking signal length. A reverse condition results from lowering the biasing current to a value below the normal value; that is, the marking signal is

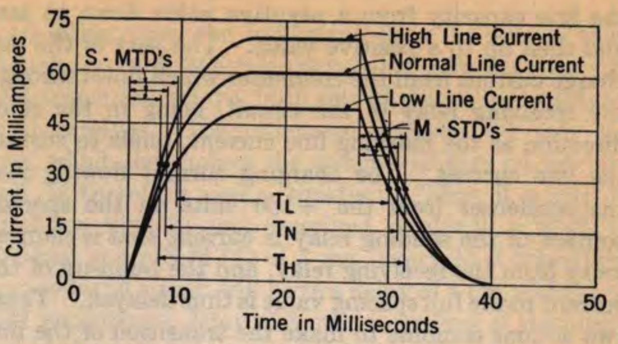


FIG. 183. EFFECT OF LINE CURRENT MAGNITUDE ON SIGNAL LENGTH

increased in length, as shown by T_c, because the S-MTD decreases while the M-STD increases.

The same effect as that obtained by raising the bias current, which shifted the relay operating points toward the narrow part of the wave, is obtained if the biasing current is held constant and the line current decreased. This in effect shifts the narrow part of the wave towards the relay operating points, and again the marking impulse is shortened. This is illustrated by the wave shape for the "Low Line Current" in Figure 183. On the other hand, increasing the line current while the biasing current remains the same, increases the current at all points on the wave shape and effectively shifts the broader part of the wave toward the operating points. This lengthens the impulse as illustrated by the wave shape for the "High Line Current" in Figure 183. In other words, increasing the line current in a neutral circuit tends to produce marking bias and decreasing it tends to produce spacing bias.

93. Wave Shapes in Polar Telegraph Systems

A one-way polar circuit using a ground return is shown schematically in Figure 161. The sending relay connects −130 volts to the line for the marking condition and +130 volts for the spacing condition. The resistance at the sending end is adjusted by means of a potentiometer (not shown in the drawing) connected in the line circuit so that the current is normally about +35 mils for the marking condition and −35 mils for the spacing condition. These are, of course, the "steady state" values.

In this, as in other circuits, the change of the line current from marking to spacing (M-S transition) and from spacing to marking (S-M transition) will be delayed because of the capacity between the line and ground. When the line current is marking, the voltage on the condenser, representing the capacity between the line and ground, is negative. On the other hand, when the line current is spacing, the voltage on this condenser is positive. The change of the line current from marking to spacing then involves a change of the voltage on the line capacity from a negative value down to zero and then up to a positive value. The part of the discharge current from the condenser which flows through the receiving relay of the circuit, being in the same direction as the marking line current, tends to sustain the line current. The charging current flowing into the condenser from the +130 volts on the spacing contact of the sending relay is current that is shunted away from the receiving relay, and the build-up of the current to the full spacing value is thus delayed. These two actions combine to make the transition of the line current from marking to spacing a gradual change which is represented by the polar wave shapes of Figure 184. The transition of the line current from spacing to marking may be analyzed in a similar manner to show the cause of the gradual change in this case.

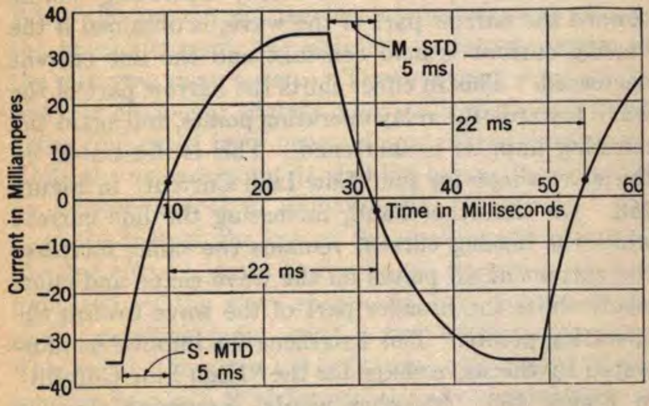


FIG. 184. SIGNAL WAVE SHAPES IN POLAR TELEGRAPH CIRCUITS

The fact that the M-S and S-M wave shapes are identical in form is a valuable feature of polar operation. To obtain the full advantage of this feature, however, the relay operating points must be symmetrically located on the wave shape. That is, the S-M relay operating point must be located the same distance from the start of the S-M wave shape, as the M-S relay operating point is located from the start of the M-S wave shape. These relay operating points will then be the same distance on each side of the zero current line of the wave shape diagrams. The S-MTD and M-STD are equal and there is no bias in the received signals.

Unbiased polar transmission thus depends upon three conditions—(1) that equal but opposite potentials be applied at the sending end; (2) that the resistance of the circuit remain constant for both positions of the sending relay armature, and (3) that the operating points of the relay be located symmetrically about the middle of the wave shape in order that equal transition delays will be secured.

Figure 185 shows a case where the steady state marking and spacing currents of a polar circuit are not equal, the marking current being +40 mils and the spacing current being −30 mils. This condition might be due to a difference in ground potential between the terminals or to an unbalance between the voltages on the contacts of the sending relay. In this case a S-M transition starts when the line current is at −30 mils and ends when the current reaches the relay operating point, while the M-S transition starts when the line current is +40 mils and ends when the current reaches the other relay operating point. When the relay is properly adjusted, the relay operating points will be the same distance on each side of the zero current line in the wave shape. The total current change of the S-M transition is slightly over 30 mils while that of the M-S transition is slightly over 40 mils. As the rate of change (slope of curves) in the two directions is still the same, the delay to the M-S transition will obviously be greater than the delay to the S-M transition. The effect of this condition on transmission is that each mark, regardless of length, will be lengthened by an amount equal to the difference between the two transition delays, and each space, regardless of length, will be shortened by the same amount.

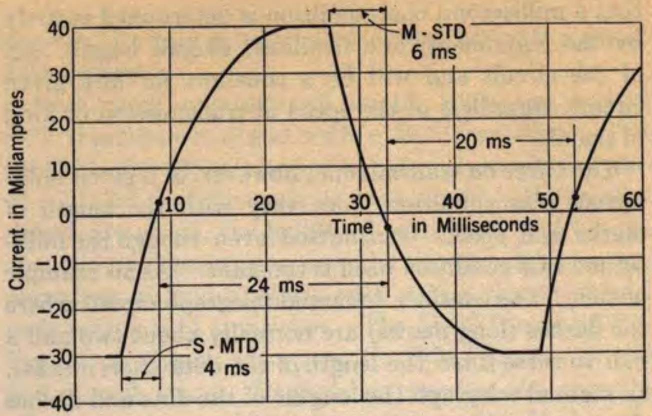


FIG. 185. EFFECT ON SIGNAL LENGTHS OF UNEQUAL POLAR LINE CURRENTS

If the bias condition of the circuit were reversed, which would be the case if the spacing current were greater than the marking current, the delay to the S-M transitions would then be greater than the delay to the M-S transitions. Under this condition all marks would be shortened and all spaces would be lengthened and a spacing bias would exist.

A situation similar to the one just described would have existed if the steady state current values had remained normal and the relay operating points had been shifted one way or the other on the wave shape. This could be caused by a biased adjustment of the relay which, if it were marking would cause the relay to operate to marking more easily than usual, and would thus shift the S-M operating point down on the wave shape. By the same token the relay would operate to spacing less readily, thus requiring more spacing current to operate it, and shifting the M-S operating point down on the wave shape also. This shifting of the operating points would once again make the transition from the marking condition to the M-S operating point on the wave shape different from the transition from the spacing condition to the S-M operating point on the wave shape. Unequal transmission delays and bias to transmission would result, just as in the previous case.

In either case, the important thing to note is that though the M-S transition delays are different than the S-M transition delays, both sets of delays are constant in themselves. The difference between the two delays, which determines the amount of bias on the circuit, is therefore also a constant. Thus if a circuit condition like the one described results in a M-STD of 5 ms. and a S-MTD of 3 ms., every M-S transition sent over the circuit will have a delay of 5 ms. and every S-M transition a delay of 3 ms., regardless of the interval of time that may exist between transitions.

CHAPTER XIV

TELEGRAPH TRANSMISSION PRINCIPLES—(*Continued*)

94. Characteristic Distortion

In the discussion so far, a transition has been always assumed to start when the line current was at the steady state (full value) marking or spacing condition. There are situations, however, where the start of the transition does not occur when the line current is at its steady state value. As we know, a definite amount of time is required for the line current to change from the steady state marking condition to the steady state spacing condition, and vice versa. Thus in Figure 184 the time required for the current to make the complete change from marking to spacing and from spacing to marking is approximately 18 ms. On each transition in this case, the line current would have plenty of time to reach the steady state value before the next transition occurred. The following transition would then start from the same current value as the preceding transitions and the transition delay would be the same as the previous delays.

In actual practice, the time required for the current to change from one steady state condition to the other is sometimes greater than the minimum time interval between transitions in the signals. Some transitions then must occur while the line current is still in the process of changing from the previous transition. These transitions have a different delay time from transitions starting when the line current is in the steady state condition and must therefore be distinguished from the latter type.

Figure 186-A illustrates a case where the line current requires 33 ms. to change from the steady state spacing condition to the steady state marking condition. Now assume that a marking impulse 22 ms. long is being transmitted. The S-M transition at the start of the marking impulse occurs when the line current is in the steady state spacing condition of −35 mils. This transition is thus a steady state current transition, and as such will have the normal S-M transition delay, which is the same for all steady state S-M transitions.

The S-M transition at the beginning of the marking impulse starts the current changing towards the steady state marking current value, an action which in this particular circuit will require 33 ms. to complete. However, the M-S transition at the end of the marking impulse occurs only 22 ms. later. At this time the line current, in the process of changing from −35 mils to +35 mils, has reached a value of +25 mils. The operation of the sending relay at the end of the marking impulse reverses the voltage applied to the line, and the line current accordingly ceases changing towards the marking condition, and starts back towards the steady state spacing condition. Since this M-S transition occurs when the line current is still in the process of changing, it is called a "changing current transition". When the line current reaches the value of −3 mils, the receiving relay operates to spacing, completing the M-S transition on the circuit.

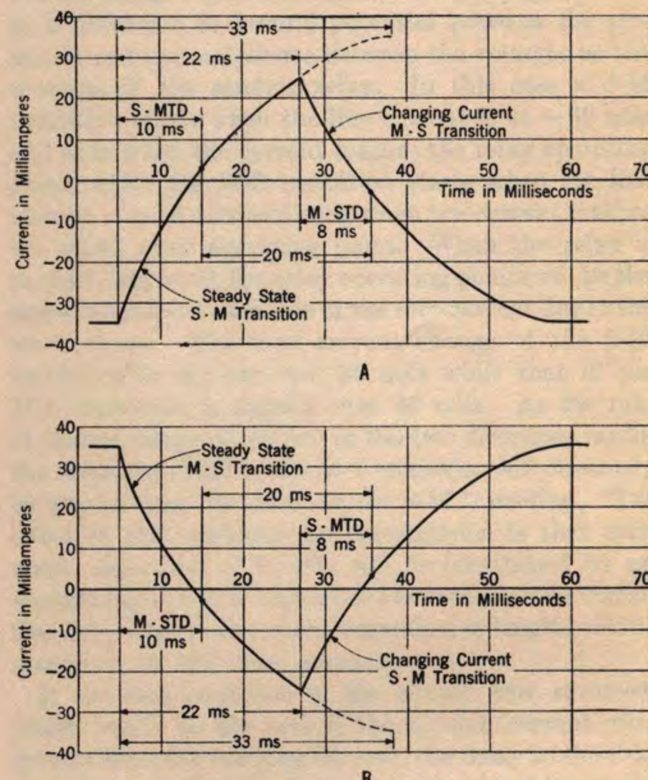


FIGURE 186

The net effect on the marking impulse being transmitted in Figure 186-A will be to shorten it 2 ms., since the transition delay at the end of the impulse (8 ms.), which adds to the impulse, is 2 ms. less than the transition delay at the start of the impulse (10 ms.), which subtracts from the beginning of the impulse.

In a polar circuit, the rate of change of the current from spacing to marking is, of course, the same as the rate of change from marking to spacing. Accordingly, since in this particular circuit 33 ms. were required for the current to change from −35 to +35 mils, 33 ms. will also be required for the current to change from +35 mils to −35 mils. It also follows, then, that if a spacing impulse only 22 ms. long is transmitted, the S-M transition at the end of the impulse will occur when the current is still in the changing condition and this transition will be a changing current transition.

Since, in the case of the current changing from spacing to marking, the value of the current at the end of 22 ms. was +25 mils, it follows that in this case the current at the end of 22 ms. will be −25 mils. This condition is illustrated in Figure 186-B. The total current change involved in the S-M transition at the end of the spacing impulse will then be from −25 to +3 mils or 28 mils, the same as the total current change that took place in the former case. Likewise, the delay to this changing current transition will be 8 ms. and the marking impulse being transmitted will then be reduced 2 ms. in length.

The magnitude of the changing current transition delays just discussed is proportional to the time required for the current to change from its value at the start of the transition to the operating point value of the receiving relay. In both Figures 186-A and -B the current change was from 25 mils to 3 mils of the opposite sign, or a total change of 28 mils. It is obvious, however, from an inspection of these figures, that if the impulse transmitted had been longer than 22 ms., the line current would have been at a higher value at the time of the transition at the end of the impulse, and the transition delay would have been greater. The limiting delay will, of course, be the steady state delay.

Also if the impulse transmitted had been less than 22 ms. in length, the line current would have been at a lower value at the time of the transition at the end of the impulse, and the transition delay would accordingly have been less. This is illustrated by Figure 187 which shows wave shapes of marking impulses for the three standard teletypewriter speeds in a circuit where the time required for the line current to change from its negative to positive value, and vice versa, is 33 ms. The marking impulses illustrated are 18 ms. long, corresponding to 75 speed operation; 22 ms. long corresponding to 60 speed operation; and 33 ms. long corresponding to 40 speed operation. Wave shapes for the spacing signals would, of course, be identical except for reversal of the current values.

In the case of the 33 ms. marking impulse, the impulse is just the required length for the current to change from one steady state condition to the other, and the transition at the end of the impulse is thus a steady state transition. In the case of the 22 ms. impulse the S-M transition at the end occurs when the line current is at +25 mil value, and this transition is thus a changing current transition starting at a current value less than the steady state value. Accordingly as we noted before, the delay is less than the delay to the steady state current transition, 8 ms. as compared to 10 ms. In the case of the 18 ms. impulse, the S-M transition occurs when the line current is only at +18 mil value. This transition is thus also a changing current transition. Due to the fact that the line current only changes 21 mils to reach the M-S operating point of the relay, as compared to the change of 28 mils for the M-S transition of the 22 ms. impulse, the delay is still less. As indicated in the figure, it is now only 7 ms.

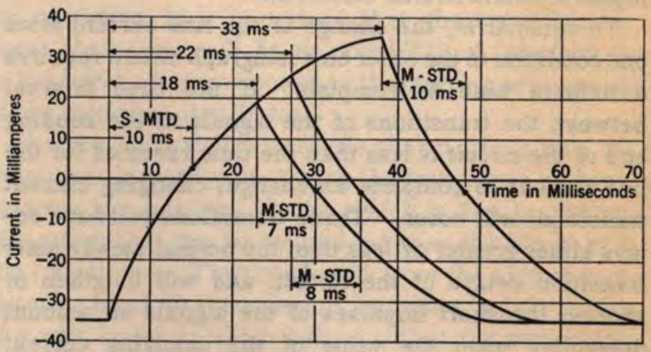


FIG. 187. CHARACTERISTIC DISTORTION EFFECTS ON SIGNAL LENGTHS AT 40, 60, AND 75 SPEED OPERATION

The amount of a changing current transition delay is thus dependent upon the value of the line current at the start of the transition. The value of the line current is dependent upon the time interval between the changing current transition under discussion and the previous transition, which started the line current to changing. Since the time interval between the beginning of these two transitions is equal to the length of the sent impulse, it is this impulse length which finally determines the transition delay under a given set of conditions.

In the condition just described, the lengths of the received signal impulses are obviously affected by the presence of the changing current transitions. This effect is called **characteristic distortion.** The magnitude of the effect is inversely proportional to the length of the sent impulses, and the nature of the effect is to shorten received **short** impulses. Since the received impulses under consideration are shortened, the effect in this case is called **negative characteristic distortion.** An opposite effect is possible. The characteristics of a circuit may be such that the line current tends to increase momentarily at the completion of each transition to a value greater than the steady state, due to transient effects. If the next transition occurs at such an instant, the transition delay will be greater than the delay on the preceding transition which means the length of the received mark or space signal, as the case may be, will be lengthened. Since the signal impulse is lengthened, this is called **positive characteristic distortion.** However, as the transient effect causes the line current to oscillate (increase and decrease) around the steady state, it is possible that the next transition might occur at the instant the line current had momentarily decreased below the steady state. In such a case the transition delay would be less which would result in negative characteristic distortion. Thus, transient conditions may cause either positive or negative characteristic distortion, but positive characteristic distortion is not so frequently encountered as negative characteristic distortion.

To summarize, **the change of the line current from one condition to the other on a telegraph circuit requires a definite time to complete. If the time interval between the transitions of the signals at the sending end of the circuit is less than the time required for the line current to complete its change, changing current transitions will occur. These transitions will have delays either greater or less than the normal steady state transition delays of the circuit, and will lengthen or shorten the short impulses of the signals an amount depending upon the value of the changing current transition delay, which in turn, is dependent upon the length of the impulse that caused the changing current transition. If the effect is to shorten the short impulses it is negative characteristic distortion. If the effect is to lengthen the short impulse it is positive characteristic distortion.**

The contrasts between characteristic distortion and bias are as follows:

1. The effect of characteristic distortion depends upon the length of the impulses transmitted. The effect of bias is independent of the length of the impulses.
2. For a given length of impulse, the effect of characteristic distortion is independent of whether it is a marking or spacing impulse. The effect of bias is always opposite on a mark to what it is on a space.
3. Characteristic distortion is related to the amount and arrangement of the capacity, inductance and resistance of a circuit. Except in neutral operation, these factors do not effect bias.
4. Bias is caused by unequal marking and spacing line current, biased relays, etc., conditions which do not effect characteristic distortion.
5. Characteristic distortion, because it is due to the capacity, inductance and resistance of a circuit, which, except for the resistance, are unchanging in value, varies only a small amount from day to day on a circuit. Bias, because it is caused by unbalanced voltages, ground potential, relays losing adjustment, etc., may vary from hour to hour on a circuit.

95. Fortuitous Distortion

The form of distortion, caused by such factors as crossfire, power induction, momentary battery fluctuations, "hits", break key operation and the like, and which displaces miscellaneous received transitions by various amounts intermittently, is known as **fortuitous distortion.** At times this effect may be large enough to produce a complete failure of the circuit. In the transmission of miscellaneous signals, the combined effect of characteristic and fortuitous distortion on the displacement of received transitions is sometimes known as "jitter".

96. Teletypewriter Margin Measurements

From the preceding discussion, it is apparent there is a need for some means of measuring the quality of telegraph signals as transmitted over various types of circuits and under varying conditions. Within certain limits, the teletypewriter itself may be used as a measuring instrument for this purpose.

As illustrated by Figure 169, only five successive equal signal intervals are required to provide combinations for all the characters normally used. These are supplemented by one equal open interval immediately preceding the group of five, for starting the rotation of the receiving distributor cam or brush arm; and, a closed interval immediately following, for the purpose of stopping the rotation of the receiving distributor after the group of five intervals have operated the selecting mechanism of the receiving machine. This closed stop interval is made equal to 1.42 times the length of each of the other six equal intervals. This longer interval insures that, under any condition normally encountered, the receiving distributor will be stopped before the next character combination is received. Using the start and selecting intervals as units, the sending distributor is so constructed that six open or closed intervals of one unit each, and one stop interval of 1.42 units are consecutively produced for each character transmitted.

The principal teletypewriter operating speeds used are 40 and 60 words per minute, with the latter predominating. The average word is assumed to consist of five letters and a space and it therefore requires six revolutions of the distributor brush arm for transmittal. At 60 speed, there are 60 times 6 or 360 revolutions per minute of the brush arm. However, as the brush arm is stopped and started once every revolution, the distributor driving shaft is operated slightly above this speed, i.e., approximately 368 instead of 360 revolutions per minute.

In one complete revolution, which requires 60 Sec./360 Rev. or .163 second per revolution (163 milliseconds), the brush arm passes over 7.42 units. Therefore when operating at 60 speed, the time for each unit signal impulse is approximately 22 milliseconds and the time for the long stop impulse is 31 milliseconds.

Figure 188-A indicates the sequence of circuit conditions produced by the sending distributor in transmitting the letter "R". Here the circular distributor is laid out as a straight line. The shaded areas represent the intervals during which the circuit is closed and the blank sections the intervals during which the circuit is opened by the sending distributor.

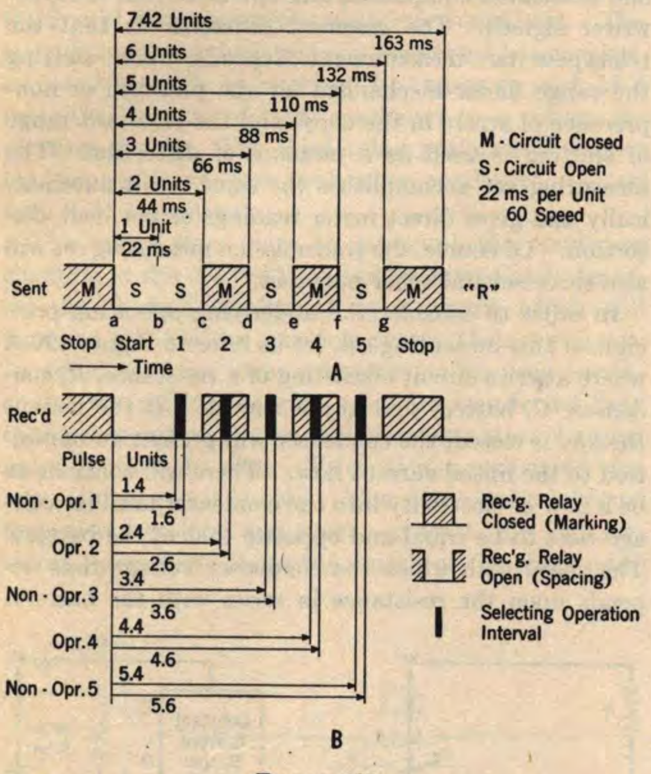


FIGURE 188

The received signals shown in Figure 188-B have the same time lengths as those produced by the sending distributor of Figure 188-A. The solid blocks superimposed upon the received signals represent those parts of the signals which are used by the selecting mechanism of the receiving machine (see Figure 169).

The selecting mechanism, when "oriented" correctly, is so arranged that it normally operates only during the central portion of the received signal impulse and requires only about twenty per cent of the unit interval. On this basis, the selection for pulse number 1 occurs during the period of time 1.4 to 1.6 units after the beginning of the received start interval, the selection for pulse number 2 occurs 2.4 to 2.6 units after the start, and the remaining pulse selections occur in a similar manner 3.4 to 3.6, 4.4 to 4.6, and 5.4 to 5.6 units, measured in each case from the beginning of the received start interval.

For the transmission of the letter "R" as shown in Figure 188-A, there are mark-to-space transitions at points *a*, *d* and *f*, and space-to-mark transitions at points *c*, *e* and *g*. For some other character combination, a transition may occur at point *b*, but in any transmitted character there can be only two, four or six transitions.

Inasmuch as the selecting functions take place only during the intervals shown by the solid blocks of Figure 188-B, and require twenty per cent of each unit interval for operation, it is important that the transitions so occur that there will be no possibility of interference to the selecting operations or to the starting or stopping of the receiving distributor.

For the ideal signal intervals shown, the above requirement is met by producing the transitions midway between the selecting blocks, which is the maximum separation that can be secured between the blocks and the transitions. The time length from the edge of each selecting block to the adjacent transition is four tenths of a unit interval, which indicates that the transitions may be shifted towards the selecting blocks as much as forty per cent of the length of a unit interval before an error is recorded on the machine. Any deviation from the ideal positions for the occurrence of transitions represents distortion and may be measured in terms of its percentage of a unit interval. Thus the above machine is able to tolerate a maximum distortion of about forty per cent.

As we have seen, distortion in the form of bias or characteristic distortion displaces the signal transitions so as to effectively shift the position of the received mark or space impulses in some definitely systematic way.

The ideal situation, of course, is for the selecting segments of the receiving distributor, which are one-fifth the length of the unit segments in the sending distributor, to be at mid-position with respect to the sending units. Under these conditions the transitions may be shifted as much as forty per cent in either direction before an error is recorded by the machine. The receiving unit of the teletypewriter machine is equipped with a mechanism whereby the latch assembly (or distributor face) may be mechanically moved through an arc corresponding to the length of a unit segment. By this means all of the selecting segments may be shifted with respect to the beginning of the start segment (receiving brush arm released) over a scale range equal to a unit segment (22 milliseconds for 60 speed). This mechanism is known as a "range finder" and is equipped with a scale graduated from 0 to 120 as indicated in Figure 189. One hundred divisions on this scale represent an arc equal to a unit segment. This arrangement provides a means of measuring the distortion on received signals.

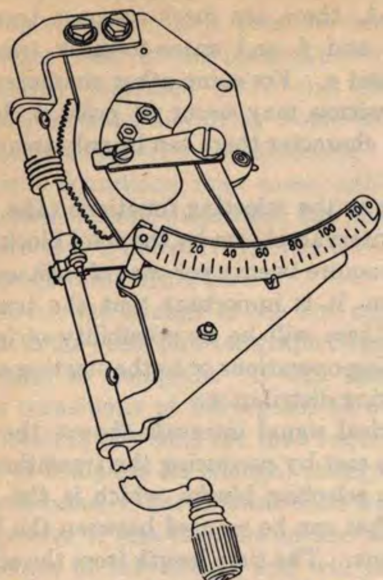


FIG. 189. TELETYPEWRITER RANGE FINDER MECHANISM

To measure the total net effect of all kinds of systematic distortion, or the position of received signals, the range finder is first moved in one direction until errors appear in the "copy" and then moved back slowly until these errors are just eliminated. Similarly, the range finder is moved the maximum distance before errors occur in the opposite direction. These two scale readings then give the operating margin of the signals under test. On perfect signals the margin would be from 10 to 90, since the effective received signal must occupy twenty per cent of the total range.

Margin measurements, in addition to showing the distortion present in the received telegraph signals, also show speed differences between the sending and receiving machines. The effect of a slow sending speed is to cause each unit to be greater than 22 milliseconds and each transition to occur progressively later than it should. The effect on the margin of operation is to raise both limits, the lower limit being raised much more than the upper limit. For example, a margin of 35 to 100 indicates the sending speed is five per cent slow. On the other hand, the effect of a fast sending speed is to cause each unit to be smaller than 22 milliseconds and each transition to occur progressively earlier than it should. The effect on the margin of operation is to lower both limits, the upper limit being lowered much more than the lower limit. For example, a margin of 5 to 60 indicates the sending speed is five per cent fast.

97. Telegraph Transmission Measuring Set

In order to make a more detailed analysis of distortion and to determine such factors as the extent of the displacement of received transitions, a time interval measuring device must be used. A telegraph transmission measuring set has been standardized for this purpose. It employs the condenser charge principle, in conjunction with a brush type distributor and associated equipment, for the measurement of maximum or total distortion, as well as bias, in terms of per cent length of a unit interval signal.

This measuring set may be compared to the teletypewriter when used as a measuring device, on the basis that each has a receiving distributor, receiving relays and associated equipment, and operates from teletypewriter signals. The essential difference is that the teletypewriter measurement depends upon shifting the range finder mechanism for the presence or nonpresence of errors in the copy, and the resultant range of shifting is used as a measure of distortion. The measuring set accomplishes the same result automatically and gives direct meter readings of per cent distortion. Of course, the transmission measuring set will also serve several other purposes.

In order to consider the underlying operating principle of this measuring set, let us refer to Figure 190-A where a series circuit consisting of a resistance, R, condenser, C, battery and key is shown. At the instant the key is closed, the condenser will present an opposition to the initial current flow. There will continue to be a flow of electricity into the condenser until its voltage rises to be equal and opposite that of the battery. The speed with which the condenser voltage rises depends upon the resistance in series with the battery.

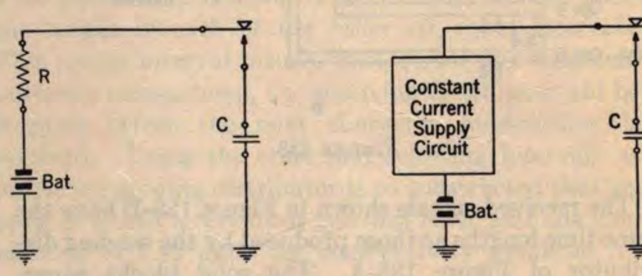


FIGURE 190

If it requires one second to rise to four-eighths of the battery voltage, it will rise to six-eighths in two seconds, to seven-eighths in three seconds, to fifteen-sixteenths in four seconds, etc. If this condenser charge were interrupted and some means used to measure the condenser voltage, it could be determined from the above just how much time had elapsed since the key was closed.

In the measuring set, the time-charge relationship is simplified by arranging a vacuum tube circuit which has a constant output current to charge the condenser, so that the voltage increase is the same for each millisecond of time that the condenser is allowed to charge. This is known as a "constant current supply circuit" and is connected in the circuit as schematically indicated in Figure 190-B.

From the simplified drawing of Figure 191 it may be noted that a polar relay is used as a master relay to repeat the incoming teletypewriter signals into a simple form of one-way polar circuit, in which are located three other polar relays. One of the three closes the start magnet circuit on every mark-to-space transition to release the brush arm of the distributor whenever the first M-S transition of a character is received. The second one of the three relays is connected to a measuring condenser, C_1, to interrupt its charge on each M-S transition and transfer the condenser to a voltage indicating circuit. The third relay performs the same function with another measuring condenser, C_2, on each S-M transition.

The relays thus serve to interrupt the condenser charging at the desired time and to immediately transfer the charged condensers to a circuit which will measure the voltage existing across their terminals. In addition, some arrangement must be used to discharge the condensers when they are transferred back to the charging circuit, so that they may start charging again at the right time. To do this, a segmented ring type of distributor is used. The ring consists of fourteen alternate long and short segments, the long segments being three times the length of the short ones. The speed of the governed motor is adjusted so that it takes 22 milliseconds for the brush arm to pass over a long segment and the adjacent short one.

The short segments are all connected together to ground, but there are no connections to the long segments with the exception of the one on which the brush rests in its stopped position. To this Stop segment is connected what is known as a "Stop Compensator Voltage" which can be varied by means of a potentiometer, the purpose of which will be indicated later. The brush arm is connected to the ungrounded side of the Constant Current Supply so that when the brush arm is resting on a short grounded segment, the condenser connected to the Constant Current Supply Circuit will be completely discharged and will be allowed to start charging as soon as the brush arm moves off the grounded segment.

The segmented ring is oriented so that, when perfect teletypewriter signals are being received, the brush arm is just half way between two successive short grounded segments when any transition occurs. Since it requires seventy-five per cent of 22 milliseconds for the brush arm to pass over a long segment, that segment is said to be seventy-five per cent in length. The charging current is adjusted to such a value that when the measuring condenser starts charging as the brush leaves a short grounded segment, the voltage across its terminals will rise to about 55 volts at the instant that the brush reaches the position midway between two successive segments.

Whenever a perfect transition occurs, a relay interrupts the charging of the measuring condenser and transfers it to the circuit containing the bias meter and

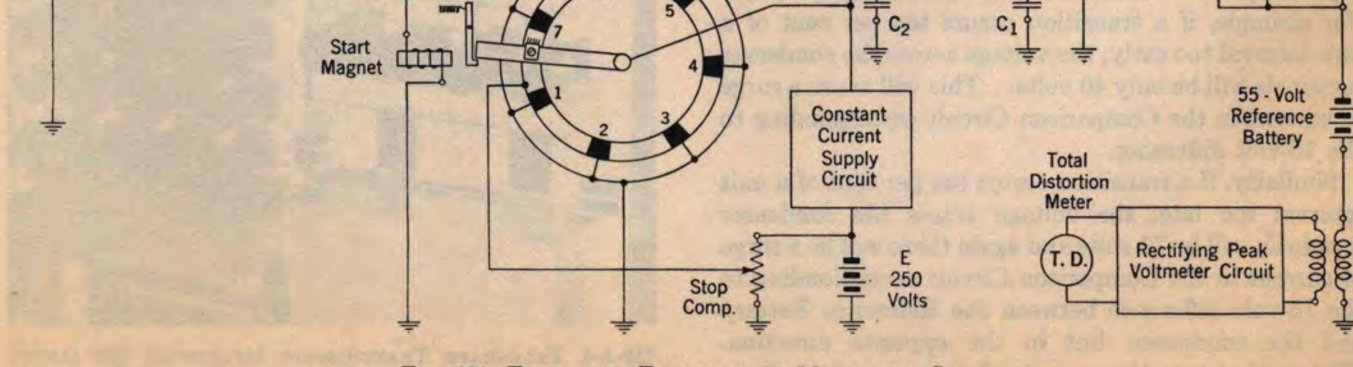


Fig. 191. Telegraph Transmission Measuring Set

reference battery just as the voltage across its terminals has reached 55 volts. The circuit containing the bias meter and 55-volt reference battery is known as the "voltage indicating circuit". If there is no transition when the brush is passing over a blank segment the condenser remains connected to the charging circuit and is charged up to 110 volts, then discharged by the short grounded segment, thereby having no effect in the voltage indicating circuit.

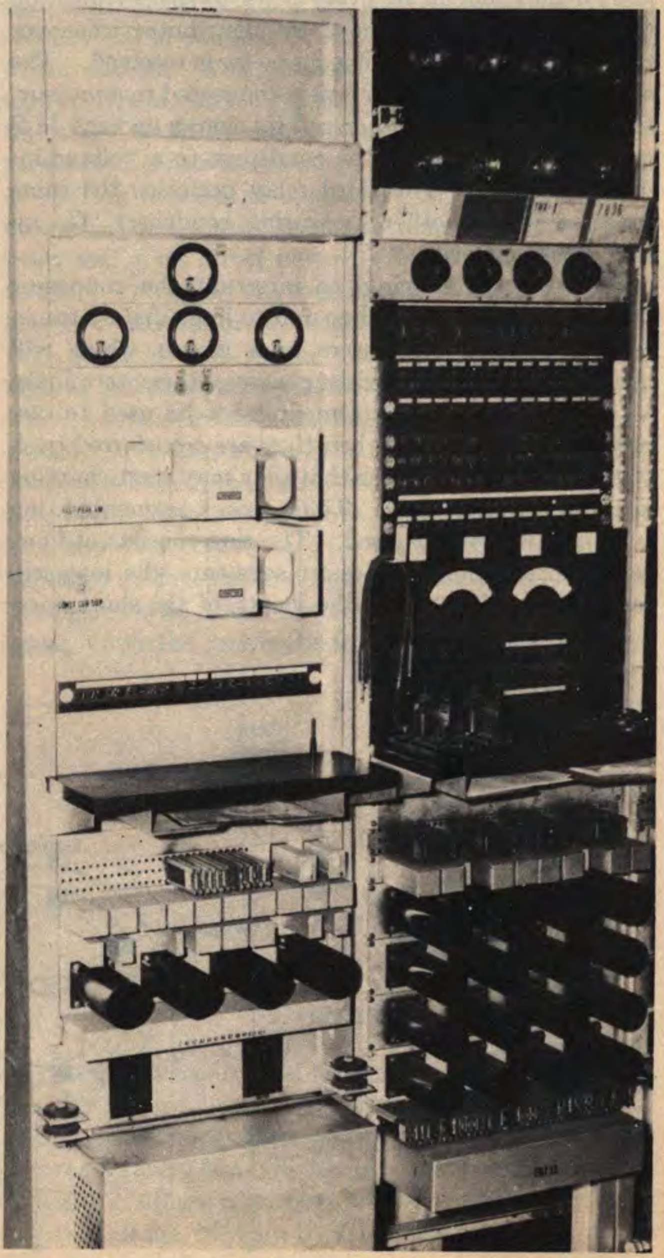
With the 55-volt battery in the voltage indicating circuit poled to oppose the condenser voltage, there will be no flow of current if the voltage of the measuring condenser is also 55 volts when transferred to this circuit by a relay. If the condenser voltage is greater than the battery voltage, there will be a discharge of electricity from the condenser through the opposing battery; but for a condenser voltage less than 55, the battery will cause a current flow into the condenser. The greater the voltage difference, the greater the surge of current. Since the condenser voltage is being compared to the battery, this voltage indicating circuit may be considered as a Comparison Circuit and the battery as a Reference Battery.

Since the reference point is the first M-S transition, and the brush does not rest on a grounded segment in its stopped position, it is necessary to control the voltage across the condenser C_1 so that there will be no indication in the comparison circuit when this first M-S transition occurs. By adjusting the "stop compensator" potentiometer associated with the battery supply E so that the voltage to ground is always equal to that of the reference battery, condenser C_1 will be charged to the reference battery voltage each time the brush passes over the "Stop" segment. Accordingly, there will be no indication in the comparison circuit for the first transition of each character received.

As we know, the effect of distortion on a circuit is to cause the received transitions to occur too early or too late. If the transition occurs too early, the measuring condenser charging is interrupted too soon and there will be less than 55 volts across its terminals when it is transferred to the Comparison Circuit. The value will depend upon how much too soon the transition occurs. For example, if a transition occurs ten per cent of a unit interval too early, the voltage across the condenser terminals will be only 40 volts. This will cause a surge of current in the Comparison Circuit corresponding to the 15-volt difference.

Similarly, if a transition occurs ten per cent of a unit interval too late, the voltage across the condenser terminals will be 70 volts and again there will be a surge of current in the Comparison Circuit corresponding to the 15-volt difference between the Reference Battery and the condenser, but in the opposite direction. With perfect transitions received, however, the voltage across the terminals of the measuring condensers will always be 55 volts at the instants that a relay interrupts the charging and transfers them to the Comparison Circuit. In this case there will be no current flow in the indicating circuit.

In order to determine the amount of any distortion present in the signals, some arrangement must be used to measure the momentary voltage differences. The ordinary voltmeter would not be satisfactory because the voltmeter needle would not have enough time to reach a steady reading. A special vacuum tube circuit known as a "rectifying peak voltmeter circuit" is used for this purpose. The indicating meter included in this



118-A-1 Telegraph Transmission Measuring Set (left) Adjacent to Voice-Frequency Carrier Bay circuit then reads the total distortion present as a percentage of unit signal.

A "bias meter" (scale 25–0–25) is connected in series with the 55-volt reference battery. With spacing bias in the received signals, all of the space-to-mark transitions occur later than they should, allowing condenser C_2 to rise to a voltage higher than 55 volts; therefore, there will be a discharge current out of the condenser of the same magnitude each time a space-to-mark transition transfers C_2 to the Comparison Circuit. These discharge currents will cause the meter needle to swing to the spacing side of zero. The larger the spacing bias the larger the discharge currents will be and the farther to the left the meter needle will swing. With marking bias in the received signals, all of the space-to-mark transitions will occur too early, preventing condenser C_2 from rising to 55 volts. Therefore, there will be a charging current in the opposite direction from the reference battery into the condenser each time a space-to-mark transition occurs, causing the meter needle to swing to the marking side of zero an amount depending upon the magnitude of the marking bias.

A variable shunt is provided across the bias meter so that the position of the meter needle can be made to read directly the per cent bias. The meter shunt is adjusted to read correctly on four transitions per character, since the average number of transitions in miscellaneous signals is four. On two transition characters, it will read half as much as it should; on four transition characters, it will read correctly; and on six transition characters, it will read one and one-half times what it should.

For miscellaneous 60 speed teletypewriter signals where only bias exists, the average indication of the bias meter will be spacing or marking and the total distortion meter will read the same magnitude as the bias meter. However, the bias meter needle fluctuates in accordance with the number of transitions in the signals while the total distortion meter reading is steady. As noted above, the reading of the bias meter will depend upon the number of transitions per character in the signals being received.

With distortion other than bias in the signals, the bias meter needle will fluctuate over a wide range but its average position will be zero. On the other hand, the total distortion meter will give a steady reading of the maximum distortion present, but the observation must obviously be made over a period of time to obtain an accurate indication.

With both bias and other forms of distortion in the miscellaneous teletypewriter signals, the bias meter needle will fluctuate over a wide range to the right or left of zero but its average position will still give a fair indication of the bias present. The total distortion meter will indicate the sum of the bias and other forms of distortion with a steady reading of the maximum distortion present. Observation over a period of time is required to estimate the bias meter average reading, as well as to obtain an accurate indication of the maximum distortion. The readings are usually recorded with the total distortion meter reading first, followed by the sign and magnitude of the average bias meter reading,—thus 15M10, meaning 15% total distortion and 10% marking bias.

CHAPTER XV

ALTERNATING CURRENTS

98. Source of Alternating E.M.F.

In taking up the study of alternating-current flow, we shall follow closely the same course as was followed in the study of direct currents. The theory will precede the applications, and step by step we shall pass from the simple circuit to the network, from the network to the transmission of electrical energy, and thence to our ultimate aim, which is the application of these to the transmission of human speech. But along with this procedure, we shall study wherein the nature of alternating-current work differs from that of direct-current work. Perhaps the first such difference lies in the source of E.M.F.

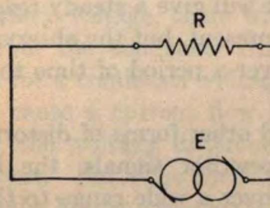


FIGURE 192

Figure 192 represents an alternating-current circuit in its simplest form. In this figure we have a new convention for source of E.M.F., which represents the collector rings of a generator. Unlike the battery or other simple form of direct E.M.F., we cannot describe such a source of E.M.F. by simply giving its voltage, for example E = 10 volts. Here we have a voltage gradually increasing to a maximum value, and then decreasing to zero, to again increase to a maximum value in the opposite direction, and again decrease to zero, where the cycle repeats itself. Even if we knew the maximum voltage value, we should not know the trend of the successive values from zero to the maximum value. Figure 193 illustrates cycles of alternating E.M.F.'s all very different in this respect.

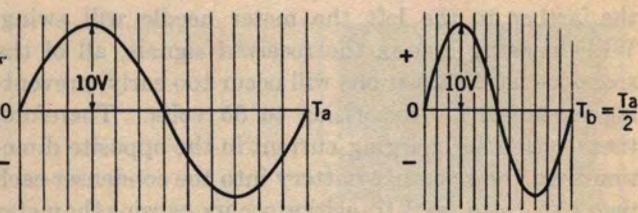


FIG. 194. SINE WAVES OF DIFFERENT FREQUENCY

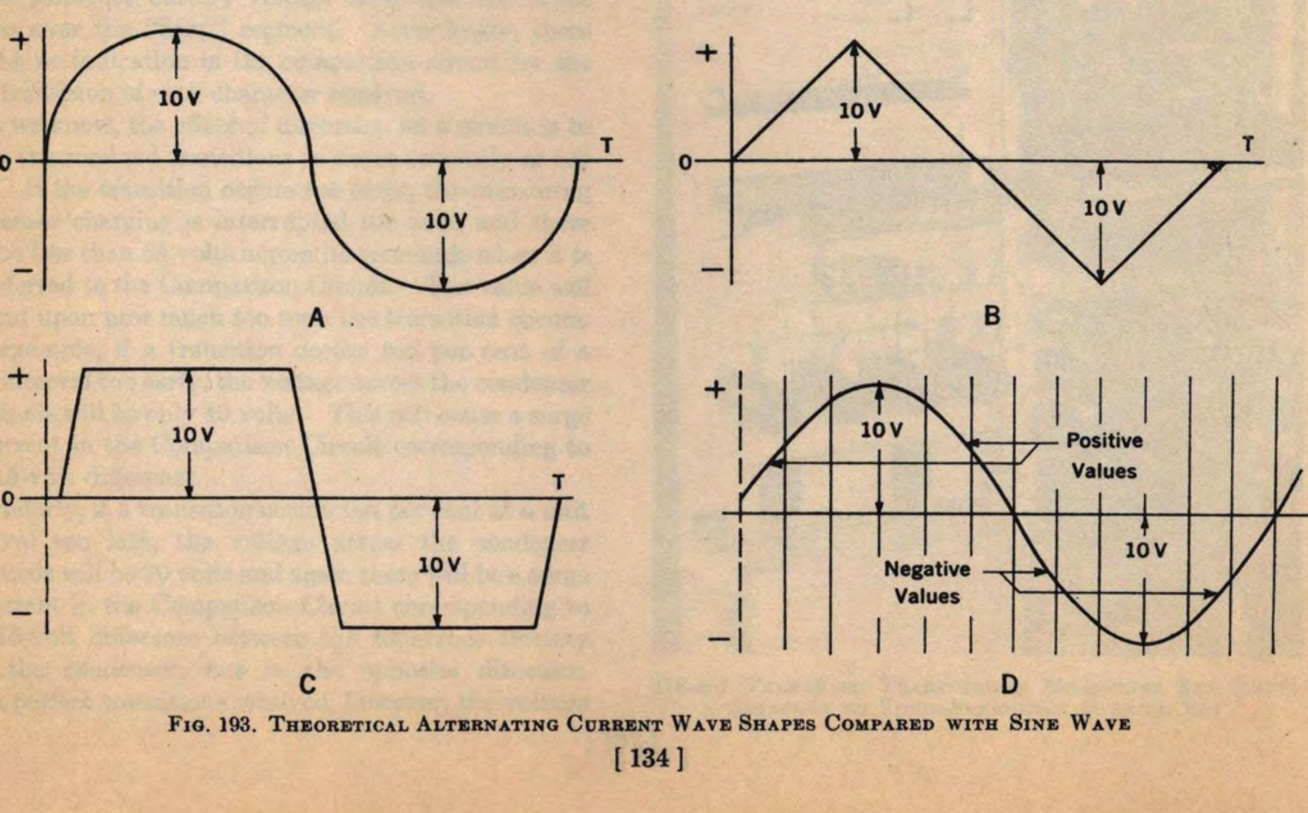


FIG. 193. THEORETICAL ALTERNATING CURRENT WAVE SHAPES COMPARED WITH SINE WAVE

Furthermore, we should not know the rapidity with which the alternations are taking place. For example, Figure 194 represents two cycles of identical E.M.F. values, but in one case the cycle is completed in one-half the time required for the other. Therefore to describe electrically a source of alternating E.M.F. we must know the following:

a. **The wave shape of the alternating cycle.**
b. **The value of the E.M.F. at some specified point in the cycle.**
c. **The length of time to complete the cycle, or the frequency.**

In classifying electrical currents in Chapter VIII, we named two steady state conditions for alternating current; one where the wave shape is a sine wave and the other where the wave shape is not a sine wave but a complex wave. The basic study of alternating-current circuits deals only with sine waves. Complex waves are analyzed into sine waves, just as complex tones are analyzed into **fundamentals and harmonics** (see Appendix IV).

99. The Sine Wave

The sine wave is named from a trigonometric function of an angle. We have learned how it may be constructed graphically, and we may treat it as a "pattern" having a name with a mathematical origin to which an E.M.F. or current may or may not conform, rather than as a mathematical expression requiring a thorough knowledge of trigonometry for interpretation. It has interesting properties and is the natural wave form in all vibratory motion. It greatly simplifies alternating-current circuits because—**a sine wave E.M.F. impressed upon a circuit having a network of any number and arrangement of resistances, induc-**

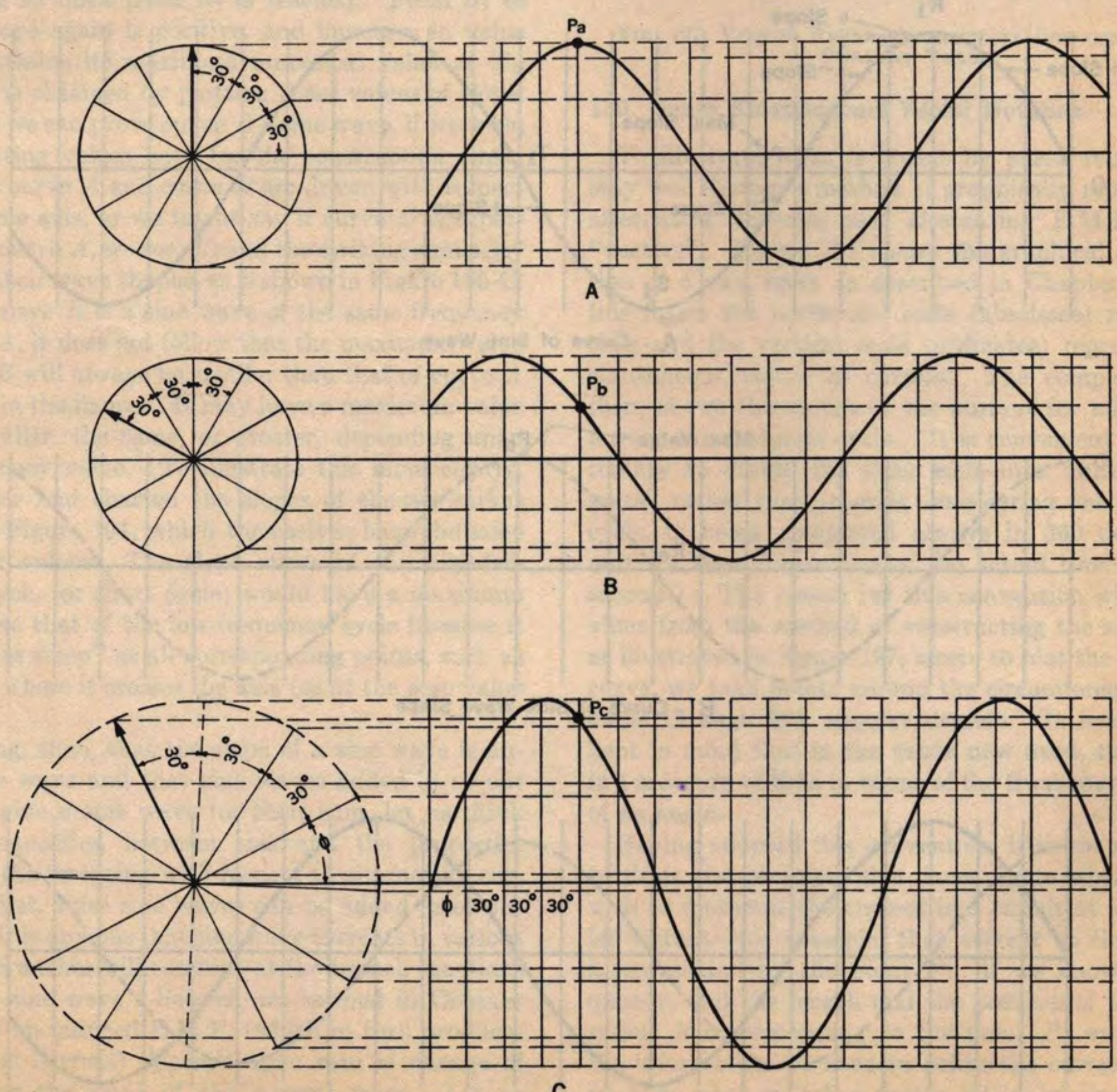


Fig. 195. Graphical Proof that the Sum of Two Sine Waves is a Sine Wave

tances, and capacities with fixed values, will set up a sine wave current in every branch of the network. No other wave shape (excepting that of direct current) will give the same wave shape for the current as that for the impressed E.M.F.

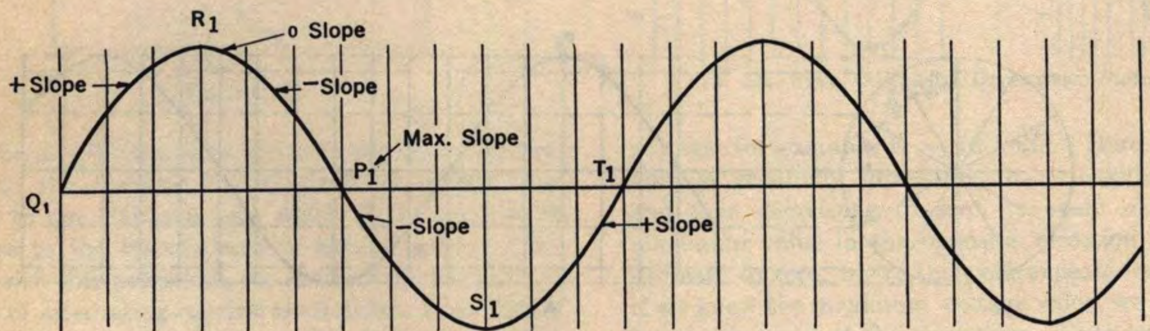
The above rule holds in all its applications since the sine wave possesses the following properties:

a. **Sine waves of the same frequency can be added (or subtracted) either in or out of "phase" and the wave shape of the result will be a sine wave.** (Phase relations are defined in the next article.)

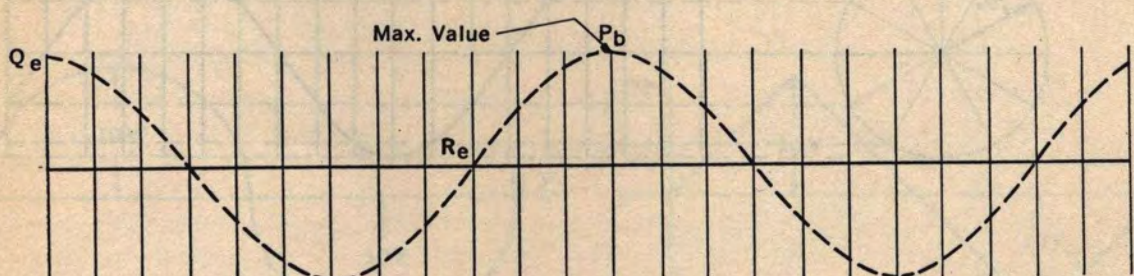
b **A sine wave E.M.F. across a resistance, inductance or capacity gives a sine wave current through the resistance, inductance or capacity (though not necessarily in phase).**

c. **Whenever an E.M.F. is induced on account of the ever-changing value of a sine wave current, this induced E.M.F. is a sine wave (though not in phase).**

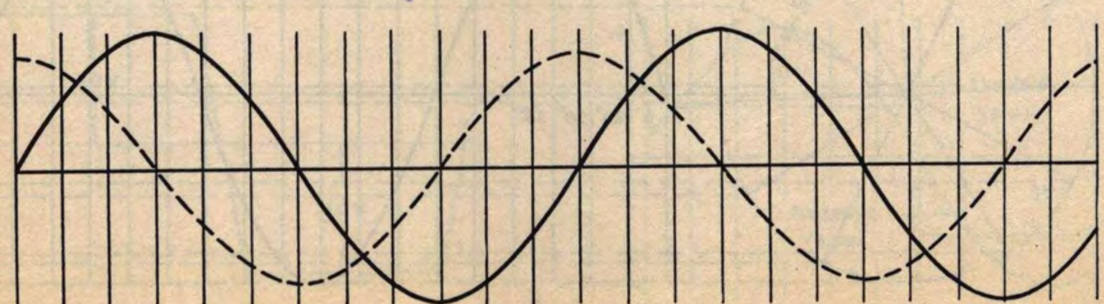
A graphical proof of property *a* may be had by referring to Figure 195. Here *A* shows a sine wave constructed graphically in the manner explained in connection with Figure 73. *B* is a similar sine wave of the same frequency but constructed independently of *A*. Now if each value on the *A* curve, for example that represented by P_a, is added to each corresponding value on the *B* curve, for example that represented by P_b, the resultant curve will be that shown as *C*. An inspection of *C* will show that it too is a sine wave and can be proven so by constructing a circle from values of the curve projected back. This is, of course, the



A - Curve of Sine Wave



B - Curve of Sine Wave Slope



C - Curves A & B Compared

Fig. 196. Graphical Proof that the Slope of a Sine Wave is a Sine Wave

converse of the construction of the sine wave and proof of the wave shape, since there can be only one circumference drawn through the projected points of intersection with the radii of the circle.

The properties of a sine wave given under *b* and *c* in the foregoing can be demonstrated graphically by determining the rate of change or "slope" of a sine wave at various points and plotting the successive values of the slope as shown in Figure 196-B. When the value shown by the curve in Figure 196-A is zero, as at point Q_1, the rate of change or slope is greatest. At any point between Q_1 and R_1, the slope is positive and is decreasing to its minimum value, zero, at point R_1. Between R_1 and P_1 the slope is negative and is increasing, attaining its maximum numerical value at P_1. After passing through P_1, the slope again decreases in magnitude, but is still negative, remaining so until point S_1 is reached. From S_1 to T_1 the slope again is positive, and increases in value until it attains its maximum numerical value at T_1. Curve B is obtained by plotting these values of slope. As before we can prove curve B a sine wave, if we wish, by projecting values back for the construction circle, but if the curve A and curve B are drawn with respect to the same axis, or we might say if curve B is superposed on curve A, we see offhand the striking similarity between their wave shapes, as is shown in Figure 196-C. Though curve B is a sine wave of the same frequency as curve A, it does not follow that the maximum value of curve B will always be smaller than that of curve A as shown in the figure. It may have a maximum value either smaller, the same, or greater, depending upon the frequency value. To illustrate this more clearly, suppose we had charted the slopes of the two curves shown in Figure 194, which themselves have the same maximum values. The slope curve of the high-frequency cycle (or short cycle) would have a maximum value twice that of the low-frequency cycle because it is "twice as steep" at all corresponding points, such as the point where it crosses the axis (or at the zero value point).

Granting, then, that the slope of a sine wave is another sine wave and that sine waves added in or out of phase give a sine wave for their sum, let us think of the connection between this and the properties stated in the foregoing with respect to alternating currents. First, since sine waves can be added in or out of phase, it is obvious that sine wave currents in various network branches will combine at the branch junctions to give a sine wave. Second, we learned in Chapter VIII that an induced E.M.F. (which in turn produces an induced current) depends upon rate of change of current, or the slope of the current wave. Let us assume a network of inductances with a sine wave E.M.F. impressed. In each individual inductance there is a sine wave induced E.M.F., if the current is a sine wave. But there will be a sine wave current because the induced current adds to or subtracts from the current due to the impressed E.M.F., and the sum or difference of two sine waves is a sine wave. Thus we can analyze all the currents in the branches of any network by either an application of sine wave addition and subtraction, sine wave slopes, or a combination of the two properties.

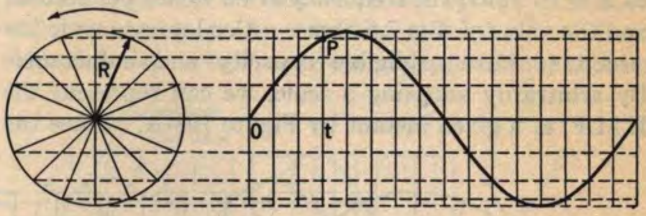


FIG. 197. VECTOR REPRESENTATION OF INSTANTANEOUS CURRENT VALUE

100. Phase Relations and Vector Notation

To illustrate what is meant by phase relation, we may well discuss a method of graphically representing alternating currents and alternating E.M.F.'s with "vectors". Figure 197 shows the graphical construction of a sine wave as described in Chapter VI. In this figure the horizontal scale (abscissae) represents time and the vertical scale (ordinates) represents instantaneous values of current. The complete curve then, shows the values of the current for all instants during one complete cycle. It is convenient and customary to divide the time scale into units of "degrees" rather than seconds, considering one complete cycle as being completed always in 360 degrees or units of time (regardless of the actual time taken in seconds). The reason for this convention will be obvious from the method of constructing the sine curve as illustrated in Figure 197, where to plot the complete curve, we take points around the circumference of the circle through 360 angular degrees. It needs to be kept in mind that **in the sense now used, the degree is a measure of time in terms of the frequency, and not of an angle.**

Having adopted this convention, it is not necessary to draw the complete sine curve figure whenever we wish to represent the current in a circuit at a particular instant—for example, that current at the instant *t*, represented by the point *P*. If we know the frequency, and the length and the position of the single radius *R* corresponding to the point *P*, we have all the information we need to define the current. Here we have what we call a vector, which we can imagine as a radius of the circle, having a length equal to the maximum current or E.M.F. value of the sine wave in question. The angle this vector makes with the horizontal gives the position of point P and if we assume a direction of rotation for the vector, we can always determine by the position of the vector whether the value of the current or E.M.F. is increasing or decreasing, and its direction. The accepted convention for direction of rotation is counterclockwise and will be understood hereafter, without the arrow being used to indicate it.

In Figure 192, let us assume that the maximum value of E is 10 volts, the frequency is 60 cycles per second, and the value of R is 7.5 ohms. Also let us assume the circuit to have negligible capacity and inductance. By arbitrarily adopting a scale, we can represent the E.M.F. at a given instant by Figure 198-A. Since the inductance and capacity of the circuit are negligible, the current at the corresponding instant will neither be retarded by inductance nor have a component part required to "charge" the circuit. It will be that determined solely by Ohm's Law. Consequently, it will change in value as the E.M.F. changes in value. In other words, it will "keep in step", becoming a maximum of 1.33 amperes at exactly the same time that the E.M.F. becomes a maximum of 10 volts, and becoming zero at exactly the same time that the E.M.F. becomes zero. **The conventional expression to describe this time relation between the voltage and the current is that the voltage and current are in "phase".**

But if, instead of a circuit such as that shown by Figure 192, we have the circuit shown by Figure 199, it will be necessary to consider the effect of the in

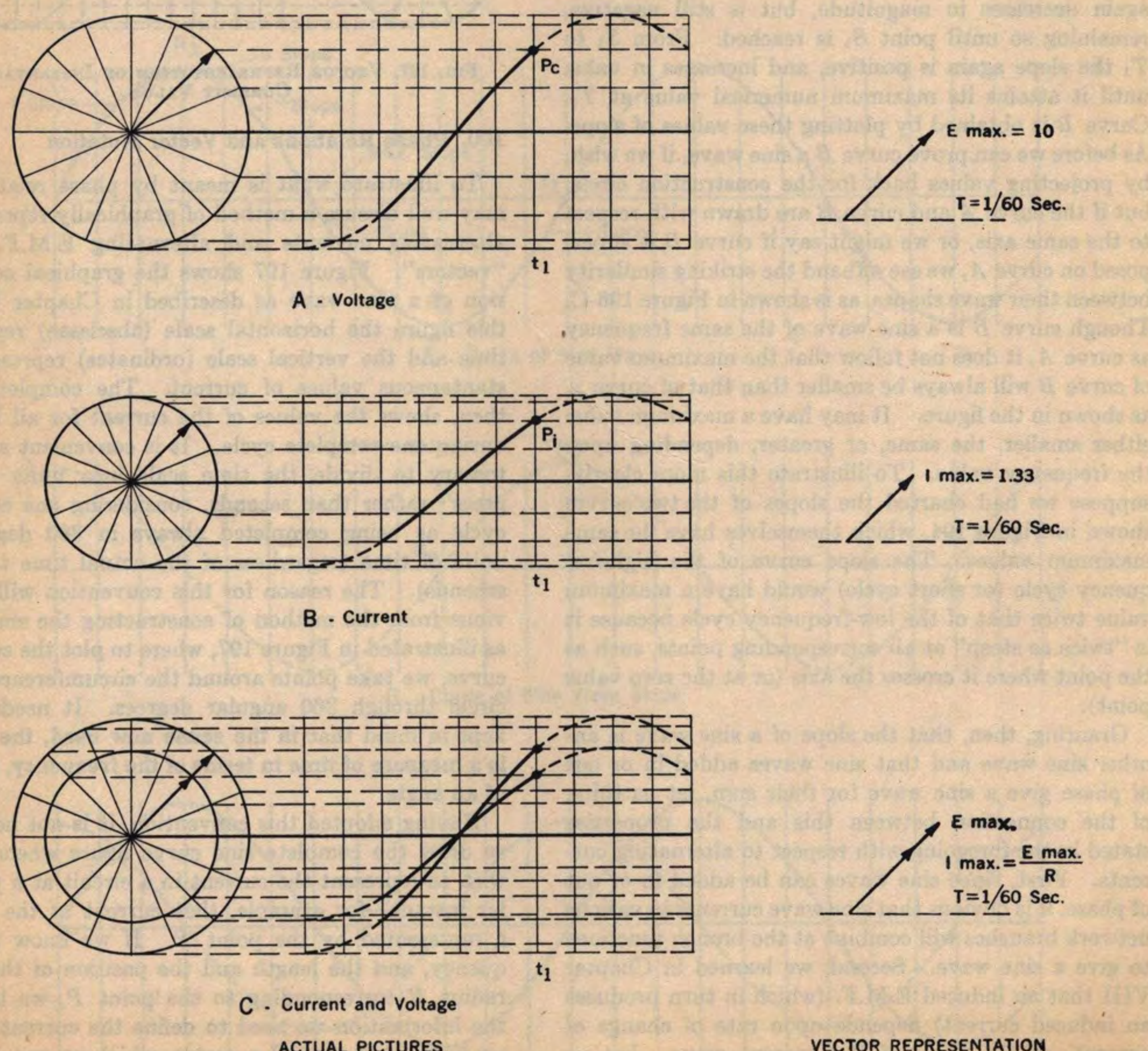


Fig. 198. Current and Voltage in Phase

ductance. This reacts to any change in current value, and an alternating current is changing in value at all times. We should therefore expect the inductance to materially affect the value of the current and to throw the maximum points out of step, or phase, because the maximum value of current will not have been established until some time after the E.M.F. has reached

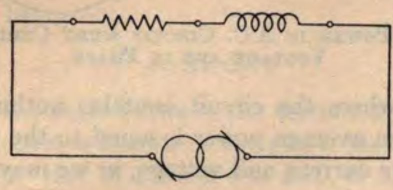


FIGURE 199

its maximum value. Figure 200 represents the relation of voltage and current that are out of phase due to the circuit having inductance. Here the vectorial representation must show the extent to which the voltage and current are out of phase. This is accomplished by having the voltage vector ahead of the current vector in its rotation by an angle which is a measure of the time by which the current "lags" behind the voltage, and whose value is obvious from the relative position of the radii of the two circles.

In the case of a circuit having a series condenser instead of an inductance, the circuit reactions are the reverse. The current vector then is ahead of, or "leads", the E.M.F. vector as shown by Figure 201. Electrical conditions in circuits containing inductance or capacity, therefore, can be represented by current and voltage vectors, which will, in general, be out of phase. Moreover, in dealing with complex networks containing inductance or capacity, we encounter current vectors which are out of phase not only with their voltage vectors, but with each other.

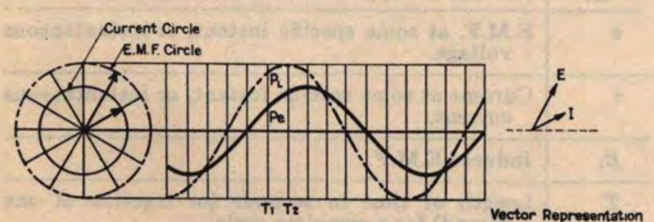


FIG. 200. CURRENT LAGGING IMPRESSED E.M.F.

In direct-current networks, we used equations based on Kirchoff's Laws which called for adding or subtracting current or E.M.F. values. In alternating-current work, we cannot accomplish this by merely adding the numerical lengths of the vectors. We must instead combine them in such a manner as to take into consideration any phase differences that may exist. This may be done graphically by placing the vectors to be added end to end, and drawing a line from the butt of the first arrow to the tip of the last. This line, called the resultant, is a vector which gives the magnitude and phase of the sum. For example, let us assume that it is desired to find the current delivered by the generator of Figure 202, when the currents in the parallel branches have the values and phase relationships indicated by vectors 1, 2, and 3. These vectors are placed end to end and the resultant drawn as indicated in 4. The length of this resultant vector gives the value of the current delivered by the generator and its angular position indicates its phase relationship with respect to the current in the parallel branches.

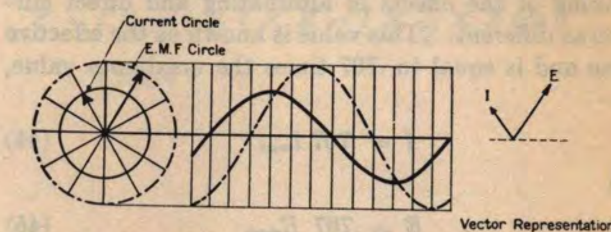


FIG. 201. CURRENT LEADING IMPRESSED E.M.F.

101. Effective E.M.F. and Current Values

In laying out current and voltage vectors thus far, we have indicated in each case the current or voltage at some particular instant of time in its cycle. The length of the vector gave the maximum value of the current or voltage and the angle that the vector made with the horizontal, in a counterclockwise sense, indicated the particular instant being considered.

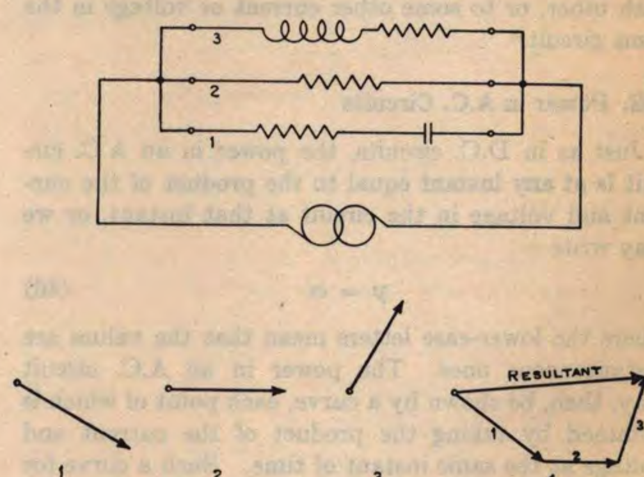


FIG. 202. GRAPHICAL ADDITION OF CURRENT VECTORS

For practical purposes, however, it would be inconvenient to be always under the necessity of stating both a value and a position in time in defining an alternating current or voltage. It is advantageous, rather, to adopt some arbitrary standard so that only the value of the current or voltage need be given to define it, its position in time being understood from the convention adopted. The maximum value would perhaps appear to be the logical choice, but this has certain disadvantages. Another, and more useful value would be the average value over a complete half-cycle, this being equal for the sine wave to .636 times the maximum value.

Still more useful is a value so selected that the heating effect of a given value of alternating current in a resistance will be exactly the same as the heating effect of the same value of direct current in the same resistance. The advantage of such a convention is apparent, since it obviates to a degree the necessity for thinking of the effects of alternating and direct currents as different. This value is known as the **effective** value and is equal to .707 times the maximum value, or—

$$I = .707\ I_{max.} \tag{44}$$

and

$$E = .707\ E_{max.} \tag{45}$$

where E and I without subscripts indicate effective values. **Unless specifically stated otherwise, values of alternating currents and voltages are always given in terms of their effective values.** Likewise, vectors representing currents and voltages give the effective value of the current or voltage by their length and, unlike the vectors we have previously considered, do not indicate by their angular position a particular instant of time within the cycle but only the time relationship of the current and voltage with reference to each other, or to some other current or voltage in the same circuit.

102. Power in A.C. Circuits

Just as in D.C. circuits, the power in an A.C. circuit is **at any instant** equal to the product of the current and voltage in the circuit at that instant, or we may write—

$$p = ei \tag{46}$$

where the lower-case letters mean that the values are instantaneous ones. The power in an A.C. circuit may, then, be shown by a curve, each point of which is obtained by taking the product of the current and voltage at the same instant of time. Such a curve for the case where the current and voltage in a circuit are in phase is shown by Figure 203.

It will be noted that, since the current and voltage are both negative at the same time, the power loops are both positive, which means that no power is being returned from the circuit to the generator. In other words, all of the power delivered by the generator is being absorbed in the resistance of the circuit. For

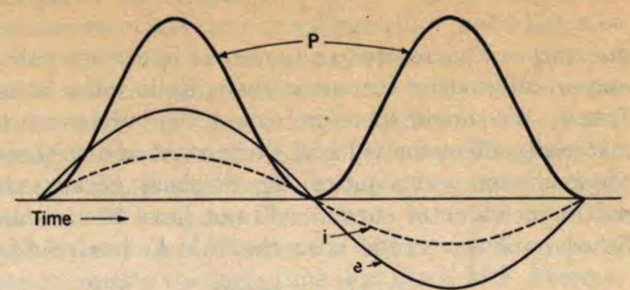


Fig. 203. Power in A.C. Circuit when Current and Voltage are in Phase

this case, where the circuit contains nothing but resistance, the average power is equal to the product of the effective current and voltage, or we may write—

$$P = EI \tag{47}$$

and, as always,—

$$P = I^2R \tag{48}$$

The condition where the circuit contains either inductance or capacity in addition to resistance, and the current and voltage are accordingly not in phase, is somewhat different. The power curve for such a case is shown by Figure 204. Here the product *ei* gives both positive and negative values and we have the

TABLE VII

Conventional Symbols Used in Alternating-Current Work

SYMBOL	STANDS FOR
P	Average power for a cycle of E.M.F. and current.
E	Effective E.M.F.
I	Effective current.
E_{ave}	Average E.M.F.
I_{ave}	Average current.
e	E.M.F. at some specific instant, or instantaneous voltage.
i	Current at some specific instant, or instantaneous current.
E_I	Induced E.M.F.
T	Length of time in seconds (or fraction of one second) for a complete cycle.
f	Frequency or the number of cycles per second.
Z	Impedance in ohms.
X_L	Inductive reactance in ohms.
X_c	Capacity reactance in ohms.
X	Total reactance in ohms.
Y	Admittance in mhos.
θ	Angle between current and impressed E.M.F., or between impedance and resistance, etc.

positive power loops A and B and the negative loops C and D. The latter loops represent power returned to the generator from the circuit. The total power absorbed by the circuit is obviously equal to the sum of A and B minus the sum of C and D. In this case, then, the power, P is no longer equal to EI but to something less than that. The factor by which EI must be reduced to obtain the true power is determined by the phase relation between the current and voltage, this power being—

$$P = EI \cos \theta \tag{49}$$

where θ is the angle between the current and voltage. The term, $\cos \theta$, is known as the **power factor** and has a maximum value of 1 when θ is zero, or the current and voltage are in phase.

It may be noted that the expression, $P = I^2R$, remains true in this case and conforms with Equation (49) because as we shall learn in the next chapter, $R = Z \cos \theta$ and $I = E/Z$, from whence—

$$P = I^2R = I \times I \times R = I \times \frac{E}{Z} \times Z \cos \theta$$

$$= IE \cos \theta.$$

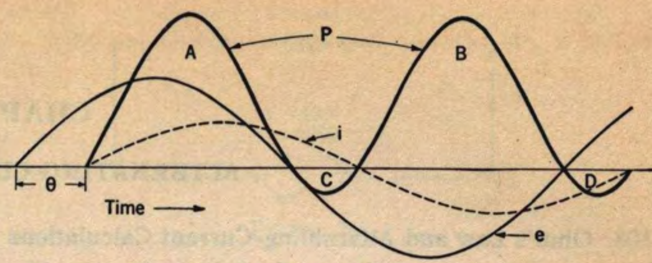


Fig. 204. Power in A.C. Circuit when Current and Voltage are not in Phase

CHAPTER XVI

ALTERNATING CURRENTS—(*Continued*)

103. Ohm's Law and Alternating-Current Calculations

In Chapter I we learned that the relation between the voltage and the current in a D.C. circuit was expressed by Ohm's Law, or

$$\frac{E \text{ (volts)}}{I \text{ (amperes)}} = R \text{ (ohms)}$$

We found this expression indispensable in our study of direct-current circuits, and certainly we shall want to apply it to alternating-current circuit calculations if we can. On the other hand, we have learned of circuit properties other than resistance that influence alternating-current flow. Moreover, these properties, viz., capacity and inductance, not only change the value of the current in amperes but introduce changes in the phase relation of the current to the voltage. Again, the effects of inductance and capacity depend entirely upon the particular frequency which we wish to consider. We must therefore introduce some new quantity that will express in ohms not only the resistance to current but the **combined effects** of resistance, capacity and inductance at a definite stated frequency. This quantity is called **impedance**, and Ohm's Law is adjusted to read—

$$Z \text{ (ohms)} = \frac{E \text{ (volts)}}{I \text{ (amperes)}} \qquad (50)$$

where Z is the symbol for **impedance** or **the combined effect of the circuit's resistance, inductance and capacity taken as a single property which can be expressed in ohms for any given sine wave frequency.** It follows, then, that if we can by certain calculations reduce a circuit's resistance expressed in ohms, its inductance expressed in henrys, and its capacity expressed in microfarads, to a single expression in ohms, we can calculate the current at a given frequency in any single branch as readily as though it were a branch of a direct-current network.

The effect of inductance or capacity in opposing the flow of current in any alternating-current circuit is known as **reactance** and is expressed in ohms the same as resistance. However, in combining resistance and reactance into a single property measured in ohms, which we have already referred to as impedance, we must add them vectorially because they do not act in phase. We shall take up the calculation of impedance after first learning how the reactance may be determined for any single frequency from the inductance and capacity values in a given circuit branch.

104. Inductive Reactance

Referring to Chapter VIII, it will be recalled that we considered two factors as being involved in the calculation of the effects of inductance; first, the physical property of the circuit called inductance and second, the rate of change of current value, which uses inductance "as a tool" in creating the reactive effects. In an **alternating-current circuit containing inductance, therefore, we should expect greater reactance for higher frequencies** because higher frequencies mean an increase in the average rate of change of current. By referring to Figure 205 this becomes apparent. Here are two current cycles of the same effective value but

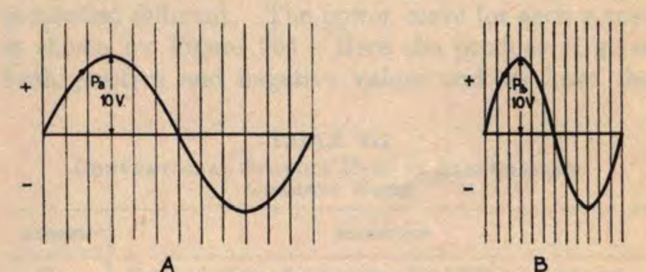


FIGURE 205

the A cycle has twice the period, or half the frequency of the B cycle. Also the slope of the A curve at any point such as P_a, is half the slope at any corresponding point such as P_b on the B curve. The slope is, as has been seen, the measure of current change and we would expect, therefore, that the induced E.M.F. of the B curve would be twice as great as that of the A curve. Thus, the reactance due to inductance depends upon first, the inductance of the circuit and second, the frequency of the current. As a matter of fact, it can be proven that the inductive reactance expressed in ohms is equal to the **inductance in henrys times the frequency in cycles per second, multiplied by 2π** or—

$$X_L = 2\pi f L \qquad (51)$$

where X_L is the inductive reactance in ohms, π is 3.1416, f is the frequency expressed in cycles per second, and L is the inductance in henrys.

For practical use this becomes—

$$X_L = 6.2832\, fL \qquad (52)$$

Example: In Figure 206 assume that the source of alternating E.M.F. is a sine wave, 10 volts, 1000 cycles per second, and the inductance shown has negligible resistance. What is the effective current through the inductance?

Note:—In practice inductance coils have appreciable resistance because any coil winding must contain a definite length of wire; the condition assumed here is that the effect of the inductance is so much greater than that of the resistance that we may neglect the value of the resistance in the calculations.

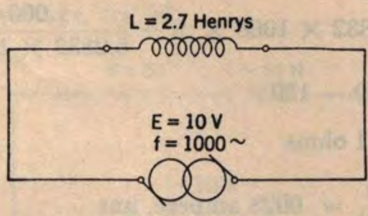


FIGURE 206

Solution:

$$X_L = 6.2832fL = 6.2832 \times 1000 \times 2.7$$
$$= 16964 \text{ ohms}$$
$$I = \frac{E}{16964}$$
$$= \frac{10}{16964}$$
$$= .00059 \text{ ampere, ans.}$$

In this example, the current will be 90° behind the impressed voltage, as shown in Figure 207, because the induced E.M.F. due to the current must be equal and opposite to the impressed E.M.F. and the induced E.M.F., as previously explained, is the rate of change or slope of current times the inductance and must, therefore, be 90° behind the current.

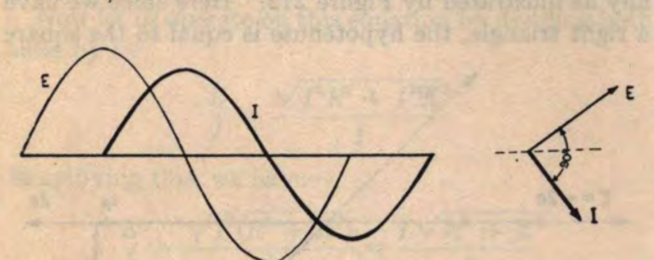


FIG. 207. EFFECT OF INDUCTIVE REACTANCE

105. Capacity Reactance

Capacity reactance has opposite effects to inductive reactance—in fact, the two tend to neutralize each other. Capacity reactance decreases with increasing frequency and capacity values. It also tends to make

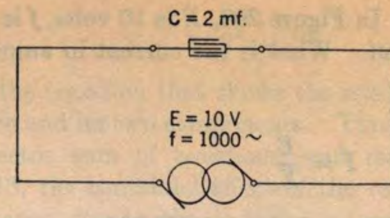


FIGURE 208

the current lead instead of lag the voltage (see Figures 207 and 209). Accordingly, if inductive reactance is assumed as positive, capacity reactance must be taken as negative.

This time relation of the voltage and current in a circuit containing capacity may be seen by referring to Figures 208 and 209. Here when the impressed voltage E is at its maximum positive value, the condenser is charged to a value equal and opposite to the impressed voltage. The current in the circuit is therefore zero. As the positive impressed voltage decreases toward zero the opposite voltage of the condenser forces current to flow in a negative direction. This negative current reaches its maximum value when the impressed voltage becomes zero. Now the impressed voltage reverses, becoming negative, and as it rises to its maximum negative value, charges the condenser in the opposite (positive) direction. During this time, the negative current decreases to zero as the condenser becomes fully charged. Then as the negative impressed voltage decreases from its maximum, the condenser voltage again takes control and causes the current to build up in the opposite direction. The relationships are therefore as shown in the figure with the current leading the voltage by 90°.

The equation for capacity reactance is as follows:

$$X_c = -\frac{1}{2\pi fC} \tag{53}$$

where C is capacity in farads. Converting C to the customary capacity unit, microfarad, we have—

$$X_c = -\frac{1{,}000{,}000}{2\pi fC} \tag{54}$$

or with 3.1416 substituted for π—

$$X_c = -\frac{1{,}000{,}000}{6.2832fC} \tag{55}$$

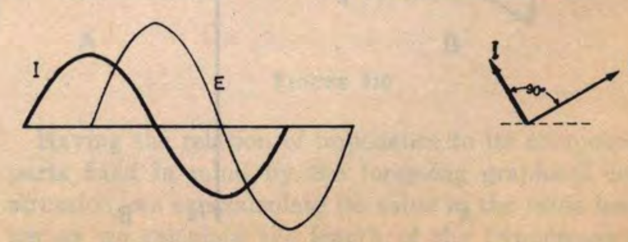


FIG. 209. EFFECT OF CAPACITY REACTANCE

Example: In Figure 208, E is 10 volts, f is 1000 and C is 2 mf. What is the current in amperes?

Solution:

$$I = \frac{E}{X_c}$$

$$X_c = -\frac{1{,}000{,}000}{6.2832 \times 1000 \times 2}$$

$$= -\frac{1{,}000}{6.2832 \times 2}$$

$$= -79.5 \text{ ohms}$$

$$I = -\frac{10}{79.5}$$

$$= -.126 \text{ ampere, ans.}$$

(minus sign here means leading current)

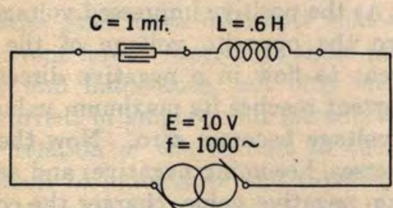


FIGURE 210

106. Combination of Inductive and Capacity Reactances

If we wish to get the combined or total reactance of an inductance in series with a capacity, such as that shown in Figure 210, we may combine the reactances as follows:

$$X = X_L + X_c$$

or, from formulas (52) and (55)—

$$X = 6.2832fL - \frac{1{,}000{,}000}{6.2832fC} \tag{56}$$

Here the signs take care of the neutralizing effect and if the calculated value of X is positive, the inductive reactance predominates; if negative, the capacity reactance predominates.

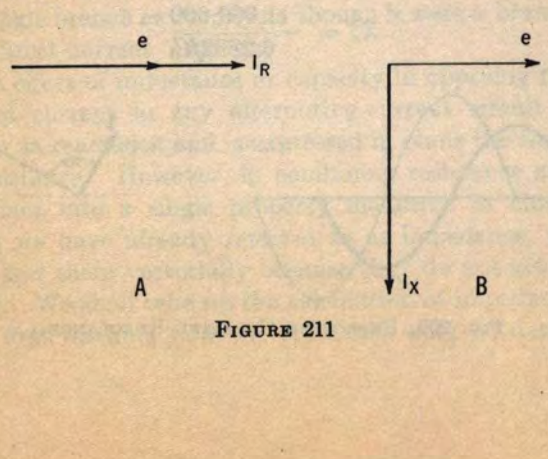


FIGURE 211

Example: Calculate the current in the circuit shown by Figure 210.

Solution:

With no resistance in the circuit—

$$I = \frac{E}{X}$$

and

$$X = X_L + X_c$$

$$= 6.2832fL - \frac{1{,}000{,}000}{6.2832fC}$$

$$X = 6.2832 \times 1000 \times .6 - \frac{1{,}000{,}000}{6.2832 \times 1000 \times 1}$$

$$= 3770 - 159$$

$$= 3611 \text{ ohms}$$

$$I = \frac{10}{3611} = .0028 \text{ ampere, ans.}$$

107. Impedance

To determine a way to combine reactance and resistance when we wish to evaluate the impedance, let us consider the relation between voltage and current under two conditions; first, when a circuit contains pure resistance, and second, when it contains pure reactance. Under the first condition, we can represent the current and voltage as shown in Figure 211-A, and for the second condition as shown in Figure 211-B. For the purpose of this discussion, the circuits are assumed to be such that $I_R = I_X$. If now we connect R and L in series and allow a voltage of $2e$ to act on the combination, we may consider the resulting current as made up of two parts, one due to a voltage e acting on R, and the other due to a voltage e acting on L. The total current will be the sum of these two components, but the addition must be made vectorially as illustrated by Figure 212. Here since we have a right triangle, the hypotenuse is equal to the square

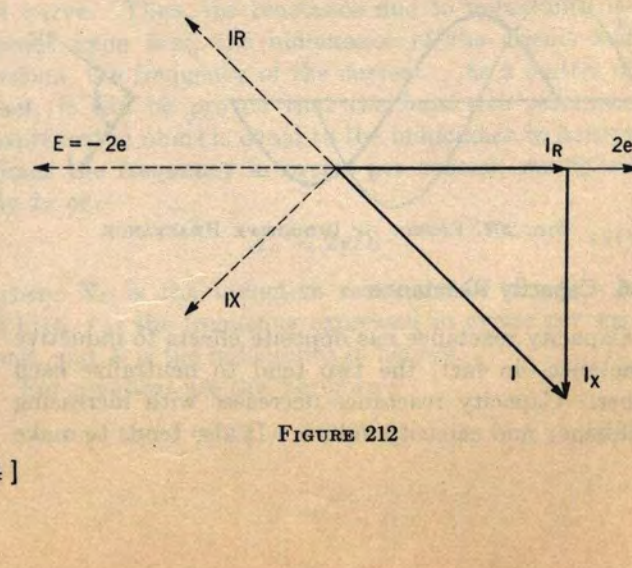


FIGURE 212

root of the sum of the squares of the two legs, or calling the components I_R and I_X, we have—

$$I = \sqrt{I_R^2 + I_X^2}$$

The voltage drop across R due to the flow of the current, I, is IR, and this drop is exactly opposite in phase with I. The drop across L is IX, with a phase relationship such that I leads IX by 90°. The latter will be clear if we refer again to the circuit of pure inductance pictured in Figure 206. Here the current lags the impressed voltage by 90°, and consequently leads the voltage drop, which is equal and opposite to the impressed voltage, by 90°.

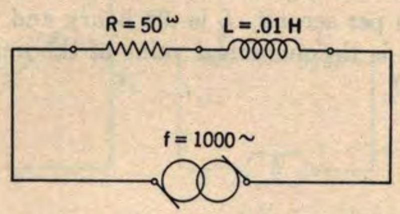


FIGURE 213

In our circuit containing both R and L, therefore, we have two component voltage drops, IR, and IX, 90° out of phase, the sum of which must be equal to the total impressed E.M.F. and exactly opposite in phase. Adding these components vectorially—

$$E = \sqrt{(IR)^2 + (IX)^2} = \text{total voltage drop.}$$

E, in this case, is also equal in value to the total E.M.F. $2e$ acting upon the combined circuit.

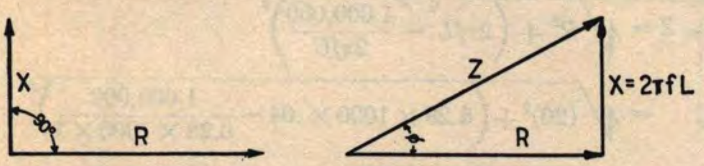


FIGURE 214

Now let us operate on this equation by dividing both sides by I—

$$\frac{E}{I} = \frac{\sqrt{I^2R^2 + I^2X^2}}{I}$$

Simplifying this, we have—

$$\frac{E}{I} = \frac{\sqrt{I^2(R^2 + X^2)}}{I} = \frac{I\sqrt{R^2 + X^2}}{I}$$

or—

$$\frac{E}{I} = \sqrt{R^2 + X^2}$$

However, $\frac{E}{I} = Z$ in ohms, from Equation (50); therefore, we have—

$$Z = \sqrt{R^2 + X^2} \quad (57)$$

which is the equation that shows the relation between impedance and its two components. Thus, impedance is the vector sum of resistance and reactance. In Figure 213, the combined effect of the resistance and the reactance due to the inductance, may be represented by the vector diagram of Figure 214 in which

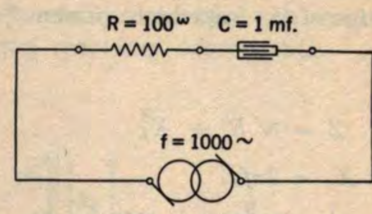


FIGURE 215

the reactance is shown as 90° ahead of the resistance. Similarly in Figure 215, the combined effect of resistance and capacity may be represented by the vector diagram of Figure 216 in which the reactance is shown as 90° behind the resistance.

In these diagrams if R is represented by the same line as the current, Z will be represented by the same line as the impressed E.M.F.; consequently the angle θ will represent the phase difference between the voltage and current, and with the adopted convention for direction of rotation and that for plotting time on the sinusoidal chart, will represent current lagging behind impressed E.M.F. for positive angle as shown in Figure 214, and current leading impressed E.M.F. for negative angle as shown in Figure 216.

We can now consider a simple series circuit with all three properties, or with resistance, inductance, and capacity, as shown in Figure 217. Here we have two reactances acting in opposite phase as shown in Figure 218-A. In constructing the impedance triangle, X_c must be considered as negative and subtracting from X_L as shown in Figure 218-B. If X_c is less than X_L, X will be positive, and if X_c is greater than X_L, as shown in Figure 219, X will be negative.

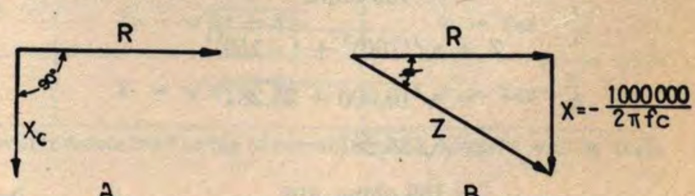


FIGURE 216

Having the relation of impedance to its component parts fixed in mind by the foregoing graphical construction, we can calculate its value in the same manner as we calculate the length of the hypotenuse of any right triangle, as has been explained. That is to say, we square both legs and take the square root of their sum. The equation then for impedance with resistance and inductance in series (as shown by Figures 213 and 214) is:

$$Z = \sqrt{R^2 + X_L^2} \tag{58}$$

Example: In Figure 213, R = 50 ohms, f = 1000 cycles per second, and L = .01 henry. What is the value of the impedance in ohms?

Solution:

$$Z = \sqrt{R^2 + X_L^2}$$

$$X_L = 2\pi f L$$

$$= 6.2832 \times 1000 \times .01$$

$$= 62.8$$

$$Z = \sqrt{(50)^2 + (62.8)^2}$$

$$= \sqrt{2500 + 3944}$$

$$= \sqrt{6444}$$

$$= 80.3 \text{ ohms, ans.}$$

Similarly, for the impedance shown by Figures 215 and 216,

$$Z = \sqrt{R^2 + X_c^2} \tag{59}$$

Example: In Figure 215, R is 100 ohms, C is 1 mf. and f is 1000 cycles per second. What is the value of the impedance in ohms?

Solution:

$$Z = \sqrt{R^2 + X_c^2}$$

$$X_c = -\frac{1,000,000}{2\pi f C}$$

$$= -\frac{1,000,000}{6.2832 \times 1000 \times 1}$$

$$= -159 \text{ ohms}$$

$$Z = \sqrt{(100)^2 + (-159)^2}$$

$$= \sqrt{10,000 + 25,281}$$

$$= \sqrt{35,281}$$

$$= 188 \text{ ohms, ans.}$$

The foregoing are special cases, but we may combine inductive reactance and capacity reactance in one general equation for impedance as shown by Equation (57) where

$$X = X_L + X_c = 2\pi f L - \frac{1,000,000}{2\pi f C}$$

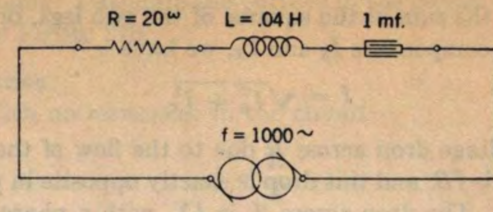


Figure 217

Therefore,

$$Z = \sqrt{R^2 + \left(2\pi f L - \frac{1,000,000}{2\pi f C}\right)^2} \tag{60}$$

Example: In Figure 217, R is 20 ohms, f is 1000 cycles per second, L is .04 henry and C is 1 mf. What is the numerical value of the impedance in ohms?

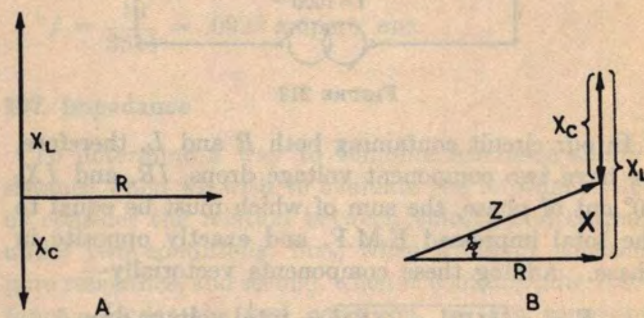


Figure 218

Solution:

$$Z = \sqrt{R^2 + \left(2\pi f L - \frac{1,000,000}{2\pi f C}\right)^2}$$

$$= \sqrt{(20)^2 + \left(6.28 \times 1000 \times .04 - \frac{1,000,000}{6.28 \times 1000 \times 1}\right)^2}$$

$$= \sqrt{(20)^2 + (251 - 159)^2}$$

$$= \sqrt{400 + 8464}$$

$$= \sqrt{8864}$$

$$= 94 \text{ ohms, ans.}$$

In these calculations we have only determined the numerical value of the impedance. This does not com-

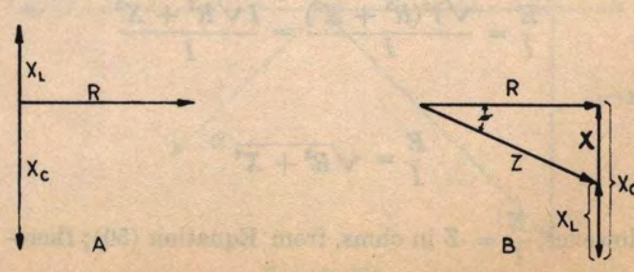
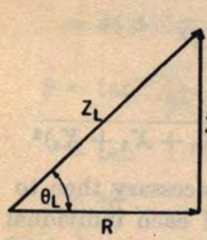


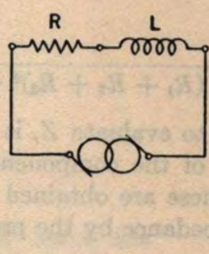
Figure 219

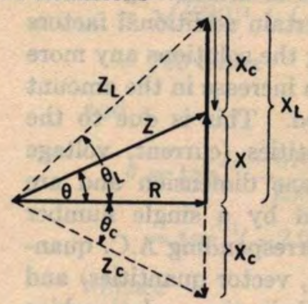
pletely describe it, however, since there could be any number of resistance, capacity and inductance combinations which would give the same numerical value. It is essential, therefore, to include an additional factor which will indicate the relative magnitudes of the resistance and reactance components of the impedance, in order to completely define it. This factor is the angle shown as θ in Figures 218 and 219. Impedance is customarily expressed, accordingly, in the form $Z/\underline{\theta}$ (Z at an angle θ) where Z is the magnitude of the impedance and θ is the angle of lag or lead between any E.M.F. impressed across the impedance and the resultant current. As may be seen from Figure 218, θ is equal to $\tan^{-1}\dfrac{X}{R}$ $\left(\text{the angle whose tangent is } \dfrac{X}{R}\right)$. Also, by simple trigonometry we know that $R = Z \cos\theta$ and $X = Z \sin\theta$. Thus, with the impedance expressed in the form $Z/\underline{\theta}$ it is completely defined and we may readily determine the magnitude of its resistance and reactance components.

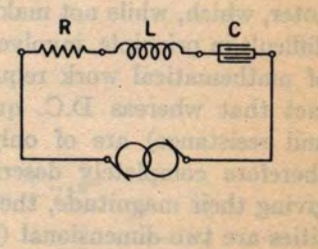
TABLE VIII

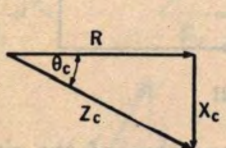
CHART OF VECTOR RELATIONS

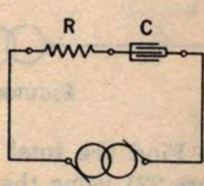


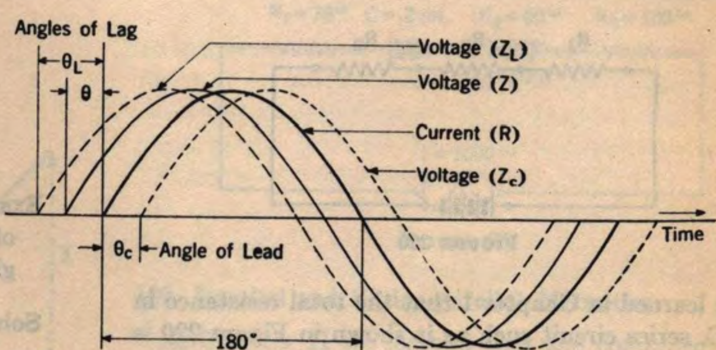












PROPERTY	REACTANCE	IMPEDANCE	PHASE ANGLE
Inductance (L)........................	$X_L = 2\pi fL$	$Z_L = \sqrt{R^2 + X_L^2}$	$\theta_L = \text{Tan}^{-1}\dfrac{X_L}{R}$
Capacity (C)........................	$X_C = -\dfrac{1{,}000{,}000}{2\pi fC}$	$Z_C = \sqrt{R^2 + X_C^2}$	$\theta_C = \text{Tan}^{-1}\dfrac{X_C}{R}$
Net Effect............................	$X = X_L + X_C$	$Z = \sqrt{R^2 + X^2}$	$\theta = \text{Tan}^{-1}\dfrac{X}{R}$

Notes: 1. If lines Z_C, Z_L or Z represent phase of voltage, line R will indicate lead or lag of current and θ_C, θ_L and θ will be angle of lead or lag

2. Power factor is cosine of phase angle (Power $= EI\cos\theta$).

3. The impedance symbol is usually written Z/θ, for example, $Z/\theta = 15'' /30°$, etc.

CHAPTER XVII

ALTERNATING CURRENTS—(*Continued*)

108. Series Networks

In Chapters I and II, means of solving direct-current networks for the current values in the various branches were described. **The same methods and formulas apply to the solution of alternating-current networks.** But in this case certain additional factors enter, which, while not making the solutions any more difficult in principle, involve an increase in the amount of mathematical work required. This is due to the fact that whereas D.C. quantities (current, voltage and resistance) are of only one dimension and are therefore completely described by a single number giving their magnitude, the corresponding A.C. quantities are two-dimensional (i.e., vector quantities) and both their magnitudes and their time relationships with some reference point must be used in making calculations with them.

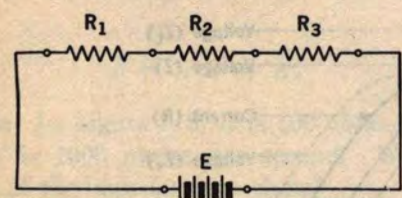


FIGURE 220

We learned in Chapter I that the total resistance in a D.C. series circuit such as is shown in Figure 220 is equal to the **arithmetic** sum of the individual resistances, or—

$$R = R_1 + R_2 + R_3, \text{ etc.} \tag{4}$$

Similarly in an A.C. series circuit, as shown in Figure 221, the total impedance is equal to the **vector** sum of the individual impedances or—

$$\overline{Z} = \overline{Z}_1 + \overline{Z}_2 + \overline{Z}_3 \tag{61}$$

the bars over the impedance symbols meaning that they are vectors and to be treated accordingly in performing the indicated additions.

To graphically illustrate the application, let us assume that $Z_1 = 10$ ohms with $\theta_1 = 30°$, $Z_2 = 15$ ohms with $\theta_2 = 45°$ and $Z_3 = 20$ ohms with $\theta_3 = 60°$; we then have the three vectors represented by Figure 222-A which, when added, give the value of Z shown in Figure 222-B. If we should represent not only the impedance vectors but the resistance and reactance components as well, we should find that each group of components adds algebraically as shown by Figure 223. By comparing Figure 223-C with Figure 223-B, we find that X is the sum of X_1, X_2 and X_3 and R is the sum of R_1, R_2 and R_3. Therefore since—

$$Z = \sqrt{R^2 + X^2}$$

we have—

$$Z = \sqrt{(R_1 + R_2 + R_3)^2 + (X_1 + X_2 + X_3)^2} \tag{62}$$

In order to evaluate Z, it is necessary then to find the values of the components of each individual impedance; these are obtained by multiplying the value of each impedance by the proper function of its angle.

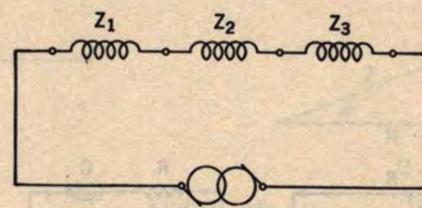


FIGURE 221

Example: Find the total impedance of the circuit of Figure 221 using the values for Z_1, Z_2 and Z_3 given above.

Solution:

$$X = Z \sin\theta$$
$$R = Z \cos\theta$$

This gives

$$X_1 = 10 \times .500 = 5 \text{ ohms}$$
$$R_1 = 10 \times .866 = 8.7 \text{ ohms}$$

The other values can be determined in the same

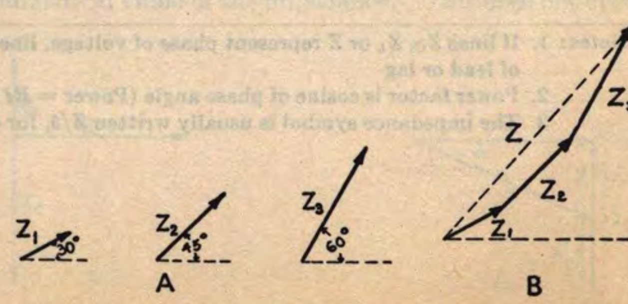


FIG. 222. GRAPHICAL ADDITION OF IMPEDANCE VECTORS

way, and we find that—

$$X_2 = 10.6 \text{ ohms}$$
$$R_2 = 10.6 \text{ ohms}$$
$$X_3 = 17.3 \text{ ohms}$$
$$R_3 = 10.0 \text{ ohms}$$

Applying Equation (62)—

$$Z = \sqrt{(R_1 + R_2 + R_3)^2 + (X_1 + X_2 + X_3)^2}$$
$$= \sqrt{(8.7 + 10.6 + 10)^2 + (5 + 10.6 + 17.3)^2}$$
$$= \sqrt{(29.3)^2 + (32.9)^2}$$
$$= 44.0 \text{ ohms}$$
$$\theta = \tan^{-1}\frac{32.9}{29.3}$$
$$= \tan^{-1} 1.12$$
$$= 48°.$$

Therefore

$$Z = 44\underline{/48°} \text{ ohms, ans.}$$

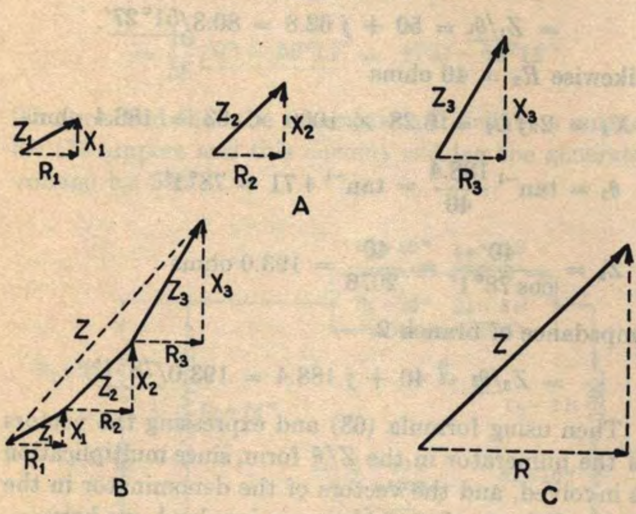


Fig. 223. Analysis of Impedance Vector Addition

The foregoing calculation covers a general case. In practice, however, we usually have given the inductance, capacity and resistance values rather than the individual impedances with their respective angles.

Example: Find the impedance of the series circuit shown by Figure 224.

Solution:

$$Z = \sqrt{(R_1 + R_2 + R_3)^2 + (X_c + X_L)^2}$$

where

$$X_c = -\frac{1{,}000{,}000}{2\pi fC}$$
$$= -\frac{1{,}000{,}000}{6.28 \times 1000 \times .2}$$
$$= -796 \text{ ohms}$$

and

$$X_L = 2\pi fL$$
$$= 6.28 \times 1000 \times .02$$
$$= 125.6 \text{ ohms.}$$

Then

$$Z = \sqrt{(70 + 60 + 100)^2 + (-796 + 125.6)^2}$$
$$= \sqrt{(230)^2 + (-670.4)^2}$$
$$= 709 \text{ ohms}$$

and

$$\theta = \tan^{-1}\frac{-670.4}{230}$$
$$= \tan^{-1}(-2.9) = -71°,$$

whence

$$Z = 709\underline{/-71°}, \text{ ans.}$$

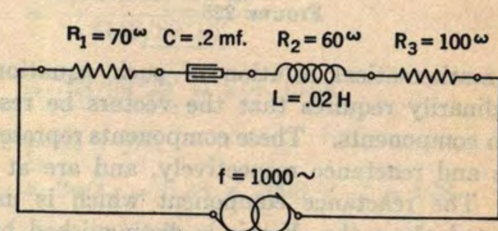


Figure 224

109. Parallel and Series-parallel Networks

In Chapter II we learned that the combined resistance of two parallel resistances was equal to—

$$R = \frac{R_1 R_2}{R_1 + R_2} \tag{8}$$

or that if more than two resistances are in parallel, the combined resistance may be found by adding together the reciprocals of each resistance (called conductance) and taking the reciprocal of this value. That is—

$$G = G_1 + G_2 + G_3 \tag{10}$$

or

$$\frac{1}{R} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}$$

Now if we substitute impedance for resistance in the above equations, they will hold for the A.C. case, providing that we remember that impedances are vector quantities. Thus for two impedances in parallel, we may write the value of the combined impedance as—

$$\bar{Z} = \frac{\bar{Z}_1 \bar{Z}_2}{\bar{Z}_1 + \bar{Z}_2} \qquad (63)$$

or, for more than two in parallel,

$$\frac{1}{\bar{Z}} = \frac{1}{\bar{Z}_1} + \frac{1}{\bar{Z}_2} + \frac{1}{\bar{Z}_3}, \text{ etc.} \qquad (64)$$

which latter may also be written—

$$\bar{Y} = \bar{Y}_1 + \bar{Y}_2 + \bar{Y}_3, \text{ etc.} \qquad (65)$$

where Y represents the reciprocal of impedance and is called **admittance.**

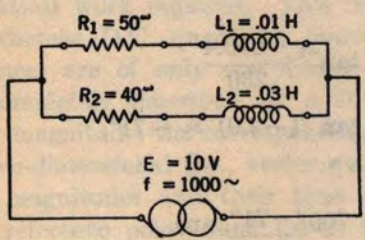


FIGURE 225

The mathematical solution of such equations as (63) ordinarily requires that the vectors be resolved into two components. These components represent resistance and reactance respectively, and are at right angles. The reactance component which is usually represented along the Y-axis is distinguished by the coefficient "j", to indicate its position relative to the resistance component along the X-axis. The vector is then expressed in the standard notation as $\bar{Z} = Z\underline{/\theta} = R + jX$ where "j" indicates a rotation of 90° in a counterclockwise direction. In the algebra of complex quantities, "j" is then handled like any ordinary coefficient. The use of this notation makes possible the direct application of the same formulas as those used in D.C. calculations to the solution of A.C. networks. As an example, let us determine the current delivered by the generator of Figure 225 and the phase angle of this current with the generator E.M.F.

By Ohm's Law we know that the total current delivered by the generator is—

$$I = \frac{E}{Z\underline{/\theta}}$$

where $Z\underline{/\theta}$ is the total impedance of the circuit and consists of the net impedance of the two parallel paths whose individual impedances may be indicated as $Z_1\underline{/\theta_1}$ and $Z_2\underline{/\theta_2}$. Then from the usual formula for parallel circuits—

$$Z\underline{/\theta} = \frac{Z_1\underline{/\theta_1} \times Z_2\underline{/\theta_2}}{Z_1\underline{/\theta_1} + Z_2\underline{/\theta_2}} \qquad (63)$$

The first step is to find the values of $Z_1\underline{/\theta_1}$ and $Z_2\underline{/\theta_2}$. We know that—

$$\theta_1 = \tan^{-1}\frac{X_1}{R_1}$$

where

$$R_1 = 50 \text{ ohms}$$

and

$$X_1 = 2\pi f L_1 = 6.28 \times 1000 \times .01 = 62.8 \text{ ohms}$$

Then

$$\theta_1 = \tan^{-1}\frac{62.8}{50} = \tan^{-1} 1.255 = 51°27'$$

Then, since $R_1 = Z_1 \cos\theta_1$—

$$Z_1 = \frac{R_1}{\cos\theta_1} = \frac{50}{\cos 51°27'} = \frac{50}{.6232} = 80.3 \text{ ohms}$$

Impedance of branch 1

$$= Z_1\underline{/\theta_1} = 50 + j\,62.8 = 80.3\underline{/51°27'}.$$

Likewise $R_2 = 40$ ohms

$$X_2 = 2\pi f L_2 = 6.28 \times 1000 \times .03 = 188.4 \text{ ohms}$$

$$\theta_2 = \tan^{-1}\frac{188.4}{40} = \tan^{-1} 4.71 = 78°1'$$

$$Z_2 = \frac{40}{\cos 78°1'} = \frac{40}{.2076} = 193.0 \text{ ohms}$$

Impedance of branch 2

$$= Z_2\underline{/\theta_2} = 40 + j\,188.4 = 193.0\underline{/78°1'}.$$

Then using formula (63) and expressing the vectors of the numerator in the $Z\underline{/\theta}$ form, since multiplication is involved, and the vectors of the denominator in the $R + jX$ form, since addition is involved, we have—

$$Z\underline{/\theta} = \frac{Z_1\underline{/\theta_1} \times Z_2\underline{/\theta_2}}{(R_1 + jX_1) + (R_2 + jX_2)}$$

$$= \frac{Z_1 Z_2\underline{/\theta_1 + \theta_2}}{(R_1 + R_2) + j(X_1 + X_2)}$$

$$= \frac{80.3 \times 193.0\underline{/51°27' + 78°1'}}{(50 + 40) + j(62.8 + 188.4)}$$

$$= \frac{15500\underline{/129°28'}}{90 + j251.2}$$

$$= \frac{15500\underline{/129°28'}}{\dfrac{90}{\cos\left[\tan^{-1}\dfrac{251.2}{90}\right]}\Big/\underline{\tan^{-1}\dfrac{251.2}{90}}}$$

$$= \frac{15500/\underline{129^\circ 28'}}{\dfrac{90}{\cos(\tan^{-1} 2.79)}\,/\underline{\tan^{-1} 2.79}}$$

$$= \frac{15500/\underline{129^\circ 28'}}{\dfrac{90}{\cos 70^\circ 17'}\,/\underline{70^\circ 17'}}$$

$$= \frac{15500/\underline{129^\circ 28'}}{\dfrac{90}{.3374}\,/\underline{70^\circ 17'}}$$

$$= \frac{15500/\underline{129^\circ 28'}}{267/\underline{70^\circ 17'}}$$

$$= \frac{15500}{267}\,/\underline{129^\circ 28' - 70^\circ 17'}$$

$$= 58.0/\underline{59^\circ 11'}$$

And

$$I = \frac{E}{Z/\underline{\theta}} = \frac{10}{58/\underline{59^\circ 11'}}$$

$$= \frac{10}{58}\,/\underline{0^\circ - 59^\circ 11'} = .173/\underline{-59^\circ 11'}$$

Thus we find that the generator will deliver a current of .173 ampere and this current will lag the generator voltage by 59°11′.

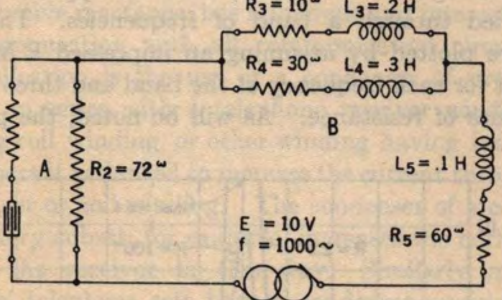


FIGURE 226

With a little practice it will be found that several of the detailed steps given above can be performed in a single operation. This may be illustrated by solving the circuit of Figure 226 to find the current delivered by the generator and its phase relationship with the E.M.F.

Solution:

$$Z_1 = R_1 + jX_1$$

$$X_1 = -\frac{10^6}{2\pi f C_1} = -\frac{1{,}000{,}000}{6.28 \times 1000 \times 1} = -159.3$$

$$Z_1 = 20 - j\,159.3 = 160.7/\underline{-82^\circ 51'}$$

$$Z_2 = R_2 + jX_2 = 72 + j\,0 = 72/\underline{0^\circ}$$

$$Z_A = \frac{Z_1 Z_2}{Z_1 + Z_2} = \frac{160.7/\underline{-82^\circ 51'} \times 72/\underline{0^\circ}}{20 - j\,159.3 + 72 + j\,0}$$

$$= \frac{11{,}570/\underline{-82^\circ 51'}}{92 - j\,159.3} = \frac{11{,}570/\underline{-82^\circ 51'}}{184/\underline{-60^\circ 1'}}$$

$$= 62.8/\underline{-22^\circ 50'} = 58.0 - j24.4$$

$$Z_3 = R_3 + jX_3$$

$$X_3 = 2\pi f L_3 = 6.28 \times 1000 \times .2 = 1256$$

$$Z_3 = 10 + j\,1256 = 1256/\underline{89^\circ 33'}$$

$$Z_4 = R_4 + jX_4$$

$$X_4 = 2\pi f L_4 = 6.28 \times 1000 \times .3 = 1884$$

$$Z_4 = 30 + j\,1884 = 1884/\underline{89^\circ 5'}$$

$$Z_B = \frac{Z_3 Z_4}{Z_3 + Z_4} = \frac{1256/\underline{89^\circ 33'} \times 1884/\underline{89^\circ 5'}}{10 + j1256 + 30 + j1884}$$

$$= \frac{2{,}365{,}000/\underline{178^\circ 38'}}{40 + j\,3140}$$

$$= \frac{2{,}365{,}000/\underline{178^\circ 38'}}{3140/\underline{89^\circ 16'}}$$

$$= 753/\underline{89^\circ 22'} = 8 + j\,753$$

$$Z_C = R_5 + jX_5$$

$$X_5 = 2\pi f L_5 = 6.28 \times 1000 \times .1 = 628$$

$$Z_C = 60 + j\,628$$

The total impedance $Z = Z_A + Z_B + Z_C$

$$Z = 58.0 - j\,24.4 + 8 + j\,753 + 60 + j\,628$$

$$= 126 + j\,1356 = 1360/\underline{84^\circ 40'}$$

$$I = \frac{E}{Z} = \frac{10}{1360/\underline{84^\circ 40'}} = .00735/\underline{-84^\circ 40'}$$

The current delivered by the generator has a value of .00735 ampere and lags the impressed voltage by 84°40′.

110. Alternating-Current Resistance

In alternating-current networks, the apparent resistance of a particular piece of apparatus is often quite different from its direct current or true resistance. As shown by Table V (Chapter VIII), the resistance offered to alternating current may be much greater than that offered to direct current; furthermore, in such cases the value of the resistance depends to some extent on the alternating-current frequency. We find, then, that **not only the reactance component of an**

impedance but its resistance component as well may be a function of the frequency.

"Alternating-current resistance", so called to distinguish it from direct current or true resistance, represents not only the actual resistance of the conductor used to wind a coil but includes also a factor due to the power losses within the iron core. That is to say, when a current flows through a coil winding and establishes a strong magnetic field in the core first in one direction and then in the other, there are certain power losses within the iron due to a heating effect. This is caused in part by hysteresis and in part by small currents induced in the iron itself as a conductor, and called "eddy currents". The total power loss in the coil includes not only the heat losses due to the resistance of the coil winding but also the core losses. Since any power loss can be expressed in the form of the Equation $P = I^2R$, we assume that the winding has in effect a resistance which satisfies this equation. But it so happens that the part of the power loss that is due to the iron core increases with the frequency. Therefore, we should expect the A.C. resistance for a high frequency to be greater than the A.C. resistance for a low frequency.

111. Resonance

In a circuit containing a given inductance, the reactance, X_L, depends upon the frequency; if the frequency is doubled, the reactance is also doubled. In the case of a given capacity value, on the other hand, the negative reactance, X_c, is reduced one-half by doubling the frequency. If a series circuit contains both inductance and capacity, as shown in Figure 227-A, there is therefore some frequency at which the negative reactance, X_c, becomes equal but opposite in value to X_L. The combined reactance is then equal to zero, as shown in Figure 227-B where the dotted line crosses the zero axis. **This is called the frequency to which the circuit is resonant,** or where—

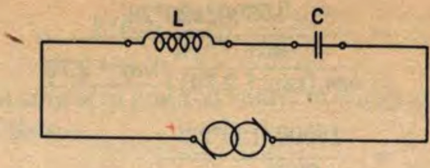
$$0 = 2\pi fL - \frac{1,000,000}{2\pi fC}$$

The value of the resonant frequency, f_r, can be determined in terms of the inductance and capacity by solving this equation for f as follows:

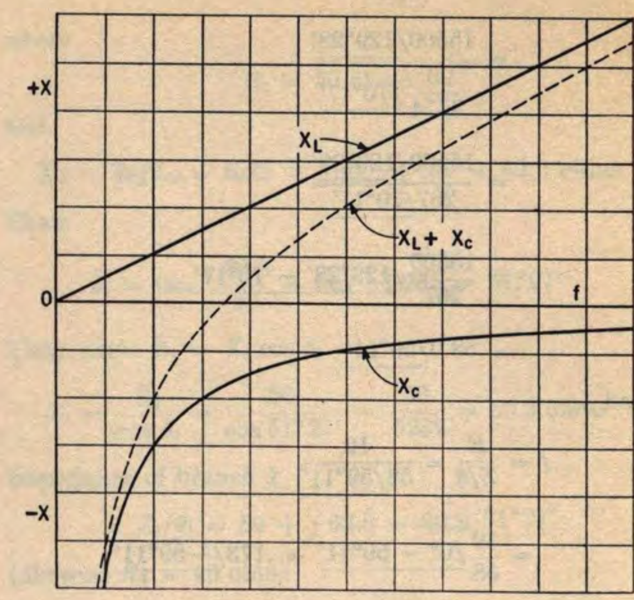
$$f_r = \frac{1,000}{2\pi\sqrt{LC}} \tag{66}$$

Since the total reactance is zero at the resonant frequency, the impedance is then equal to the resistance of the circuit and the current flow is determined solely by this resistance value.

Figure 228 illustrates the behavior of a series resonant circuit similar to that shown in Figure 227-A, but including some resistance, when the applied E.M.F. is varied through a band of frequencies. The curves were plotted by assuming an impressed E.M.F. of 1 volt for each frequency of the band and three different values of resistance. As will be noted, the peak cur-



A



B

Fig. 227. Series Resonant Circuit

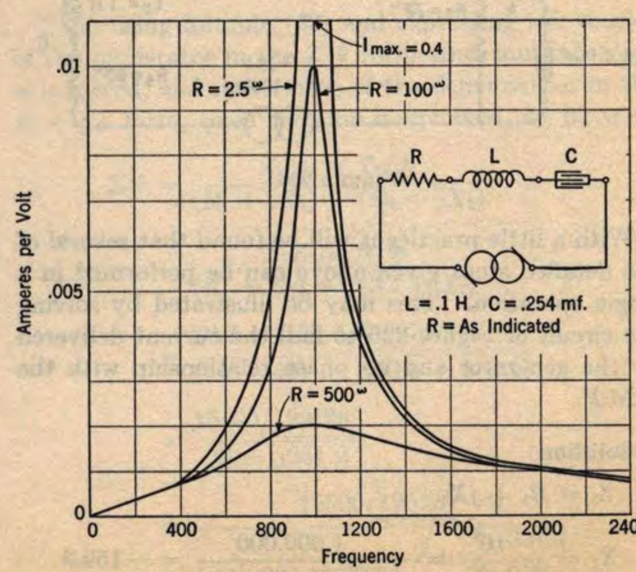


Fig. 228. Curves of Current Values in Series Resonant Circuit

rent values depend entirely upon the resistance values, for at the peak the positive and negative reactances exactly neutralize each other and the current is determined solely by the resistance. Accordingly, the addition of resistance to the series resonant circuit reduces the selectivity or sharpness of the resonance peak. That is, the ratio of the current at the resonant frequency to the current at frequencies near the resonant frequency is reduced.

Example: To what frequency is the circuit shown by Figure 228 resonant if C is .254 mf., L is .10 H, and what current will flow at resonance when R is 4 ohms and E is 1.0 volt?

Solution:

$$f_r = \frac{1{,}000}{6.28\sqrt{.10 \times .254}}$$

$$= \frac{1{,}000}{6.28\sqrt{.0254}}$$

$$= \frac{1{,}000}{6.28 \times .159}$$

$$= 1{,}000 \text{ cycles per sec.} \quad \text{Ans.}$$

$$I = \frac{E}{R} = \frac{1.0}{4} = .25 \text{ amp.} \quad \text{Ans.}$$

The resonance principle in its broadest applications, or rather the practice of neutralizing capacity reactance with inductive reactance, has numerous and interesting uses in connection with all communication circuits. One application is the use of a condenser of proper capacity in series with a telephone receiver winding, repeating coil winding, or other winding having inductance, where it is desired to increase the current through the receiver or coil winding. The condenser of a common battery subset, for example, increases the current through the receiver in this way. Similarly most operators' telephone sets have a condenser associated with the induction coil.

A second application of the resonance principle is the so-called "tuned" circuit, or the resonant circuit used for selectivity. It is an arrangement whereby the circuit has a much lower impedance to some particular frequency than to any other frequency; if a band of frequencies is impressed, it selects, so to speak, a high current for the particular frequency but permits only a negligible current for any other frequency. Figure 228 illustrates this principle.

Another connection of the tuned circuit, shown in Figure 229-A, is called the **anti-resonant** connection. For this condition, when the positive reactance is equal and opposite to the negative reactance, the combined impedance presented to the generator is extremely great and there is a minimum load on the generator. In other words, the generator circuit is practically open. Figure 229-B shows the combined reactance, X_t, presented to the generator by this circuit. It can be seen that at the resonant frequency the two parallel reactances combine to give an extremely high value. At the same time, there must be a current through the inductance, determined by dividing the voltage of the generator by the impedance of this branch. Similarly, there must be a current through the condenser which can be determined in the same way. These currents are equal in value, but are flowing in opposite directions, thereby neutralizing each other in the lead to the generator. Effectively, this gives an open circuit in so far as the generator is concerned, but a circuit equal to either the inductance or capacity alone connected to the generator in so far as either of the branches is concerned. The physical explanation here is that a current is oscillating around through the inductance and condenser, with the E.M.F. of the generator merely sustaining this oscillation. Of course, since the inductance must have some resistance, there will be an I^2R loss in the inductance, and it would never be possible to have the theoretical case where

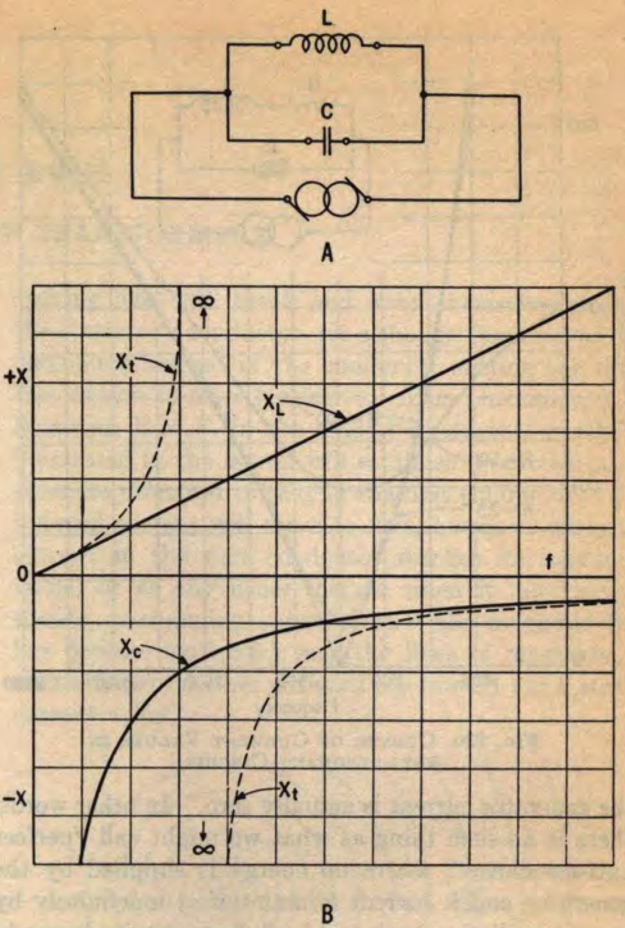


FIG. 229. ANTI-RESONANT CIRCUIT

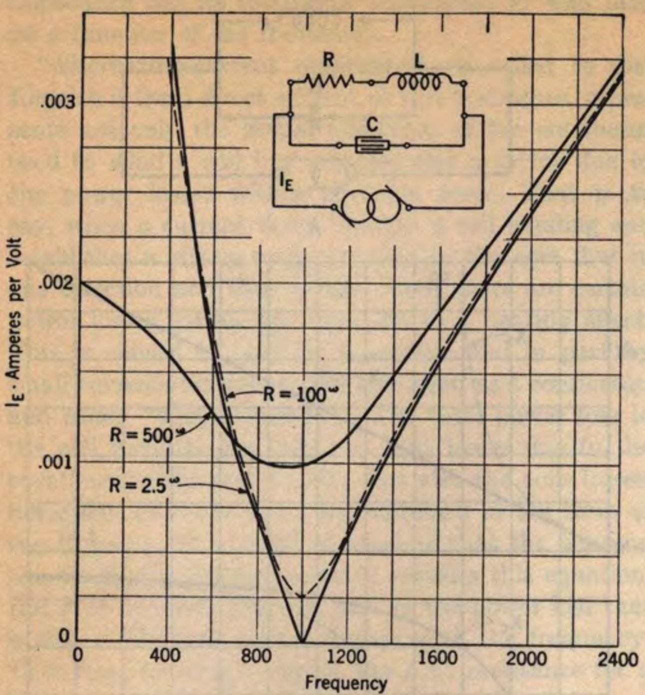


FIG. 230. CURVES OF CURRENT VALUES IN ANTI-RESONANT CIRCUIT

the generator current is actually zero. In other words, there is no such thing as what we might call "perfect anti-resonance", where no energy is supplied by the generator and a current is maintained indefinitely by energy oscillating back and forth from the condenser to the coil.

Figure 230 illustrates the selectivity of an anti-resonant circuit made up of the same units as were used in the series resonant circuit. It will be noted that the selectivity of the anti-resonant circuit is also decreased as the resistance is increased. Indeed, there is a value of resistance beyond which the circuit loses its resonant characteristics altogether. Moreover, in this case, the resistance may be seen to have some effect on the value of the resonant frequency.

Perhaps the most widely known application of both resonant and anti-resonant circuits is in connection with radio sending and receiving sets. From our view-point, the most interesting, perhaps, is the carrier application which is discussed in a later chapter.

Numerous other applications are possible. Thus a simple series resonant circuit may be used as a substitute for a step-up transformer, where an E.M.F. greater than the impressed E.M.F. but with little current drain, is desired. An example of such use may be found in certain vacuum tube circuits, where the operation depends upon the value of the impressed E.M.F. on the grid (which is practically an open circuit) rather than upon the current strength of the incoming energy.

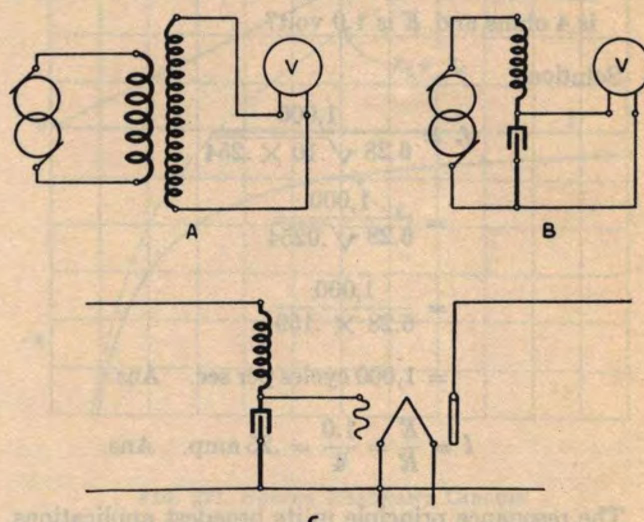


FIG. 231. USE OF RESONANCE PRINCIPLE TO INCREASE VOLTAGE

Figure 231 illustrates this principle. Here Sketch A illustrates the step-up transformer while Sketch B shows how a resonant circuit can be used to greatly increase the E.M.F. of the generator at a single frequency, thereby accomplishing the same result as the transformer. If a voltmeter is connected across the condenser alone, it will be found that the voltage is many times that of the generator at the resonant frequency. Sketch C illustrates the connection of the grid circuit of a vacuum tube for securing a higher potential than that which is impressed from the line.

CHAPTER XVIII

REPEATING COILS AND TRANSFORMERS

112. Mutual Induction

The inductive effects discussed in Chapter VIII dealt with the magnetic interlinkages from one turn of a coil winding to the other turns of the same winding. We defined the effects coming from such magnetic interlinkages as "self-inductance". The current resulting from the **induced** E.M.F. was superposed upon the current resulting from the **impressed** E.M.F.

In practice, we may experience inductive effects in circuits other than the one in which the current due to the impressed E.M.F. is flowing. **That is to say, two coils may be so related that the lines of magnetic induction established by a current in the first coil may cut the turns of the second coil (which may be connected to an entirely different circuit) in the same way that similar lines established by any one turn of a single coil cut the other turns of the same coil.** This effect is called **mutual induction** and the property of the electrical circuit that is responsible for the effect is known as its **mutual inductance.**

113. Theory of the Transformer

In the study of magnetism we found that a wire in which there is a current is always surrounded by a magnetic field. This field, when created by a current establishing itself in the conductor, grows outwardly from the wire as the current increases. Figure 232

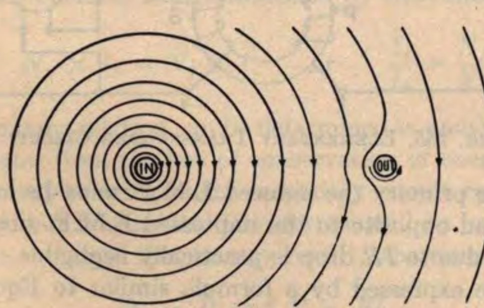


FIGURE 232

shows a group of lines of magnetic induction around a conductor (shown in cross-section) in which the current is increasing in value. If a second conductor is in the vicinity, it will be cut by these lines moving outward from the current-carrying conductor. In the same manner that stationary lines seem to break and wrap themselves around a moving conductor (Figure 70), the moving lines will break and wrap themselves aroune the stationary conductor, for although the lines cut thd conductor instead of the conductor cutting the lines, the motion is merely relative. This phenomenon induces an E.M.F. in the second conductor, which, as illustrated in the figure, will establish a current in the opposite direction to that in the first conductor. The induced current will cease to flow, however, when the current in the first conductor reaches its maximum value, or at any other instant when it may have a steady, unchanging value because the magnetic field has become stationary and the lines of magnetic induction move neither outward nor inward for a steady current value.

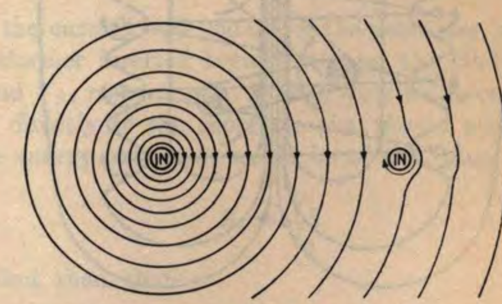


FIGURE 233

If the current in the first conductor is decreased, we have the reverse condition, or that shown in Figure 233. Here the lines, instead of expanding and moving outward, are contracting and moving inward, cutting the second conductor as formerly, but now the current induced is in the opposite direction. It is now in the same direction in the second conductor as in the first. This law for induced E.M.F. may be expressed as follows: **For any two parallel conductors, a current in one increasing in value induces an E.M.F. in the other, tending to establish a current in the opposite direction, and a current decreasing in one will induce an E.M.F. in the other, tending to produce a current in the same direction.**

Instead of the two single conductors shown in Figures 232 and 233, let us consider two separate coils, one inside the other, as in Figure 234. If we call the one carrying the original current the "primary", which in this case we may represent by the inside coil, and the other the "secondary", we shall find that a strong magnetic field is established by a changing current in the primary. This will cut the entire group of conductors represented by the turns of the secondary, thereby inducing appreciable potential in the secondary. The ordinary telephone induction coil operates in this manner. The primary, when connected in series with the transmitter, carries a current which decreases and increases in value in response to the varying resistance of the transmitter. Consequently, an alternating current is induced in the secondary of the coil.

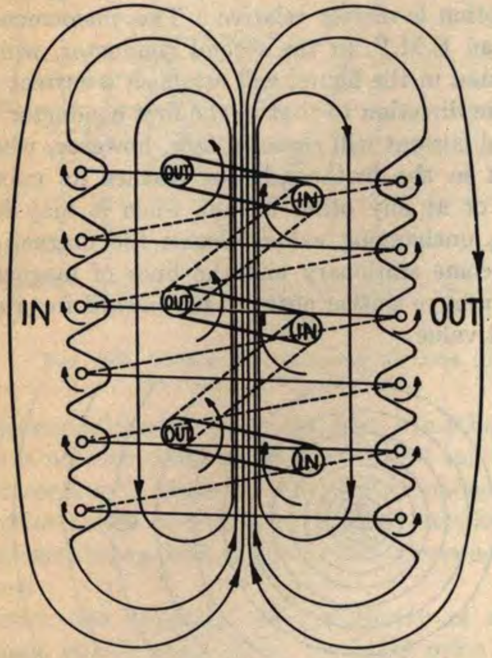


Fig. 234. Principle of Induction Coil

If now the two separate coils of Figure 234 are wound on the same iron core in the manner indicated by Figure 235, the effect will be intensified. Because the iron offers a path of low reluctance to the magnetic flux, the total number of lines will be greatly increased and all of the lines set up by the primary winding, P, will cut all of the secondary winding, S.

If the windings, P and S have the same number of turns, and both the coils and core are constructed so as to have negligible energy losses, we shall find that the voltmeter reading is the same when connected across the terminals of S as when connected across the terminals of P. In other words, the induced E.M.F. of the secondary winding is equal to the impressed E.M.F. of the primary winding. Such a device is called an ideal transformer of unity ratio.

If, now, we should increase the number of turns of the secondary winding S, we would find that the voltmeter reading would be greater on the secondary than on the primary side of the transformer. If we should decrease the number of turns of the winding S, the effect would be reversed. We have here a means, therefore, of controlling the voltage applied to a load; we may effectively increase or decrease the generator voltage by a proper choice of transformer. If a transformer has a greater number of turns on the secondary than on the primary so that the voltage is increased, it is called a "step-up" transformer; if it has a lesser number of turns on the secondary than on the primary so that the voltage is decreased, it is called a "step-down" transformer. The voltage across the two windings is directly proportional to the number of turns. This relation is expressed by the equation:

$$\frac{V_P}{V_S} = \frac{N_P}{N_S} \tag{67}$$

We may explain this relation between the number of turns and voltage by our original law governing inductive effects, which states that **the induced voltage is proportional to the rate of cutting lines of magnetic induction.** Each time the alternating E.M.F. in the primary completes a cycle, it establishes a magnetic flux in the iron core which collapses to be established in the opposite direction, to again collapse, etc. This flux must cut each and every turn about the iron core. In doing so, for the ideal case where there is no loss due to magnetic leakage, etc., the same voltage is induced in each individual turn. This voltage may be represented by the symbol v. Now, the voltage measured across the secondary (with no load connected) is merely the sum of these individual turn voltages or—

$$V_S = N_S \times v \tag{68}$$

where N_S is the number of turns on the secondary.

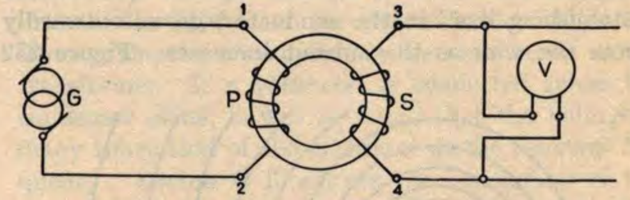


Fig. 235. Elementary Transformer Circuit

In the primary the induced E.M.F. must be exactly equal and opposite to the impressed E.M.F. since the E.M.F. due to IR drop is practically negligible. This could be expressed by a formula similar to Equation (68), thus—

$$V_P = N_P \times v \tag{69}$$

Since v is the same in both Equations (68) and (69), we may derive Equation (67) by dividing (69) by (68).

In Figure 235 the current being supplied by the generator is negligible, inasmuch as we have considered the transformer as having no energy losses. If, however, a load in the form of a shunting impedance is connected to its secondary as shown by Figure 236, the induced E.M.F. in the winding S causes a current to flow through the impedance Z_s, which from Equation (50), can be expressed as follows—

$$I_s = \frac{V_s}{Z_s}$$

When this current starts to flow through the load Z_s, and through the winding S, it will establish other lines of magnetic induction in the transformer core, which oppose those established by the current in the winding P. This will tend to neutralize the magnetic field in the iron core, thereby tending to counteract the inductance of the winding P and to make it more nearly like a plain resistance. With the induced E.M.F. in the winding P reduced, a greater current will flow from the generator through this winding, thus again increasing the flux in the iron core, so that finally there are produced the same induced E.M.F. effects as in the case of the transformer on open circuit. We therefore find that the transformer adjusts itself to any load that may be connected to the secondary just as if an equivalent load were connected directly to the generator, i.e., the current supplied by the generator increases with an increase of current in the secondary of the transformer.

This current, however, is not necessarily of the same value in the primary as in the secondary, but like the voltage, depends upon the ratio of the number of turns of the primary to the number of turns of the secondary. The relation between current values is the inverse ratio of the number of turns. In other words, the winding having the greater number of turns has a proportionately smaller current. This is seen when we consider that the flux in the core depends upon the current value times the number of turns, and the flux established by one coil balances that established by the other—

$$N_p \times I_p = N_s \times I_s \quad \text{or} \quad \frac{I_s}{I_p} = \frac{N_p}{N_s} \qquad (70)$$

The same relation can be determined in another way. We know from the law of conservation of energy that the energy existing in the secondary circuit can never exceed, but for an ideal transformer will be just equal to, the energy of the primary circuit, where since—

$$P_p = P_s \text{ and } P = EI,$$

we have—

$$V_sI_s = V_pI_p$$

from which—

$$\frac{I_s}{I_p} = \frac{V_p}{V_s} \quad \text{or} \quad \frac{I_s}{I_p} = \frac{N_p}{N_s}. \qquad (70)$$

Though we find that connecting the load Z_s to the secondary of the transformer causes the generator to furnish a current output in much the same way as if a load were connected across the generator, it does not follow that the same current flows through the load Z_s with the transformer inserted between the generator and Z_s as would flow if Z_s were connected directly to the generator without the transformer. We have

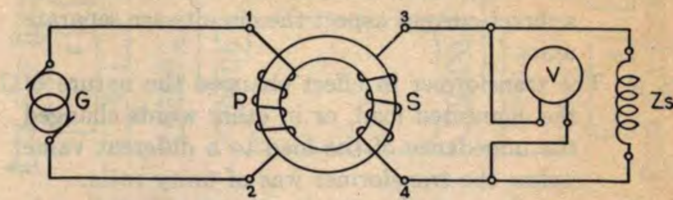


FIGURE 236

just seen that the voltages measured on the two sides of the transformer are directly proportional to the number of turns, and we know, moreover, from Equation (50) that—

$$Z_s = \frac{V_s}{I_s}$$

But the current and voltage of the generator with the transformer inserted between it and the load Z_s are I_p and V_p, respectively, so that were we to connect a load directly to the generator that would absorb the same energy output, it would be of the value—

$$Z_p = \frac{V_p}{I_p}$$

We find, then, that—

$$\frac{Z_p}{Z_s} = \frac{V_p}{I_p} \div \frac{V_s}{I_s} = \frac{V_p}{I_p} \times \frac{I_s}{V_s} = \frac{V_p}{V_s} \times \frac{I_s}{I_p} = \frac{N_p}{N_s} \times \frac{N_p}{N_s}$$

or—

$$\frac{Z_p}{Z_s} = \left[\frac{N_p}{N_s}\right]^2 \qquad (71)$$

Inequality ratio transformers may be rated either according to their voltage ratios, step-up or step-down as the case may be, or in accordance with their impedance ratios. In power work where transformers are primarily used to change the voltage of the system, the rating is on the voltage basis. In telephone work where inequality transformers are used in most cases primarily to match unequal impedances, as will be explained later, they are usually rated in accordance with their impedance ratios.

Before taking up specific uses of the transformer, let us review in general what its presence in Figure 236 has or may have accomplished:

a. The characteristics of the electrical energy may have been changed, or we might say its state may have been "transformed", inasmuch as in the primary circuit we may have had high current and low voltage, while in the secondary circuit we may have had low current and high voltage, or vice versa, depending upon whether the transformer was step-up or step-down.

b. The electrical energy was transferred from one circuit to another without any metallic connection being made between the two circuits; from a direct-current aspect the circuits are separate units.

c. The transformer in effect changed the nature of the connected load, or in other words changed the impedance of the load to a different value unless the transformer was of unity ratio.

In power work the principal use of a transformer is to accomplish the result given in *a* above, whereas in telephone work we are directly concerned with *b* and *c*.

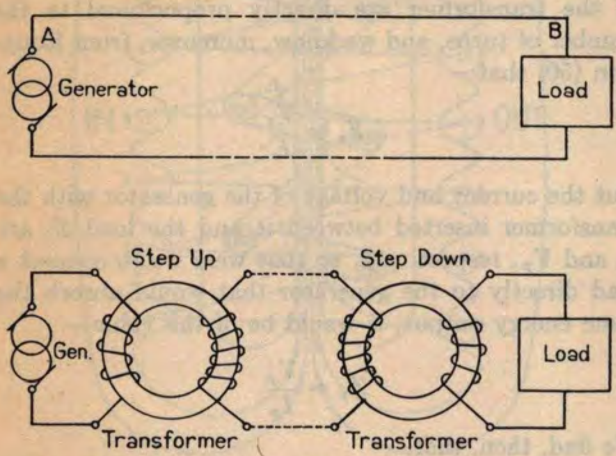


Fig. 237. Use of Transformers in Power Circuit

First, let us illustrate the power case by referring to Figure 237 which shows the use of transformers in a simple power transmission circuit. Let us assume that a 110-volt alternating-current generator at station *A* is to be used to supply a load several miles away. The load is of such nature that it must have 100 amperes at an impressed voltage of 100 volts. Transmission from *A* to *B* must, therefore, be accomplished with a loss of 10 volts for a current of 100 amperes and this means that the IR drop of the line must not exceed 10 volts. Therefore, the resistance of the line, from the equation—

$$R = \frac{E}{I} = \frac{10}{100} = \frac{1}{10}\text{ ohm}$$

must not exceed 1/10th of an ohm, requiring extremely large copper conductors. If, however, a step-up transformer of 1-to-20 voltage ratio is inserted at the generator, and a step-down transformer of 20-to-1 ratio is inserted at the load, we shall find from the relation between current, voltage and power, that the current in the transmission line will be equal to 5 amperes instead of 100 amperes. It will then be possible to have a 200 volt drop in the line and still have a voltage of 2000 on the primary of the transformer at the distant end, or the required 100 volts when stepped down. Since the current in the line will now be 1/20th of 100, or 5 amperes, the resistance of the line in this case will be—

$$R = \frac{200}{5} = 40\text{ ohms}$$

We find, then, that the size of the conductors for the transmission line where the transformers are used, must be such that the resistance will not exceed 40 ohms, whereas in the first case it must be such that the resistance will not exceed 1/10th ohm. The amount of copper required in the second case is 1/400th or only 1/4th of one per cent of that required in the first case. The economy due to the copper saving is apparent.

114. Transformer Applications to Telephone Circuits

The applications of transformers to telephone circuits are numerous and varied. The reduction of energy losses in alternating-current transmission, as illustrated in Figure 237, has an application to telephone transmission but is not so important as other uses. One very general use is to accomplish the result given as *b* above. In this case, the primary function of the coil is to transfer energy to another circuit rather than to change the voltage and current values. When so used in telephone work, they are generally called "repeating coils" rather than "transformers" because their primary function is to "repeat" the energy into a different circuit rather than to transform it into a different state. There are, however, inequality ratio repeating coils which perform both functions. On the other hand, in connection with telephone repeater circuits and certain other telephone apparatus, input and output coils are used primarily to change the voltage and current characteristics. These are accordingly called "transformers", and not "repeating coils".

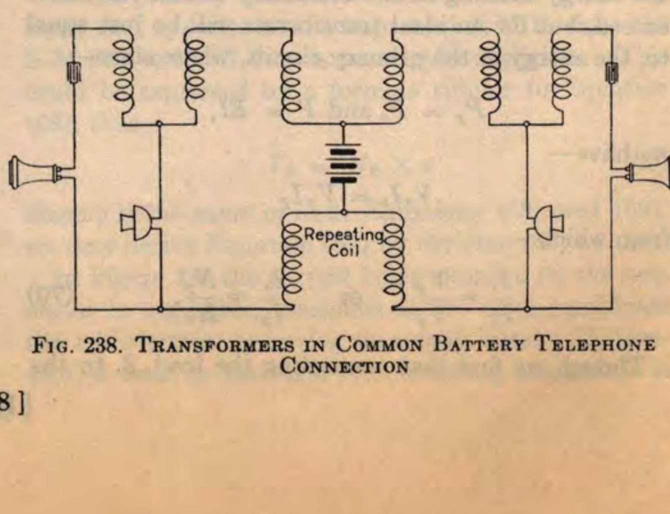


Fig. 238. Transformers in Common Battery Telephone Connection

Perhaps the most common application of the repeating coil in telephone work is in connection with the common battery cord circuit, as illustrated by Figure 238. Here the alternating-current flow in one subscriber's line is repeated into the other subscriber's line with little energy loss, and at the same time the windings of the coils afford the proper direct-current connections for each subscriber's station to receive a superposed D.C. current for transmitter supply. A similar use of the same type of coil in connection with a toll switching trunk circuit is illustrated in Figure 140. Here only one side of the coil is used for battery supply while a condenser is bridged at the mid-point of the other winding, which prohibits the flow of direct current from that side of the circuit. Here again the repeating coil accomplishes the transmission of voice current from one side to the other without its being appreciably affected by the direct-current features of the circuit.

Another very general use of repeating coils in the telephone plant is for deriving "phantom" circuits. Here the coils serve a unique purpose which has no counterpart in electrical power work, and is not included in the classification of transformer functions given above. We shall therefore need to consider this application more fully. However, it may be noted that the coils, while serving this particular purpose, may also function effectively as impedance matching devices.

115. The Phantom Circuit

Figure 239 is a simplified diagram of two adjacent and similar telephone circuits arranged for phantom operation. By means of repeating coils installed at the terminals of the wire circuits, a third telephone circuit is obtained. This third circuit is known as the phantom and utilizes the two conductors of each of the two principal, or "side" circuits, as one conductor of the third circuit. The two side circuits and the phantom circuit are together known as a phantom group. These three circuits, employing only four line conductors, can be used simultaneously without interference with each other, or without crosstalk between any combination, provided the four wires have identical electrical characteristics and are properly transposed to prevent crosstalk.

The repeating coils employed at the terminals are designed for voice-current and ringing-current frequencies, and do not appreciably impair transmission over the principal or side circuits. The third or phantom circuit is formed by connecting to the middle points of the line sides of the repeating coil windings, as shown in the figure. Since the two wires of each side circuit are identical, any current set up in the

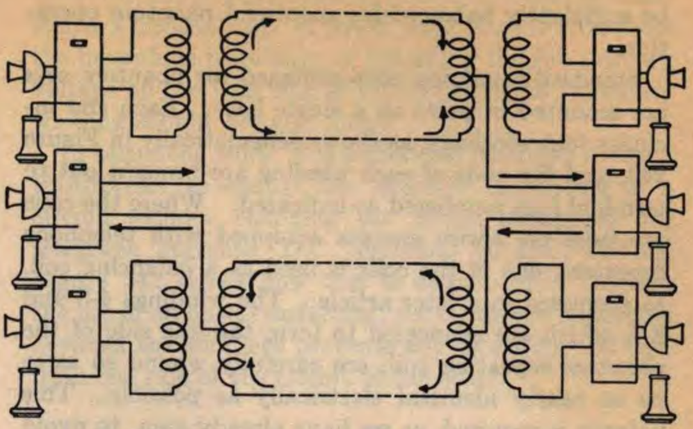


FIG. 239. PRINCIPLE OF THE PHANTOM CIRCUIT

phantom circuit will divide equally at the mid-point of the repeating coil line windings. One part of the current will flow through one-half of the line winding, and the other part of the current will flow in the opposite direction through the other half of the line winding. The inductive effects will be neutralized, and there will be no resultant current set up in the drop or switchboard side of the repeating coil. Since the phantom current divides into two equal parts, the halves will flow in the same direction through the respective conductors of one side circuit, and likewise return in the other side circuit. At any one point along a side circuit, there will be no difference of potential between the two wires due to current in the phantom circuit, and a telephone receiver bridged across them will not detect the phantom conversation.

Since there is no connection, inductive or otherwise, between the two circuits at the terminals, it is equally true that a conversation over a side circuit cannot be heard in the phantom. This can be understood by imagining a flow in the closed side circuit through the line wires and the windings of the repeating coils at each end. With the side circuit conductors electrically equal, there can be no difference of potential between the mid-point of the repeating coil line winding at one end and the mid-point of the repeating coil line winding at the other end because the drops of potential for the two parts of the side circuit are equal and opposite. If the side circuit, therefore, impresses no difference of potential on any part of the phantom circuit, the side circuit conversation cannot be heard over the phantom.

In the theory of the phantom it should not be forgotten that the conductors are assumed to be electrically identical, or in other words, the conductors are perfectly "balanced". The phantom is very sensitive to the slightest upset of this balance, and circuits that are sufficiently balanced to prevent objectionable crosstalk or noise in physical circuit operation, may not be sufficiently balanced for successful phantom operation.

Standard repeating coils designed for phantom sets are mounted in pairs on a single base. Each coil includes four windings, as shown schematically in Figure 240, and the ends of each winding are brought out tc terminal lugs numbered as indicated. Where the coils are used on 2-wire circuits equipped with telephone repeaters, one of the coils is used as a balancing coil, as discussed in a later article. The windings 4-3 and 8-7, which are connected to form the line side of the phantom repeating coil, are carefully wound so as to be as nearly identical electrically as possible. This balance is required, as we have already seen, to avoid crosstalk within the phantom group. The windings 2-1 and 6-5 are not so well balanced and are therefore always connected to form the drop side of the repeating coil.

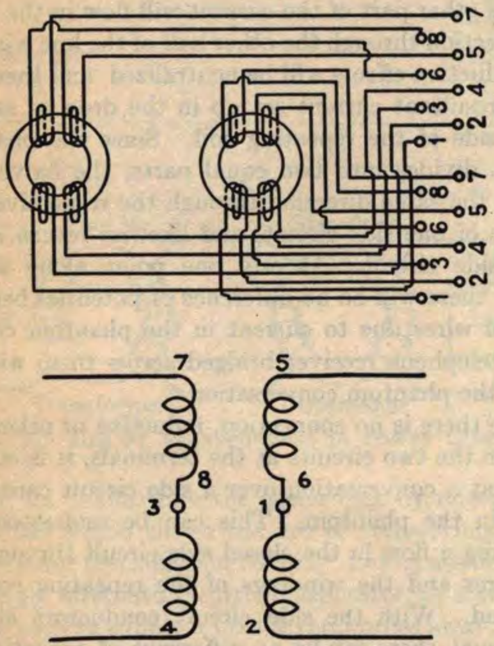


Fig. 240. Wiring Diagram of Standard Repeating Coil

The code numbers of representative standard phantom repeating coils, having various impedance ratios as shown, are given in Table IX. The 93 and 75 type coils are identical except for different mounting arrangements. The same is true of the 62 and 85 type coils. In the manufacture of the 93 and 75 type coils, a core made of many turns of fine gaged silicon-steel wire is sawed so as to introduce a gap in the magnetic circuit. This gap is filled with a compressed powdered magnetic material, which while increasing slightly the core's reluctance, gives it a high degree of magnetic stability, preventing permanent magnetization under abnormal service conditions. The two windings for the line side are wound on the core together to give the

TABLE IX

Phantom Repeating Coils

Impedance Ratio Line to Drop 4-3 & 8-7 to 2-1 & 6-5	Suitable for 20-cycle Signaling		Not Suitable for 20-cycle Signaling	
	Relay Rack	Coil Rack	Relay Rack	Coil Rack
1:1	93-A	75-A	62-A	85-A
1:1.62	93-B	75-B	62-B	85-B
1.62:1	93-F	75-F	62-C	85-C
2.66:1	93-G	75-G	62-E	85-E
1.24:1	93-H			
2.28:1	93-J			
1:1.28			62-F	
1:2.34			62-G	

required high degree of balance, but the drop windings are wound individually. The iron core of the coil is wrapped with cotton tape to protect the windings, and after the windings are put in place, the coil itself is given a wrapping of cotton tape. It is then impregnated with a moisture proof compound, placed in its case, and melted resin is poured around it until it is firmly imbedded. The leads are then brought out to the terminal punchings.

The 85 and 62 type coils are made in the same ratios as the 93 and 75 type coils, and have approximately the same electrical characteristics. Their cores are made in the same way as described above, except that the gap in the magnetic circuit is not filled with compressed iron powder. This feature makes these types of coils somewhat more stable and they are therefore especially well adapted for use in circuits on which rapidly changing direct currents are superposed, such as those involved in high speed teletypewriter service. The same feature, however, tends to make these coils very inefficient at low frequencies and they cannot be used on circuits employing 20-cycle signaling.

116. Autotransformers

There is a type of transformer which may step-up or step-down the voltage, but does not employ a secondary circuit that is electrically separated from the primary circuit. It is called the "autotransformer", and the manner in which the primary and secondary windings are connected together electrically, as well as magnetically, is shown in Figure 241-A. As in any transformer, the primary and secondary windings are both wound on the same iron core so that any lines of magnetic induction that thread one winding also thread the other. The fact that the windings are directly connected does not prevent the voltages on the primary and secondary sides from being proportional to the number of turns in the windings, just as in the regular transformer. Thus, in Figure 241-A if there are 800 turns between *a* and *d*, 200 turns between *b* and *c*, and 110 volts are connected to *bc*, the primary, the secondary voltage will be—

$$110 \times \frac{800}{200} = 440 \text{ volts}$$

This is a one-to-four step-up transformer, and if we should reverse the primary and secondary connections, we would have a four-to-one step-down transformer.

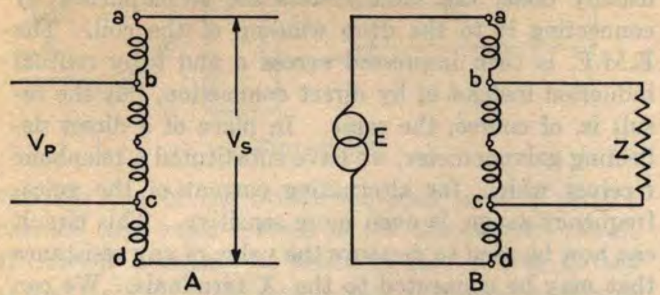


FIG. 241. THE AUTO-TRANSFORMER

Figure 241-B shows a step-down autotransformer with a voltage E connected to its primary and a load of impedance Z connected to its secondary. Since the winding *bc* is shunted by the load Z, it will be apparent that only part of the current in Z will flow through the secondary winding. The remainder will flow through the portion of the primary winding represented by *ab* and *cd*, and the generator. In other words, the primary current will flow through only a portion of the primary winding, and only a portion of the load current will flow through the secondary winding. In an ordinary transformer having entirely separate primary and secondary windings, on the other hand, all of the load current flows in the secondary winding, and all of the generator current flows in the primary winding. In a practical transformer the currents cause I^2R losses in the windings in which they flow in any case. But because of the fact noted above, the I^2R losses in the autotransformer are lower than in a transformer of the usual type, it being understood that the same size wire is used on each. This means that an autotransformer can be designed to have the same losses as a regular transformer, and still have less copper in the windings, thereby effecting a saving.

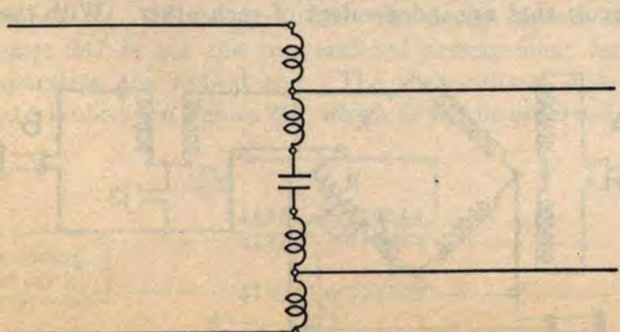


FIGURE 242

However, the direct electrical connection between windings is a disadvantage in power work since it introduces the hazardous possibility of obtaining the full primary voltage on the load side of the device if the secondary winding should become open.

In telephone work, it is sometimes desirable to use a transformer to change the effective impedance of a circuit without changing the circuit continuity for telegraph currents, D.C. signaling or D.C. testing. Figure 242 shows how this is accomplished through the use of an autotransformer; the condenser connected between the windings prevents any direct-current flow from one side of the circuit to the other.

117. The Hybrid Coil

In telephone repeater operation, as in duplex telegraphy, we must receive incoming energy and direct it into a receiving circuit (input) which is separate and distinct from the sending (output) circuit. This is essential inasmuch as the device used for amplifying voice-frequency currents can operate in one direction

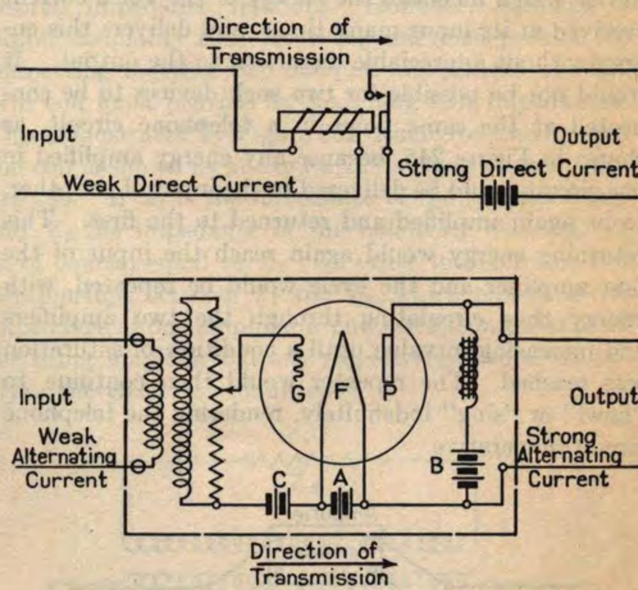


FIGURE 243

only. Its limitations in this respect are analogous to the telegraph relay, which repeats a direct-current signal from a circuit having a small amount of energy into one having a greater amount of energy (see Figure 243). The use of such one-way amplifiers, without some device for securing transmission in both directions, would be restricted to such a layout as is shown in Figure 244. This would require not only twice the circuit facilities for each long distance connection, but also special telephones at each terminal. We learned in Chapter XI how duplex telegraphy is accomplished over a single wire by application of the Wheatstone

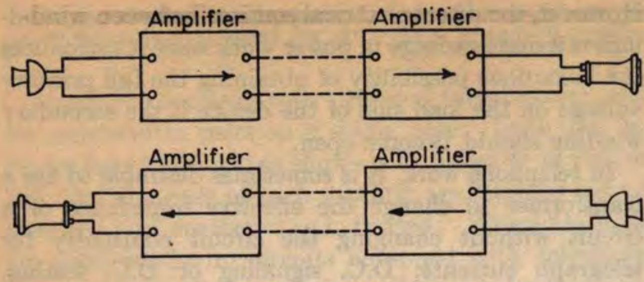


FIGURE 244

bridge principle and the use of an artificial line. The problem in the case of the telephone repeater is somewhat more complicated, but its solution is effected by employing the same principle of bridge balance, using an artificial line called a "balancing network".

In later chapters, we shall discuss the vacuum tube and its application in the telephone repeater. We may take up at this time the "hybrid coil", which makes two-way transmission over circuits equipped with one-way amplifying devices possible. For this discussion we may consider the amplifier circuit as a device which increases the energy of the voice current received at its input many times, and delivers this energy without appreciable distortion to the output. It would not be possible for two such devices to be connected at the same point in a telephone circuit, as shown in Figure 245, because any energy amplified in one circuit would be delivered to the input of the other, to be again amplified and returned to the first. This returning energy would again reach the input of the first amplifier and the cycle would be repeated, with energy thus circulating through the two amplifiers and increasing in value until a condition of saturation was reached. The repeater would then continue to "howl" or "sing" indefinitely, rendering the telephone circuit inoperative.

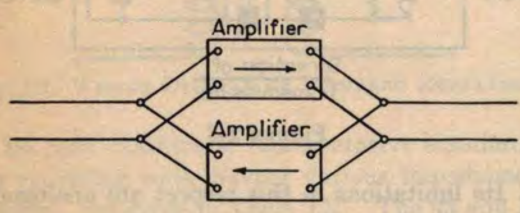


FIGURE 245

To eliminate the possibility of repeater singing, we must convert the ordinary telephone circuit into a receiving and a sending circuit which are independent of each other. That is, as in the case of the duplex set, the two circuits must be connected to the same line, yet any current flowing in one must not in any way affect the other. We can obtain this desired result by applying the principle of bridge balance, but the application is now to alternating currents. A Wheatstone bridge with proper modifications, however, can be operated with alternating current as well as direct current. To illustrate, in Figure 246 we have a repeating coil connected as an alternating-current Wheatstone bridge. Here the source of voltage is an A.C. generator instead of a battery, and instead of connecting the voltage to the points *a* and *b* as is usually done, the same results are accomplished by connecting it to the drop winding of the coil. The E.M.F. is then impressed across *a* and *b* by mutual induction instead of by direct connection, but the result is, of course, the same. In place of a direct deflecting galvanometer, we have substituted a telephone receiver which, for alternating current of the voice-frequency range, is even more sensitive. This circuit can now be used to measure the value of any resistance that may be connected to the *X* terminals. We can

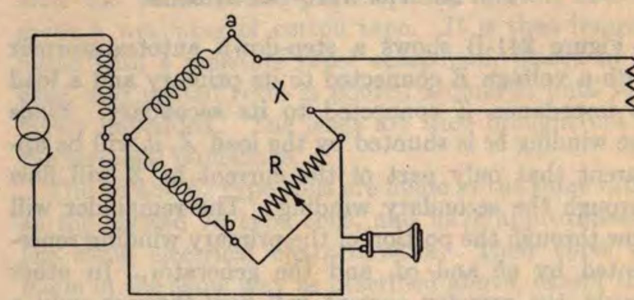


FIGURE 246

also use this circuit to measure any impedance that might be connected to the *X* terminals, provided the variable arm *R* has in series with it a variable reactance for balancing the reactive component of the unknown impedance.

Let us now assume that an alternating-current bridge circuit, such as that shown in Figure 246, but arranged to measure impedance as well as non-inductive resistance, has a transmitter substituted for its A.C. generator, and a telephone line terminating in a subset at the distant end, connected to the *X* terminals. Such an arrangement is illustrated by Figure 247. Here we have a device for terminating an ordinary telephone circuit so as to provide a receiving and a sending circuit that are independent of each other. With the

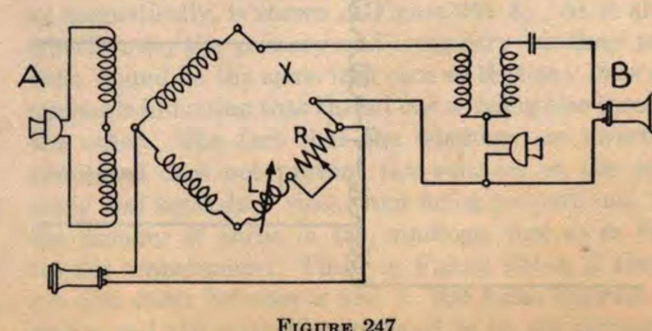


FIGURE 247

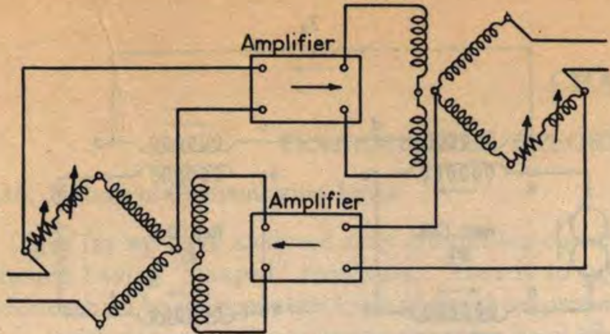


FIGURE 248

variable arm of the bridge adjusted to give perfect balance, any voice current in the transmitter circuit at Station *A* cannot be heard in the receiver circuit at that station, for the same reason that a galvanometer needle is stationary in any balanced bridge. We have double-tracked, so to speak, the ordinary 2-way telephone circuit.

If we now take two such circuits, and introduce two amplifiers as shown in Figure 248, we have a device that may be used as a telephone repeater at some intermediate point in a 2-wire telephone circuit. Here the energy coming from one amplifier cannot find its way into the input of the other and cause singing.

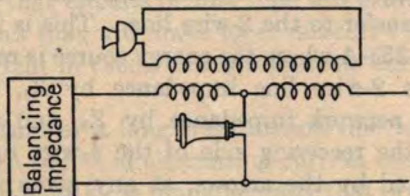


FIGURE 249

The coil that takes the place of the bridge mechanism in Figures 247 and 248, is known as a hybrid coil, sometimes called bridge transformer, three-winding transformer, repeater output transformer, etc. In the actual coil, there are a few additional details of design that do not permit the identity of the simple A.C. bridge circuit to be so readily recognized. These are not difficult to follow, however, after having been once pointed out. In the first place, the design shown in Figure 247 is not the conventional arrangement for illustrating the hybrid coil. The conventional schematic is shown in Figure 249, which, it will be observed

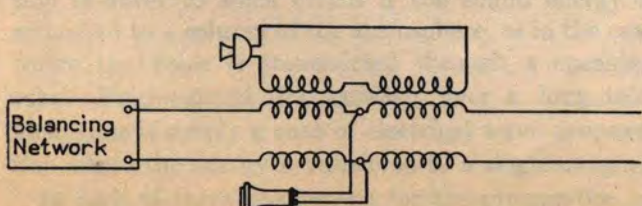


FIG. 250. THE HYBRID COIL

shows the same circuit connections as Figure 247 but is less similar to the standard convention for the Wheatstone bridge. In the actual hybrid coil, the line coils are divided and connected on both sides of the line as shown by Figure 250, in order that perfect symmetry in the wiring of the talking circuit may be maintained. Both sets of windings, of course, are inductively coupled to the third winding. Figure 251

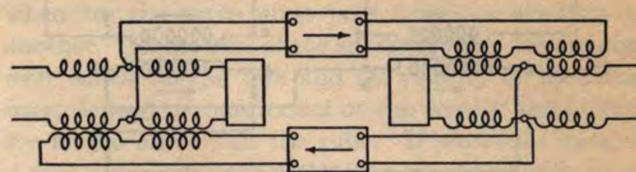


FIG. 251. TWO-WIRE TELEPHONE REPEATER CIRCUIT

shows the revised schematic of the amplifier connections to two hybrid coils in a 2-wire telephone repeater circuit.

In the hybrid coil, as in other transformers or repeating coils, the design must be such as to give the desired impedance relations. However, although a simple inequality ratio repeating coil must provide for connecting together two unequal impedances, the hybrid coil must provide for matching four impedances. This is illustrated by Figure 252, where for convenience the coil is shown as in Figure 249 instead of as in Figure 250. If Z_1 is the impedance of the telephone line and Z_2 the impedance of the balancing network, Z_1 is, of course, equal to Z_2. In order to determine the relationships between Z_3 and Z_4, which represent the impedance of one amplifier input and the impedance of the other amplifier output, respectively, we must analyze the electrical conditions.

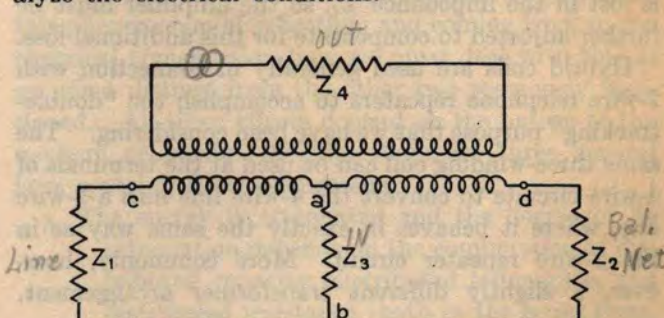


FIGURE 252

If we represent the source of voltage in the output circuit by a generator connected in series with Z_4, the energy supplied to the coil will obviously divide equally at the bridge, one-half going to each of the two equal impedances, Z_1 and Z_2. None will get to Z_3. The part going to Z_2, which represents the impedance of the network circuit, accomplishes no useful purpose and is lost. For this reason alone, the amplifier must

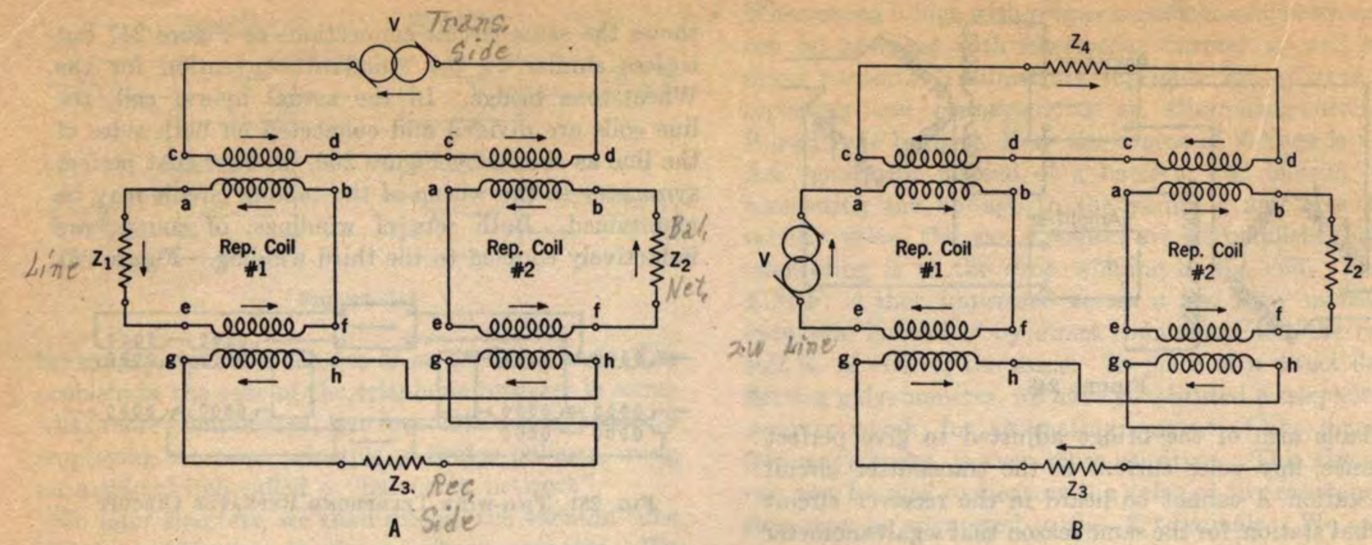


FIG. 253. 4-WIRE TERMINATING SET

be adjusted to supply twice the energy that is required for actual transmission. If, now, we simulate the conditions for inward transmission, connecting the generator in series with Z_1, the coil relations are such that half the energy goes to Z_3 and half is dissipated in Z_4, but none reaches Z_2. The voltage induced between c and a is equal to the voltage induced between a and d because the windings have the same number of turns and are on the same magnetic core. The turn ratio of the coil is fixed at such a value that the voltage induced in the latter winding is just equal to the voltage drop across Z_3. Consequently, points b and d are at the same potential. There is therefore no current flow between these points, and Z_2 consumes no energy. As before, however, half the incoming energy is lost in the impedance Z_4, so the amplifier must be further adjusted to compensate for this additional loss.

Hybrid coils are used generally in connection with 2-wire telephone repeaters to accomplish the "double-tracking" purpose that we have been considering. The same three-winding coil can be used at the terminals of 4-wire circuits to convert the 4-wire line into a 2-wire line, where it behaves in exactly the same way as in the 2-wire repeater circuit. More commonly, however, a slightly different transformer arrangement, known as a "4-wire terminating set", is used for this purpose. This consists of two ordinary repeating coils connected with one winding reversed, as shown in Figure 253.

The principle involved here is the same as for the hybrid coil proper, as may be seen by analyzing the circuit. Thus, we may consider first the case of energy coming from the transmitting side of the 4-wire line for transfer to the 2-wire line. This is illustrated by Figure 253-A where the energy source is represented by V, the 2-wire line impedance by Z_1, the equal balancing network impedance by Z_2, and the impedance of the receiving side of the 4-wire line by Z_3. As indicated by the arrows, at any given instant V sets up equal voltages in Z_1 and Z_2, but because the winding $g-h$ of repeating coil 2 is reversed, the voltage set up in this winding is opposed by the equal voltage set up in winding $g-h$ of repeating coil 1. As a result, no current is established in Z_3. Similarly, where the energy comes from the 2-wire line, as illustrated in Figure 253-B, equal voltages are set up in Z_3 and Z_4 and there is no current in the network, Z_2. This is because the direction of the voltage set up in winding $e-f$ of repeating coil 2 is such as to oppose the equal voltage set up in winding a–b.

CHAPTER XIX

TRANSMISSION THEORY OF LONG TELEPHONE LINES

118. Nature of Transmission Lines

Thus far we have analyzed only alternating-current circuits having "lumped" constants. That is to say, whenever we have encountered one of the three properties, resistance, capacity or inductance, we have considered it as pertaining to a specific piece of apparatus having a definite location in space. The only capacity we have known has been that which was a property of some device of definite size such as a condenser, and we have been able in every case to connect directly to the terminals of such a device. The same may be said of each resistance and inductance. This has simplified the make-up of the networks we have considered. In taking up the long transmission line, however, we shall find a different set of conditions. Though we shall not encounter any properties other than capacity, resistance and inductance, these will not be lumped. They will be more or less uniformly distributed along the entire length of the line, in fact they will be almost inseparably distributed. We can naturally expect, therefore, that circuits of this type will exhibit certain peculiarities that will make more difficult the analysis of the current in them, which represents energy transmission.

The nature of a long transmission line to which is connected a source of alternating-current energy, or an alternating E.M.F., is fundamentally that of a **medium** for wave propagation. It is another manifestation of the various forms of energy we have about us in all nature such as sound, heat, and light, being transmitted through some medium, though in this case we are dealing with electrical waves rather than sound, heat, or light waves. We speak of this form of transmission as "propagation". In all forms of propagation, the energy is in the form of moving waves and encounters opposition at every point in the medium. This tends to dissipate or cause the energy to die out, and we speak of this as the "attenuation" of the energy. A typical illustration is the case of sound energy being transmitted through the atmosphere. The attenuation is lower to some extent if the sound energy is restricted to a **column** of the atmosphere, as in the case where the voice is transmitted through a speaking tube. Voice-current transmission over a long telephone line is simply a case of electrical wave propagation where the energy is restricted to a single channel.

In each of these phenomena for the propagation of the various forms of energy, both the degree of attenuation and the speed at which the wave travels through the medium depend upon the nature of the medium. Furthermore, there are certain reactions that take place whenever the wave must pass from one medium to another. In the case of the speaking tube, the distance over which we can talk and the velocity of the sound wave depend to some extent on the density and humidity of the air within the tube. If we could imagine a case where one end of the tube was filled with air of a different density and degree of moisture saturation from that at the other end, we might hear an echo at the speaking end, due to a part of the energy being reflected back at the junction of the two transmitting mediums.

Perhaps a better illustration of the reflection phenomenon is the case of light, which has a definite velocity through the atmosphere but when it strikes a clear body of water such as a still lake, travels slower in the water than in the atmosphere. By our own observation, we know on the one hand, that this light may continue through the water until it illuminates pebbles on the bottom of the lake, while on the other hand, we find a mirrorlike reflection on the surface of the still water and know that some light is being reflected as it strikes the surface, in the same way that light is reflected when it strikes the surface of a mirror. It is only reflected in part, however, as we have evidence that some of the light has penetrated the more difficult medium. In all forms of wave motion we may have this phenomenon of reflection, and coming back to our electrical transmission line, we must deal with this as an effect distinct from the other two previously mentioned. All three effects depend on the nature of the medium or media. Briefly, there are three general laws covering these phenomena:

a. The energy is attenuated and the degree of its attenuation depends on the combination of distributed capacity, distributed inductance, and distributed resistance (both in the series form, as that of the conductors and in the shunt form, as that of leakage).

b. There is a definite speed at which the wave travels, which depends upon the electrical characteristics of the transmission line as established by the properties mentioned in *a* above.

c. There is a reflection of energy whenever the wave passes the junction of one transmission line with another, if the two lines have different electrical characteristics.

To analyze alternating-current flow to the most accurate degree under such conditions, where we have wave propagation rather than simple flow through a localized circuit with lumped properties, and must take into consideration the circuit properties as they exist and the conditions brought about by the uniform distribution of these properties over great lengths, would naturally involve the higher branches of mathematics. For most practical purposes, however, and for the applications that we meet in everyday telephone work, we may closely simulate or approximate the electrical make-up of any transmission line by some form of circuit having lumped properties.

In order that we may obtain a clear idea of the processes involved in as simple a manner as possible, we may profitably first consider this general problem on a direct-current basis. In doing this we will need to remember that such a treatment is largely hypothetical, as both telephone and telegraph transmission are essentially alternating-current phenomena; but we will, nevertheless, be able to establish certain general principles more easily than by handling the problem as an alternating current one from the beginning. Then having established these principles, we may revert to our alternating-current transmission problem and make such modifications as are necessary in order that they may apply equally well to the alternating-current case.

TABLE X

The Comparative Percentages of Power Delivered to a Receiving Device for Various Ratios of Its Resistance to the Internal Resistance of the Supply System, and the Efficiency at Which Power Is Supplied to the Receiving Device for the Same Ratios

VALUE OF R_2	% OF MAXIMUM $P_2 = 100 \times \frac{4R_2R_1}{(R_1 + R_2)^2}$	% EFFICIENCY $= \frac{100}{\frac{R_1}{R_2} + 1}$
2.0 R_1	88.9	66.7
1.1 R_1	99.8	52.4
1.0 R_1	100.	50.0
.9 R_1	99.7	47.4
.5 R_1	88.9	33.3
.2 R_1	55.6	16.7

119. The Transmission System

Any transmission system consists of three essential parts; a source of energy, a medium over which it is desired to transmit energy to a receiving device, and the receiving device itself, which usually converts the electrical energy into some form more useful. In a power transmission line, an electrical generator may be the source of energy; high voltage lines with transformers at either end may be the transmitting medium; a motor, lamp, or heater may be the receiving device for converting electrical energy into some other useful form. In a long distance telephone connection, a transmitter may be considered as the source of energy; the line from the speaking party to the listening party with all of its associated conductors, coils, and connections, may be thought of as the transmission medium, and the telephone receiver at the distant end may be considered as the third part of the transmission system, or the device which converts tiny electrical currents into audible vibrations of air called sound waves. Regardless of the kind of system, we must have these factors.

120. Transfer of Power

If a transmission system is to accomplish its purpose, it must be so designed that the energy transmitted from the source to the receiving device is sufficient to successfully operate the receiving device. As a secondary consideration it may be designed for power efficiency that is, regardless of the magnitude of the power delivered to the receiving device, the power lost in transmitting the energy from the source must be kept at a minimum. Although this is important in any transmission system, its special importance is in power transmission. In telephone work we probably think more of the primary purpose, that is, the **system's effectiveness in transferring the maximum quantities of power to the receiving device,** regardless of what percentage may be lost.

To illustrate the principles of power transfer and power efficiency, let us consider a small direct-current power distribution system. Such a system is usually a complicated network, consisting of a combination of many series and parallel resistances. When connecting a lamp to the lighting mains, we are concerned primarily in the transfer of power to the lamp. The lamp then, is the receiving device; the wiring from the lamp to the mains is the transmitting medium, and the mains are the source of energy. Looked at in this manner, the source of energy is no longer a simple device such as a battery, but is itself an energized network of complex make-up. Moreover, the current and voltage distribution in this energized network are influenced by the presence or absence of the lamp; current and voltage values elsewhere in the system will change as the lamp is connected to or disconnected from the mains. We know that our receiving device has a constant resistance, and for a constant voltage, draws a definite current. We further know that the power that is expended in the device is equal to its resistance multiplied by the current squared. This we may call the **useful expended power.** But if the current coming from the source of electromotive force, must traverse other resistances or other parts of the complicated network,

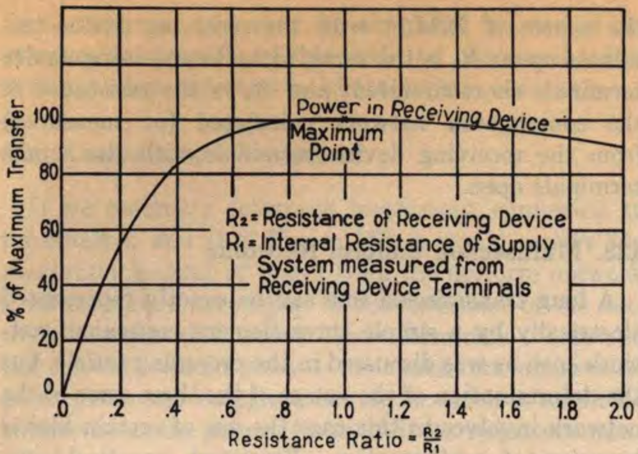


FIG. 254. POWER TRANSFER AS A FUNCTION OF RESISTANCE RATIO

as is the usual case, part of the power which is actually delivered by the source because of the connection of the particular receiving device, will be lost in the distribution system. The ratio obtained by dividing the power received by the device by the power expended by the source on account of its connection, is called the **power efficiency.** This will increase with increase in resistance of the receiving device. Other things being equal, therefore, we have the most efficient operation when the receiving circuit is one of very high resistance.

On the other hand, we may be interested in receiving all of the power possible, regardless of whether the operation under such circumstances is efficient or not. In the case of a telephone receiver at the end of a long transmission line, we are primarily interested in the receiver taking from the electrical system the maximum amount of power. **The condition for maximum transfer of power is obtained when the resistance of the receiving circuit is equal to the resistance of the network to which it is connected, as measured across the receiving terminals.** The simplest application of this is secured by connecting to a battery a resistance equal in magnitude to the internal resistance of the battery. In this case the battery will transfer to the external

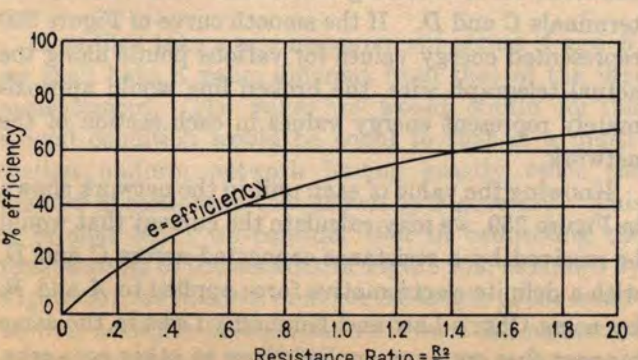


FIG. 255. EFFICIENCY AS A FUNCTION OF RESISTANCE RATIO

circuit the maximum amount of power, but in doing so will operate at an efficiency of only 50 per cent.

Figure 254 shows a curve which represents the power of the external circuit for various ratios of the resistance of the external circuit to that of the internal circuit. Figure 255 shows the efficiency for the same conditions. Table X gives the values from which these curves are plotted.

121. Pollard's Theorem

For the purpose of simplifying electrical calculations, we can consider any electrical system as one network supplying energy to another. One or the other of these networks may then be replaced by an equivalent circuit of maximum simplicity. **For every energized network there is an equivalent simple electrical circuit which consists of an E.M.F. and a resistance in series.**

This means that regardless of how complicated an electrical circuit may be, its effect in supplying current to any other circuit connected to it at two designated terminals, is equivalent to some source of electromotive force in series with a resistance. In other words, it is equivalent to a source of electromotive force, such as a battery, having an internal resistance of a definite value. This principle is called **Pollard's Theorem** and Figure 256 illustrates its use. E is a source of electromotive force connected to a complicated network; A and B are terminals to a particular branch of the complicated network. If it is desired to connect some receiving device to these terminals, the effect of this electrical system on the receiving device will be the same as that of the electrical system shown by Figure 257 where E' is the electromotive force measured across the terminals A and B of Figure 256, and R' is the resistance measured or calculated from the same terminals with the electromotive force E short-circuited. Pollard's Theorem may be briefly stated as follows:

> **The current supplied to an electrical device connected to two terminals of any electrical system is equal to the potential measured across these terminals before the device is connected, divided by the resistance measured or calculated across these terminals with the source of E.M.F. short-circuited, plus the resistance of the receiving device.**

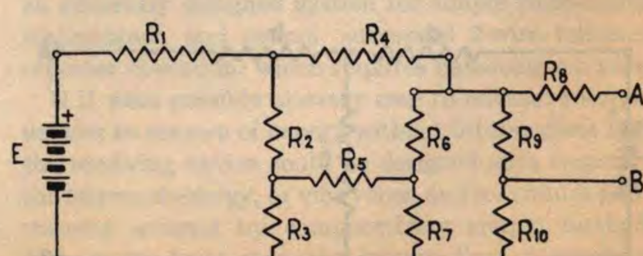


FIGURE 256

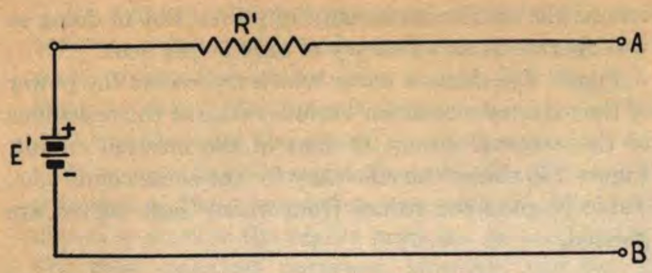


FIG. 257. APPLICATION OF POLLARD'S THEOREM TO THE NETWORK OF FIG. 256

122. Equivalent Networks

Pollard's Theorem gives us a method of substituting an equivalent circuit for any complicated electrical system, but in so doing we are required to replace the source of electromotive force with one having another value. It is often desired to determine **the simplest equivalent network for a complicated electrical system which will supply to some receiving device the same current as the electrical system and will take from the same source of electromotive force the same current as the electrical system.** A network consisting of three resistances of proper value arranged in the form of a **T** as shown by Figure 258 can always be substituted for any network, regardless of how complicated, and fulfill these conditions. For example, the circuit in Figure 258 may be substituted for that shown by Figure 256, and the current supplied to this system by the electromotive force E will remain unchanged, and the current received by a device connected to the terminals A and B will be the same. As we shall see in a later chapter, this same result can also be effected by means of a simple network having three arms arranged in the form of a π.

In determining values for the three resistances in an equivalent **T**-network such as is shown by Figure 258, the following equations may be used:

$$\text{Resistance of } a = R_1 - c \tag{72}$$

$$b = R_3 - c \tag{73}$$

$$c = \sqrt{(R_1 - R_2)R_3} \tag{74}$$

where R_1 is the calculated (or measured) resistance of the complicated network at the terminals connected to the source of E.M.F. with the receiving device terminals open; R_2 is the same with the receiving device terminals short-circuited; and R_3 is the resistance of the complicated network calculated (or measured) from the receiving device terminals with the source terminals open.

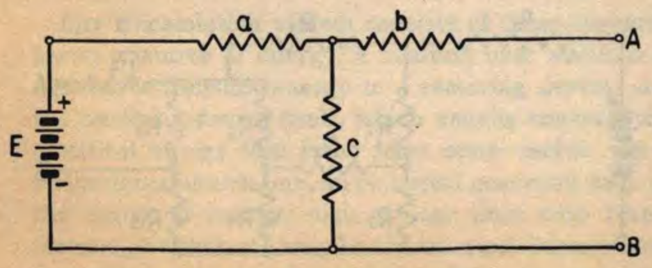


FIG. 258. T-NETWORK

123. Multisection Uniform Networks

A long transmission line can be exactly represented electrically by a simple three-element equivalent network such as was discussed in the preceding article, but the determination of the values of the three arms of the network involves in this case the use of certain higher branches of mathematics. For most practical purposes, we may deal with the transmission line by considering it as consisting of a number of separate sections. Treating at this time the direct-current case, we shall assume an approximately equivalent network for a transmission system such as a grounded telegraph wire 50 miles in length, and having a uniform leakage to ground throughout. We can imagine such a circuit as ten uniform sections, and for our purpose, may consider the leak to ground in each five-mile section as concentrated at the middle point. With these assumptions, our circuit may be represented by the network shown in Figure 259. Such a network is

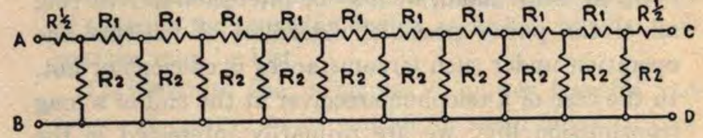


FIG. 259. MULTISECTION UNIFORM NETWORK

called a **multisection uniform network** because it consists of a number of identical units joined together. While a network thus constructed is not identical to the actual telegraph wire, we can construct one as nearly identical as we may desire by making our sections shorter in length. In this particular case, a 10-section uniform network would give a very small error in calculations for a receiving instrument connected across terminals C and D. If the smooth curve of Figure 260 represented energy values for various points along the actual telegraph wire, the broken line would approximately represent energy values in each section of the network.

Knowing the value of each unit in the network shown in Figure 259, we may calculate the current that would be received by a resistance connected across C and D, with a definite electromotive force applied to A and B, by using Ohm's Law and Kirchoff's Laws in the same manner that we have applied them to other networks. This procedure is rather laborious for long transmission lines, however, and may be simplified by use of an **attenuation** formula. This will be discussed after defining what is meant by characteristic resistance.

124. Characteristic Resistance

If we assume a telegraph instrument connected to terminals C and D of Figure 259, it would receive the maximum amount of power from the uniform network if its resistance were equal to the resistance of the network as measured across these terminals. Likewise the network would receive the maximum amount of power from any energized circuit to which it might be connected at the points A and B, if its resistance measured across A and B were equal to the resistance of the energized circuit. This we learned from the principle of maximum power transfer. But before connecting an energized circuit for sending to A and B or a circuit for receiving to C and D, let us measure the resistance of the network at A and B, and then construct a simple resistance R_0 of this measured value, and connect it to C and D.

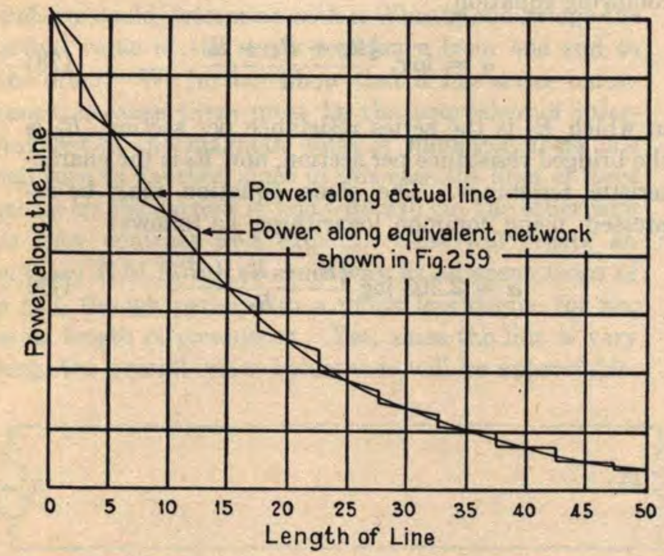


FIG. 260. COMPARISON OF POWER IN MULTISECTION UNIFORM NETWORK AND LINE WHICH IT SIMULATES

If we then take a new measurement across A and B, we shall have a value different from that of the first measurement. The value we would obtain for this second condition would be equal to that of a multisection uniform network having exactly twice the number of sections of that shown by Figure 259. This is evident when we consider that in connecting the resistance R_0 to the network of Figure 259, as shown by Figure 261, we in effect doubled the length of the multisection network, because the resistance R_0, connected to the terminals C and D, is equal to the resistance of the network measured from A and B when the ter-

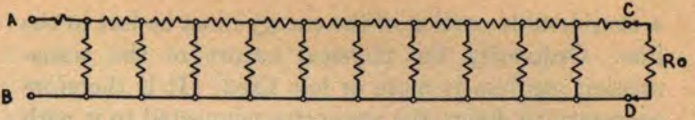


FIGURE 261

minals C and D are open. Therefore, Figure 261 is equivalent in all respects to a 20-section network open at the distant end, instead of the 10-section network shown by Figure 259. If the new measured resistance value is now substituted for the resistance R_0 in Figure 261, we shall have a network equivalent to a 30-section uniform network. If we again take measurements and replace R_0 with the still newer value and continue this practice indefinitely, each time in effect increasing the length of the network by ten sections, we will eventually have the equivalent of a line so long that anything connected to its distant end will have no effect upon the current which the sending element delivers to the line at the terminals A and B. Either short-circuiting or opening the distant end will not affect the equivalent resistance of the line.

This equivalent resistance, or the resistance of a network having an infinite number of sections, is called the **characteristic resistance.** Its value can be calculated from the relationship—

$$R_0 = \sqrt{\tfrac{1}{4}R_1^2 + R_1R_2} \tag{75}$$

where R_1 and R_2 are the elements of a network as shown in Figure 259, and R_0 is its characteristic resistance. If the resistance of the receiving device is made equal to R_0 and if the sending circuit supplying energy to the line is so designed as to have the same resistance as R_0, we shall have the conditions of maximum energy transferred both from the sending circuit into the line at A and B, and from the line into the receiving telegraph instrument at C and D. While in practice this condition may not be generally applied to telegraph operation, it does apply to long distance telephone circuit operation. Characteristic resistance bears the same relation to a direct-current transmission line as characteristic impedance bears to an alternating-current transmission line. In both cases, the principle is the same. It is paramount in the operation of long distance telephone circuits from two viewpoints—first, an efficiently designed system for simple voice-current transmission and second, successful 2-wire telephone repeater operation, which requires balancing networks.

If it were possible in every case to connect receiving devices to sources of energy without intermediate lines, the receiving device could be designed with respect to the source of energy, or vice versa, and maximum power transfer secured by comparatively simple methods. But, as we have seen, the intermediate transmission line complicates the problem; especially when at best a considerable portion of the energy must be lost in the line. Ordinarily the physical nature of the transmission medium is more or less fixed. It is therefore necessary to design the apparatus connected to it with respect to the characteristic resistance (or impedance) of the line rather than to design one unit with respect to the other.

125. Attenuation

This subject, too, has little importance when applied to direct-current circuits such as the telegraph circuit previously discussed, but it likewise may be treated more simply from the direct-current aspect. We shall examine at this time, therefore, what is meant by attenuation, and the use of attenuation formulas in calculating current or voltage values at points along, or at the distant end, of a transmission line. If in Figure 261 **the multisection uniform network has an infinite number of sections or is terminated in its characteristic resistance, R_0, the ratio of the current leaving any one section to that entering the section will be the same, regardless of what section is considered.** That is,

$$\frac{I_2}{I_1} = \frac{I_3}{I_2} = \frac{I_4}{I_3} = \text{a constant} \tag{76}$$

To illustrate this, let us assume that the current entering the network at A and B is decreased at the end of the first section to a given fractional value, for example $\frac{1}{2}$; the remaining current will be likewise decreased one-half to a value of one-quarter of the original at the end of the second section. In the same way, the current will be reduced to one-eighth of the original value at the end of the next section, to one-sixteenth at the end of the following section, and so on indefinitely. This "dying out" or attenuation is due to a part of the current in each section returning through the shunting resistance instead of flowing toward the receiving end, and thereby becoming lost in so far as transmission from one end of the network to the other is concerned. If, for example, we desire to calculate the current value at the distant end of a 10-section uniform network we must multiply the ratio of the current entering each section to that leaving each section by itself 10 times, or take the 10th power of the fraction I_2/I_1.

Such calculations are usually made by the use of logarithms. This permits an equation to be written giving the ratio of the current at the receiving end, I_n, to the current at the sending end, I_1 as follows:

$$\frac{I_n}{I_1} = \frac{1}{e^{n\alpha}} \tag{77}$$

where n is the number of sections, e is the base of the Naperian logarithm system, and α is the **attenuation constant.** The value of α can be calculated from the following equation:

$$\alpha = \log_e \frac{\frac{1}{2}R_1 + R_2 + R_0}{R_2} \tag{78}$$

in which R_1 is the series resistance per section, R_2 is the bridged resistance per section, and R_0 is the characteristic resistance. The same equation may be expressed, using common logarithms, as follows:

$$\alpha = 2.303 \log \frac{\frac{1}{2}R_1 + R_2 + R_0}{R_2} \tag{79}$$

CHAPTER XX

TRANSMISSION THEORY OF LONG TELEPHONE LINES—(*Continued*)

126. The Transmission Line as a Multisection Network

Now having made use of a direct-current analysis of the general problem of transmission over long distances to establish certain definitions and methods of attack, we may turn our attention to the less artificial but somewhat more complex problem of alternating-current transmission. Let us assume that Figure 262 represents a very long telephone line with alternating-current energy (such as that coming from a telephone transmitter) applied at one end, and some receiving device of impedance Z_R connected at the distant end. We know that such a line has series resistance. By short-circuiting the receiving device at the distant end, we could determine with a Wheatstone bridge the actual value of the series resistance from one end to the other. We further know that it has series inductance, because there must be the equivalent of interlinkages of the magnetic lines of induction from one coil turn to another, since in this case the lines of force set up by the current in one wire will cut the other wire as they contract and expand. This will create an induced E.M.F. in the same way as adjacent loops of a coil, though perhaps to a much less degree for the same length of conductor. Yet, since the line is very long, the overall series inductance will be appreciable.

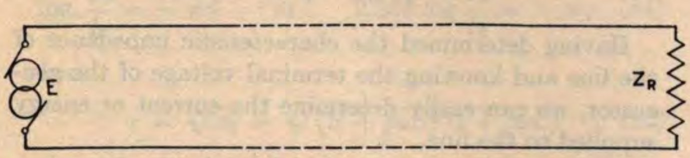


FIGURE 262

We further know that the line has some leakage which in any practical case will depend upon atmospheric conditions, but the insulation will never be so perfect that some leakage cannot be detected with a sufficiently sensitive instrument. There is one other property of the line. If it is open at the distant end, it will be found to act very much like a condenser. When a battery in series with a sensitive meter is connected to it, there will be a throw of the needle showing that the line temporarily is taking current to charge the two wires as though they were plates of a condenser.

Now let us assume that we know the resistance, inductance, leakage, and capacity of each mile of the circuit, and also its total length. If we evaluate the constants of the circuit in its entirety and attempt to use these values directly to build a simple network that will simulate the line, we will find the task impossible. Even a T-network made up of these "nominal" values will fail to simulate the line if the latter be of any great length; and the greater the length, the greater will be the electrical dissimilarity between line and network. We could, of course, construct an equivalent T-network, using constants determined by measurements as explained in the preceding chapter, which would exactly simulate the line, but we should find this **T** quite unlike the nominal **T**. The relationship between the two networks would not be a simple one and would necessitate the use of "hyperbolic trigonometry" for its determination. However, by taking shorter sections of line to simulate, we find a closer agreement

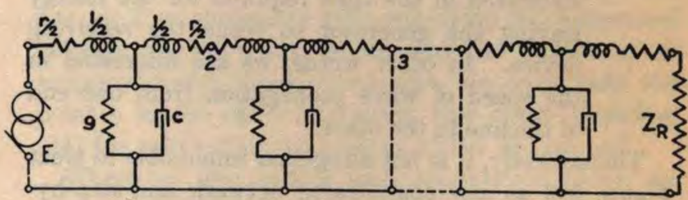


FIGURE 263

between the line and the nominal **T**, so that by considering the line as made up of a large number of extremely short sections, and constructing a multisection uniform network as illustrated in Figure 263, we can approximately simulate the line.

The degree of accuracy to which we approximate the line is going to depend on how far we go in breaking up the quantities into smaller parts. To begin with, we shall take an extreme case. Let us assume, for instance, that we have a circuit 1,000 miles long and are going to construct a network section for each foot, giving more than 5,000,000 sections for the network. Certainly we could not question the accuracy with which such a multisection uniform network would approximate the actual line. Assuming, therefore, that we have succeeded in so breaking up our distributed properties into tiny lumped properties which can be connected into a form of network, let us now accept this network, as illustrated by Figure 263, as equivalent for all practical purposes to the actual transmission line illustrated by Figure 262.

Now our interest in this network lies in—

a. The current that will leave the generator at the sending end and flow into the network. This will be determined solely by the impedance the network presents at the "voltaged" end, when connected to the generator. If V_0 is the terminal voltage of the sending device, and I_0 the entering current, then $I_0 = V_0/Z_0$ where Z_0 is called the "sending end impedance" of the line (when the line is infinitely long it is the "characteristic impedance" of the line). Since our line is 1,000 miles long and will for all practical purposes draw the same sending current from the generator as an infinite line, we can consider Z_0 in this case as the characteristic impedance.

b. We are next interested in what part of the energy leaving the generator will eventually reach the receiving device at the distant end. Or, since energy depends on both voltage and current, we are interested in what part of the generator's voltage will be impressed across the terminals of Z_R or what part of the generator's current will flow through Z_R.

c. For many transmission considerations we are also interested in the time required for the energy leaving the generator to reach the receiving device. In other words, we are interested in the speed of wave propagation from one end of the line to the other.

Theoretically, it is not altogether impossible to treat Figure 263 as any complicated network and step-by-step to calculate the impedance Z_0, as long as the number of sections is finite. In this case we are dealing with 5,000,000 sections, and it would be possible to calculate the current in each branch of the network or even through the distant receiving device, but certainly such extended computations would be impracticable and almost endless. The calculations for uniform multi-section networks are never made in this laborious manner. By a certain mathematical analysis, we derive short-cuts, based upon the following:

Knowing the make-up of the network sections, we might describe each as a series impedance z representing the series resistance and inductance of one foot of line, and a bridged impedance z_s, representing the bridged leakage and capacity of one foot; or instead of using z_s, we may for convenience use its reciprocal, which is called admittance and designated by the symbol y. We may express z and y in terms of the resistance, inductance, conductance (or leakage) and capacity for one foot of line. Let us represent these latter four quantities by R, L, G and C respectively. Here it should be noted that C is in farads and not microfarads. Now, the series impedance contains R and L and is given by the equation—

$$z = R + j\omega L \tag{80}$$

where z, R and L are as defined above, ω is equal to $2\pi f$, and j is an operator indicating 90° rotation as discussed in Article 109.

In the same manner, since G and C are bridged properties, we may write

$$y = G + j\omega C \tag{81}$$

An inspection of the make-up of each section would lead us to expect the characteristic impedance, Z_0, to become greater as z becomes greater, for the series impedance is tending to decrease the current which the generator attempts to establish. We should also expect an increase in the impedance of the shunt across each tiny section, the admittance of which we have designated as y, to permit less current to be shunted at each section and returned to the generator, thereby in its overall effect decreasing the amount of current that the generator would feed into the network. In other words, we should expect the quantity Z_0 to become greater as z_s becomes greater, or as y, which is the reciprocal of z_s, becomes smaller.

The value of Z_0 will tell us something of the nature of our transmitting medium, and since it is called characteristic impedance, it corresponds to the term "characteristic resistance" of the D.C. line, as discussed in Article 124. It can be shown that the value of Z_0 for an infinite line may be determined from a relatively simple equation, as follows:

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

Having determined the characteristic impedance of the line and knowing the terminal voltage of the generator, we can easily determine the current or energy supplied to the line.

Our next interest, as stated by b in the foregoing, is the part of this current or energy which will eventually reach the receiving device. Clearly it would be endless to proceed with ordinary network calculations, but again the calculations are simplified if we know **the degree to which each section of the network causes the current wave propagated along the line to die out.** Knowing this, we may say that the same attenuation when the line is treated as infinite applies to the voltage, because **the impedance of an infinite line is always the same when looking away from the sending end regardless of what junction of sections may be considered.** To illustrate, if we should open the multi-section network of Figure 263 and measure the impedance looking away from the generator at point 3, we would get the same result as if we measured the impedance connected to the generator. We would get the value Z_0, which is the characteristic impedance of the line. Since Z_0 always remains the same and must always be equal to V/I at any point along the network, V and I must be attenuated in the same ratio. If I becomes one-half of its value at some point along the line, then V must become one-half of its value, etc.

Now as we noted in Article 125, inasmuch as all sections are identical in their make-up, it can be seen that the loss or attenuation in each section will be the same, so that if the ratio of entering current to leaving current for the first section is nine-tenths, the ratio of currents for the second section and any succeeding section will be nine-tenths. Thus if we know the ratio of the current at point 2 to the current supplied the section by the generator, which we may represent by I_1/I_0, and wish to find the current at some point along the line, we can multiply this ratio by the succeeding ratios for each section as follows:

$$\frac{I_n}{I_0} = \frac{I_1}{I_0} \times \frac{I_2}{I_1} \times \frac{I_3}{I_2} \times \cdots \frac{I_n}{I_{n-1}}$$

or since

$$\frac{I_1}{I_0} = \frac{I_2}{I_1} = \frac{I_3}{I_2} \text{ etc.}$$

$$\frac{I_n}{I_0} = \left[\frac{I_n}{I_{n-1}}\right]^n \qquad (83)$$

This is sometimes written k^n where k is the value of any one of these ratios, but for convenience in computation, the ratio is usually expressed by logarithms—

$$\log_e \frac{I_n}{I_0} = -n\gamma \qquad \text{or} \qquad 2.303 \log \frac{I_n}{I_0} = -n\gamma \qquad (84)$$

where

$$\gamma = \sqrt{zy} = \sqrt{(R + j\omega L)(G + jwC)} \qquad (85)$$

and n denotes the number of sections traversed by I_n. Here we have a mathematical short-cut for our network calculations, which expressed in words, is as follows: **If we wish to know the relation between the current at any point along a transmission line and that delivered by the generator at the sending end, we can multiply the propagation constant of one section, γ, by the number of sections traversed, and the product taken negatively is 2.303 times the logarithm of the current ratio.**

But the quantity γ is more than a constant that gives the mere dying out effect of the current. The ratio of current I_n to I_0 is a relation of both effective values and phase difference, as both I_n and I_0 are vectors and are not necessarily in phase. This must be taken care of by treating γ as we treat all vectors; and γ is a vector quantity because it is equal to $\sqrt{zy}$ and both z and y are vectors. We must, therefore, separate the constant γ into two components, one of which applies to attenuation alone, and the other of which has to do with speed of propagation. We may write then, that—

$$\gamma = \alpha + j\beta \qquad (86)$$

where α is the symbol for the **attentuation constant** and β is the symbol for the **wave length constant.** Although it is practically always easier to evaluate α and β by making use of Equations (85) and (86), it is possible to write equations giving their values directly in terms of the primary constants — R, L, G and C. These equations are as follows:

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

$$\beta = \sqrt{\tfrac{1}{2}\sqrt{(R^2+\omega^2L^2)(G^2+\omega^2C^2)} - \tfrac{1}{2}(GR - \omega^2LC)} \qquad (88)$$

In the foregoing discussions of Figure 263, we have in each case designated some point along the line such as point n, at which we wish to determine the current. This applies to an infinite line. If, however, Z_R is equal to Z_0, or in other words, if the line at the distant end is terminated in a receiving device having an impedance equal to the characteristic impedance of the line, or if an inequality ratio repeating coil is inserted between the receiving device and the line so as to properly match these impedances, we could take the point n as the distant terminal and apply Equation (84) for calculating the current at the distant end.

Where we are dealing with attenuation alone, we may express (84) as follows:

$$-n\alpha = 2.303 \log \frac{I_n}{I_0} \qquad (89)$$

where I_n/I_0 is the ratio of current magnitudes only. In other words, the ratio is now an arithmetic comparison between current delivered and current sent, ignoring the fact that there may be some phase difference between the two currents. To convert this to power, we can use the expressions—

$$P_0 = E_0 I_0 \cos \theta$$

and

$$P_n = E_n I_n \cos \theta$$

Now, for the infinite line—

$$\frac{E_0}{I_0} = Z_0 \qquad \text{and} \qquad \frac{E_n}{I_n} = Z_0$$

whence

$$\frac{E_0}{I_0} = \frac{E_n}{I_n} \qquad \text{or} \qquad \frac{E_n}{E_0} = \frac{I_n}{I_0},$$

so that

$$\frac{P_n}{P_0} = \frac{E_n I_n \cos\theta}{E_0 I_0 \cos\theta} = \frac{E_n I_n}{E_0 I_0} = \left[\frac{I_n}{I_0}\right]^2 \qquad (90)$$

Now Equation (89) can be squared to give—

$$2.303 \log \left[\frac{I_n}{I_0}\right]^2 = -2n\alpha \qquad (91)$$

and combining this with (90), we have—

$$2.303 \log \frac{P_n}{P_0} = -2n\alpha \qquad (92)$$

also since

$$I_n = I_0 e^{-n\alpha} \qquad \text{and} \qquad E_n = E_0 e^{-n\alpha}$$

therefore

$$P_n = E_n I_n \cos\theta = E_0 I_0 e^{-2n\alpha} \cos\theta \qquad (93)$$

The power, therefore, is seen to die out or attenuate in a ratio which is the square of the current ratio.

In the foregoing we find for the most part a mathematical significance of α and β. Let us now analyze the physical circuit to determine what actually happens as the current is sent from point to point. In order to simplify the analysis, we shall start with an actual cycle of E.M.F. impressed on the sending end of a multisection network, and consider separately the effects of inductance and capacity on the propagation of this wave.

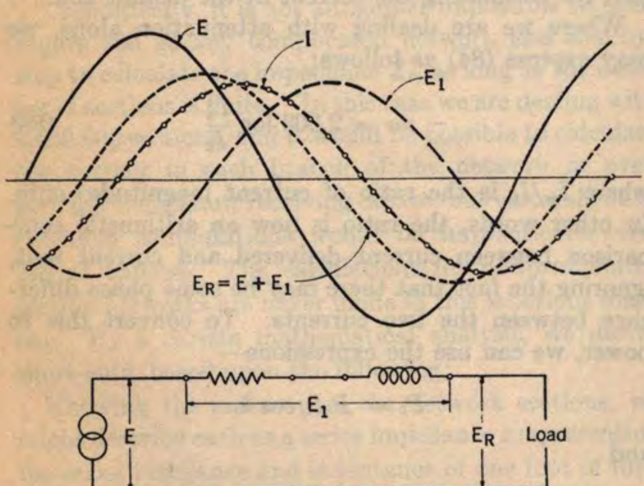


Fig. 264. Voltages and Current in an Inductive Circuit

From our previous study, we know that inductance acts to cause the current to lag behind the impressed voltage, so that in a circuit made up of resistance and inductance we would expect a lagging current. Figure 264 shows the time relationship between voltage and current in such a circuit, where E is the voltage curve, and I the current curve. This current sets up a back or induced E.M.F. E_1, which is the sum of the IR drop across the resistance and the IX drop across the inductance. It combines with the original voltage E to give the resultant voltage E_R on the load side of the inductance. The curve E_R is obtained by adding E and E_1 and it will be observed that the resulting curve lags E, the original voltage. A circuit containing resistance and capacity, on the other hand, produces a leading current as shown by Figure 265, and this current produces an IR drop which is opposite in phase with the current. Now if we combine the IR drop and the voltage, we obtain the resultant voltage E_R, which exists across the condenser and the load. This voltage likewise lags E, the original voltage.

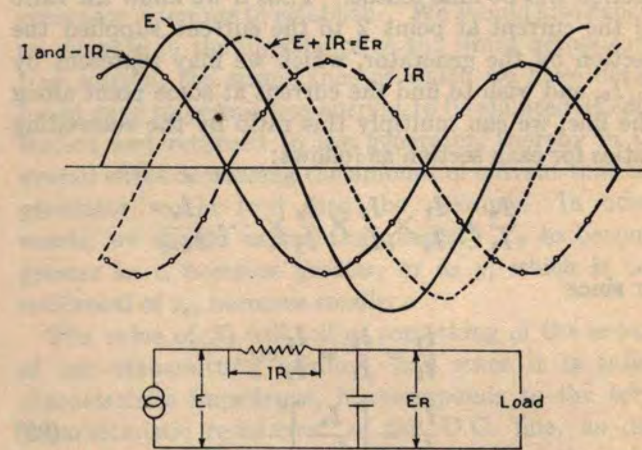


Fig. 265. Voltages and Current in a Capacitive Circuit

In both cases we have obtained a resultant voltage which lags behind the impressed voltage. Bridged capacity assists series inductance in the phase retarding effect. Due to the presence of reactance, therefore, the voltage has been "held back", so that the maximum voltages act later than they would if the reactance were removed. In other words, **the voltage wave has been slowed down.** Here, then, we have an explanation of the significance of the wave-length constant; it is merely an index figure to show how much the wave is retarded. Let us now apply our knowledge to the further study of the transmission line which we have represented by a series of **T**-sections. Each section, due to resistance and leakage, absorbs energy and therefore reduces the voltage which can act on the next section. Further, the voltage available at the next section lags behind the voltage impressed on the section, so that as we move away from the generator, the acting voltages are lagging farther and farther behind the generator voltage. Here we have a connecting link between geographical **distance** travelled along the line and **time.**

To bring this out clearly, let us assume that we take

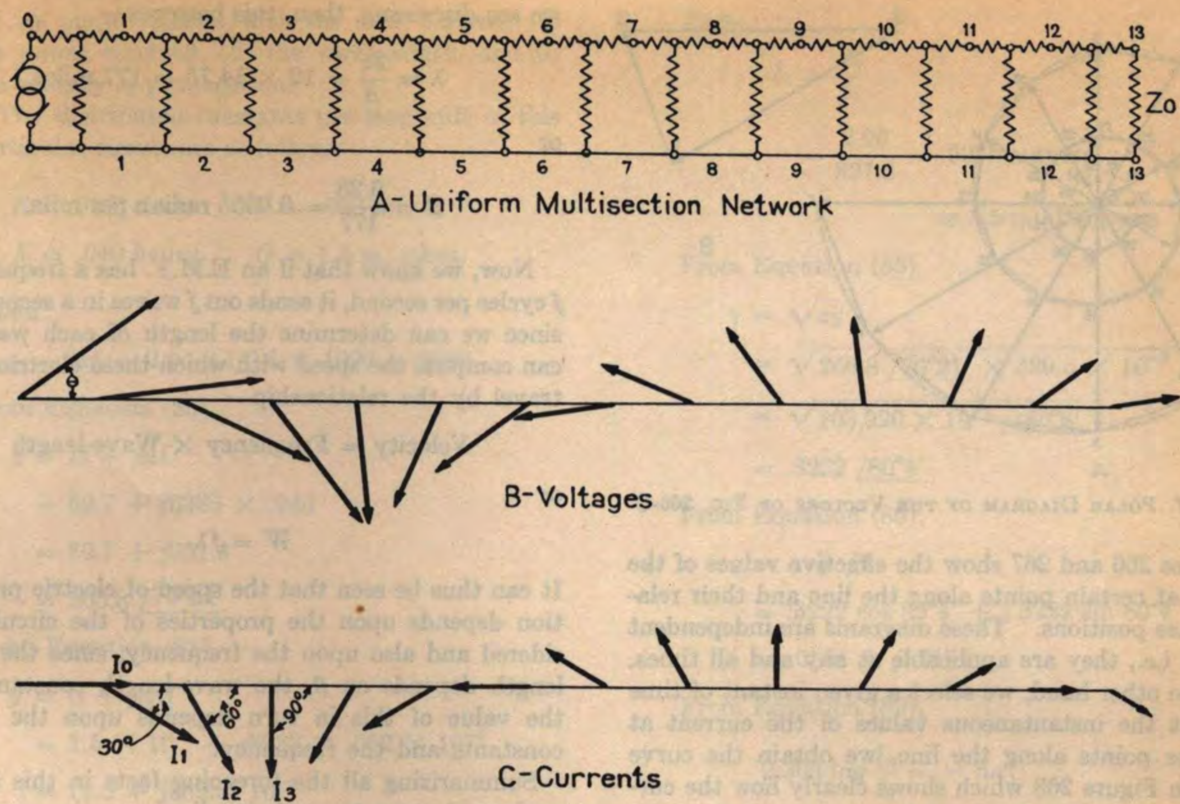


FIG. 266. TRANSMITTED CURRENTS AND VOLTAGES AT JUNCTIONS OF A MULTISECTION UNIFORM NETWORK

our sections of such a length that, for a frequency of 1000 cycles, the time lag between voltages can be represented by 30 degrees per section on the **time-voltage** diagram; if we simulate by each section fourteen and three-quarters miles of 104 open wire circuit, we will obtain such a relationship. In order to make the story complete, we will also assume the reduction in voltage magnitude due to resistance and leakage loss to be 0.895 per section. If we assume the original voltage E_0 to be 10 volts, the voltage at the end of the first section, E_1, will be 8.95 volts, lagging 30° behind E_0. E_2, at the end of the second section, will be 0.895×8.95 or 8.01 volts, lagging 30° behind E_1 or 60° behind E_0. If we represent the voltages at various points by vectors, we will obtain a system of vectors as shown in Figure 266-B, where the multisection network is shown as Figure 266-A and the voltage acting at each junction is directly below.

Since the ratio of current to voltage is constant, it follows that the chart representing currents will have the same form, with each vector proportional and removed by an angle θ from the corresponding voltage vector, where θ is the angle of the characteristic impedance Z_0. Thus we may treat a similar figure such as 266-C as a "distance-current diagram" where the vectors, I_0, I_1, I_2, etc., show the magnitude and relative phase of the currents at the network junctions. If now we refer all the current vectors to a common reference point, we will obtain a broken curve such as that of Figure 267-A, which shows graphically how the currents at various points are related. In this figure the vector $I_0 = G-0$ is the current entering the first section and $I_1 = G-1$, the current leaving that section. Then the vector 1-0 must be the current that passes through the shunt in the first section, because the sum of the current through the shunt and the current going ahead gives 1-0 as the resultant of the vector diagram. This is perhaps more clearly illustrated by Figure 267-B. For the same reason 2-1 will be the current passing through the second shunt, etc.

We may, therefore, conceive of the total entering current as the resultant of a number of component currents which flow from the generator through the various shunt paths and back to the generator, each component of a different magnitude and phase. The effect of these components can be observed, since at certain junctions the line current is flowing in the opposite direction to that taken by the entering current; at other points there is a 90° phase difference between the two; and at still other points there is no phase difference. In other words, the current vector may be considered as moving about G, rotating through 30° for every section traversed and diminishing in value about 10% in each section.

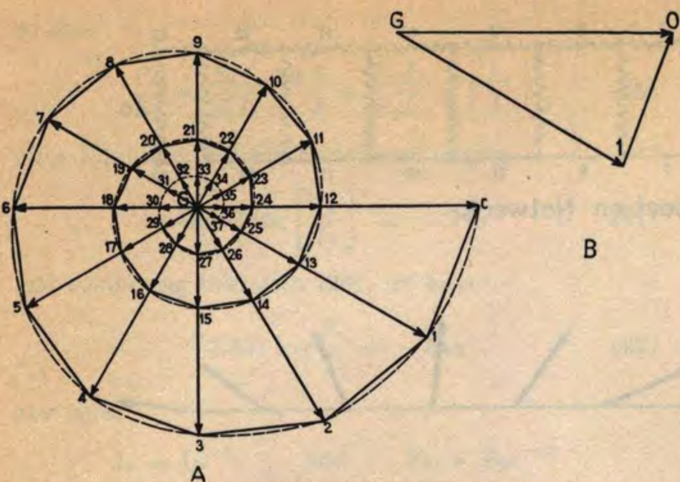


FIG. 267. POLAR DIAGRAM OF THE VECTORS OF FIG. 266-C

Figures 266 and 267 show the effective values of the current at certain points along the line and their relative phase positions. These diagrams are independent of time, i.e., they are applicable at any and all times. If on the other hand, we select a given instant of time and plot the instantaneous values of the current at the same points along the line, we obtain the curve shown in Figure 268 which shows clearly how the current reverses in direction as it passes through the various sections. A little study of this curve suggests that it is related to the sine curve, and such is actually the case. Due to the decay of current from section to section, the sine wave is somewhat distorted, but if the decay were eliminated, the curve would be a pure sine wave. A comparison of the method used to obtain Figure 268 with the method of deriving the sine curve will show this clearly.

Figure 268 shows graphically both ways in which the line has affected the propagation of the wave. The decrease in the height of each successive cycle illustrates the attenuation of the current. The fact that we have a succession of cycles plotted against distance instead of time also shows how there has been established by the medium a definite speed of propagation. For the particular frequency there is a definite length, viz., 12, which as expressed here is the number of sections for one complete wave. We may call this wave-length λ, and indicate a definite relation between λ and β as follows:

$$\lambda = \frac{2\pi}{\beta} \qquad (94)$$

since β is a constant for the line at a given frequency and is a measure of the amount of phase shift per section, and when obtained from Equation (86) or (88), is on the basis of radian measure. In other words, there are 360 degrees or 2π radians in one wave-length (or one cycle) λ. For the particular network we are discussing, then, this becomes—

$$\lambda = \frac{2\pi}{\beta} = 12 \times 14.75 = 177 \text{ miles}$$

or

$$\beta = \frac{6.28}{177} = 0.0355 \text{ radian per mile.}$$

Now, we know that if an E.M.F. has a frequency of f cycles per second, it sends out f waves in a second, and since we can determine the length of each wave, we can compute the speed with which these electric waves travel by the relationship—

$$\text{Velocity} = \text{Frequency} \times \text{Wave-length}$$

or

$$W = f\lambda \qquad (95)$$

It can thus be seen that the speed of electric propagation depends upon the properties of the circuit considered and also upon the frequency, since the wave-length depends on β, the wave-length constant, and the value of this in turn depends upon the circuit constants and the frequency.

Summarizing all the foregoing facts in this article, we have for our transmission line—

a. Both current and voltage are retarded.
b. Both current and voltage are attenuated.
c. The amount of attenuation and the amount of "slowing down" are determined by the physical properties of the circuit and by the frequency of the applied voltage.

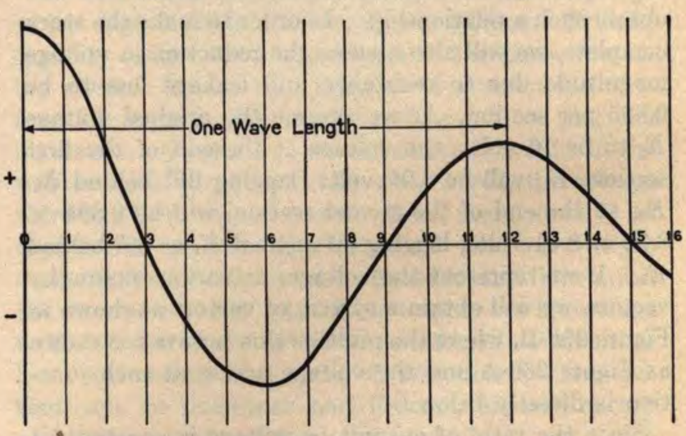


FIG. 268. STANDING WAVE OF CURRENT ALONG UNIFORM TRANSMISSION LINE

Example: Assuming a 50-mile, 19-gage H-44 side circuit terminated in its characteristic impedance and with an input power at the sending end of 10 milliwatts, calculate at 1000 cycles per second (1) the characteristic impedance, (2) the magnitude of the received current at the distant end and

(3) its phase relation with the sent current, (4) the power received, (5) the wave-length, and (6) the velocity of propagation.

The distributed constants per loop mile of this particular circuit are as follows:

$R = 89.7$ ohms $\quad C = .062$ mf.

$L = .040$ henry $\quad G = 1.5$ m. mhos.

Solution:

$$\omega = 2\pi f = 2 \times 3.1416 \times 1000 = 6283$$

From Equation (80),

$$z = R + j\omega L$$
$$= 89.7 + j6283 \times .040$$
$$= 89.7 + j251.3$$
$$= 266.8 \underline{/70°21'}$$

From Equation (81),

$$y = G + j\omega C$$
$$= 1.5 \times 10^{-6} + j6283 \times .062 \times 10^{-6}$$
$$= (1.5 + j389.5)\ 10^{-6}$$
$$= 389.5 \times 10^{-6} \underline{/89°47'}$$

From Equation (82),

$$Z_0 = \sqrt{\frac{z}{y}} = \sqrt{\frac{266.8 \underline{/70°21'}}{389.5 \times 10^{-6} \underline{/89°47'}}}$$
$$= \sqrt{684{,}980 \underline{/-19°26'}}$$
$$= 827.5 \underline{/-9°43'} \quad \text{Ans. (1).}$$

The input power, P_0, is

$$P_0 = E_0 I_0 \cos\theta$$

Substituting

$I_0 = E_0/Z_0$ in the above,

$$P_0 = \frac{E_0^2 \cos\theta}{Z_0}$$

or

$$E_0^2 = \frac{P_0 Z_0}{\cos\theta}$$

(**Note:** When Z_0 is a pure resistance, θ is zero and its cosine is one. Therefore, when θ is small in value it may, for all practical purposes, be disregarded.)

$$E_0^2 = \frac{.010 \times 827.5}{.9856} = 8.396$$

$$E = 2.90 \text{ volts}$$

Then

$$I_0 = \frac{E_0}{Z_0}$$
$$= \frac{2.90}{827.5} = .0035 \text{ ampere}$$

or 3.5 milliamperes

From Equation (85)

$$\gamma = \sqrt{zy}$$
$$= \sqrt{266.8 \underline{/70°21'} \times 389.5 \times 10^{-6} \underline{/89°47'}}$$
$$= \sqrt{103{,}920 \times 10^{-6} \underline{/160°8'}}$$
$$= .3222 \underline{/80°4'}$$

From Equation (86),

$$\gamma = \alpha + j\beta$$
$$= .3222 \cos 80°4' + j.3222 \sin 80°4'$$
$$= .0556 + j\ .3174$$

From Equation (89),

$$2.303 \log \frac{I_n}{I_0} = -n\alpha$$

or

$$2.303 \log \frac{I_0}{I_n} = n\alpha$$

$$\log \frac{3.50}{I_n} = \frac{50 \times .0556}{2.303} = 1.207$$

$$\frac{3.50}{I_n} = 16.11$$

$$I_n = \frac{3.50}{16.11} = .22 \text{ milliampere} \quad \text{Ans. (2)}$$

Phase shift per mile $= \beta$

$$\beta = .3174 \text{ radian or } 18.2°$$

Total phase shift for 50-mile circuit is

$$50 \times 18.2 = 910° \quad \text{Ans. (3).}$$

Then $\quad I_n = .22 \underline{/910°}$ milliampere

From Equation (89),

$$2.303 \log \frac{E_0}{E_n} = n\alpha$$

$$\log \frac{2.90}{E_n} = \frac{50 \times .0556}{2.303} = 1.207$$

$$\frac{2.90}{E_n} = 16.11$$

$$E_n = \frac{2.90}{16.11} = .18 \text{ volt}$$

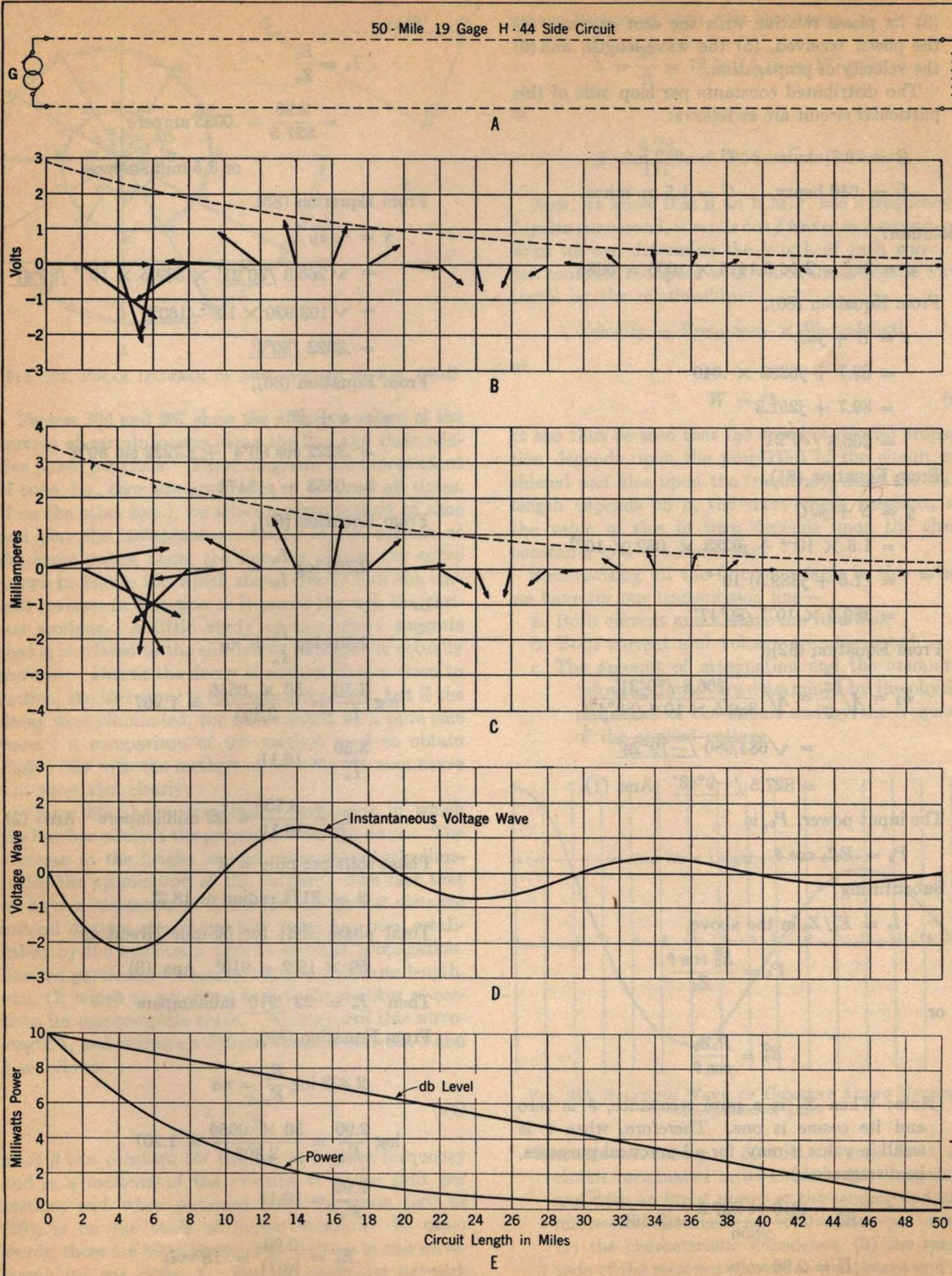

50-Mile 19 Gage H-44 Side Circuit
G
Z_o
A
Volts
B
Milliamperes
C
Voltage Wave
Instantaneous Voltage Wave
D
Milliwatts Power
Level in Decibels
db Level
Power
Circuit Length in Miles
E

FIGURE 269

From Equation (93),

$$P_n = E_n I_n \cos\theta$$
$$= .18 \times .00022 \times .9856$$
$$= .000039 \text{ watt}$$

or .039 milliwatt. Ans. (4)

From Equation (94),

$$\lambda = \frac{2\pi}{\beta}$$
$$= \frac{6.283}{.3174} = 19.79 \text{ miles.} \quad \text{Ans. (5)}$$

From Equation (95),

$$W = f\lambda$$
$$= 1000 \times 19.79$$
$$= 19{,}790 \text{ miles per second.} \quad \text{Ans. (6)}$$

The conditions along the line are graphically illustrated by Figure 269. The ordinates of the dashed curves in B and C represent the magnitudes of effective voltage and current at all points throughout the length of the circuit. The voltage and current vectors represent both magnitude and phase relation at the end of each 2-mile section. The standing voltage wave on the line is shown in D. This curve represents the instantaneous voltage values at all points on the line, i.e., the sine of the voltage vectors in B. The power at all points along the line is shown by E. As the power is proportional to the product of $EI \cos\theta$, it decreases faster than either the effective voltage or current values illustrated by the dashed curves in B and C.

127. Reflection and Transition Loss

We have noted that it is a characteristic of wave motion that in passing from one medium to another, a certain amount of the energy propagated by the wave is lost. For instance, light waves striking a pane of glass, water, or some denser medium are in part transmitted and in part reflected. The amount of energy reflected depends on the physical properties of the media through which the wave passes, the greater the dissimilarity, the greater the reflection. We may consider that such reflection is due to the different velocities with which the dissimilar media propagate energy, so that at the junction some interaction takes place, the result of which produces reflection, i.e., a change in the amount of energy propagated.

In transmitting electric waves, this reflection phenomenon is frequently met with, and it causes a "reflection loss". The amount of loss can actually be computed or measured, and if we take the case of two unequal impedances Z_1 and Z_2, the ratio of power received to the power that would be received on a smooth circuit ($Z_1 = Z_2$) is given by—

$$\frac{P_2}{P_1} = \frac{4Z_1Z_2}{(Z_1 + Z_2)^2} \times \frac{\cos\theta_2}{\cos\theta_1} \quad (96)$$

where Z_1 is one vector impedance with an angle θ_1, Z_2 is the other vector impedance with an angle θ_2, and the direction of propagation is from Z_1 to Z_2. In practical telephone work the factor $\frac{\cos\theta_2}{\cos\theta_1}$ is usually neglected, due to the fact that in the ordinary connection, substation to substation, this factor cancels out when considering the total reflection loss on the circuit. If it is remembered that reflection loss is a reduction in energy which is met with in all forms of wave propagation, a clearer conception of this phenomenon is obtained.

There is another so-called loss met with in transmission work which is known as "transition loss". In the preceding chapter we learned that if the load resistance was not equal to the resistance of the system to which it was connected, the power received by the load would not be a maximum. Similarly, in A.C. circuits, certain conditions must be met in order that the load may receive maximum power. Briefly stated, these conditions are that **the resistance of the load must equal the resistance of the generator and the reactance of the load must be of the same magnitude as the reactance of the generator but of opposite sign.** When these conditions prevail the two reactance components will cancel one another so that the circuit will behave as a D.C. circuit. It naturally follows that the resistance components must follow the D.C. law given in Chapter XIX. The transition loss, so-called, is in effect a comparison of the power that is received by a load under any given circuit conditions with the power that could be received if conditions permitted the maximum transfer of power. Usually this reduction in power is given in the form of a ratio in the same way that reflection loss is given by a ratio. If we designate by P_2 the power that is actually received by the load and by P_1 the maximum power that could be received, the ratio of the two is given by—

$$\frac{P_2}{P_1} = \frac{4R_1R_2}{(Z_1 + Z_2)^2} \quad (97)$$

Transition loss is not a true physical loss, nor is it a measure of the efficiency of the circuit; it is merely an indication of what percentage of the maximum power possible of utilization is being utilized.

128. Units for the Measurement of Transmission Losses and Gains

As in dealing with any other quantity, we require some unit of measurement when dealing with the energy losses due to attenuation in the transmission of human speech, or in the transmission of any alternating current from a sending device to a receiving device over a long line or through complicated circuits. Without some such unit we would be handicapped in giving any scientific expression to the grade of telephone transmission under various conditions. It would be natural for us to say that sufficient energy had been transmitted from the speaking station to the listening station for the listener to hear distinctly every spoken word, or to say that the sound coming from the receiver at the receiving station was so faint as not to be intelligible, but this would be a crude method of comparison. For the same reason that we need some adopted standard as a unit of length, such as the foot or the meter to measure distance, we require some standard for the measurement of transmission loss or transmission gain in telephone work.

For many years the unit used for this purpose was the "standard cable mile". This represented the loss due to one mile of an old type of standard 19-gage cable, having a resistance of 88 ohms per loop mile and a capacity of .054 mf. per mile. In this cable the series inductance and the shunt leakage were negligible, while the bridged capacity of .054 mf. was appreciable. It therefore attenuated the various frequencies that make up the band for telephone transmission unequally, attenuating the higher frequencies more than the lower frequencies. To illustrate, the attenuation constant α was equal to .109 for 800-cycle frequency and .122 for 1000-cycle frequency, etc.

This meant that the percentage reduction in power caused by inserting a mile of standard cable between a sending and receiving element was different for different frequencies. Under these conditions, to say that a telephone circuit had an equivalent of a certain number of miles of standard cable was largely meaningless unless the frequency at which the equivalent was computed or measured was stated at the same time. This rather confusing situation led to the dropping of the mile of standard cable altogether as a unit of measurement and the substitution of an arbitrarily selected unit not differing greatly in magnitude from the standard cable mile through the voice range, but having exactly the same significance at any and all frequencies. That is to say, **the new unit, called the decibel** (abbreviated "db"), **represents always a fixed percentage reduction in power no matter what frequency is involved.** Its magnitude may perhaps be best grasped by remembering that in a circuit equating to ten db the output power will always be one-tenth of the input power. Mathematically, the power ratio for **one** db may be expressed as—

$$\frac{P_1}{P_0} = 10^{.1} \tag{98}$$

where P_1 is input power and P_0 is output power. This corresponds to a current ratio of $10^{.05}$ and to an attenuation constant value of $\alpha = .115$. Table XI showing the power ratios for several values of decibels will aid in forming a clear conception of the magnitude of the unit.

TABLE XI

Relation between Decibels and Power Ratios for Gains and Losses

Decibels	Approximate Power Ratio: For Losses — Fractional	For Losses — Decimal	For Gains — Decimal
1	4/5	.8	1.25
2	2/3	.63	1.6
3	1/2	.5	2.0
4	2/5	.4	2.5
5	1/3	.32	3.2
6	1/4	.25	4.0
7	1/5	.2	5.0
8	1/6	.16	6.0
9	1/8	.125	8.0
10	1/10	.1	10.0
20	1/100	.01	100.0
30	1/1000	.001	1000.0

For any given power ratio the number of *db* corresponding can be determined by the following simple formula—

$$\text{No. of db} = N = 10 \log \frac{P_1}{P_0} \tag{99}$$

or, if the current ratio rather than the power ratio is known—

$$N = 20 \log \frac{I_1}{I_0} \tag{100}$$

or from Equation (77)—

$$N = 20 \log \frac{I_1}{I_0} = 20 \log e^{\alpha} = 20 \times \alpha \times \log e$$

$$= 20 \times .434 \times \alpha = 8.68\alpha \tag{101}$$

Although in the above we have been considering the decibel in connection with measurements of "loss" or attenuation, it is equally useful in the measurement of "gain" such as that given by a telephone repeater. A telephone repeater would be said to have a gain of so many db if the circuit in connection with which it were used was effectively shortened or had its net attenuation reduced to that extent.

CHAPTER XXI

LOADING

129. The Effect of Line Characteristics on Attenuation

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

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

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

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

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

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

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

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

130. Loading as a Means of Reducing Attenuation

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

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

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

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

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

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

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

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

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

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

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

131. Loading to Reduce Distortion

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

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

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

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

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

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

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

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

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

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

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

On the uniform line, current passes continuously from the positive voltaged wire to the negative voltaged

Development of the Loading Coil

wire through the small capacity elements distributed along the circuit, whereas in the network this passage of current can take place only at the middle of the T-sections. Thus, the inductive effect in the network will be somewhat greater than that in the uniform line due to the larger currents traversing the inductances, while the capacity effect will be correspondingly less. Now, of course, the greater the number of sections in the network, the less will be the difference in behavior of the uniform line and the network. But since any difference is increased with higher frequency, any network analysis of a uniform line really resolves itself into a determination of how many sections shall be taken in order that the simulation of the line by the multisection network may be satisfactory over a given frequency range. Similarly, on loaded circuits, the coil spacing (which in effect merely defines the length of each section of the multisection network), is so chosen that the change produced by frequency, over a given range, does not differ greatly from the effect that would be produced on a continuously loaded circuit of about the same average constants, R, L, G, and C. It should be noted, however, that either the value of the loading inductance will be a factor in determining the spacing since its inductance will be a part of the simulating network, or any chosen standard spacing will be a factor in determining the inductance values the coils may have.

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

132. The Cut-Off or Critical Frequency

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

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

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

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

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

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

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

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

and since

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

we have—

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

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

134. Mechanical Analogy of Loading

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

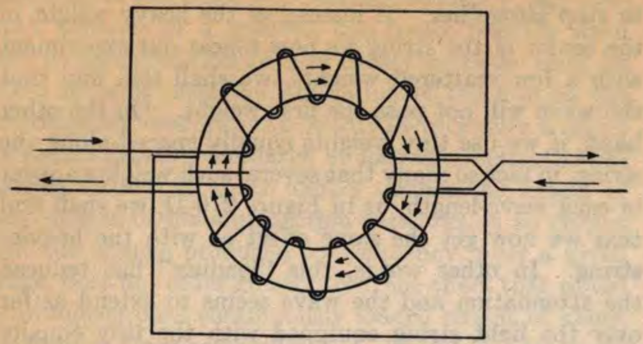


FIG. 271. WINDING OF SINGLE LOADING COIL

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

135. Features in Loading Coil Design

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

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

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

136. Important Considerations in Loading Practices

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

a. In connecting a loading coil, care must be exercised to prevent a reversal of one winding, thereby neutralizing the coil's inductance.

b. The inductance values of loading coils should be kept within 2% where the circuit is used in connection with telephone repeaters.

c. The loading coil spacing should be accurate to within 2% where the circuit is used in connection with telephone repeaters.

d. To prevent loading coil magnetization, the line current used for telegraph operation should not exceed the specified limits for the particular type of loading.

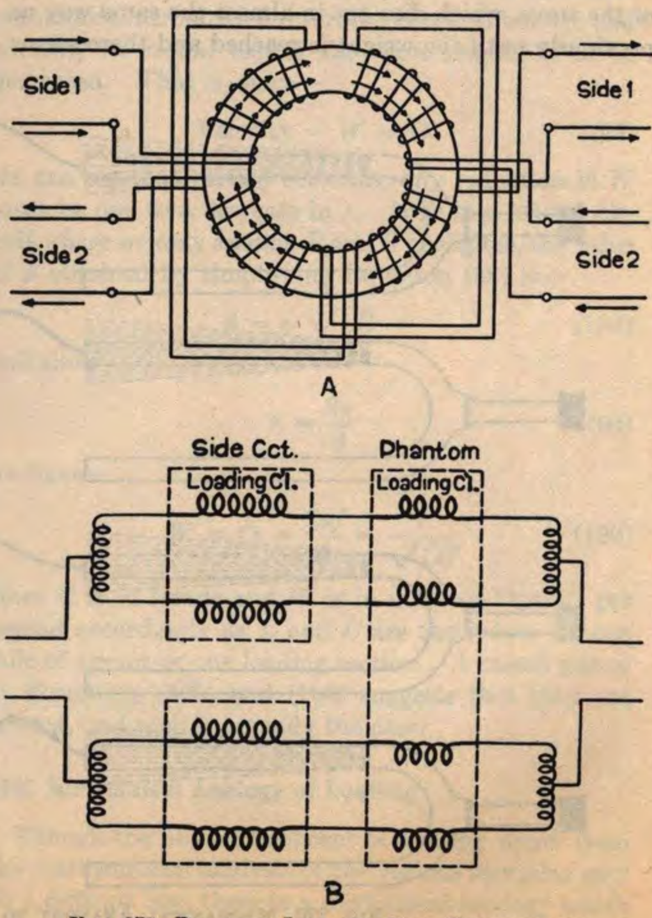


FIG. 272. PHANTOM LOADING COIL CONNECTIONS

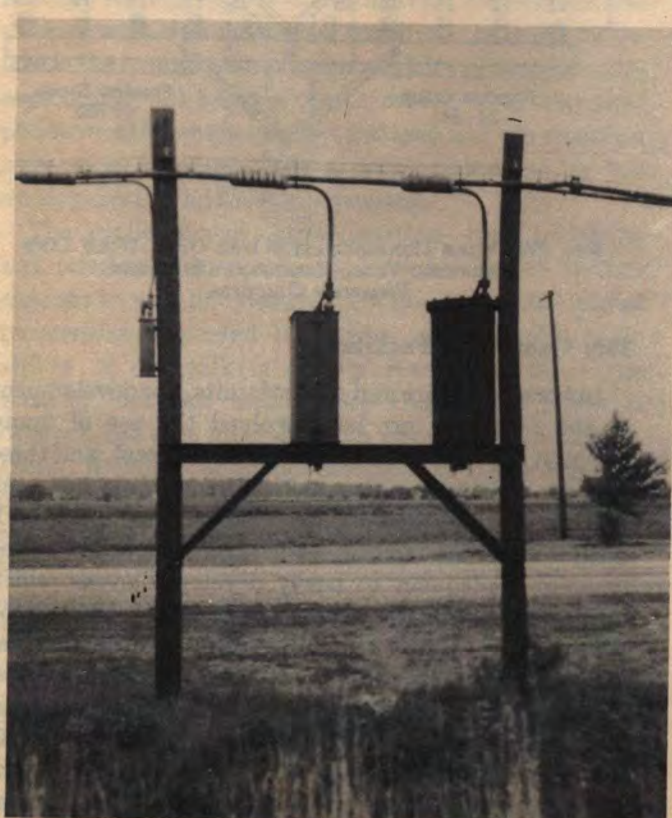
e. It is practicable to load the side circuits of a phantom group without loading the phantom or to load the phantom without loading the side circuits, but one side circuit cannot be loaded without the other.

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

137. Building-out Short Loading Sections

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

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



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

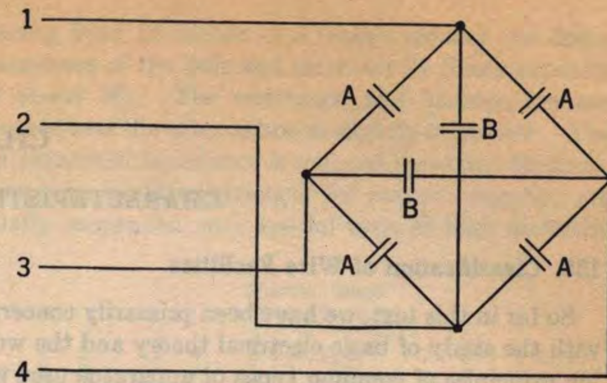


Fig. 273. Phantom Group Building-out Condenser

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

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

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

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

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

and

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

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

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

CHAPTER XXII

CHARACTERISTICS OF CIRCUIT FACILITIES

138. Classification of Wire Facilities

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

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

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

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

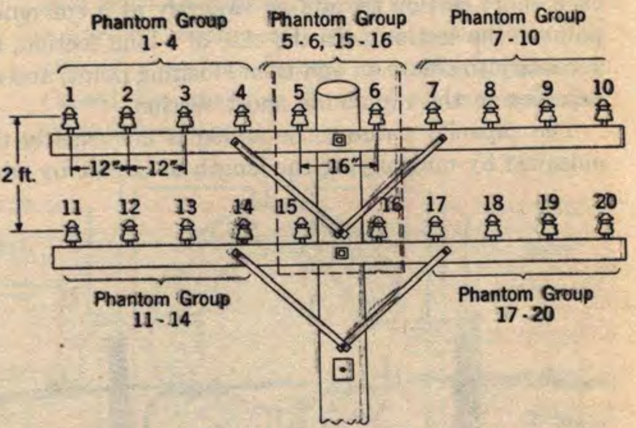


FIG. 274. WIRE CONFIGURATION FOR OPEN WIRE LINE CARRYING VOICE-FREQUENCY SIDE AND PHANTOM CIRCUITS

139. Open Wire Facilities

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

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

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

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

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

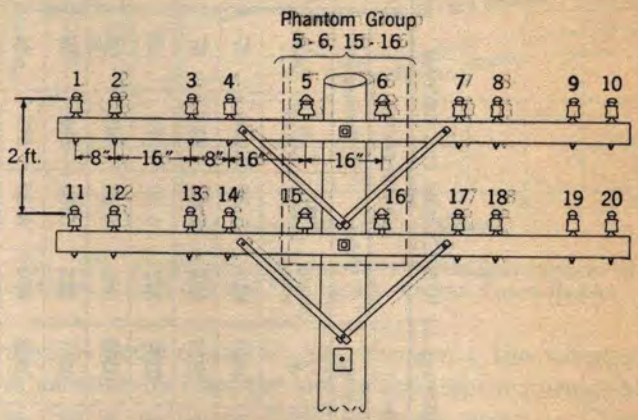


Fig. 275. Wire Configuration for Open Wire Line on which Type-C Carrier Systems are Superimposed

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

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

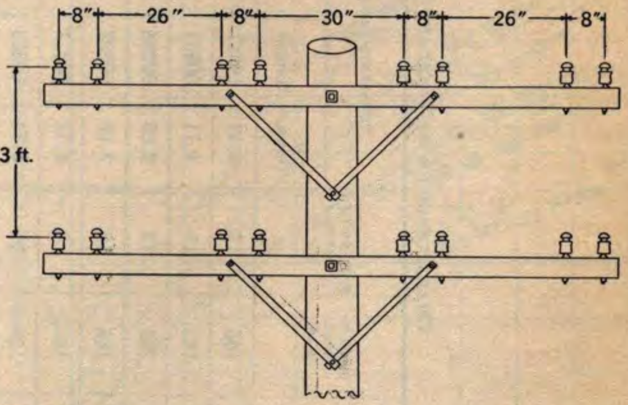


Fig. 276. Wire Configuration for Open Wire Line on which Type-J Carrier Systems are Superimposed

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

TABLE XII

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

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

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

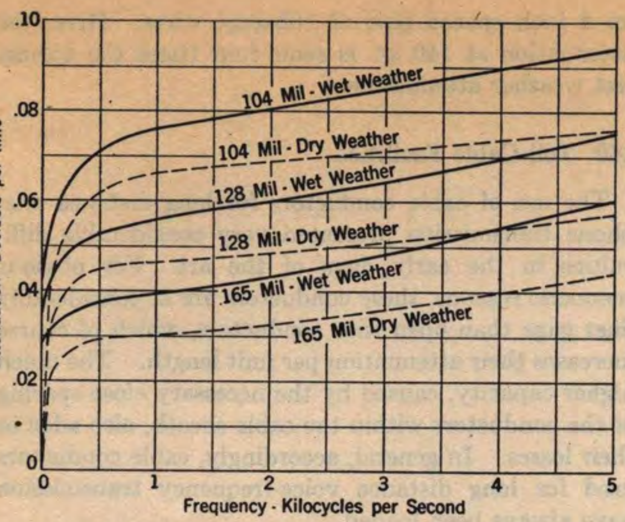


FIG. 277. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE SIDE CIRCUITS OVER THE VOICE RANGE

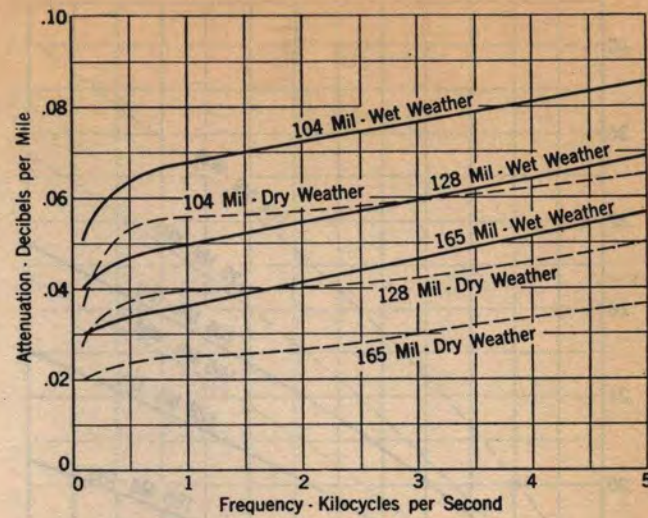


FIG. 278. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHANTOM CIRCUITS OVER THE VOICE RANGE

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

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

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

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

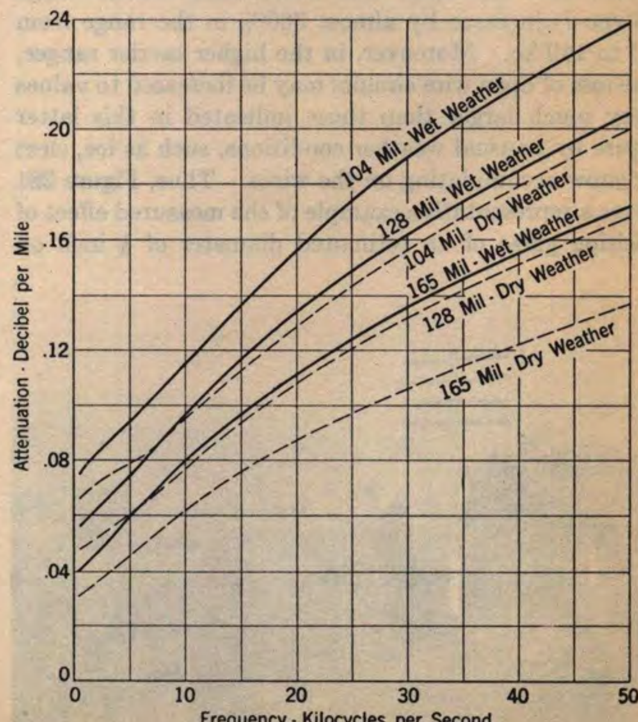


FIG. 279. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHYSICAL CIRCUITS OVER THE TYPE-C CARRIER RANGE

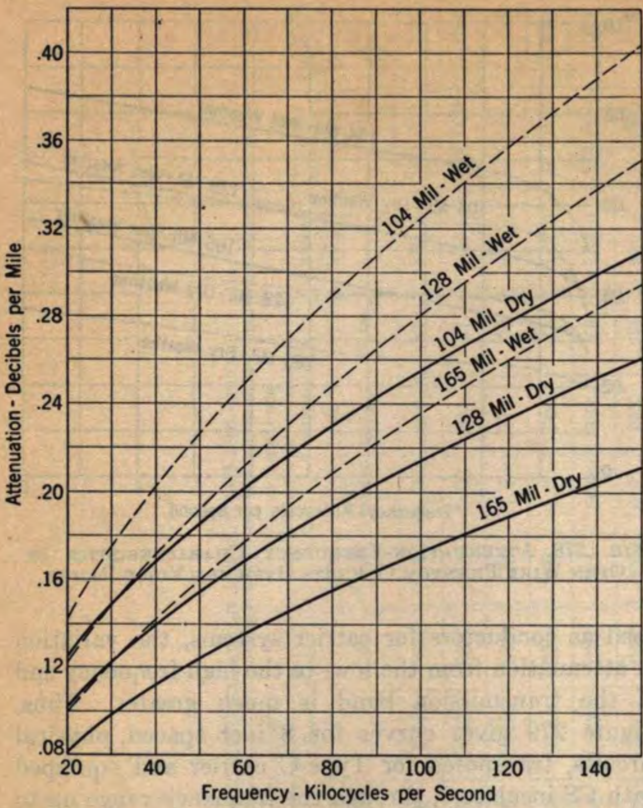


FIG. 280. ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHYSICAL CIRCUITS OVER THE TYPE-J CARRIER RANGE

losses over the open wire broad-band carrier range (Type-J) increase by almost 300% in the range from 20 to 140 kc. Moreover, in the higher carrier ranges, the loss of open wire circuits may be increased to values very much larger than those indicated in this latter figure by unusual weather conditions, such as ice, sleet or snow accumulating on the wires. Thus, Figure 281 gives a representative example of the measured effect of melting glaze of an estimated diameter of ½ inch on an 8 inch spaced pair of 165-gage wires. Here, the attenuation at 140 kc. is some four times the normal wet weather attenuation.

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

140. Toll Cable Facilities

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

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

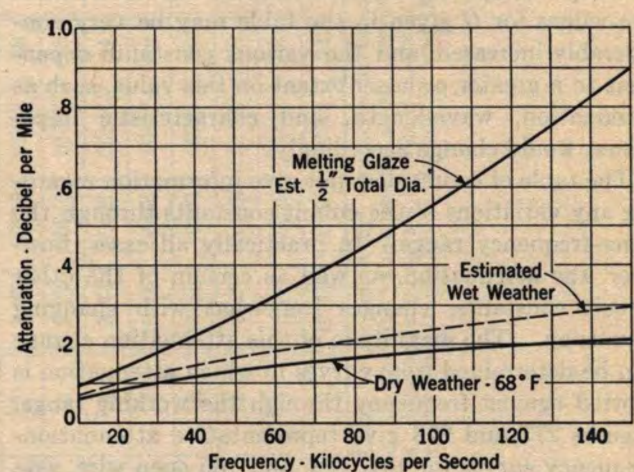


FIG. 281. CURVE SHOWING THE EFFECT OF SLEET DEPOSIT ON ATTENUATION OF OPEN WIRE CIRCUIT

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

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

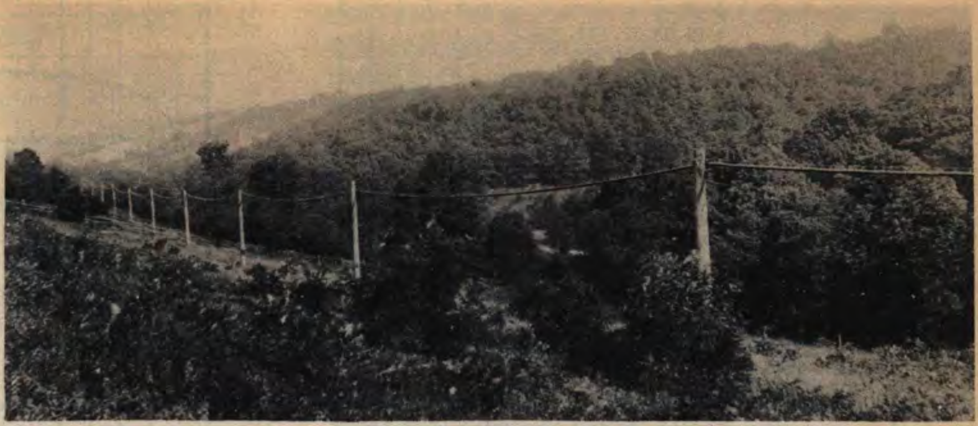
TABLE XIII

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

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

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

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



Toll Cable

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

TABLE XIV

Loading Coil Spacing Code

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

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

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

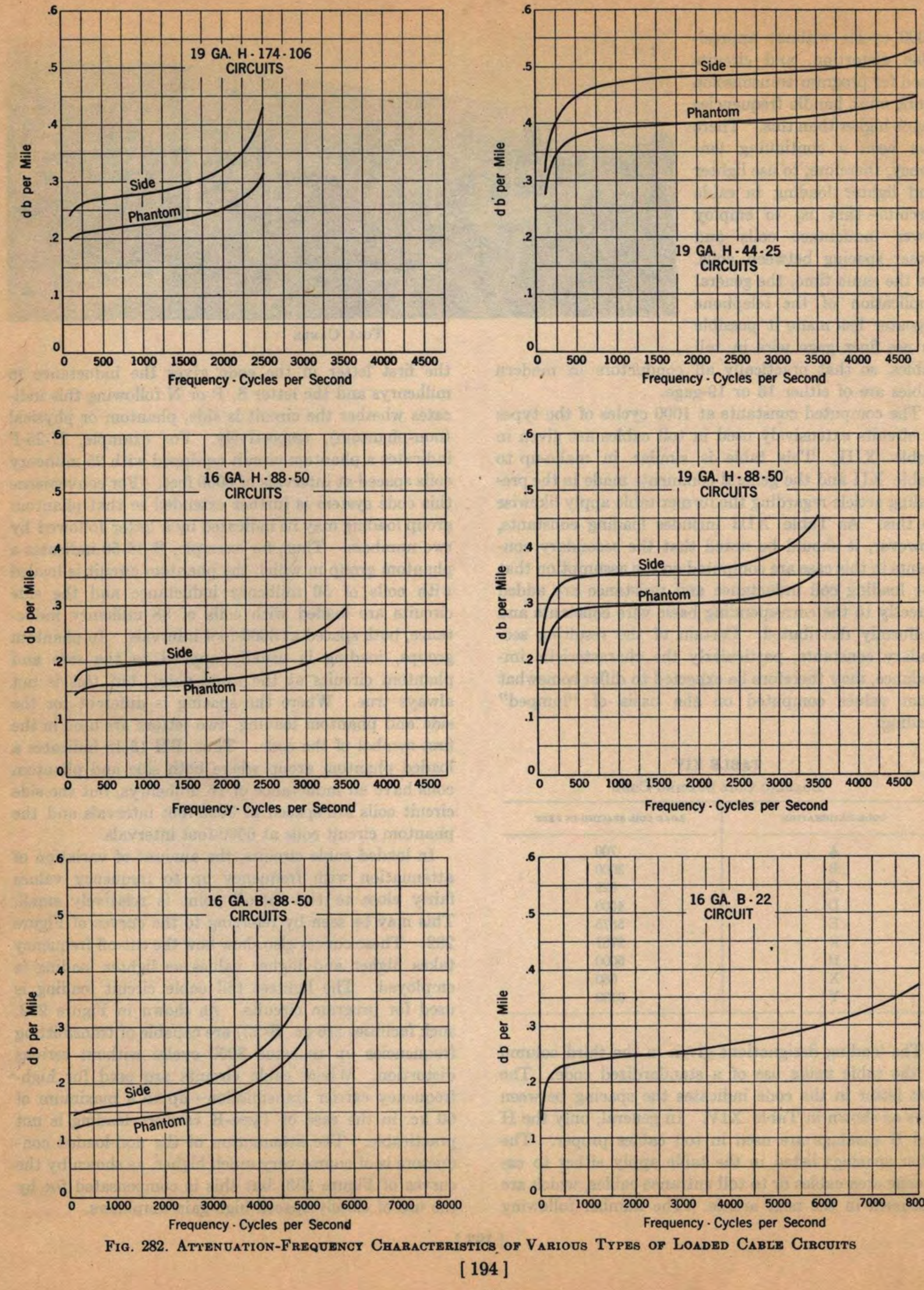


FIG. 282. ATTENUATION-FREQUENCY CHARACTERISTICS OF VARIOUS TYPES OF LOADED CABLE CIRCUITS

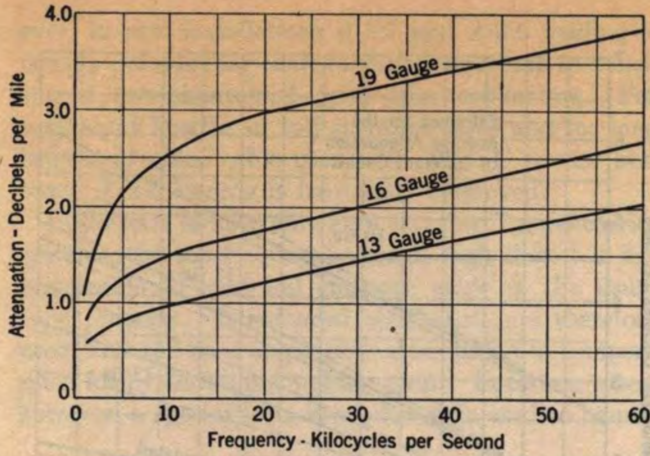


FIG. 283. ATTENUATION-FREQUENCY CHARACTERISTICS OF NON-LOADED CABLE CIRCUITS OVER THE TYPE-K CARRIER RANGE

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

COAXIAL CABLE

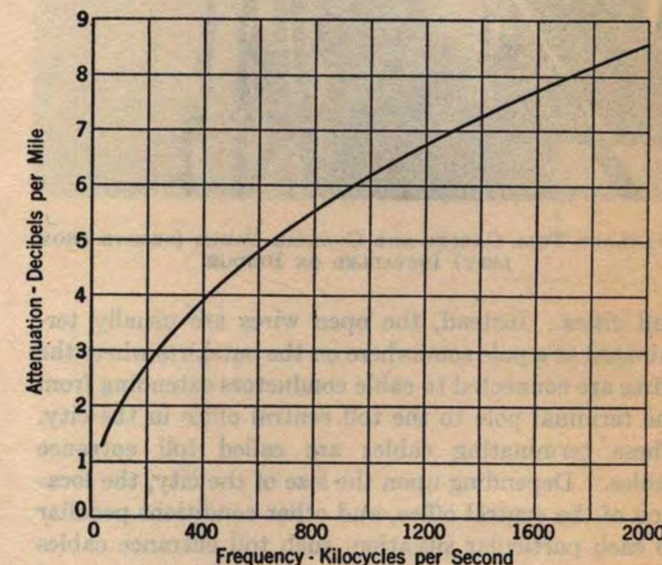


FIG. 284. ATTENUATION-FREQUENCY CHARACTERISTIC OF COAXIAL CIRCUIT

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

141. Toll Entrance Cable Facilities

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

STANDARD TOLL CABLES AND COAXIAL CABLE (SECOND FROM LEFT) INSTALLED ON BRIDGE

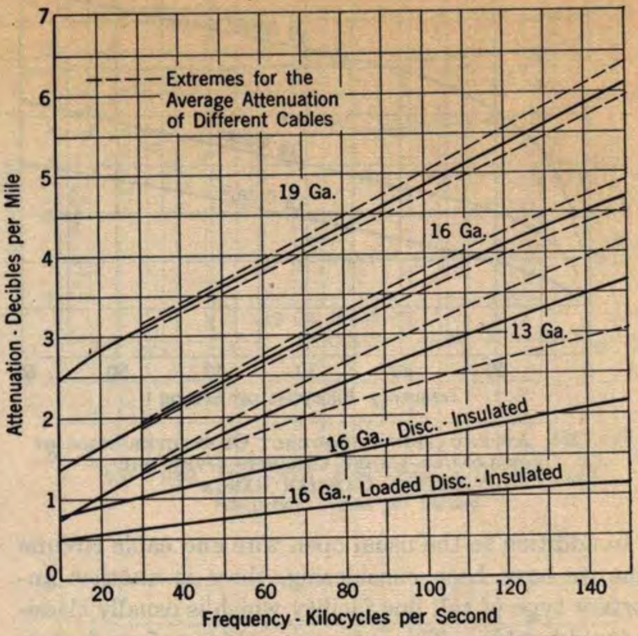
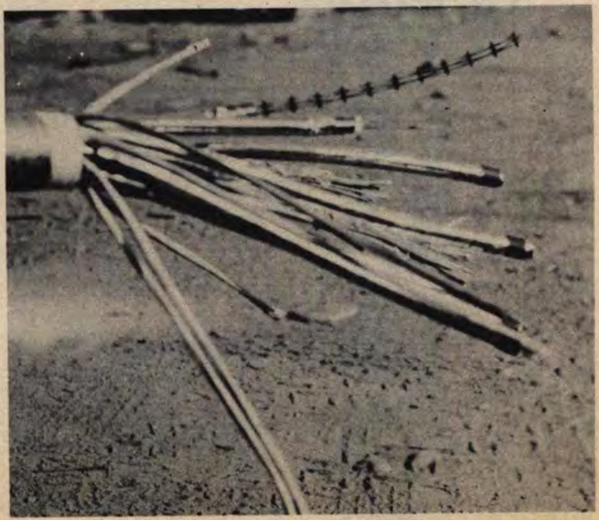


FIG. 285. ATTENUATION-FREQUENCY CHARACTERISTICS OF DISC-INSULATED TOLL ENTRANCE CABLE CIRCUITS COMPARED WITH NON-LOADED PAPER INSULATED CIRCUITS

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

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

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



SPIRAL-FOUR, DISC-INSULATED TOLL ENTRANCE CABLE FOR TYPE-J CARRIER SYSTEMS ever, in new installations A-2.7 and A-3.0 loading is usually provided for cable facilities connected to 8 inch spaced non-phantomed open wire conductors. For very short lengths of toll entrance cable and for long lengths of office cable used with Type-C carrier systems, X-2.5 loading is frequently employed.

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

CHAPTER XXIII

ATTENUATING, EQUALIZING AND TIME-DELAY CORRECTING NETWORKS

142. Pads or Attenuators

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

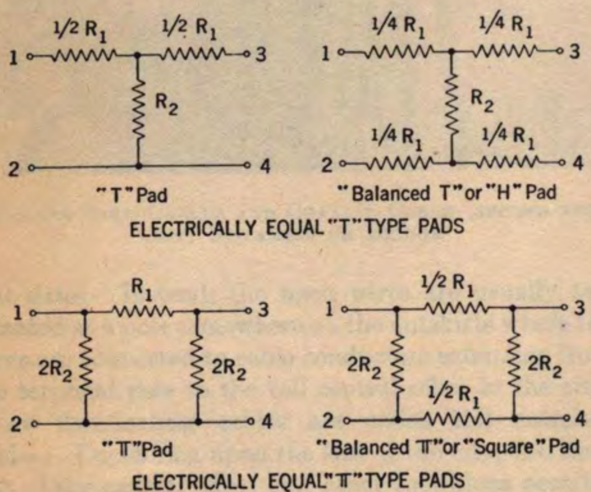


FIGURE 286

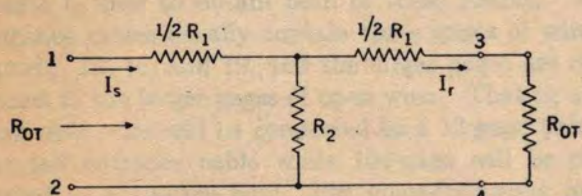
resistances. By arranging appropriate resistances in a network of series and shunt paths, any specified value of attenuation may be obtained without introducing any impedance irregularities in the circuit in which the network is connected. Such resistance networks are usually called **pads** and the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

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

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

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

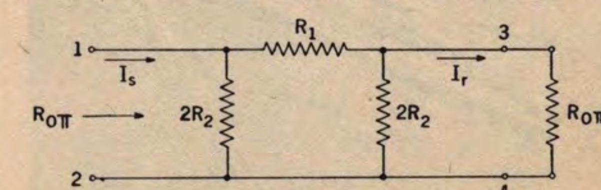
When a symmetrical pad is terminated in its characteristic resistance, the resistance it presents at the other terminal is still, of course, its characteristic resistance. In view of this it is a simple matter to solve for the characteristic resistance of the pad in terms of its series and shunt resistances.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



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

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

FIGURE 287



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

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

FIGURE 288

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

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

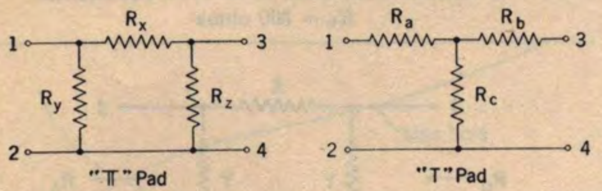


FIGURE 289

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

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

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

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

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

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

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

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

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

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

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

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

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

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

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

This becomes—

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

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

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

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

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

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

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

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

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

or

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

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

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

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

Also from Figure 287

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

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

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

Substituting R_2 in the first equation above—

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

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

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

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

From this

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

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

and

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

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

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

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

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

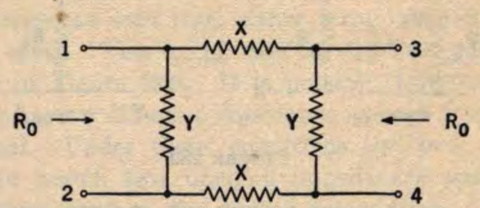
143. Attenuation Equalizers

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

TABLE XV

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

$R_0 = 300$ ohms



Resistance Values for Square Pad

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

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

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

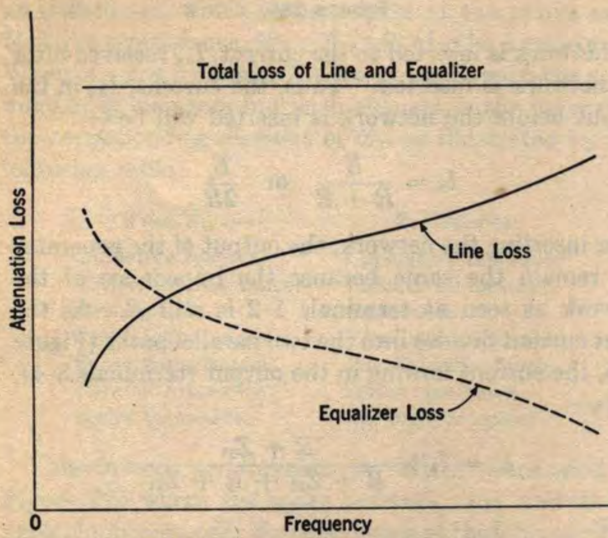


FIG. 290. PRINCIPLE OF ATTENUATION EQUALIZER

In long circuits equipped with telephone repeaters, the desired equalizing effects can be obtained without introducing an appreciable impedance irregularity by inserting equalizing networks at the mid-point of the primary sides of the repeater input transformers. Such applications are illustrated in Figures 336 and 337. Instead of changing the net loss of the line, however, this arrangement changes the overall gain-frequency characteristic of the repeater to match reasonably closely the loss-frequency characteristic of the line. That is, for the frequencies where the line loss is high the repeater gain is also high and vice versa. The overall loss-frequency characteristic of the line and repeater together is then reasonably uniform over the transmitted frequency band.

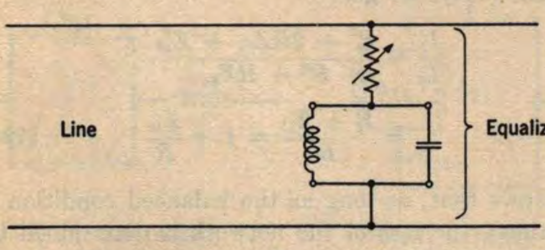


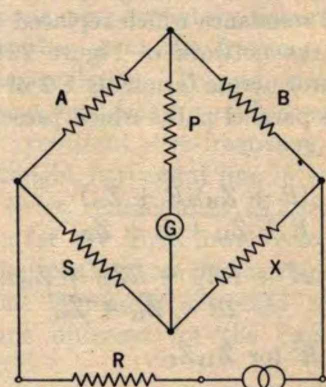
FIG. 291. SIMPLE BRIDGED EQUALIZER

144. Bridged T-Equalizer

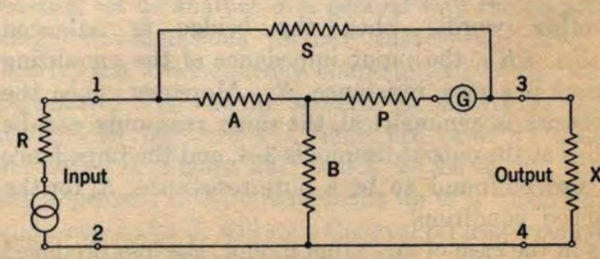
Both of the above methods of equalization give satisfactory results where the amount of attenuation distortion to be corrected is relatively small. To use either of these methods to correct a large attenuation distortion, might result in an impedance irregularity of such a magnitude as to more than offset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

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

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



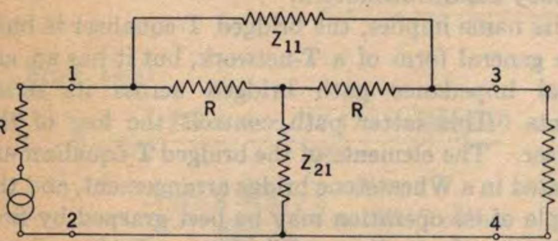
A. WHEATSTONE BRIDGE CIRCUIT



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

FIGURE 292

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



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

Fig. 293. Bridged T-Equalizer

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

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

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

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

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

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

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

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

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

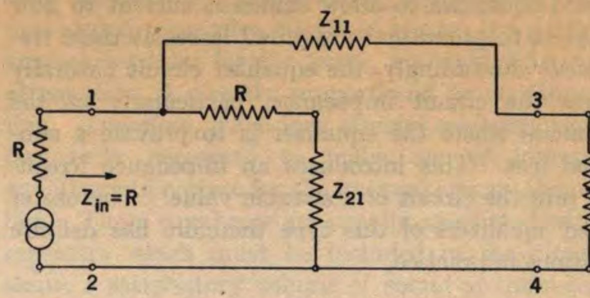


Figure 294

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

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

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

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

Then

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

or

$$\frac{I_b}{I_a} = \frac{2R + Z_{11} + Z_{21}}{R + Z_{21}} \qquad (124)$$

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

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

$$= \frac{R + Z_{11}}{R} = 1 + \frac{Z_{11}}{R} \qquad (125)$$

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

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

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

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

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

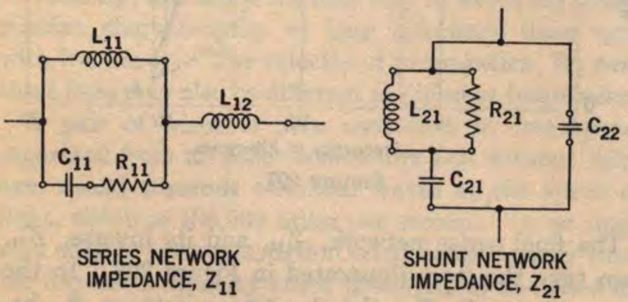


FIG. 295. INVERSE SERIES AND SHUNT NETWORKS

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

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

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

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

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

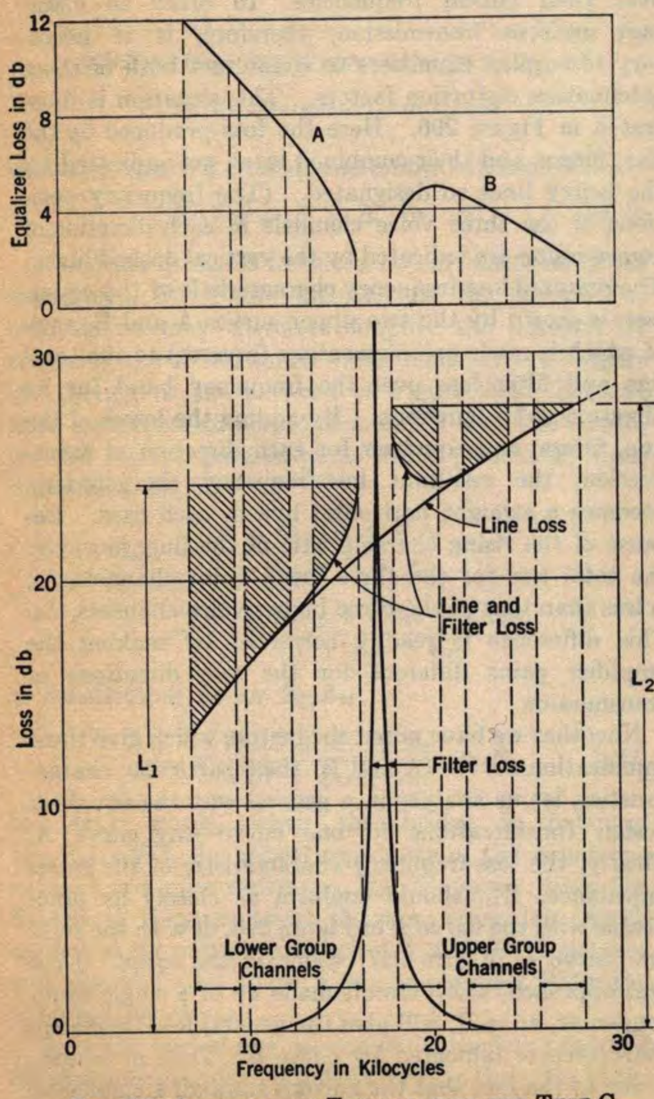


Fig. 296. Attenuation Equalization for Type-C Carrier System

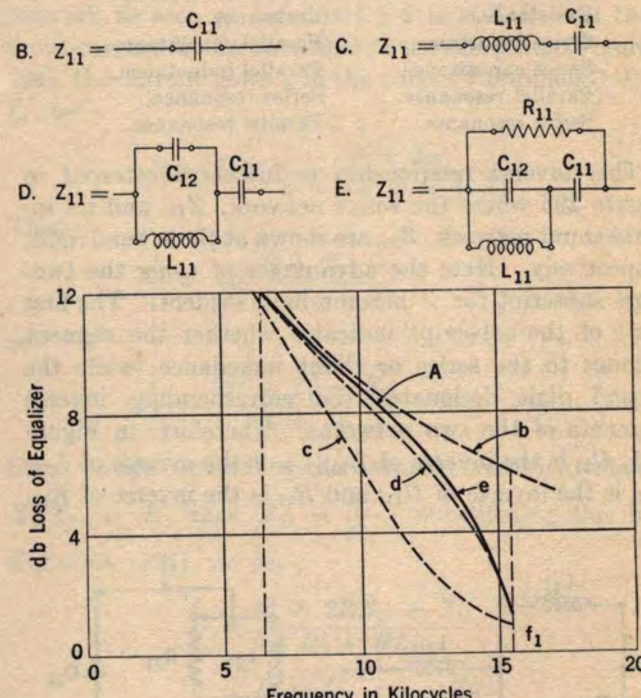


Figure 297

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

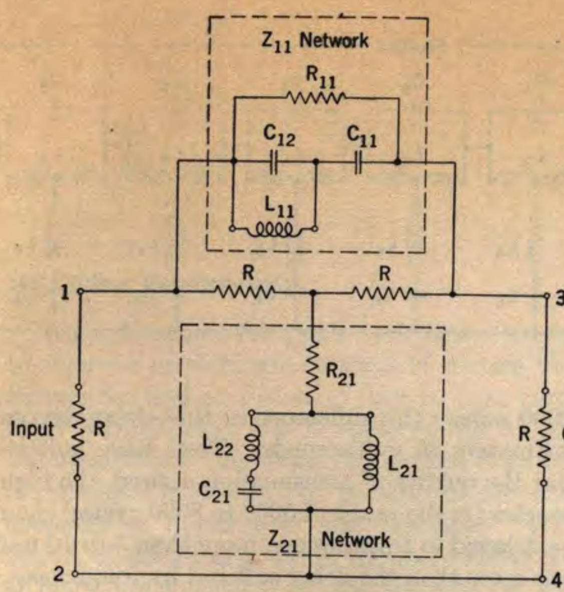


FIGURE 298

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

145. Time-Delay or Phase Equalizers

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

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

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

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

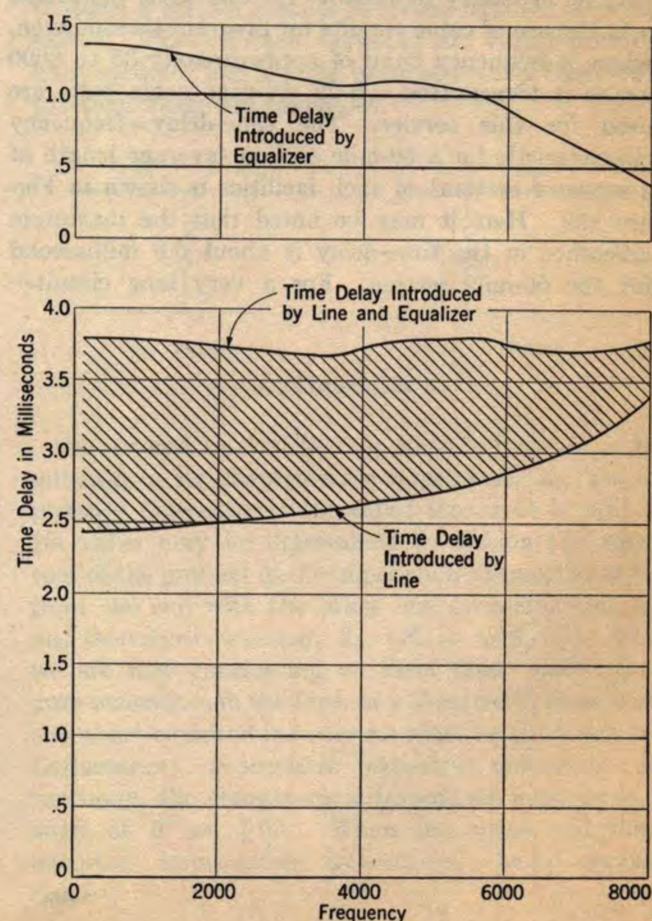


FIG. 299. PRINCIPLE OF TIME-DELAY OR PHASE EQUALIZING NETWORK

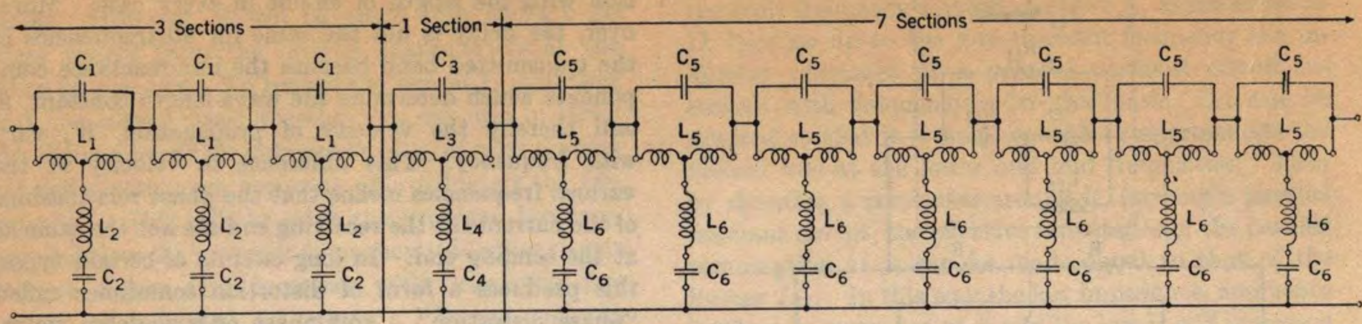


FIGURE 300

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

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

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

CHAPTER XXIV

FILTERS

146. Filter Requirements

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

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

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

147. Low-and High-Pass Filter Sections

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

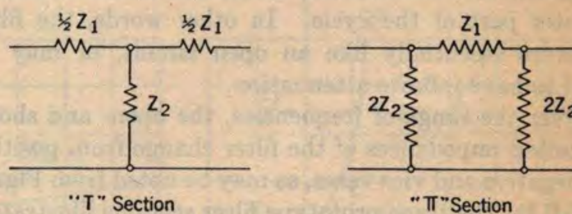


FIGURE 301

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

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

or

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

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

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

or

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

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

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

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

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

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

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

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

and

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

which, simplified, is—

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

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

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

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

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

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

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

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

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

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

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

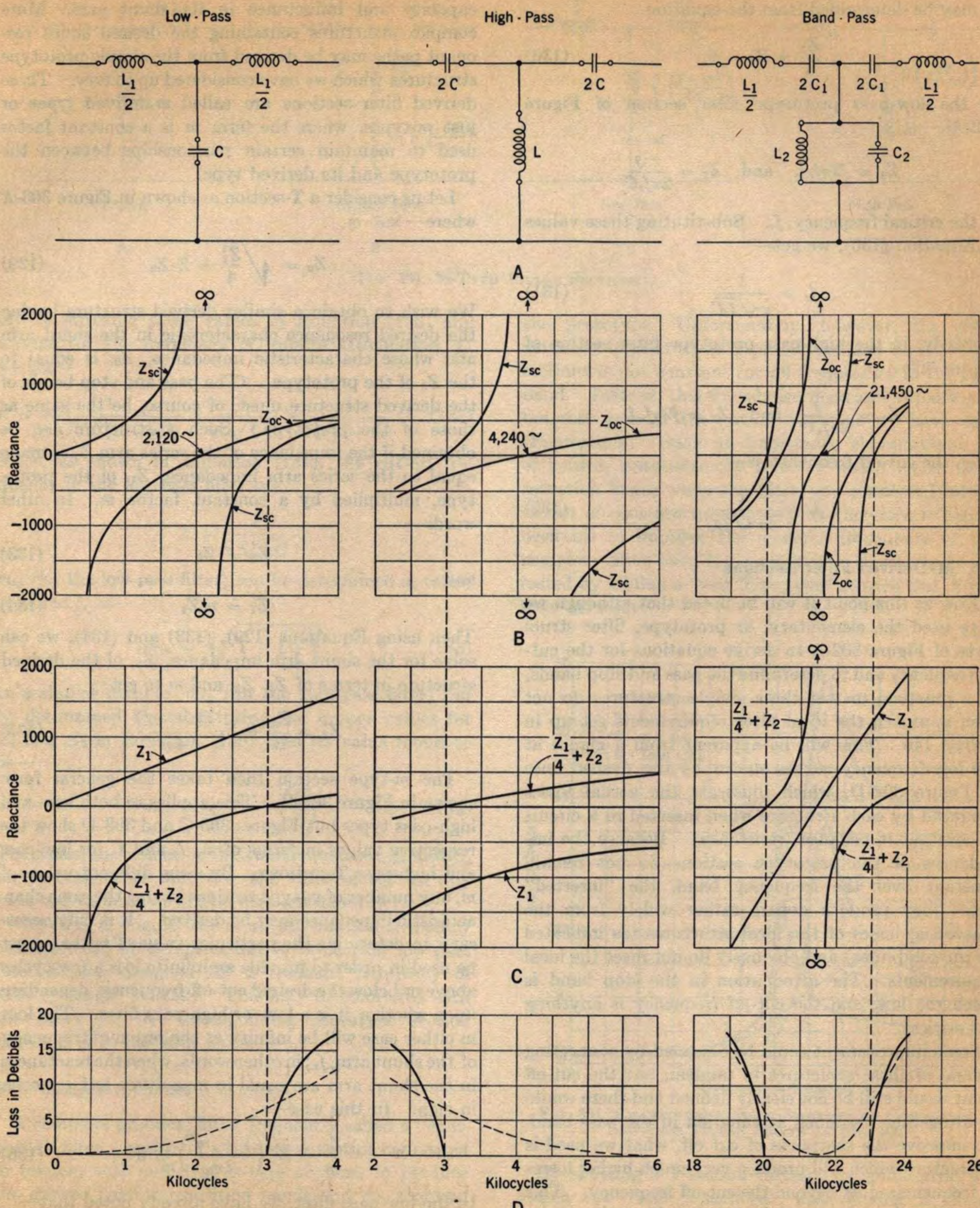


FIG. 302. REACTANCE AND LOSS CHARACTERISTICS OF ELEMENTARY PROTOTYPE FILTER SECTIONS

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

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

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

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

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

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

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

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

Then the cut-off frequency is—

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

148. M-Derived Filter Sections

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

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

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

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

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

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

and

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

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

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

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

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

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

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

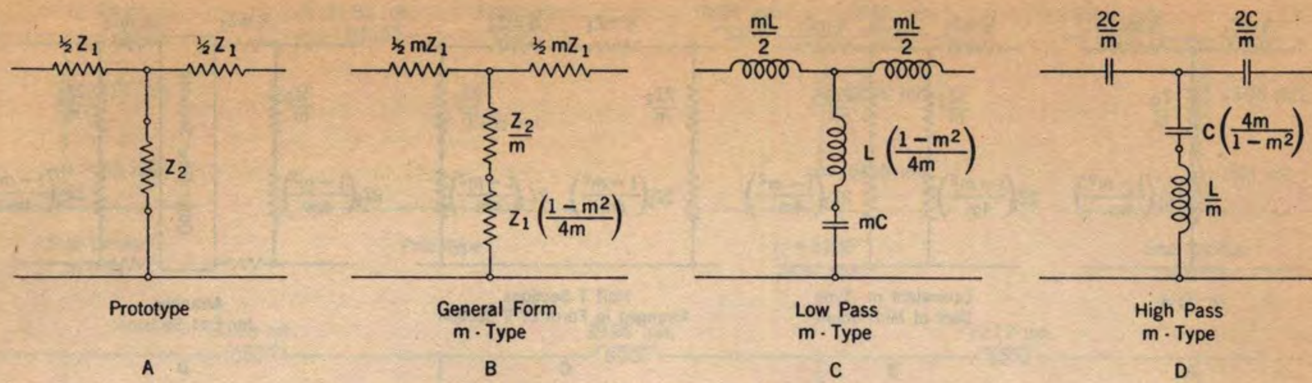


FIG. 303. M-TYPE FILTER SECTIONS

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

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

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

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

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

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

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

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

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

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

149. Composite Filters

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

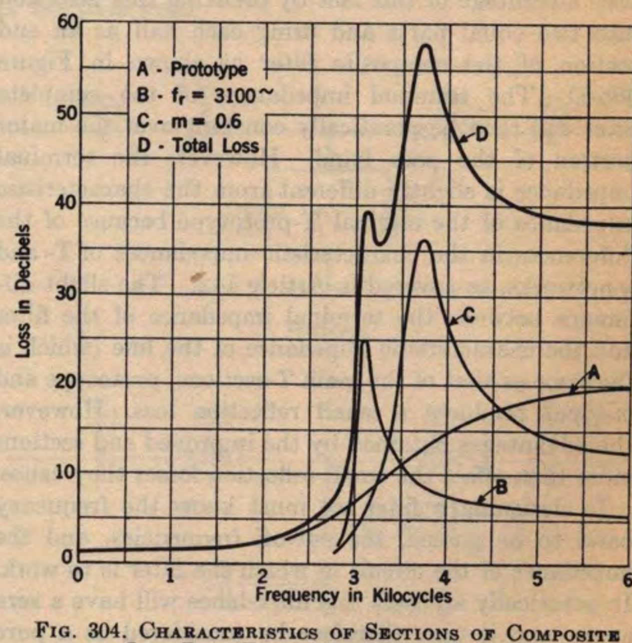


FIG. 304. CHARACTERISTICS OF SECTIONS OF COMPOSITE LOW-PASS FILTER SHOWN IN FIG. 306

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

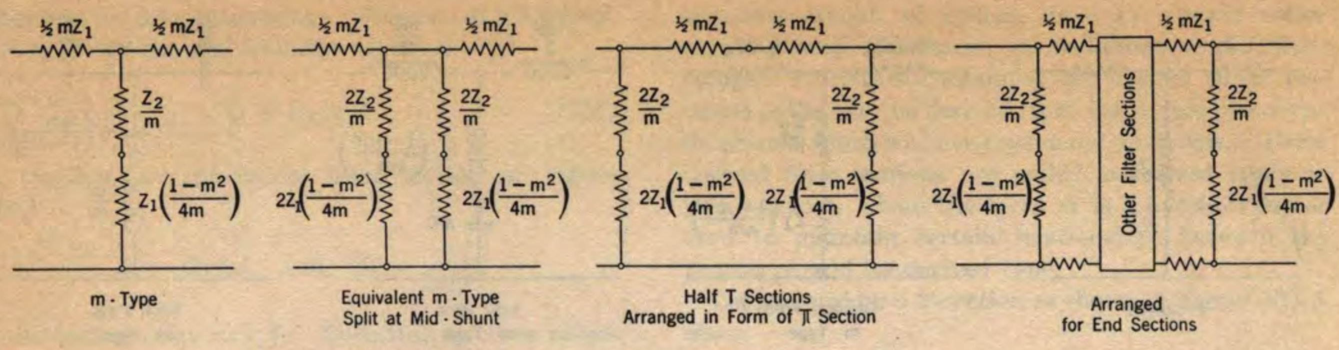


Fig. 305. Half T M-Type Filter End Sections

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

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

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

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

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

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

and

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

and for the high-pass filter—

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

and

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

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

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

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

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

$$C = \frac{1}{\pi f_c R} = \frac{1}{\pi 3000 \times 600}$$
$$= .1768 \times 10^{-6} \text{ farad or } .1768 \text{ mf.}$$

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

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

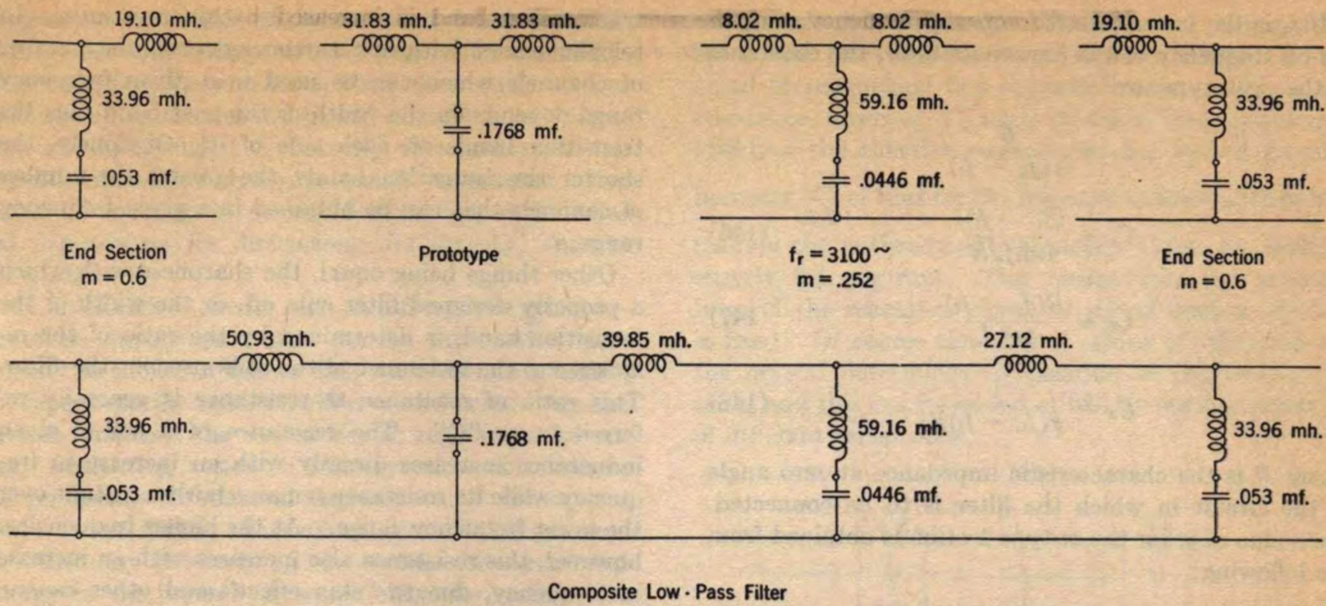


FIG. 306. MAKE-UP OF TYPICAL COMPOSITE LOW-PASS FILTER

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

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

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

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

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

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

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

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

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

150. Band-Pass Filters

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

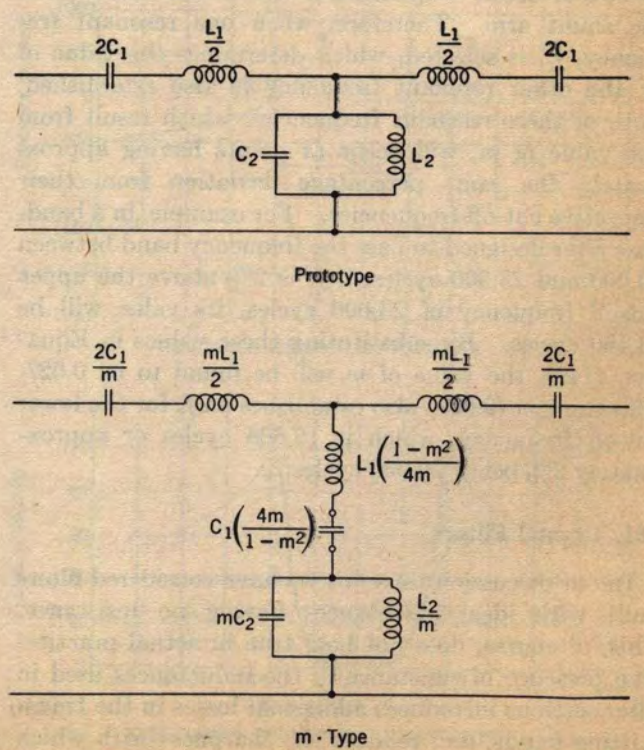


FIG. 307. BAND-PASS FILTER SECTIONS

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

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

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

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

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

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

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

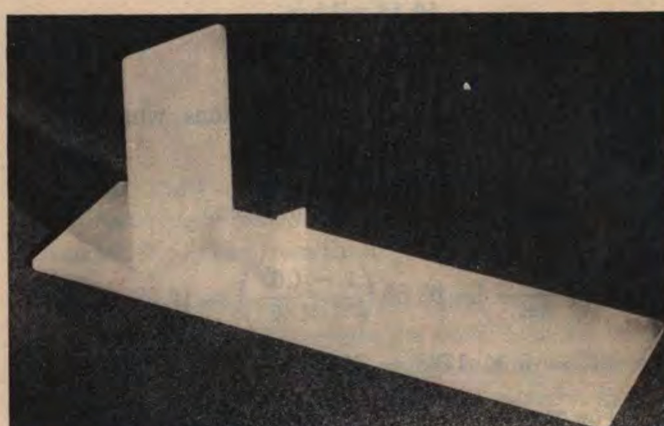
where f_r is the resonant frequency of infinite attenuation.

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

151. Crystal Filters

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

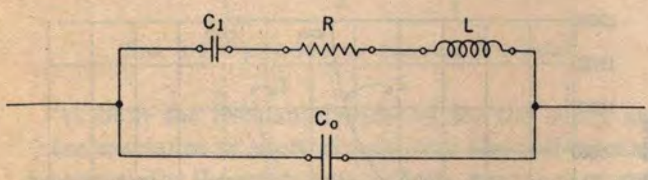
Other things being equal, the sharpness with which a properly designed filter cuts off, or the width of the transition band, is determined by the ratio of the reactance to the resistance of the coils used in the filter. This ratio of reactance to resistance is generally referred to as "Q". The reactance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



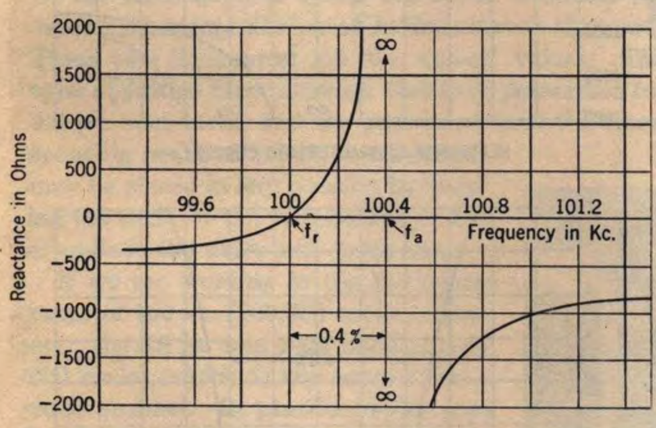
Quartz Crystals

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

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



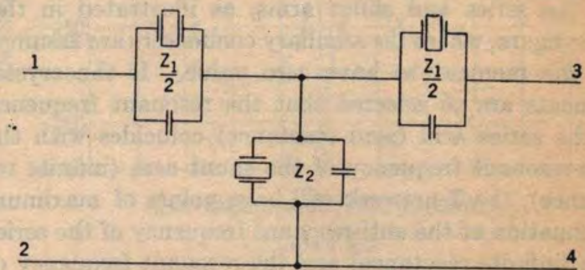
EQUIVALENT ELECTRICAL CIRCUIT OF A QUARTZ CRYSTAL



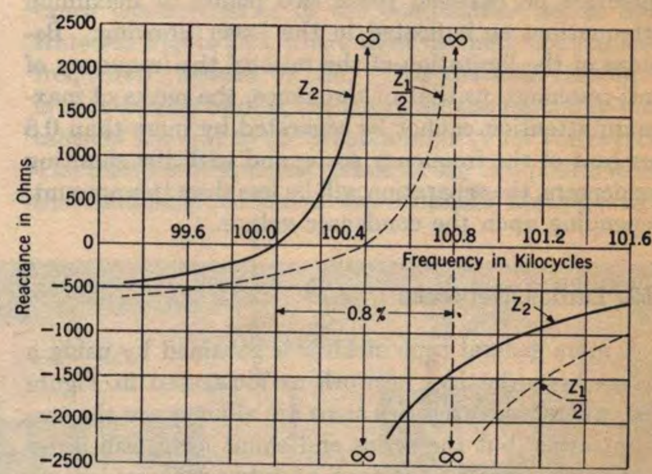
REACTANCE CHARACTERISTIC OF A QUARTZ CRYSTAL

FIGURE 308

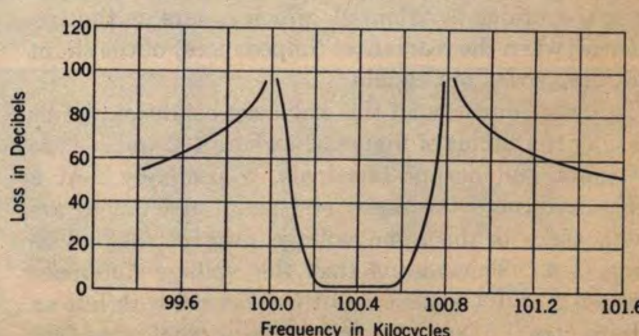
The ratio of C_0 to C_1 is a constant for any given crystal material. For quartz it is 125 to 1. The Q for quartz crystals is of the order of 20,000 or more. The general reactance characteristics of such a crystal are also shown in Figure 308 where the resonant frequency is f_r, and the anti-resonant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



"T" FILTER SECTION USING CRYSTALS



REACTANCE CHARACTERISTIC CURVES



ATTENUATION · FREQUENCY CHARACTERISTIC

FIGURE 309

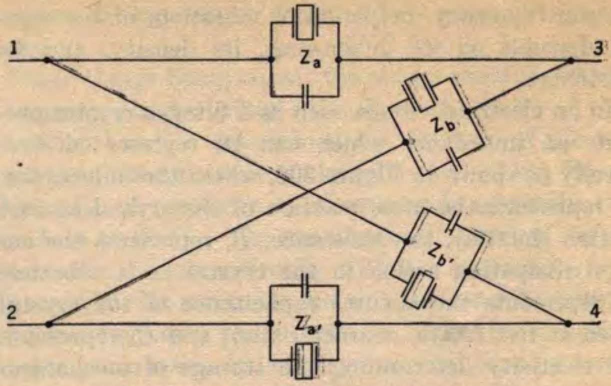
If an auxiliary condenser is placed in parallel with the crystal, however, the effective capacity of C_0 is increased, and as a result the anti-resonant frequency can be made less than 0.4 per cent above the resonant frequency. This can be done without detracting from the favorable characteristics of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

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

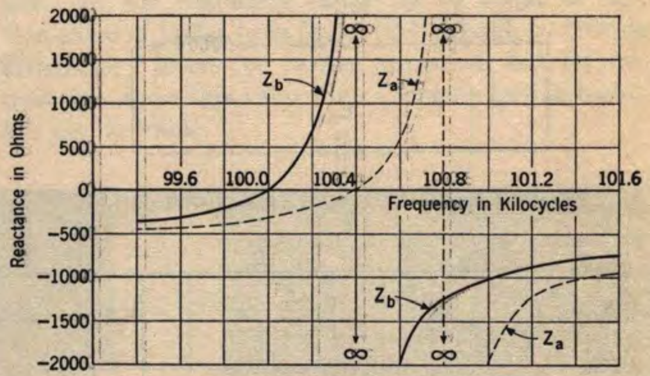
152. Lattice Networks

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

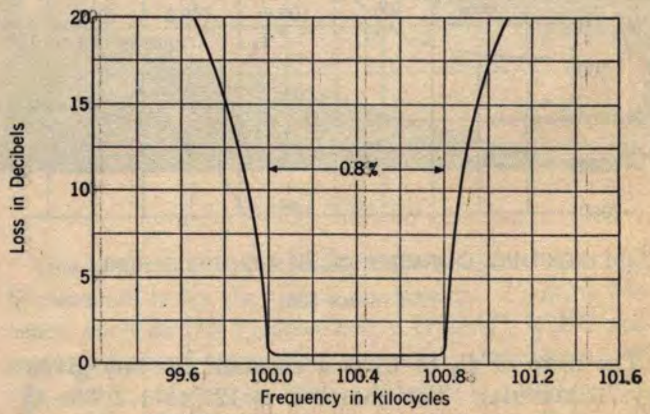
To better understand this balanced condition, let us consider the circuit of Figure 310 where 1–2 and 3–4 are the input and output terminals, respectively. At a frequency where the series and shunt reactances are equal, there is the same voltage drop across 1–3 as across 1–4. This means that the voltage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



LATTICE FILTER SECTION USING CRYSTALS



REACTANCE CHARACTERISTIC CURVES



ATTENUATION - FREQUENCY CHARACTERISTIC

FIGURE 310

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

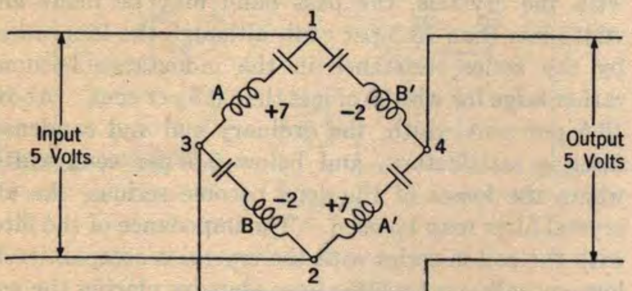
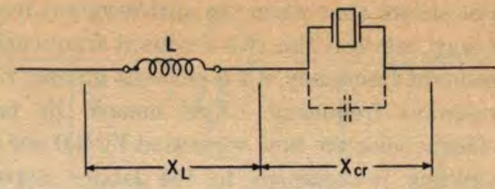


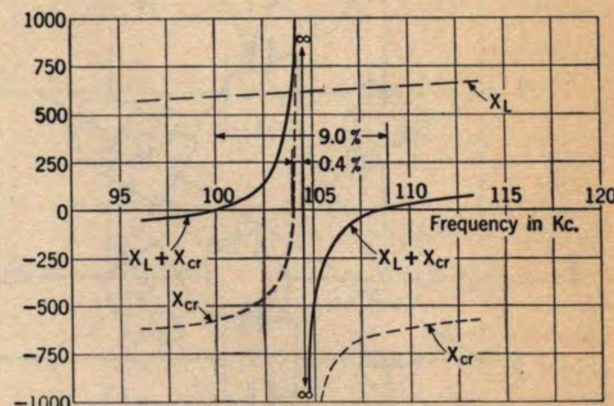
FIG. 311. PRINCIPLE OF LATTICE NETWORK

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

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



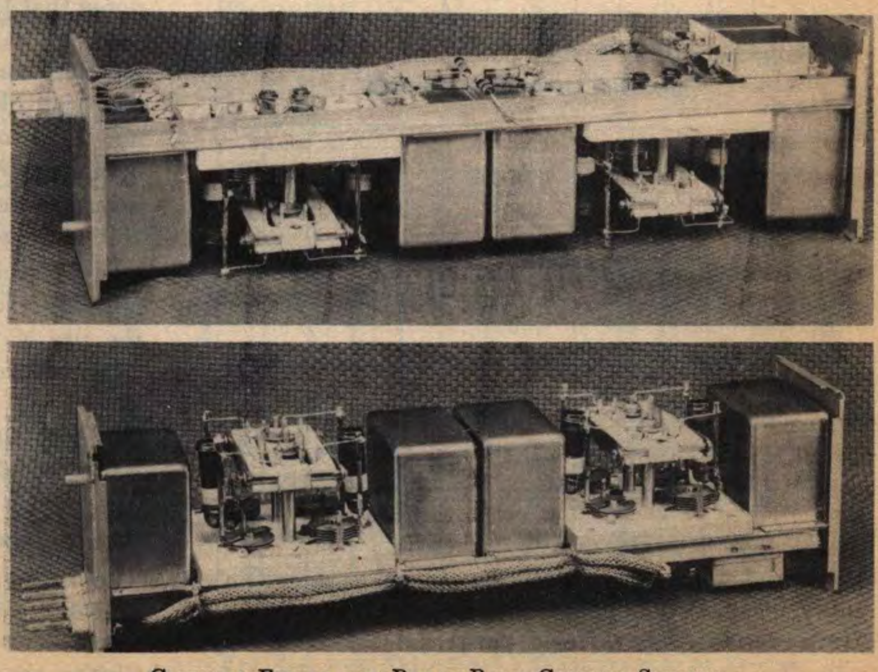
INDUCTANCE IN SERIES WITH CRYSTAL



REACTANCE CHARACTERISTIC CURVES

FIGURE 312

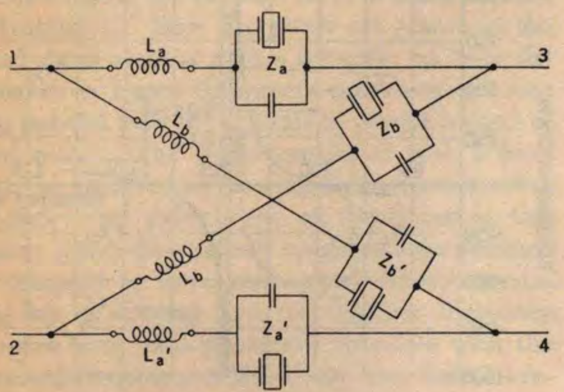
If a coil is placed in series with a crystal, the reactance characteristics of the combination are as illustrated in Figure 312, where there are now two resonant frequencies instead of one. This characteristic curve is obtained by adding the curve for the inductive reactance of the coil to the reactance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



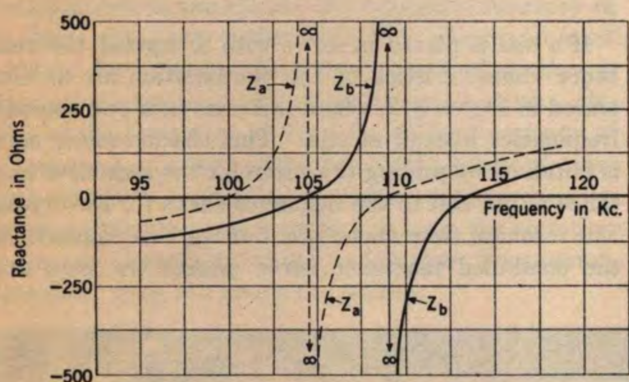
CHANNEL FILTER FOR BROAD-BAND CARRIER SYSTEMS

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

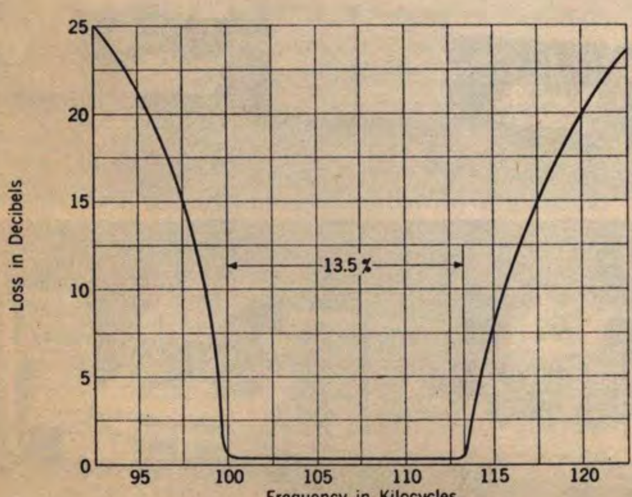
By adding inductances to the lattice network of Figure 310 we have the filter illustrated in Figure 313,



LATTICE FILTER SECTION WITH SERIES INDUCTANCE



REACTANCE CHARACTERISTIC CURVES



ATTENUATION - FREQUENCY CHARACTERISTIC

FIGURE 313

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

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

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

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

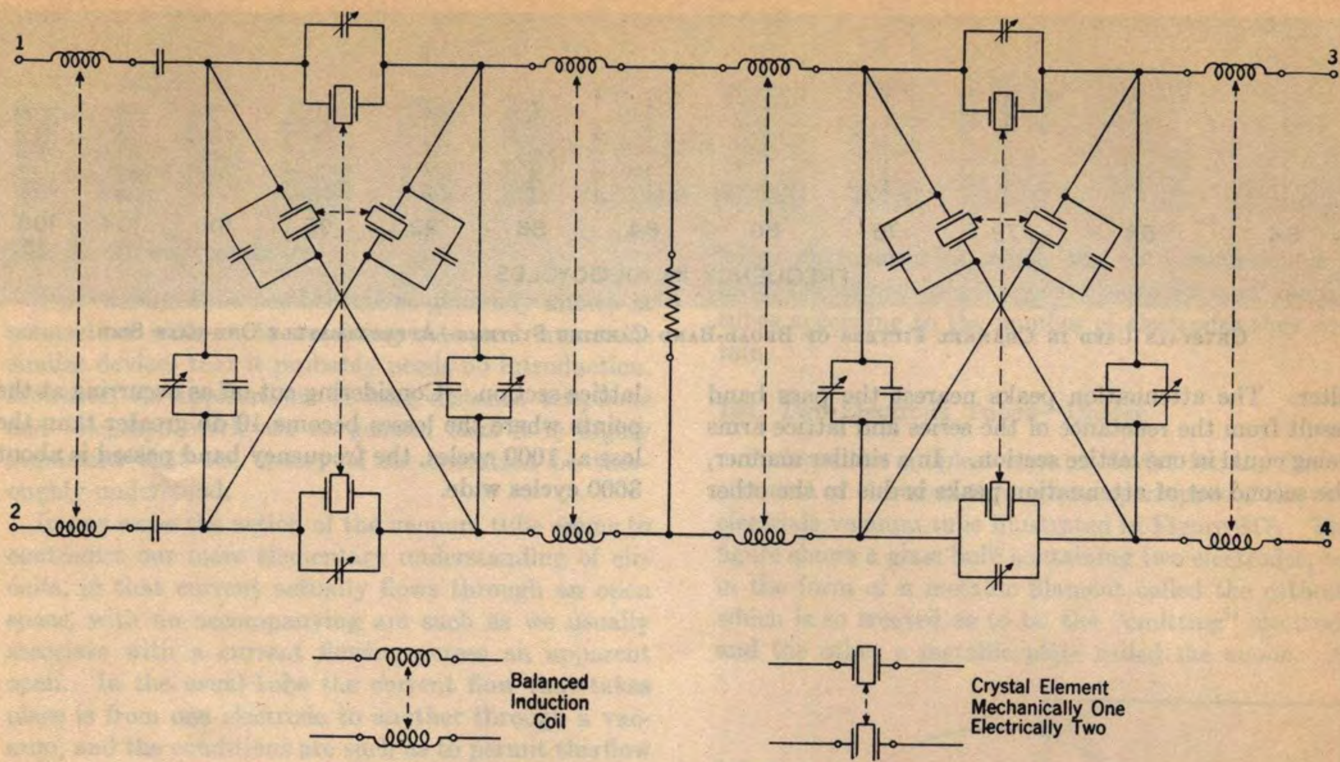


Fig. 314. Typical Band-Pass Crystal Filter

layer of aluminum which is divided along the center line lengthwise of the crystal to form two electrically independent crystal units from a single crystal. Since the crystals vibrate longitudinally with a node across the middle, they are clamped at this node in mounting as illustrated in Figure 315. The four crystals used in

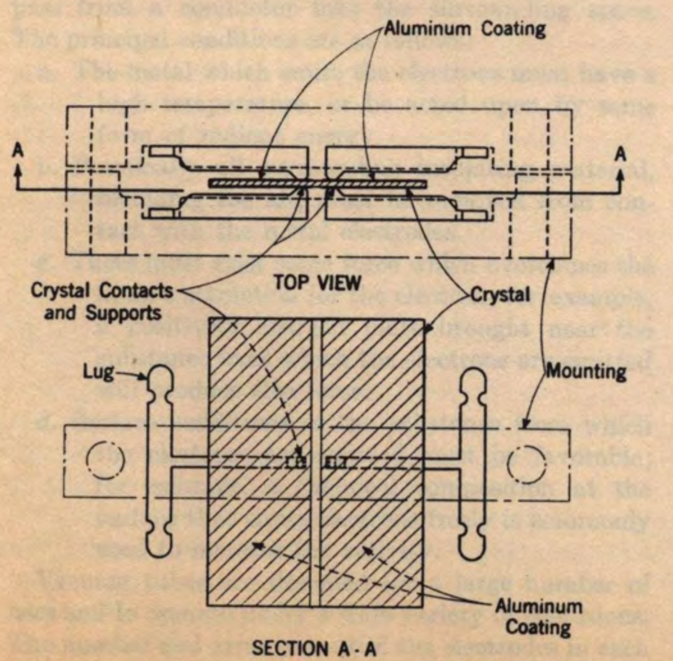


Fig. 315. Method of Mounting Quartz Crystal

the lowest frequency channel (64 kc.) of the Type-K carrier system range in length from about 1.65 to 1.58 inches, while those in the highest frequency channel (108 kc.) range from about 0.96 to 0.94 inch. These crystals vary in thickness from about 0.0248 inch to 0.0433 inch, while the widths are in the order of 50 to 80 per cent of their lengths.

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

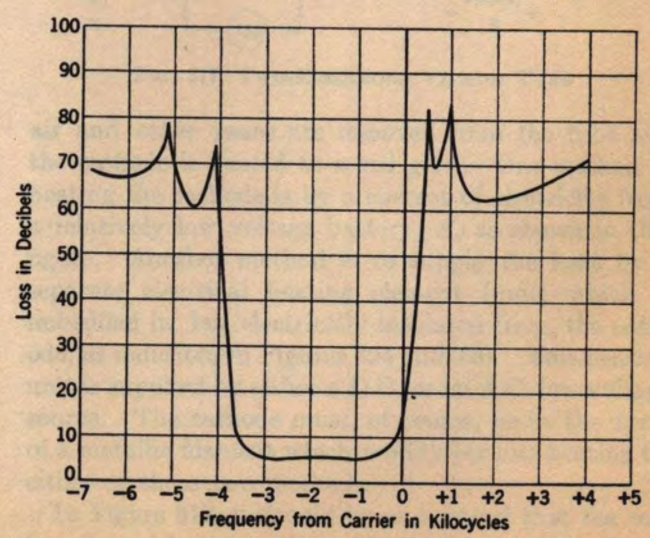
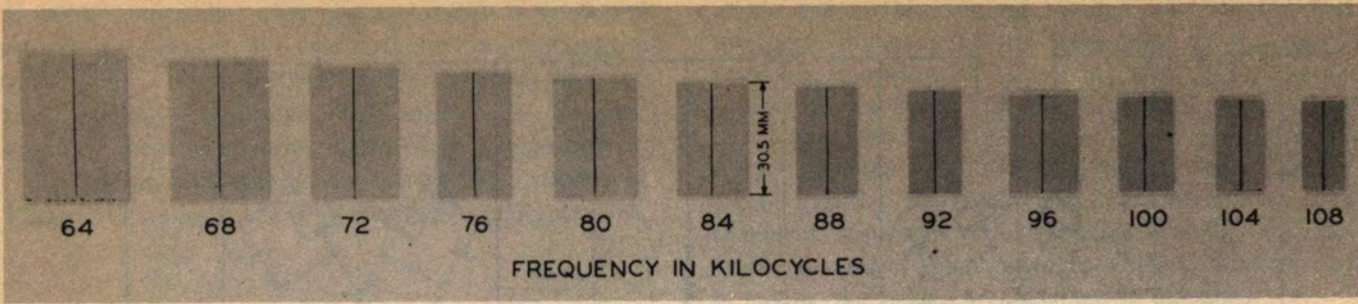


Fig. 316. Loss-frequency Characteristic of Filter Shown in Fig. 314



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

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

CHAPTER XXV

VACUUM TUBES

153. Thermionic Emission

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

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

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

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

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

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

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

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

154. Two-Electrode Tubes—Diodes

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

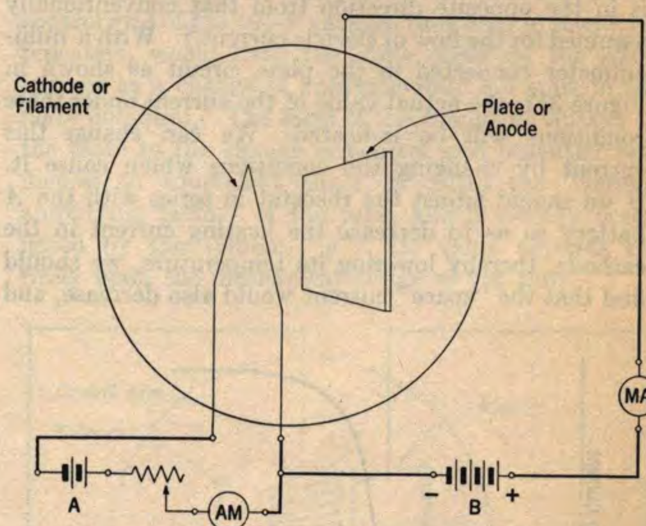


Fig. 317. Two-Electrode Vacuum Tube

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

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

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

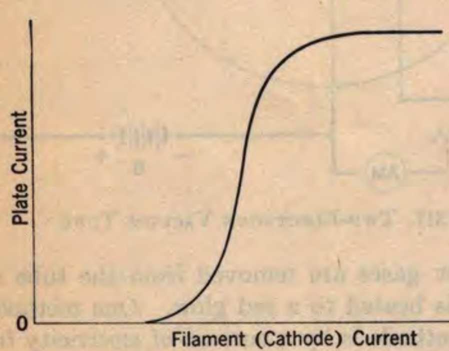


FIG. 318. CATHODE CURRENT VS. PLATE CURRENT CHARACTERISTIC

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

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

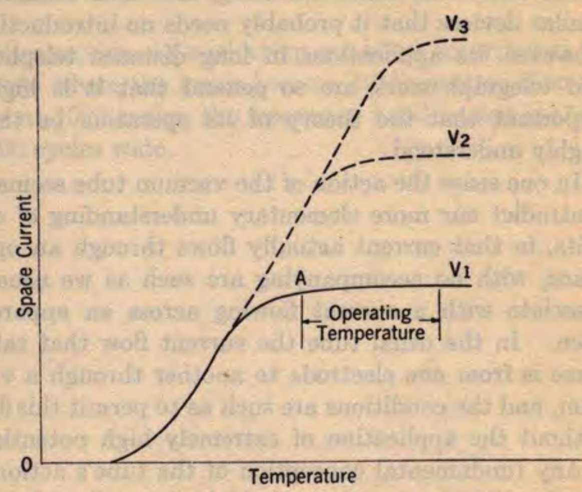


FIG. 319. CATHODE TEMPERATURE VS. PLATE CURRENT CHARACTERISTIC

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

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

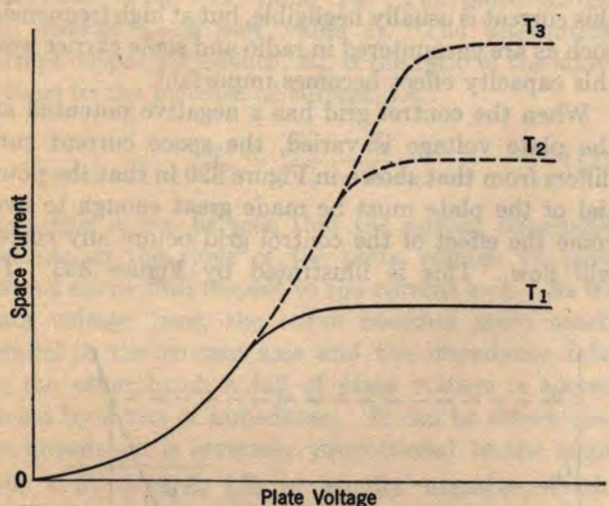


FIG. 320. PLATE VOLTAGE VS. PLATE CURRENT CHARACTERISTIC

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

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

155. Three-Electrode Tubes—Triodes

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

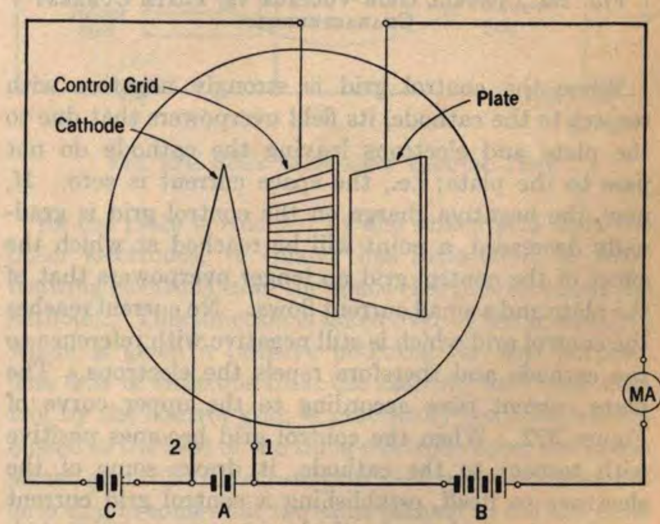


FIG. 321. THREE-ELECTRODE VACUUM TUBE

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

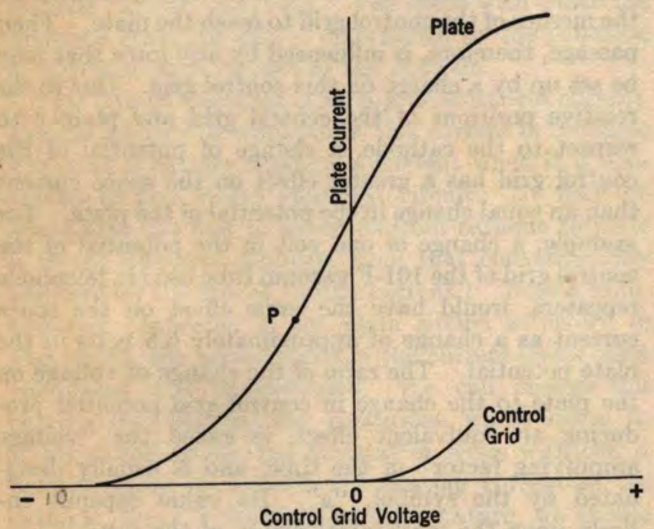


FIG. 322. CONTROL GRID VOLTAGE VS. PLATE CURRENT CHARACTERISTIC

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

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

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

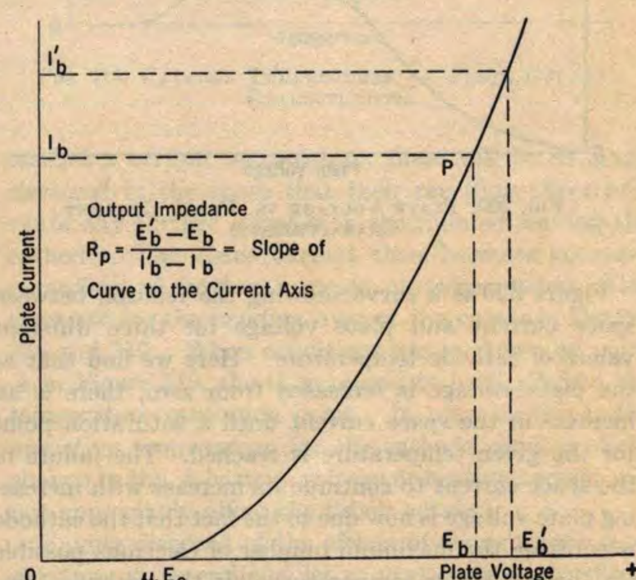


FIG. 323. PLATE VOLTAGE VS. PLATE CURRENT CHARACTERISTIC OF THREE-ELECTRODE TUBE

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

$$R_{d.c.} = \frac{E_b}{I_b} \tag{150}$$

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

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

$$R_p = \frac{E_b' - E_b}{I_b' - I_b} \tag{151}$$

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

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

156. Four-Electrode Tubes—Tetrodes

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

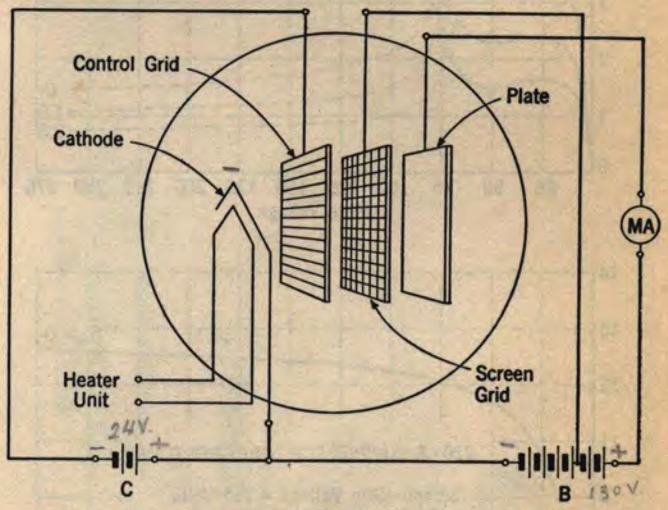


Fig. 324. Four-Electrode Vacuum Tube

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

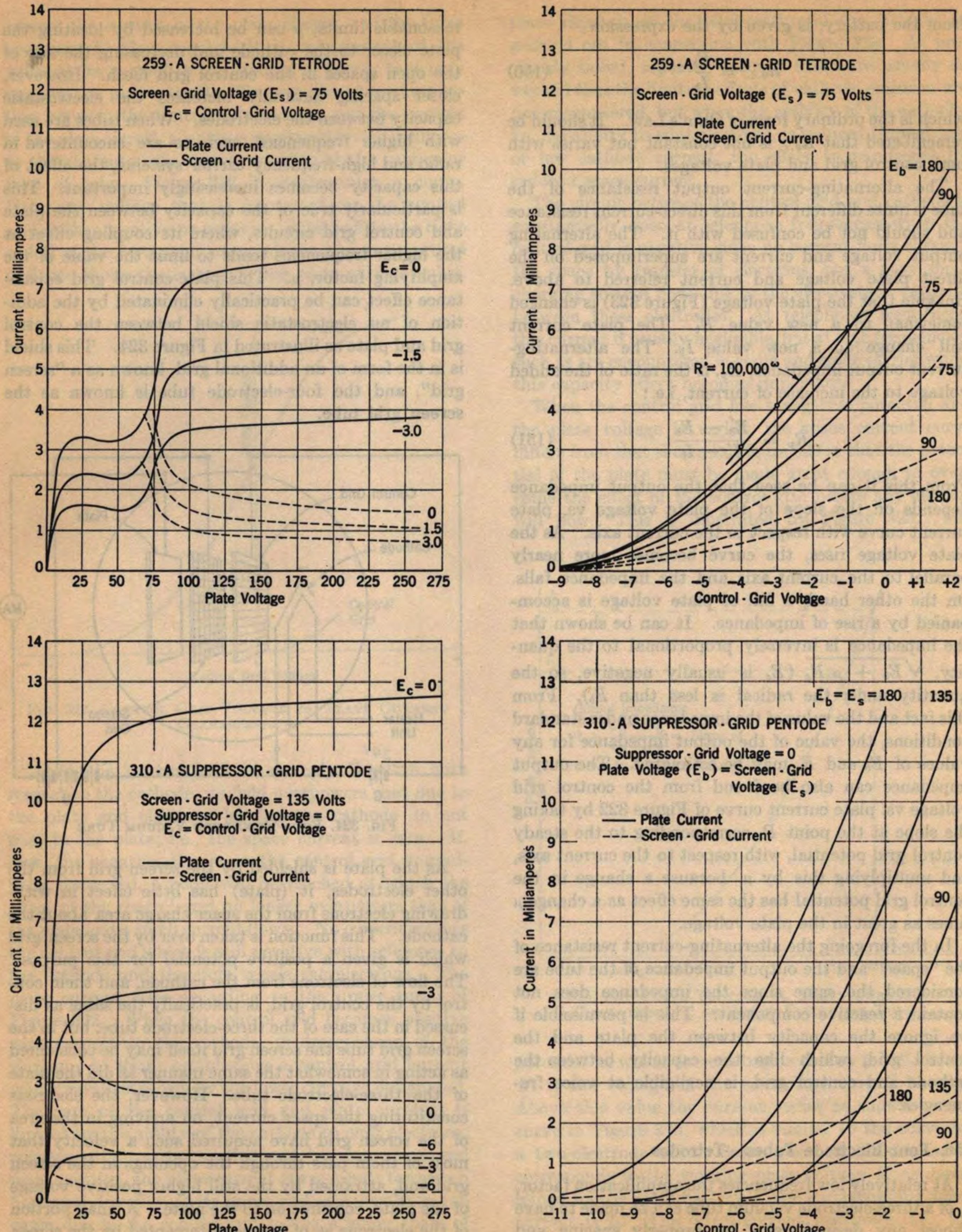


FIG. 325. CHARACTERISTICS OF STANDARD FOUR- AND FIVE-ELECTRODE VACUUM TUBES

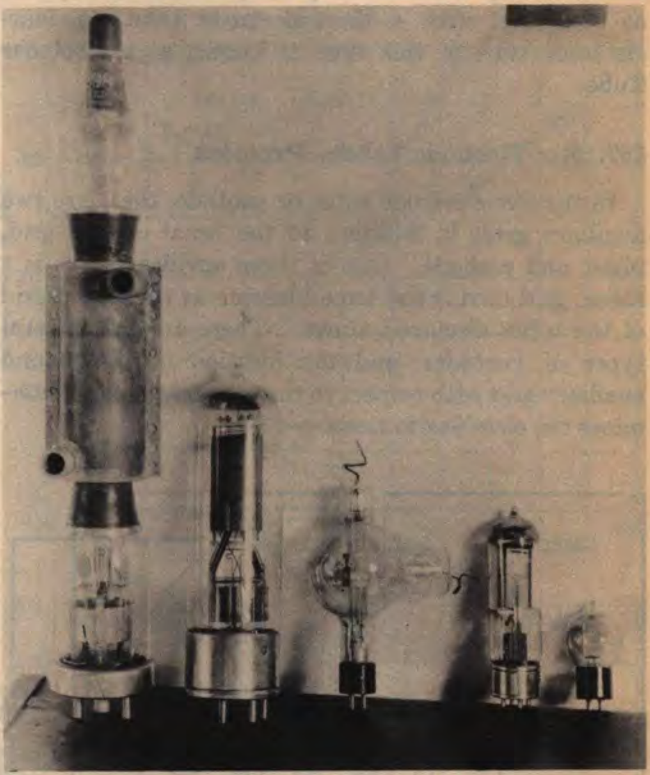
by the plate voltage vs. plate and screen grid current curves in Figure 325 for a representative Screen-Grid Tetrode.

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

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

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

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



Radio Transmitting Tubes

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

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

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

157. Five-Electrode Tubes—Pentodes

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

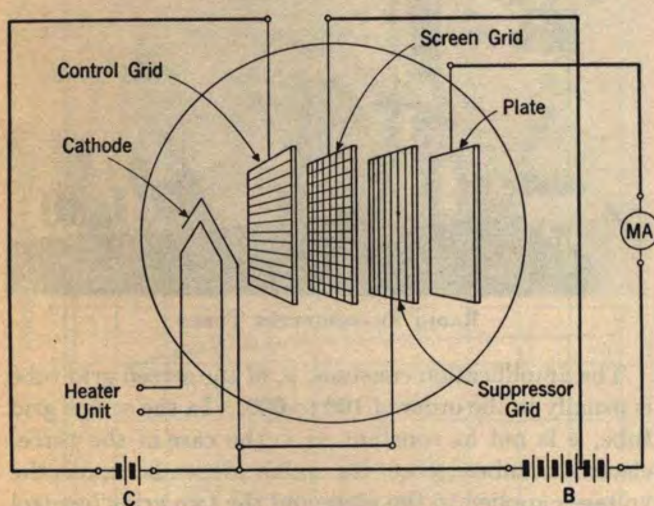
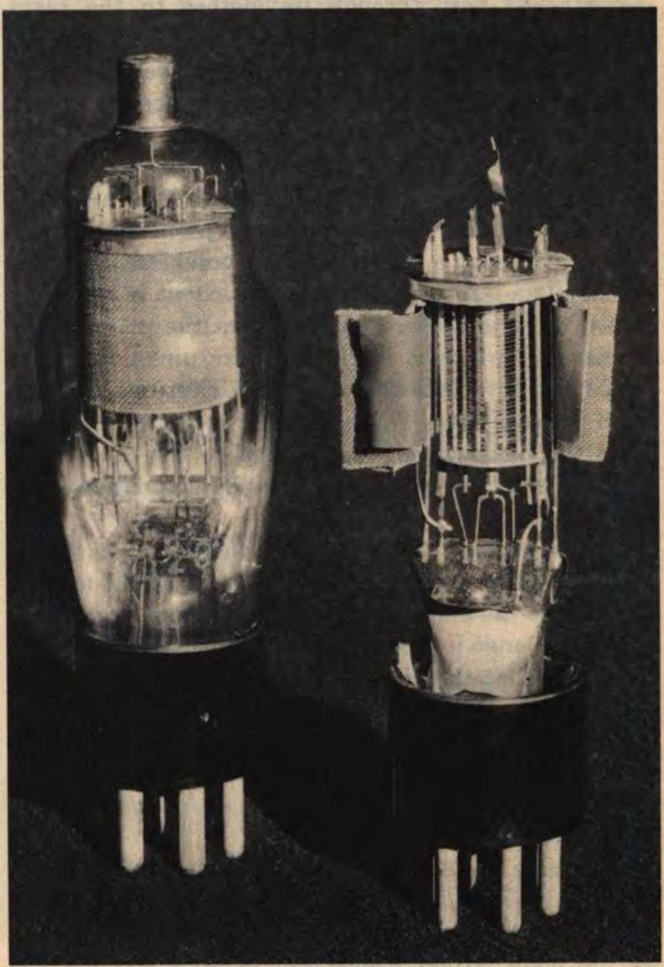


Fig. 326. Five-Electrode Vacuum Tube

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



310-A Pentode Vacuum Tube

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

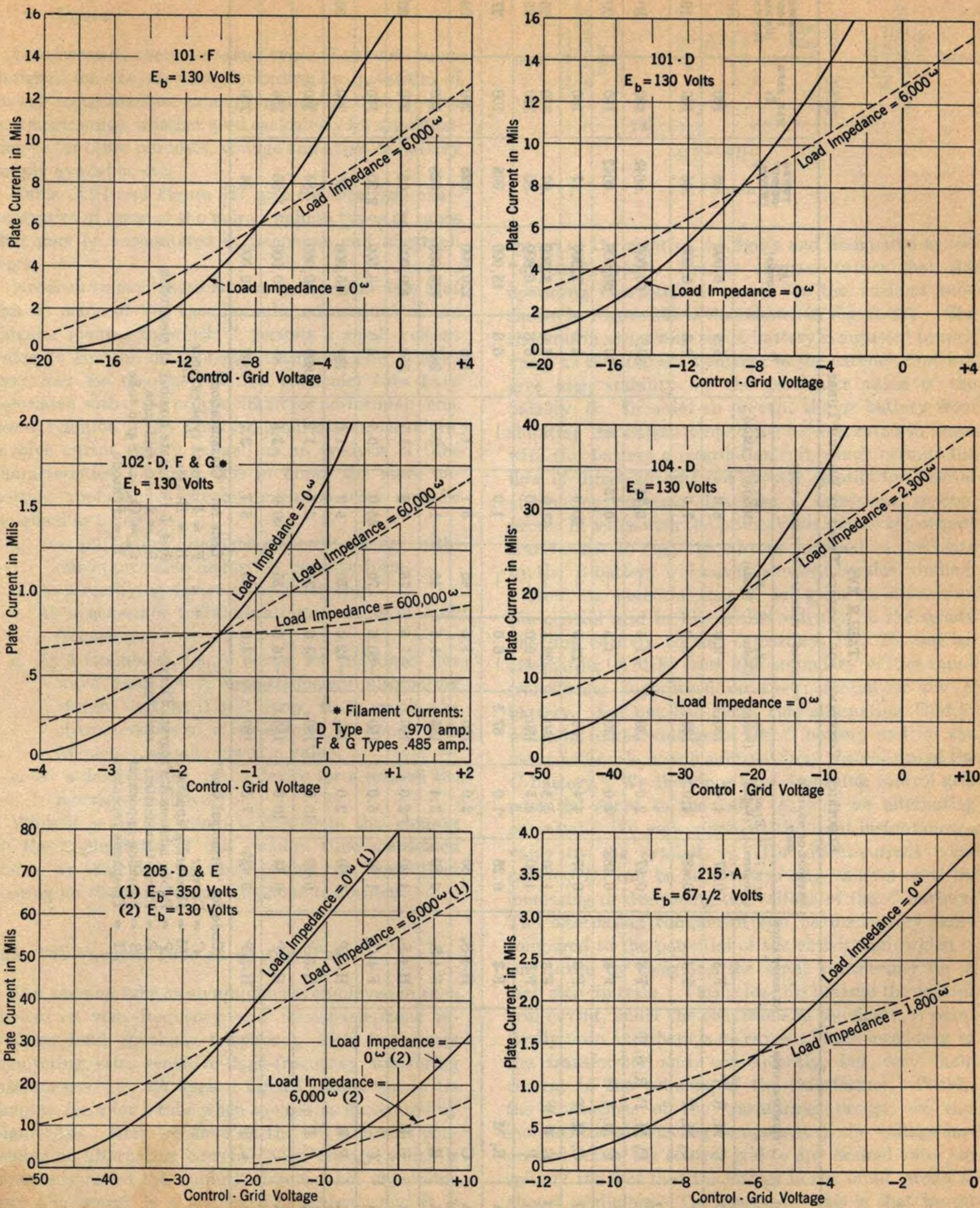


Fig. 327. Characteristics of Standard Three-Electrode Vacuum Tubes

TABLE XVI
Electrical Constants for Various Types of Vacuum Tubes

Code No.	Principal Uses	Classification	Cathode or Heating Unit		Voltages & Currents of Plate & Grids					μ	R_p (Ohms)	Power Output (Watts)	Max. Safe E_b	Microphonic Noise (DB below 10^{-6} watts)
			I_f or I_h (Amperes)	E_f or E_h (Volts)	E_b (Volts)	E_c (Volts)	E_s (Volts)	I_b (Mils)	I_s (Mils)					
101-D	C, M, R, T	F-3	0.97	4.4	130.	−9.0	—	8.0	—	5.9	6,000	.06	160.	
101-F	C, M, T	F-3	0.485	4.0	130.	−8.0	—	7.0	—	6.5	6,000	.06	160.	19
102-D or G	C, M, P, R, S, T	F-3	0.97	2.0	130.	−1.5	—	0.75	—	30.	60,000	.0042	160.	30
102-F	C, M, T, S	F-3	0.485	2.0	130.	−1.5	—	0.75	—	30.	60,000	.0042	160.	20
104-D	C, M, T	F-3	0.97	4.4	130.	−22.5	—	20.0	—	2.5	2,300	.17	160.	
205-D	C, P, R	F-3	1.60	4.5	350.	−22.5	—	33.0	—	7.0	3,500	.07	350.	25
205-E	P	F-3	1.60	4.5	130.	−9.0	—	5.0	—	6.8	6,600	.89	350.	25
215-A	R, M	F-3	0.25	1.0	67.5	−6.0	—	1.0	—	6.0	18,000	.008	100.	27
245-A	C	H-4-S_c	1.60	2.0	180.	−1.5	45.	5.1	1.4	170.	220,000	.025	200.	29
246-A	R	F-4-S_c	0.10	3.4	135.	−1.5	45.	1.5	0.2	285.	725,000	.028	180.	
259-A	C, R	H-4-S_c	1.60	2.0	180.	−1.5	75.	5.5	1.2	550.	400,000	.05	200.	20
281-A	C	F-4-C_o	1.60	5.0	150.	−60.0	70.	35.0	0.7	5.0	3,500	2.20	170.	
285-A	P, R	H-5-P	1.60	2.0	180.	−12.0	150.	9.0	1.6	130.	145,000	.65	200.	20
310-A	C, P	H-5-P	0.32	10.0	180.	−3.0	135.	5.4	1.2	1800.	40,000	.34	250.	
311-A	C	H-5-P	0.64	10.0	155.	−16.0	155.	40.0	3.5	85.	30,000	1.80	250.	
328-A	C, P	H-5-P	0.425	7.5	180.	−3.0	135.	5.4	2.2	1800.	40,000	.34	250.	

Code for "Principal Use"

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

Code for "Classification"

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

158. Characteristics of Various Standard Vacuum Tubes

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

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

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

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

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

159. The Vacuum Tube as an Amplifier

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

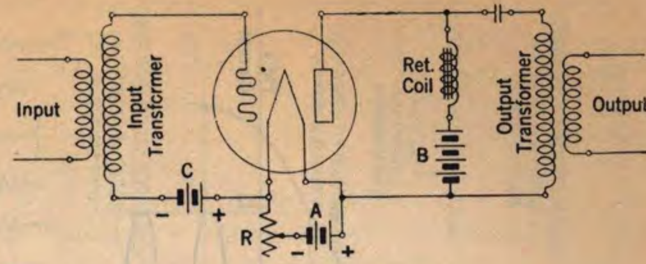


FIGURE 328

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

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

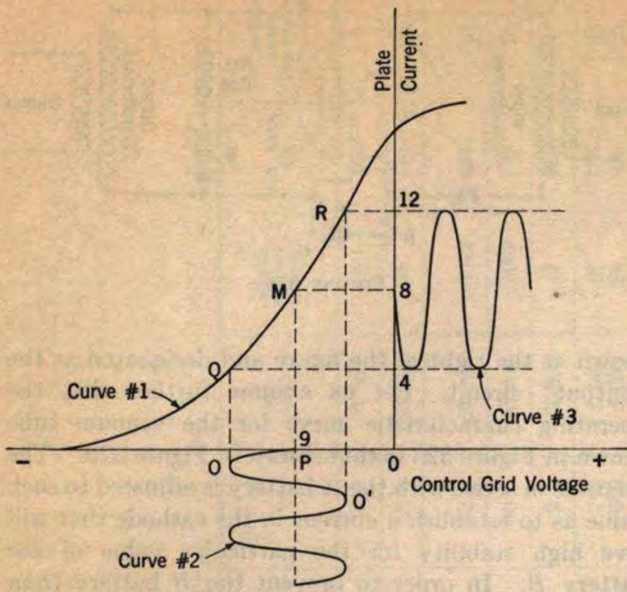


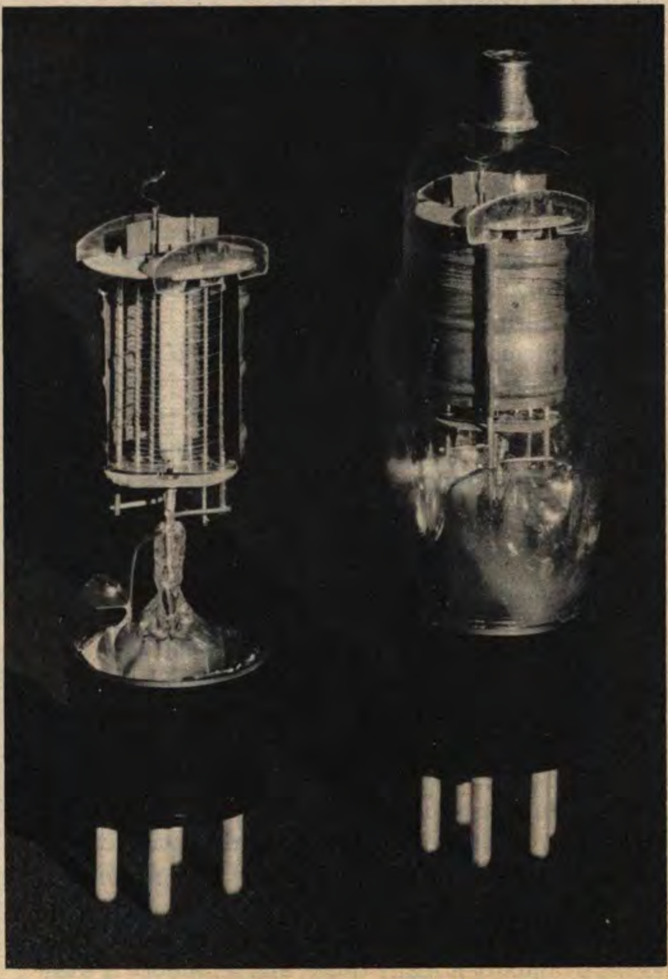
Fig. 329. Vacuum Tube as Amplifier

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

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

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

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



311-A Pentode Vacuum Tube

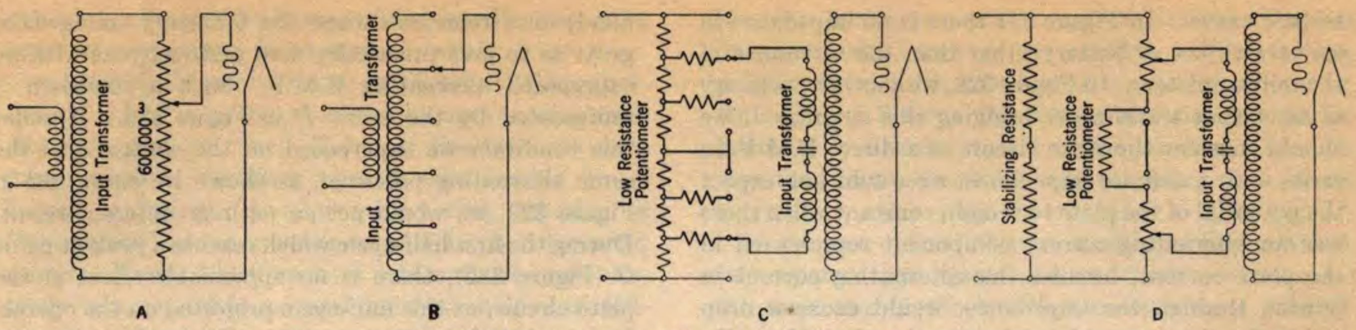


Fig. 330. Methods for Varying Amplifier Gain

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

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

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

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

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

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

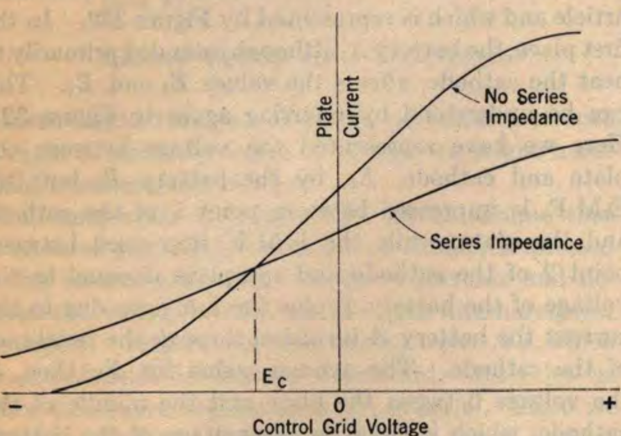


FIG. 331. EFFECT OF OUTPUT IMPEDANCE ON TUBE CHARACTERISTICS

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

160. The Vacuum Tube as a Rectifying Device

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

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

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

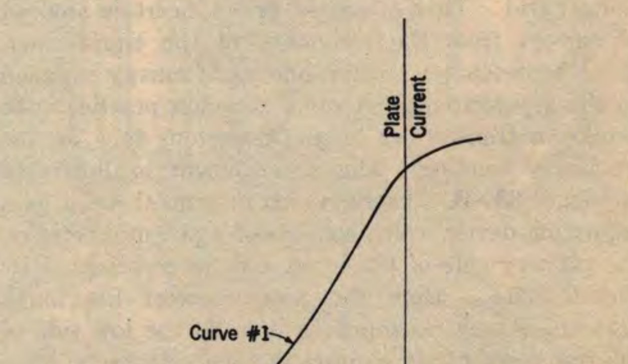
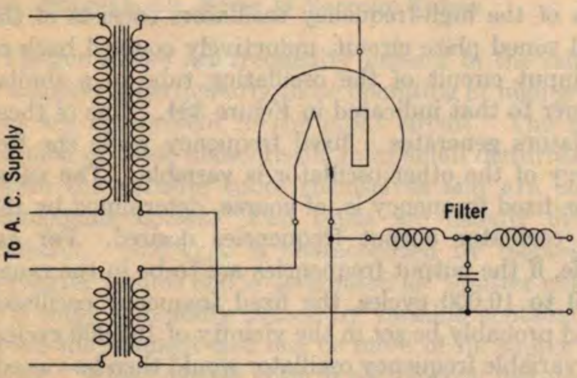


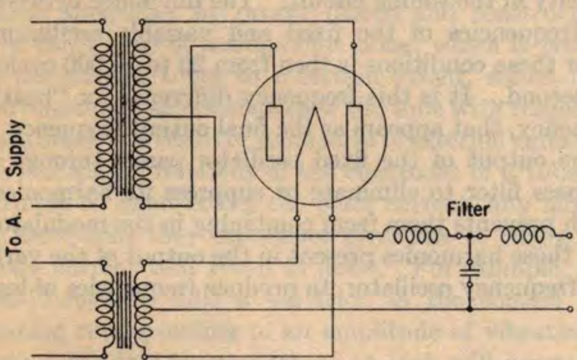
FIG. 332. VACUUM TUBE AS RECTIFIER

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

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



A · HALF WAVE RECTIFIER



B · FULL WAVE RECTIFIER

FIGURE 333

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

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

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

161. The Vacuum Tube as a Generator (Oscillator)

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

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

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

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

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

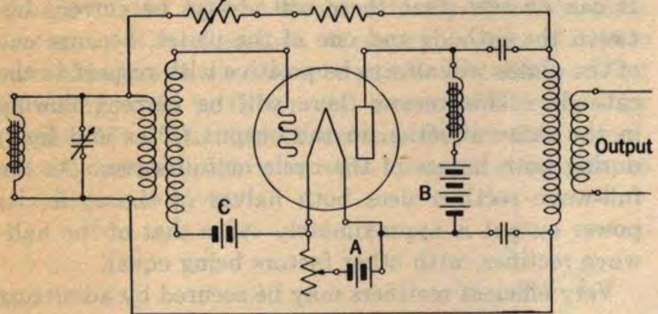


Fig. 334. Vacuum Tube Oscillator Circuit

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

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

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

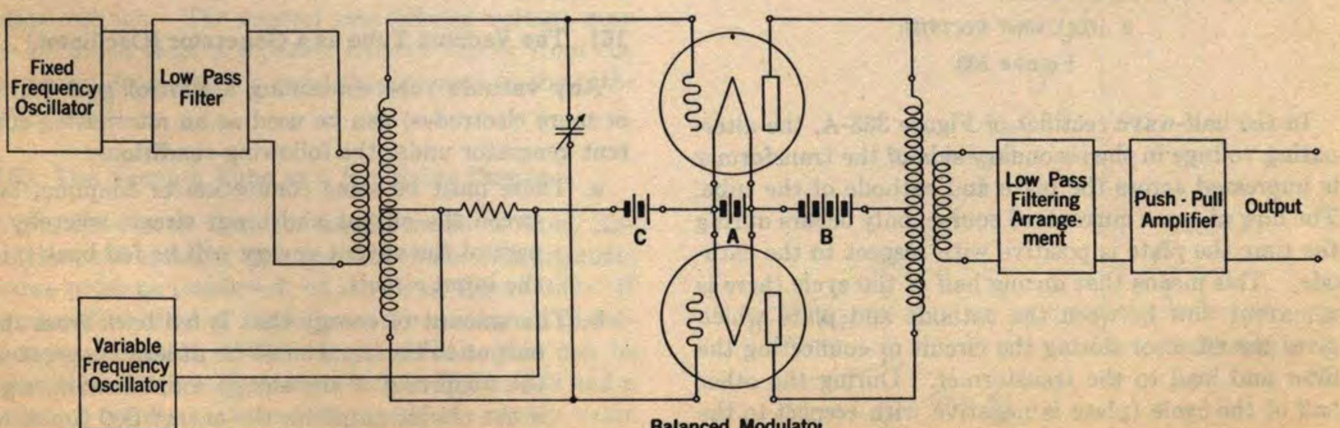


Fig. 335. Beat-Frequency Vacuum Tube Oscillator Circuit

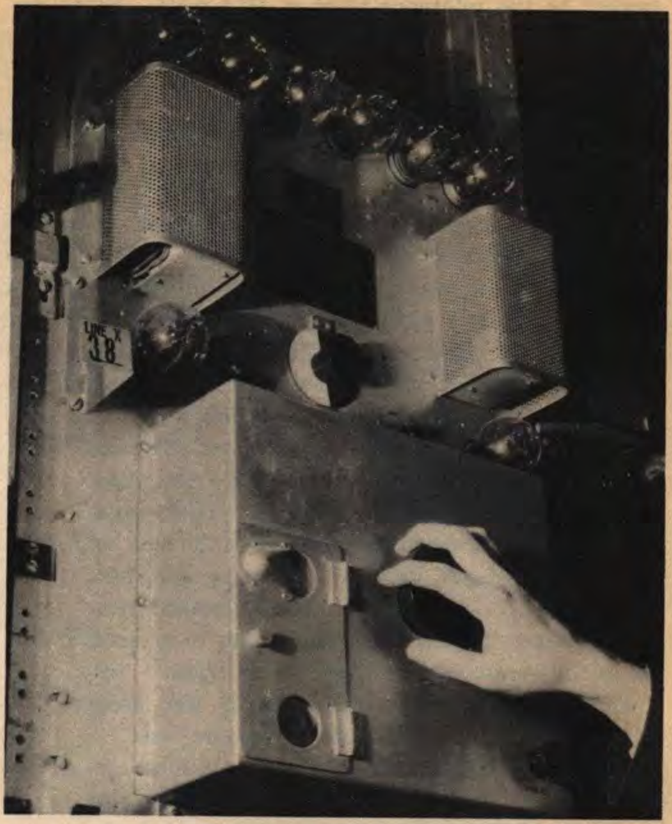
than 100,000 cycles per second. In addition, a low-pass filtering arrangement is inserted between the modulator and the push-pull amplifier which prevents the fundamental frequencies of the two oscillators, and any other high-frequency products of modulation, from passing into the amplifier. Therefore,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

162. Microphonic Noise in Vacuum Tubes

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

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

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



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

With the increasing requirements in vacuum tube performance, it becomes necessary to reduce as far as practicable these internal effects which result in microphonic noises. This is especially true where amplifiers are used on high grade program circuits. Obviously, there must be some means provided to measure the microphonic noise output of a vacuum tube for a given amount of external agitation. Such a testing arrangement is used by the manufacturer in checking the noise characteristics of the various tubes manufactured. The amount of microphonic noise a vacuum tube produces is, of course, determined to a certain extent by the severity of the external agitation to which it is subjected. Therefore, there must be some standard agitation arrangement for the tubes tested before any weight can be given the amount of microphonic noise measured, and then the measurements are primarily for comparison. It is, however, a simple matter to establish a limit for the amount of permissible microphonic noise in various types of amplifiers for a given amount of agitation. By using a fixed reference point this noise can be measured in terms of decibels with respect to the reference level of 10^{-5} watt. Since no ordinary vacuum tube produces a microphonic noise output as high as this level, for the given standard amount of agitation used in the manufacturer's test, the measurements are in db below this reference level. The agitation the vacuum tube is subjected to in this test is, however, much more severe than would ordinarily occur in service. This mechanical agitation would have to be reduced in the order of 40 db to duplicate approximately the maximum agitation that a vacuum tube is likely to be subjected to in a telephone repeater station.

CHAPTER XXVI

TELEPHONE REPEATERS AND AMPLIFIER CIRCUITS

163. Uses of Telephone Repeaters and Amplifiers

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

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

164. 22-Type Repeaters

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

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

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

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

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

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

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

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

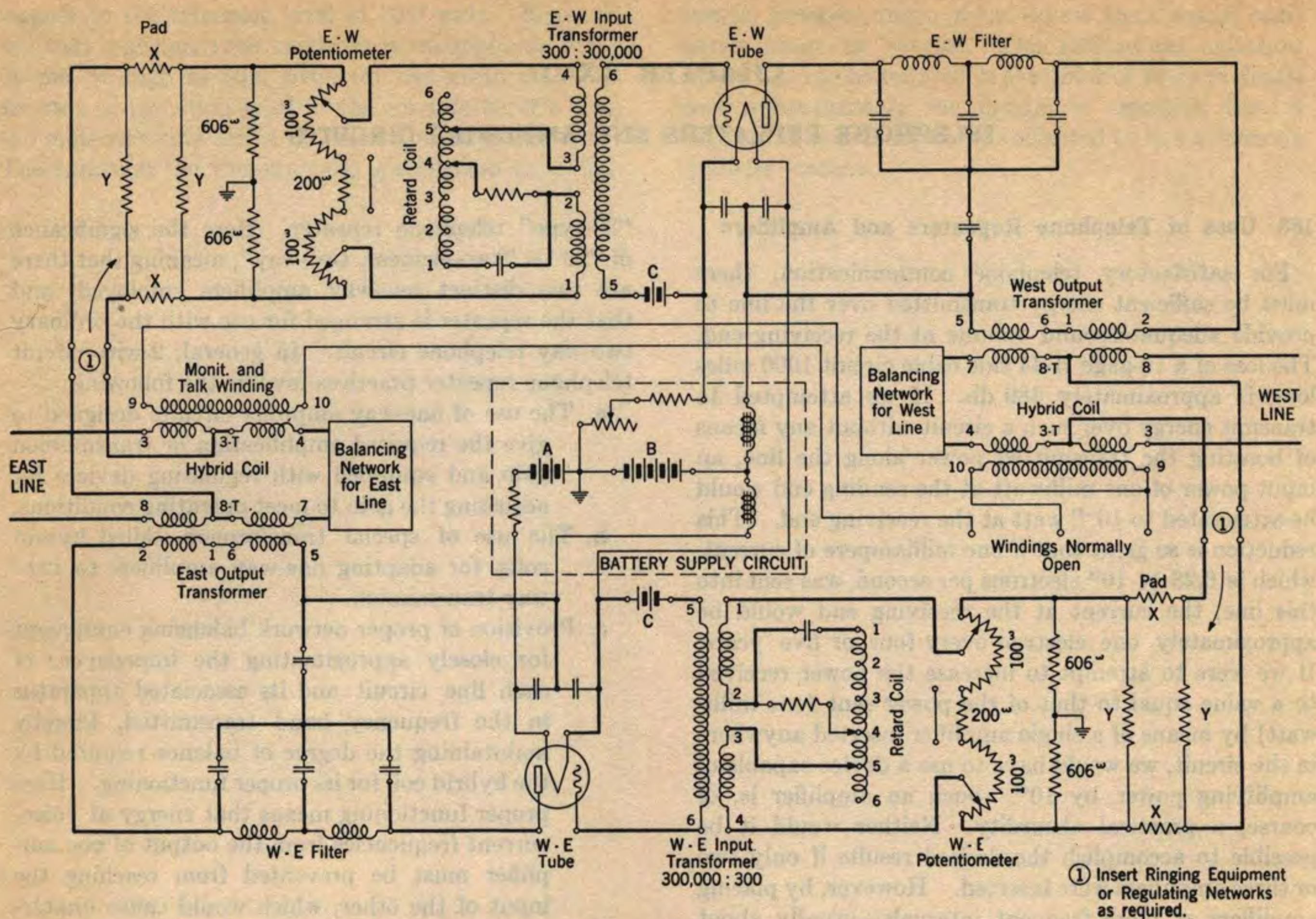
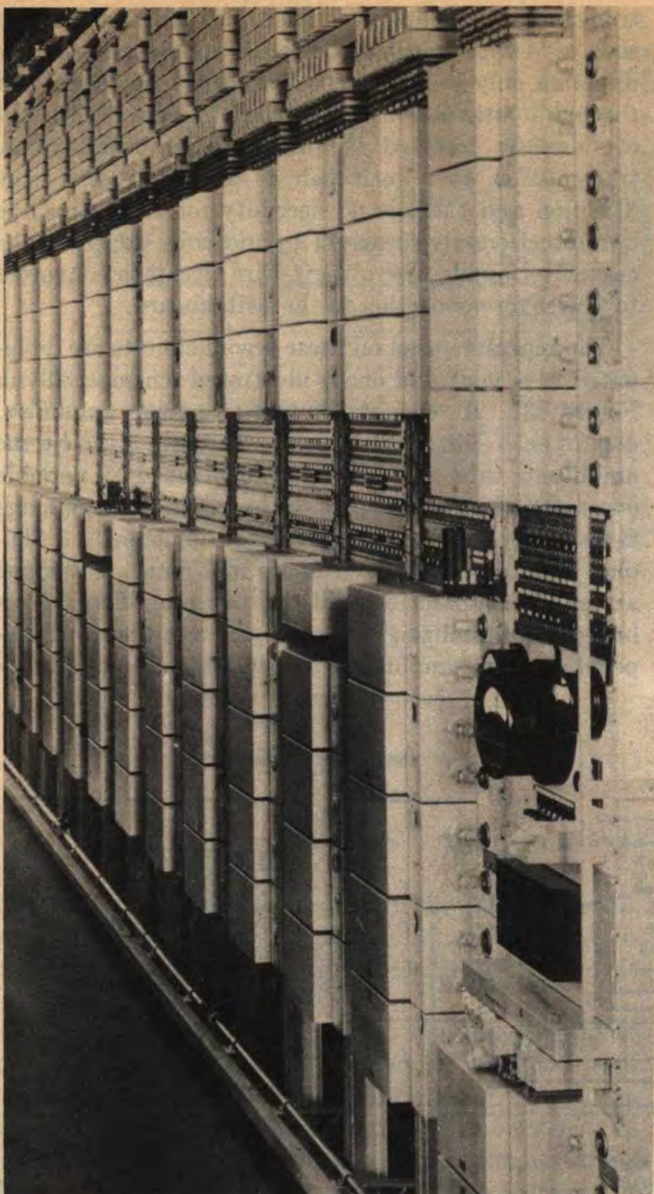


Fig. 336. 22-Type Telephone Repeater Circuit

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

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

To compensate, or equalize, for the difference in transmission characteristics of various lines, equalizing networks are connected in series at the mid-point of the low impedance side of the input transformers, as



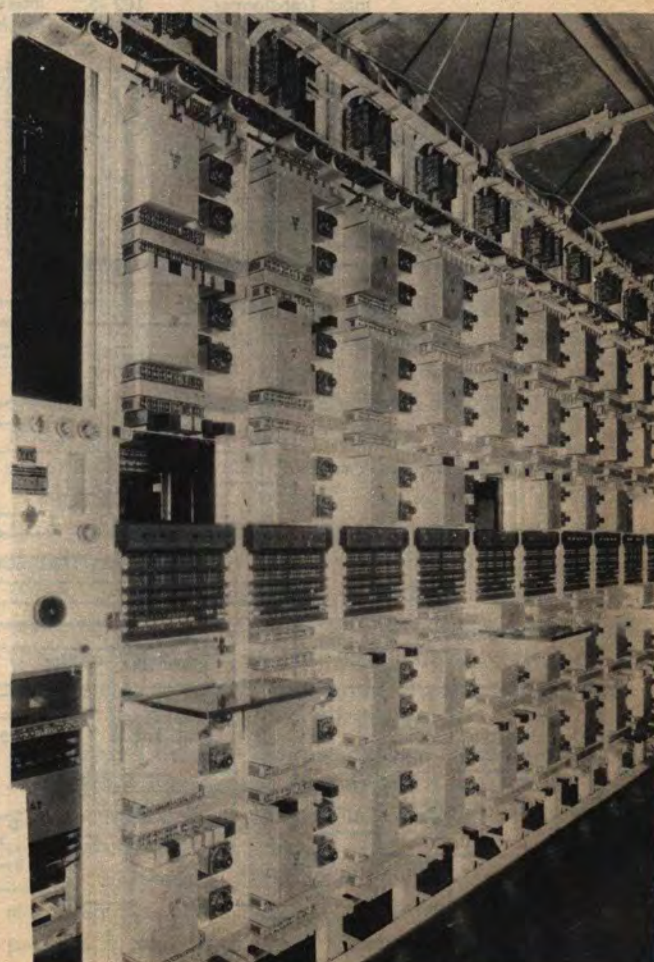
22-Type Telephone Repeaters

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

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

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

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



44-Type Telephone Repeaters

repeating coils (transformers) of the proper impedance ratio.

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

165. 44-Type Repeaters

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

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

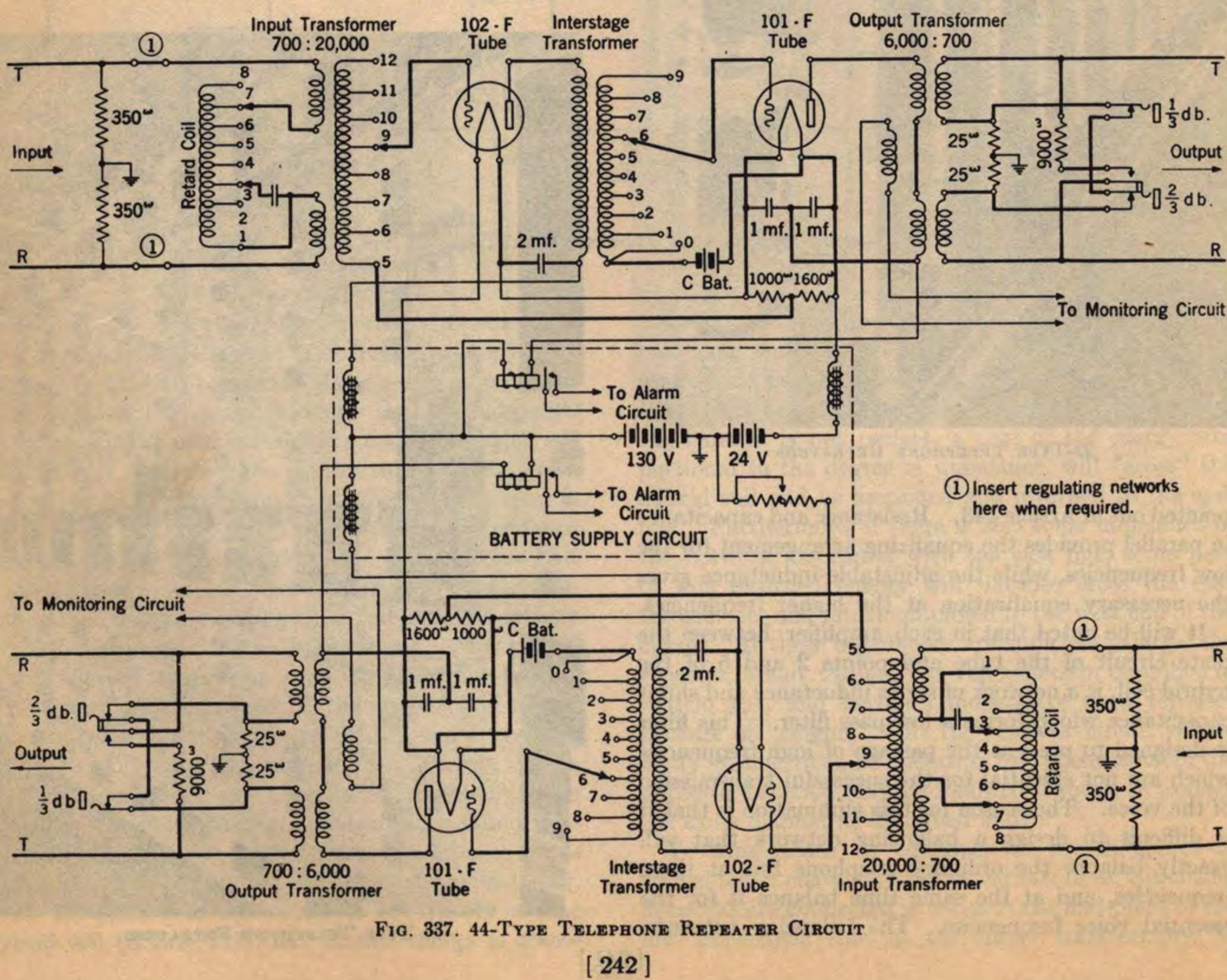


FIG. 337. 44-TYPE TELEPHONE REPEATER CIRCUIT

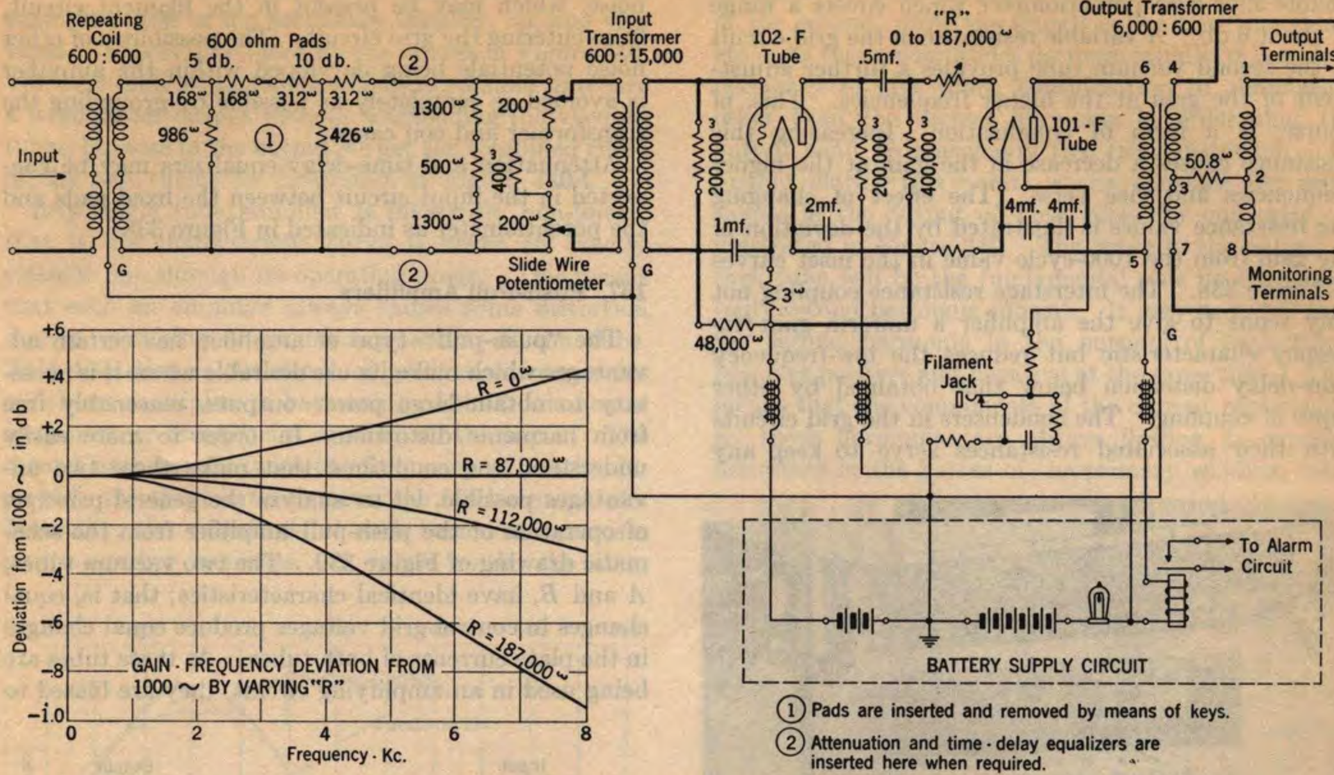


Fig. 338. Amplifier Used in Program Transmission Circuits

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

The overall gain of the amplifier may be adjusted at three places. Steps of 4.7 db may be obtained from taps on the secondary of the input transformer. (Once the repeater has been initially adjusted, however, it is not ordinarily necessary to change these connections.) Finer gain adjustments are obtained from the taps on the secondary of the interstage transformer, which are operated by a dial switch. Each tap on this switch corresponds to .95 db. Still smaller steps, of either ⅓ or ⅔ db, may be obtained by inserting dummy plugs in either of the two jacks in the output circuit. In one case, a 50-ohm resistance is put in series with the output coil and in the other, a 9000-ohm resistance is bridged across the output coil in addition. The shape of the gain-frequency characteristic is controlled by an equalizing network of inductance and capacitance connected in series at the mid-point of the primary of the input transformer.

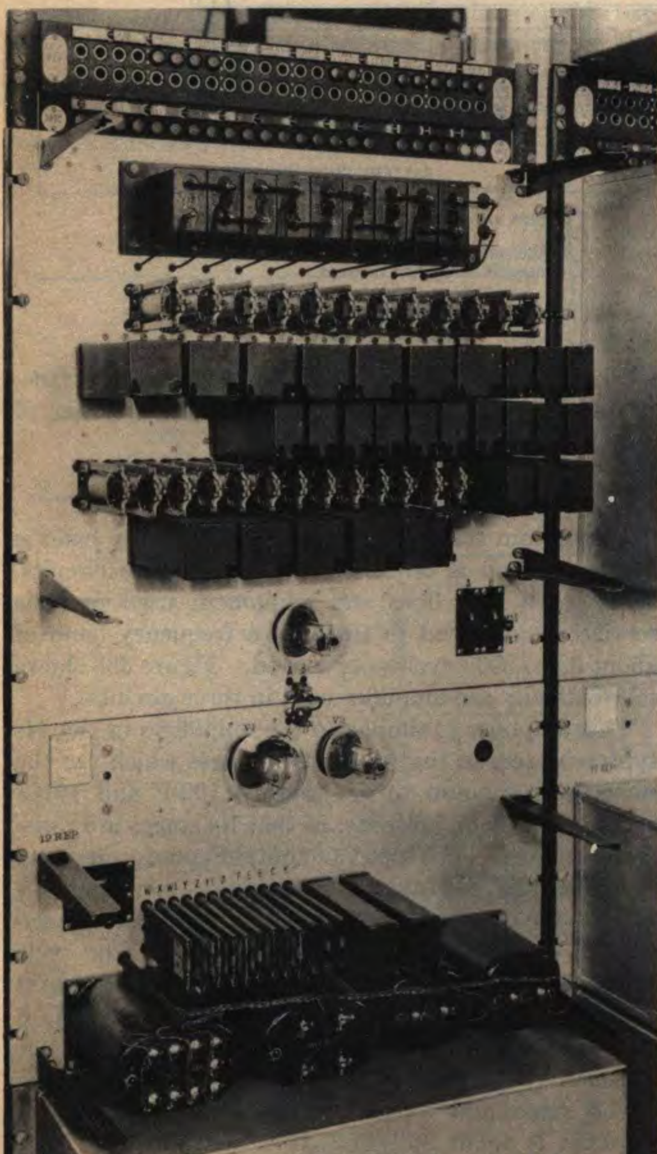
The first stage of amplification in this repeater is a voltage amplifier using a 102-F type vacuum tube. The output of this tube operates into the interstage transformer which has a voltage ratio of 1.25:1. The second stage, which employs a 101-F type tube, is a power amplifier. The output impedance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

166. Program Amplifiers

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

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

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



Cable Program Repeater and Associated Regulating Network

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

167. Push-Pull Amplifiers

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

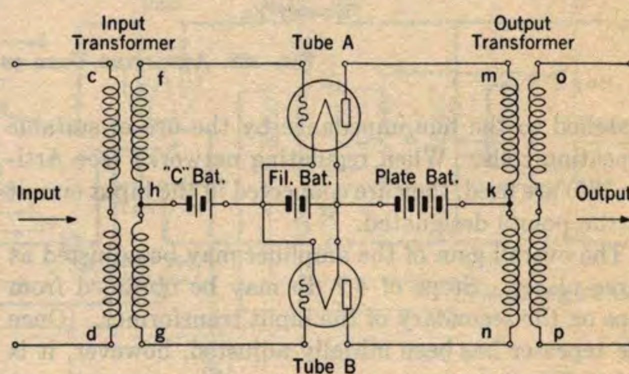


Fig. 339. Push-pull Amplifier

operate on the approximate straight line portion of their characteristic curve (control grid voltage vs. plate current curve). An alternating voltage applied at the input, *cd*, impresses voltages of equal magnitude but opposite polarity upon the control grids of tubes *A* and *B*. As the control grid of one tube becomes less negative (more positive), its plate current increases; at the same time, the control grid of the other tube becomes equally more negative, which decreases its plate current; and vice versa. Since the plate battery is connected to the mid-point, *k*, of the primary winding, *mn*, of the output transformer, the plate currents flow in opposite directions in each half of the primary winding. When the two plate currents are equal, therefore, there is no current in the secondary winding, *op*. On the other hand, **a decreasing plate current in one half of the primary winding, and an increasing plate**

current in the other half, induce equal currents in the same direction in the secondary winding, *op*, and vice versa. This is illustrated by the curves of Figure 340. The push-pull arrangement makes it possible to secure a large power output without overloading the vacuum tubes, because in the output we get the combined effect of the two tubes, as illustrated in Figure 340-C.

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

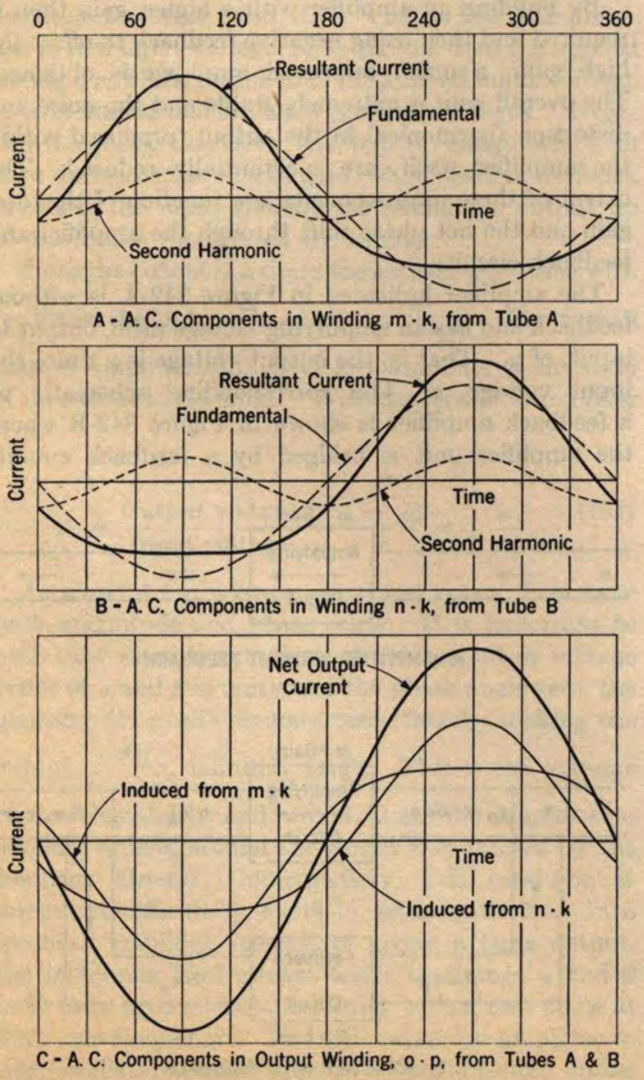
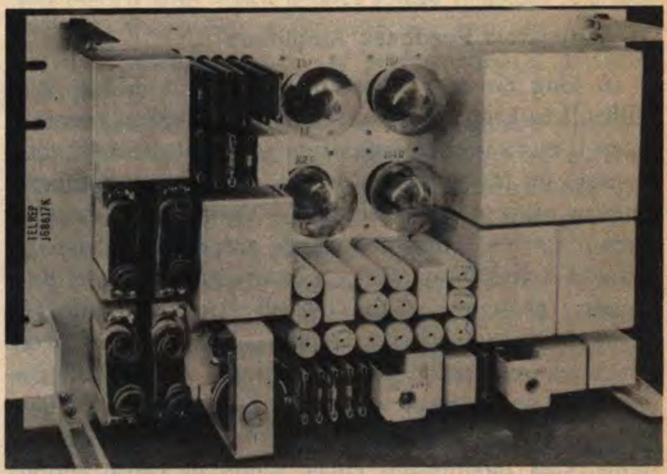


Fig. 340. Principle of Push-pull Amplifier



Push-pull Amplifier Used on Open Wire Program Circuits

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

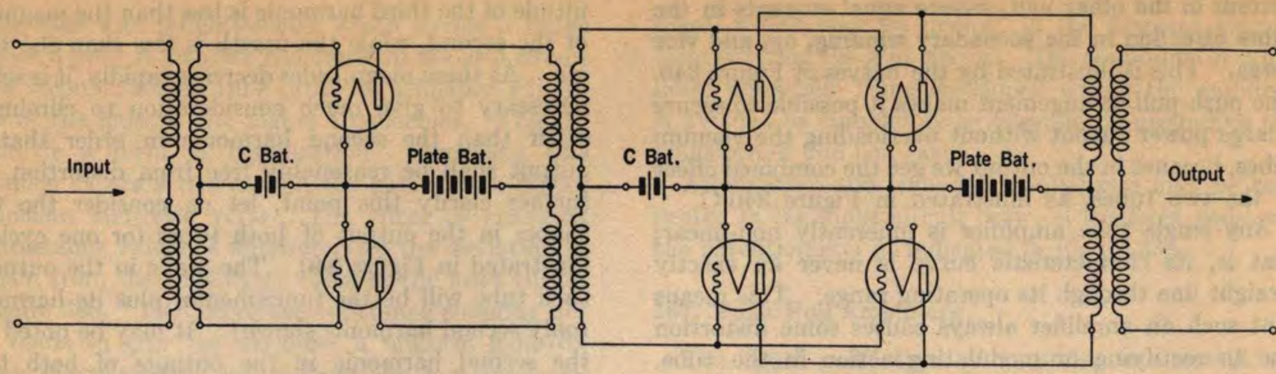


Fig. 341. Push-pull Amplifier with Second Stage Tubes in Parallel

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

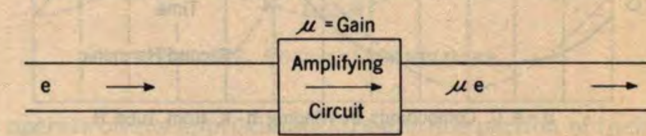
168. Negative Feedback Amplifiers

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

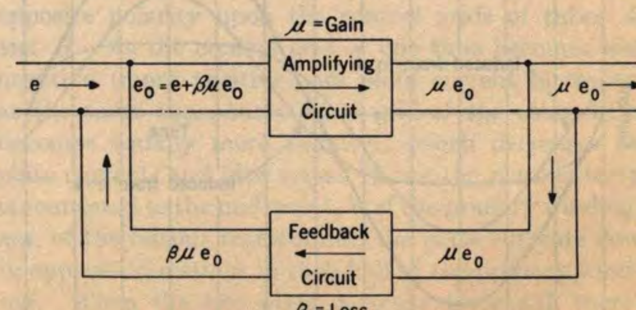
All of the above problems may be overcome by using amplifiers of the "negative feedback type". The principle of the "feedback" amplifier is to return a portion of the output back to the input. The gain of the amplifier is then increased or decreased, depending upon the magnitude and phase relation of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

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

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



A. AMPLIFIER WITHOUT FEEDBACK



B. AMPLIFIER WITH FEEDBACK

Figure 342

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

$$e_0 = e + \mu\beta e_0 \tag{152}$$

Solving this for e_0, we get—

$$e_0 = \frac{e}{1 - \mu\beta} \tag{153}$$

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

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

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

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

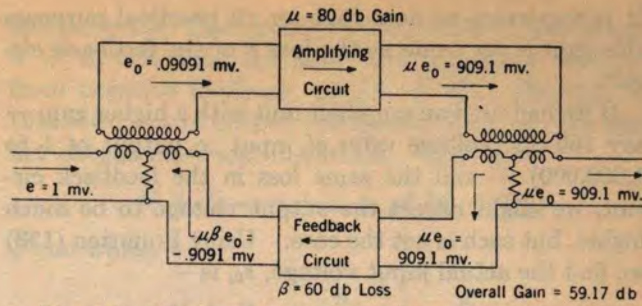


FIG. 343. EXAMPLE OF NEGATIVE FEEDBACK PRINCIPLE

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

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

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

The output voltage is—

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

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

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

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

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

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

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

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

The output voltage, μe_0, would now be—

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

The overall gain of the amplifier is now—

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

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

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

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

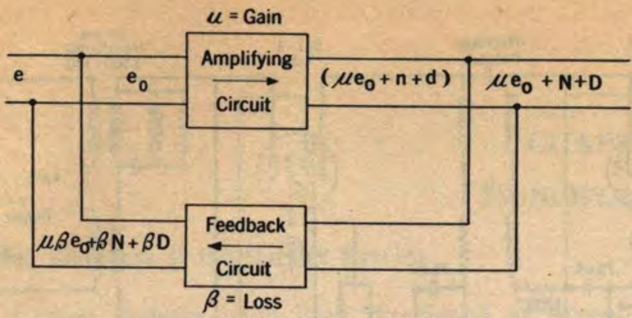
TABLE XVII

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

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

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

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



AMPLIFIER WITH FEEDBACK

FIGURE 344

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

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

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

From which—

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

or—

$$N = \frac{n}{1 - \mu\beta} \tag{155}$$

and—

$$D = \frac{d}{1 - \mu\beta} \tag{156}$$

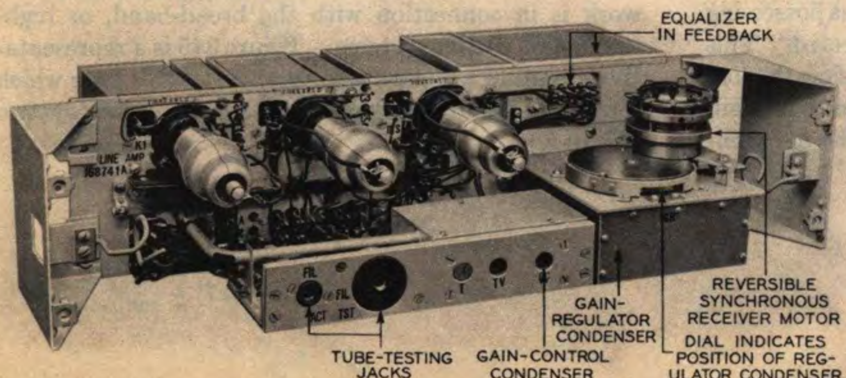
In other words, negative feedback reduces the noise and distortion by the factor $\frac{1}{1 - \mu\beta}$. As an example, let μ = 100 db (voltage ratio of 100,000 to 1) and β = 60 db (voltage ratio of 1 to 1,000). Then with negative feedback, the noise and distortion becomes—

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

and

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

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



LINE AMPLIFIER USED IN TYPE-K CARRIER SYSTEMS

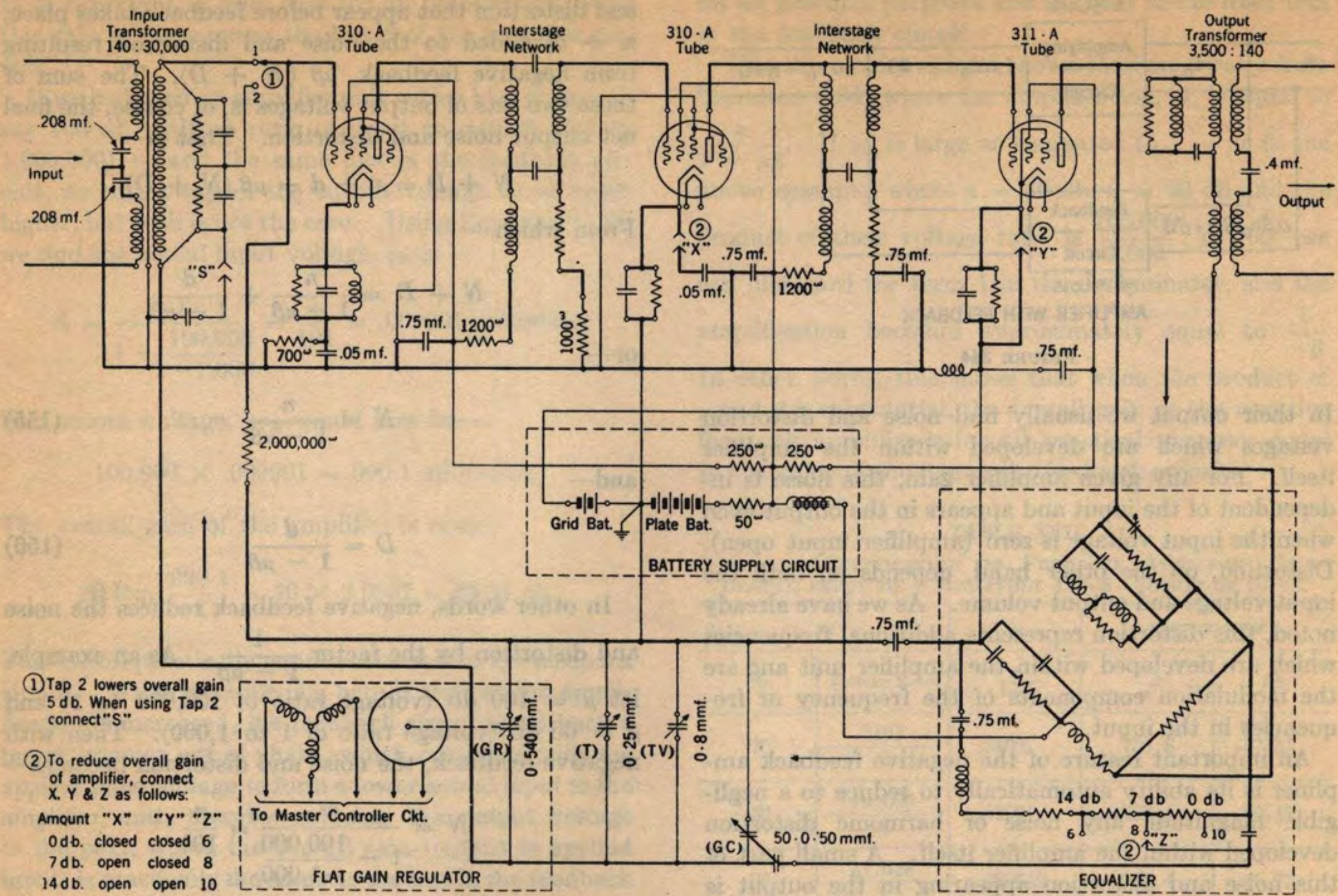


Fig. 345. Three-stage Negative Feedback Line Amplifier Used with Type K Carrier Systems

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

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

CHAPTER XXVII

PRINCIPLES OF CARRIER

169. Elements of the Carrier System

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

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

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

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

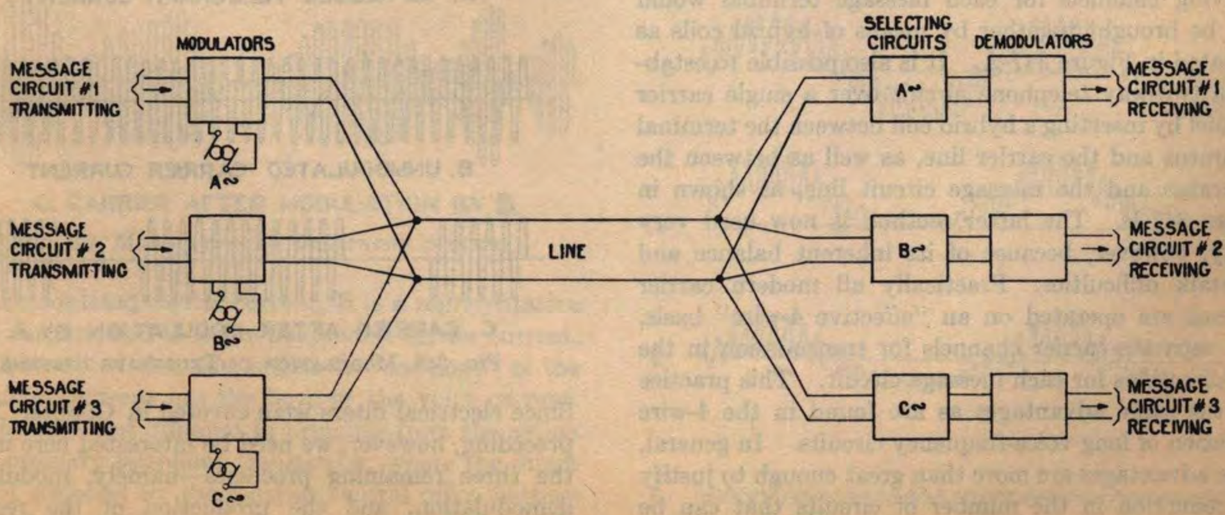


Fig. 346. Principle of Carrier Operation

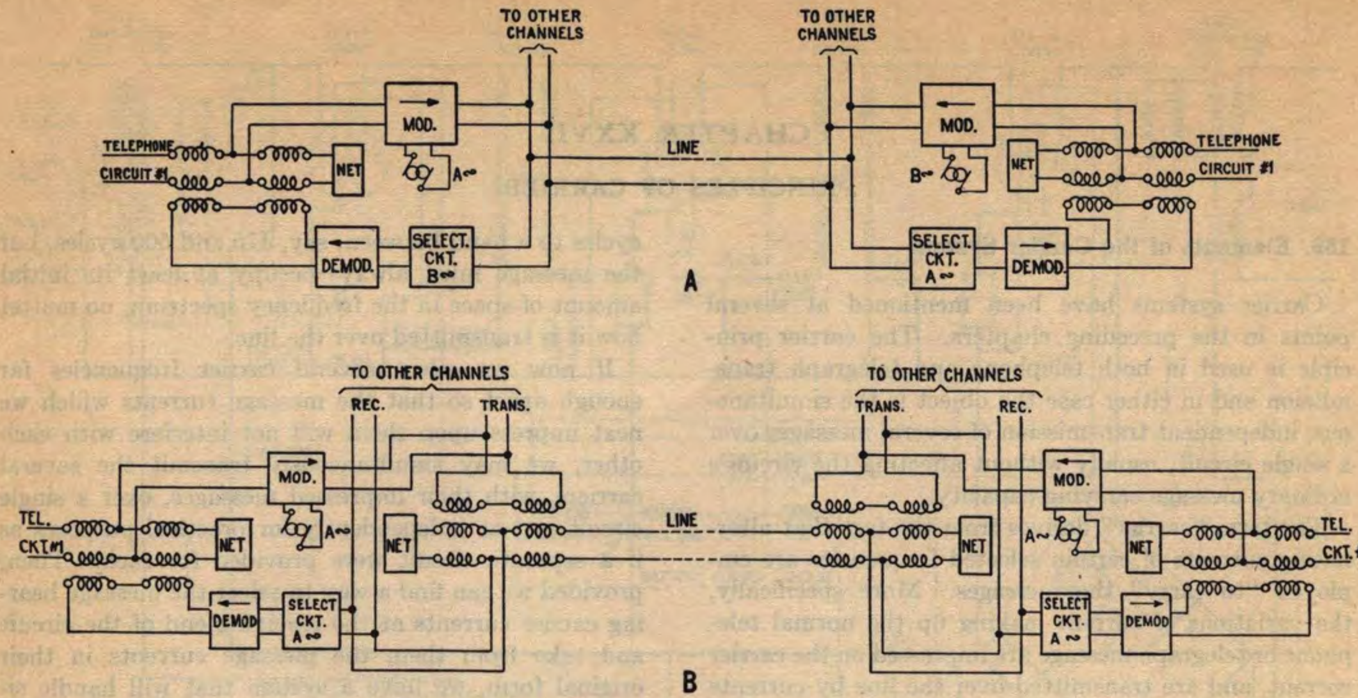


FIG. 347. METHODS OF OBTAINING TWO-WAY OPERATION

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

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

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

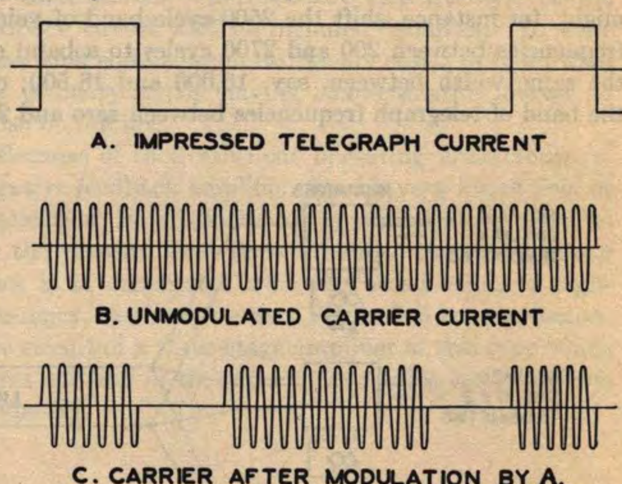


FIG. 348. MODULATION IN TELEGRAPH SYSTEMS

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

170. Modulation

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

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

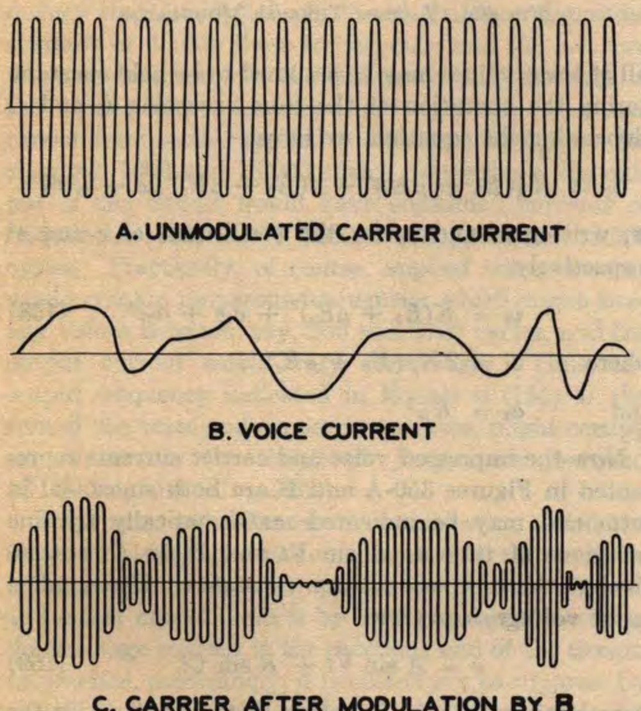


FIG. 349. MODULATION IN TELEPHONE SYSTEMS

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

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

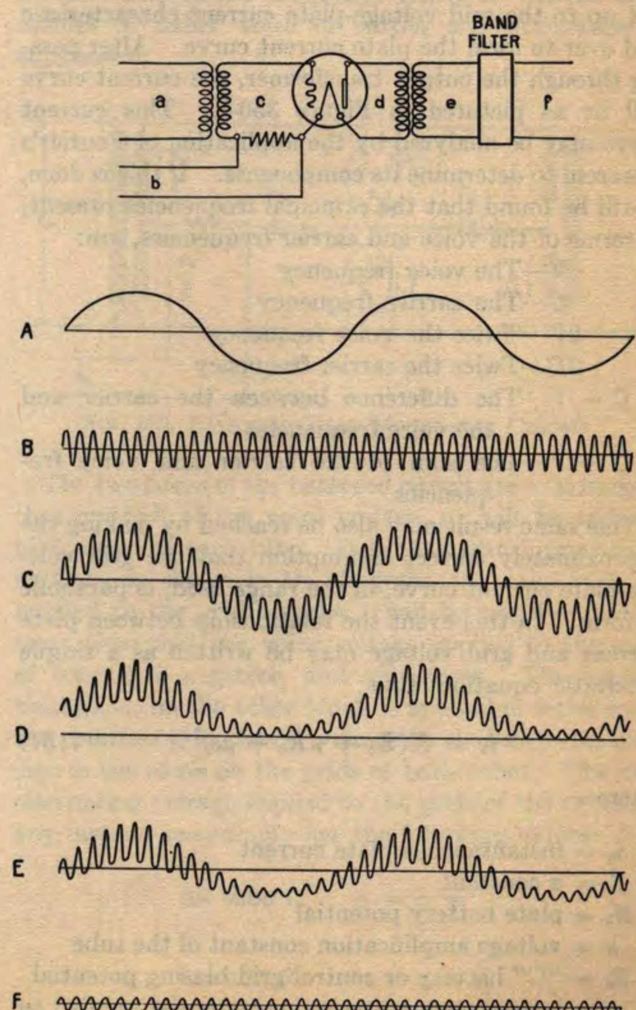


FIG. 350. CURRENTS IN MODULATOR CIRCUIT

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

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

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

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

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

where—

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

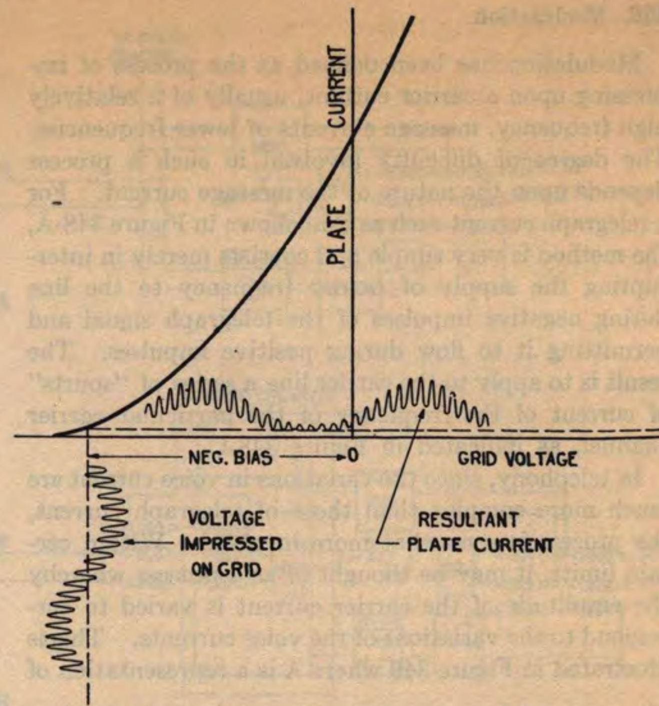


Fig. 351. Vacuum Tube as Modulator

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

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

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

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

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

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

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

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

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

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

and, expanding—

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

Making use of the trigonometric relationships—

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

and

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

we may expand further to obtain—

$$\begin{aligned} i_b = {} & K(E_b + \mu E_c)^2 + a_1 A \sin Vt + a_1 B \sin Ct \\ & + \tfrac{1}{2}a_2A^2 - \tfrac{1}{2}a_2A^2\cos 2Vt + a_2AB\cos(C - V)t \\ & - a_2AB\cos(C + V)t + \tfrac{1}{2}a_2B^2 - \tfrac{1}{2}a_2B^2\cos 2Ct \\ = {} & K(E_b + \mu E_c)^2 + \tfrac{1}{2}a_2(A^2 + B^2) + a_1A\sin Vt \\ & + a_1B\sin Ct - \tfrac{1}{2}a_2A^2\cos 2Vt - \tfrac{1}{2}a_2B^2\cos 2Ct \\ & + a_2AB\cos(C - V)t - a_2AB\cos(C + V)t. \quad (161) \end{aligned}$$

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

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

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

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

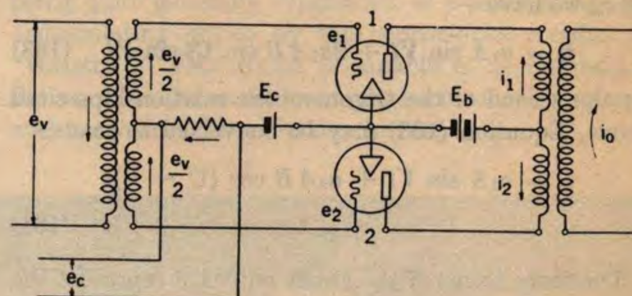


FIG. 352. BALANCED TUBE MODULATOR CIRCUIT

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

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

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

These values of input voltage may be substituted in

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

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

and, for tube 2—

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

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

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

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

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

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

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

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

$$i_0 = a_1A \sin Vt + a_2AB \cos (C - V)t - a_2AB \cos (C + V)t \qquad (164)$$

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

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

171. Demodulation

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

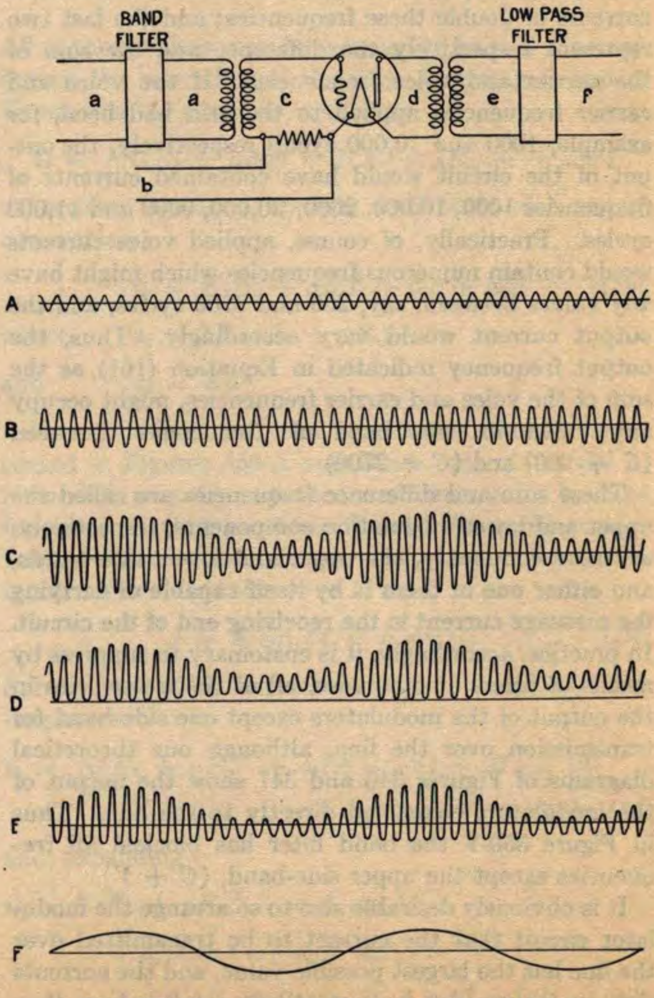


Fig. 353. Currents in Demodulator Circuit

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

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

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

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

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

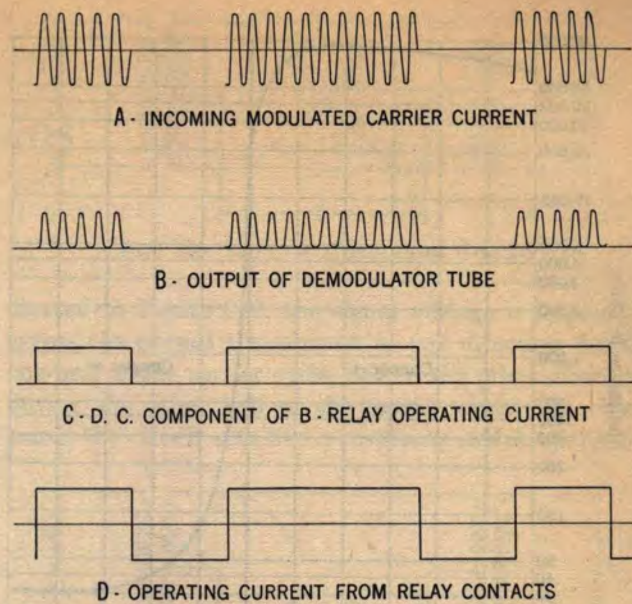


FIG. 354. DEMODULATION IN TELEGRAPH SYSTEMS

172. Copper-Oxide Varistors as Modulators and Demodulators

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

COPPER-OXIDE VARISTOR UNIT USED AS MODULATOR AND DEMODULATOR

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

For use as modulators and demodulators in carrier

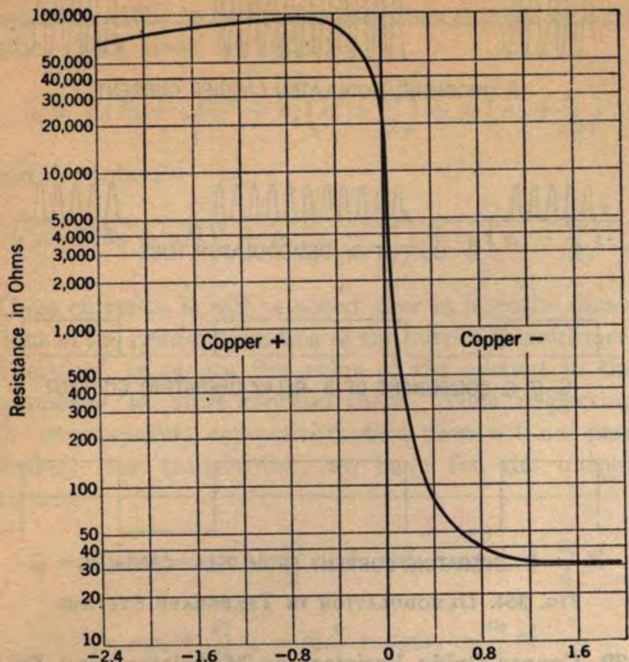


FIG. 355. RESISTANCE CHARACTERISTIC OF COPPER-OXIDE VARISTOR

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

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

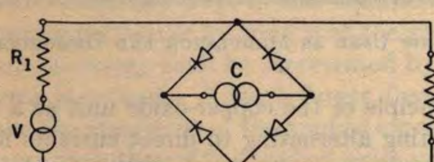


FIG. 356. BALANCED BRIDGE MODULATOR CIRCUIT

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

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

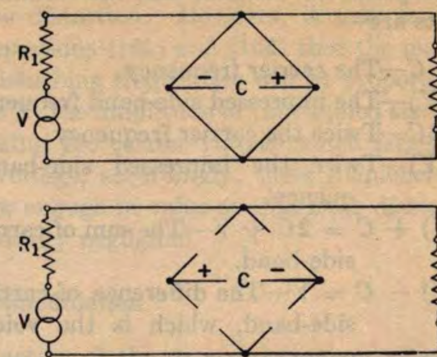


FIG. 357. OPERATING PRINCIPLE OF CIRCUIT OF FIG. 356

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

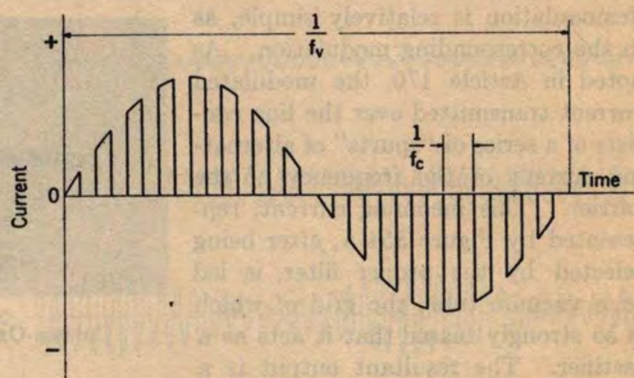


FIG. 358. OUTPUT CURRENT OF BALANCED BRIDGE MODULATOR

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

$$e = A \sin Vt$$

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

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

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

Making use of the trigonometric relationship—

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

the above equation may be rewritten as—

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

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

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

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

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

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

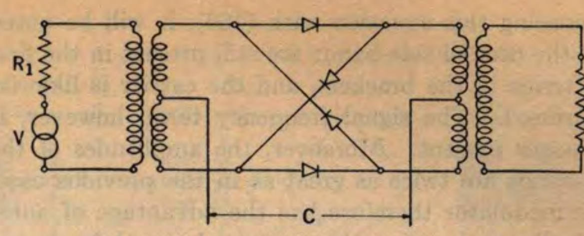


FIG. 359. LATTICE MODULATOR CIRCUIT

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

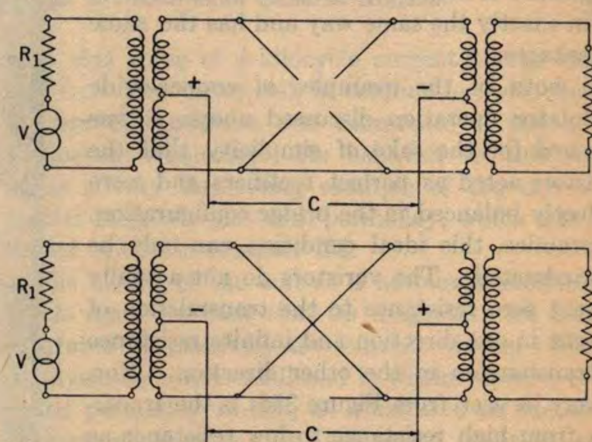


FIG. 360. OPERATING PRINCIPLE OF CIRCUIT OF FIG. 359

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

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

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

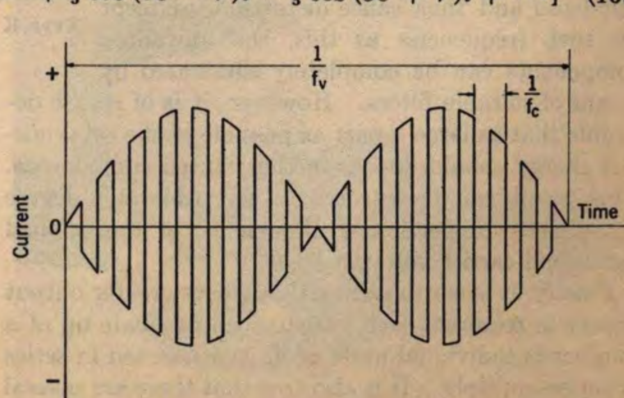


FIG. 361. OUTPUT CURRENT OF LATTICE MODULATOR

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

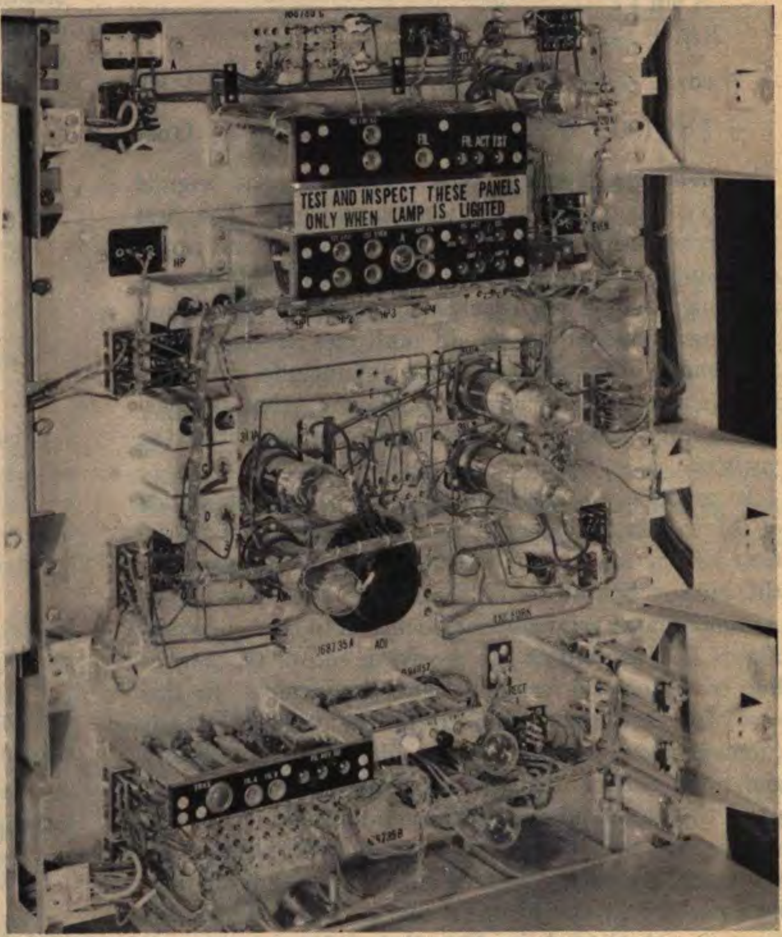
In both of the examples of copper-oxide modulator operation discussed above, it was assumed for the sake of simplicity that the varistors acted as perfect rectifiers and were perfectly balanced in the bridge configuration. In practice, this ideal condition can only be approximated. The varistors do not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

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

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

173. Sources of Carrier Frequencies

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



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

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

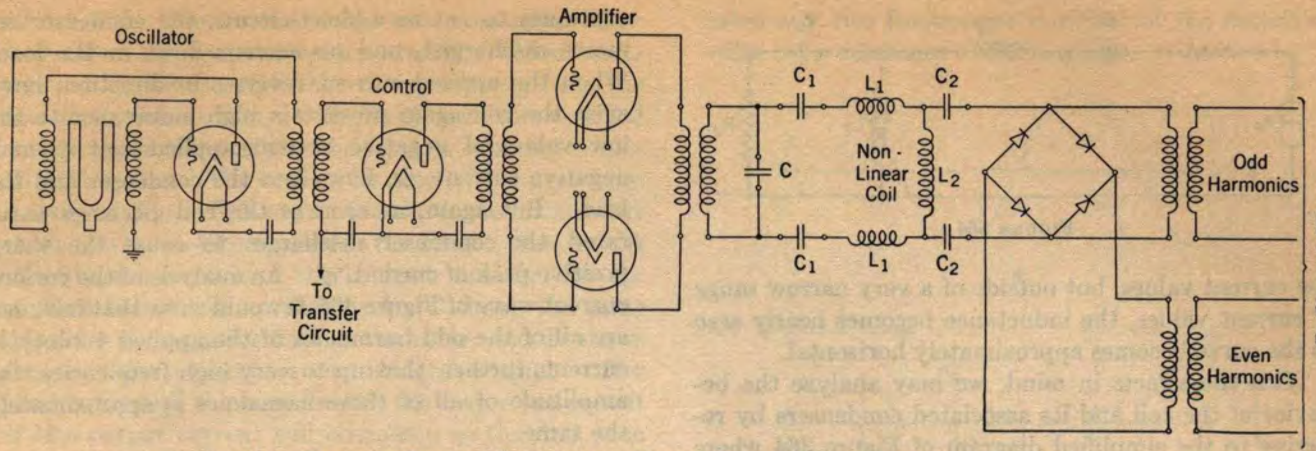


FIG. 362. CARRIER SUPPLY CIRCUIT FOR TYPES J AND K BROAD-BAND CARRIER SYSTEMS

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

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

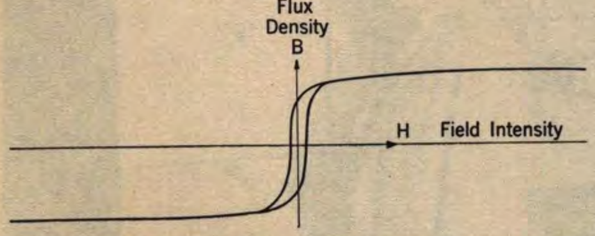


FIG. 363. B–H CURVE OF CORE OF HARMONIC PRODUCING COIL

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

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

HARMONIC PRODUCER COIL SHOWING PERMALLOY TAPE CORE

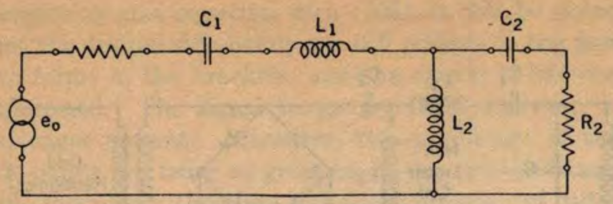


FIGURE 364

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

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

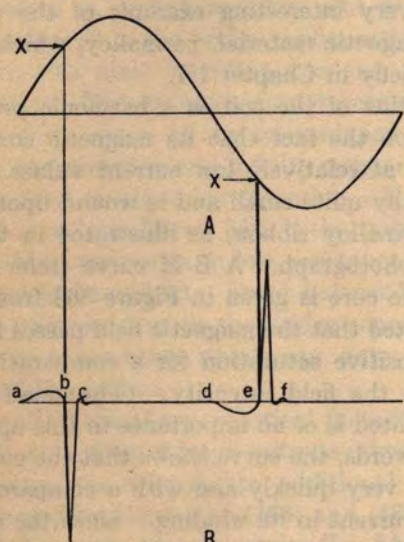
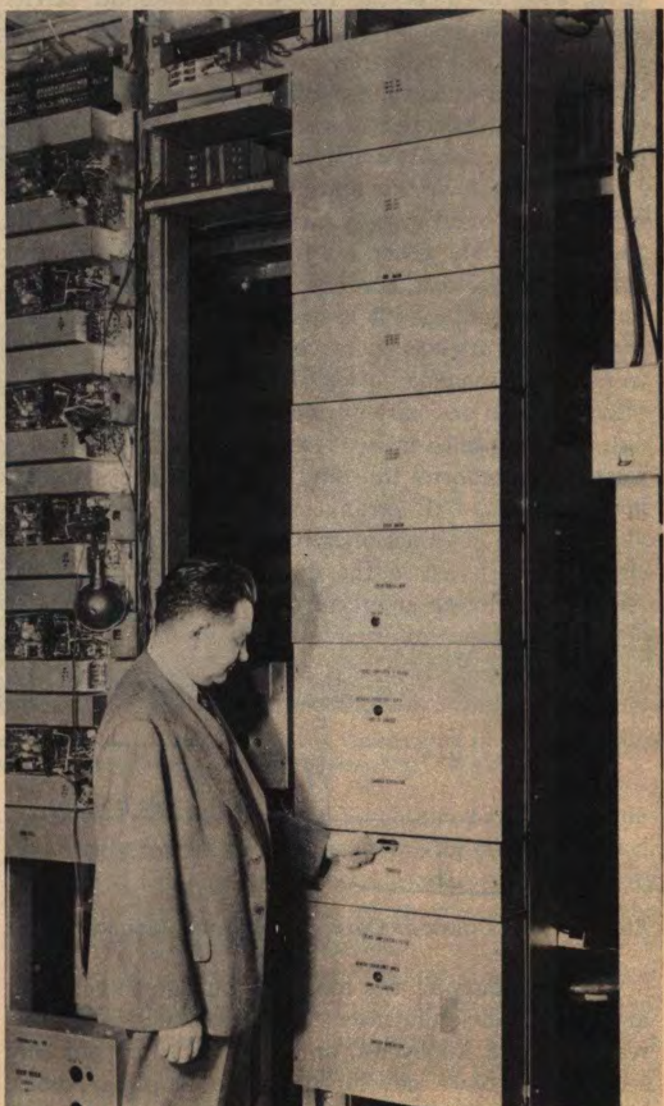


FIG. 365. OUTPUT CURRENT OF HARMONIC PRODUCING COIL CIRCUIT

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

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



TYPE-K CARRIER SUPPLY BAY

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

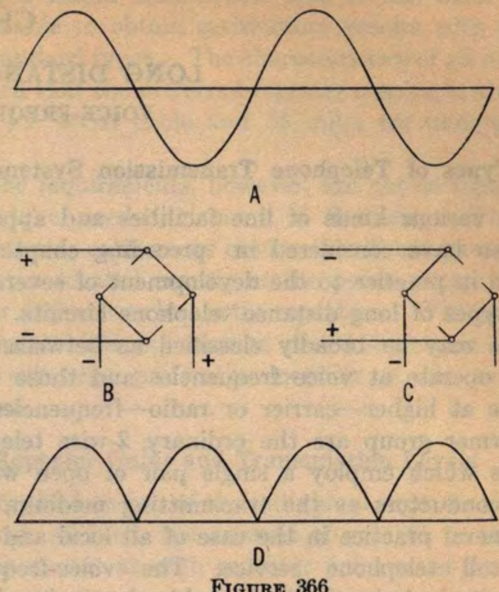


FIGURE 366

CHAPTER XXVIII

LONG DISTANCE TRANSMISSION SYSTEMS

VOICE-FREQUENCY TELEPHONE CIRCUITS

174. Types of Telephone Transmission Systems

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

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

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

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

175. Repeater Spacing

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

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

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

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

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

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

176. Repeater Gains and Transmission Levels

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

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

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

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

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

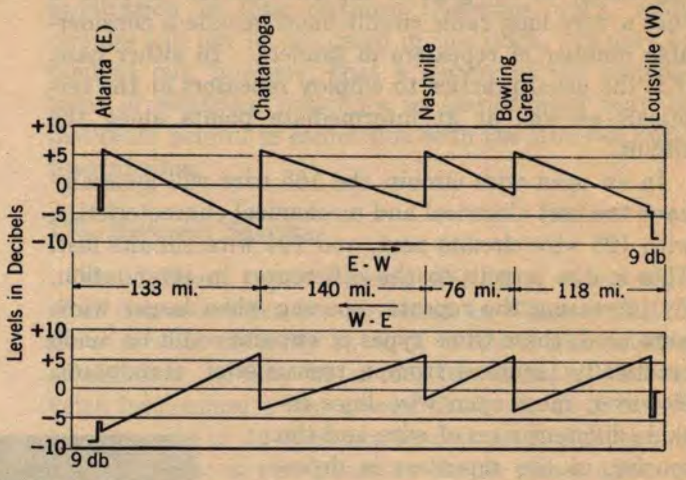


Fig. 367. Energy Level Diagram for Open Wire Circuit

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

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

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

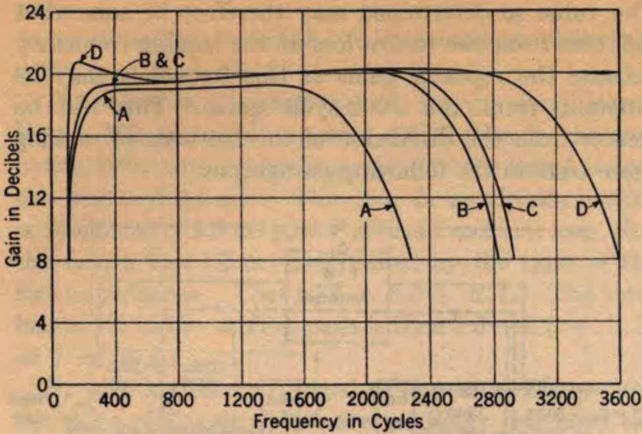


FIG. 368. GAIN-FREQUENCY CHARACTERISTICS OF 22-TYPE REPEATERS

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

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

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

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

177. Return Loss

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

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


CABLE REPEATER STATION AT WAUKEGAN, ILLINOIS

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

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

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

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

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

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

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

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

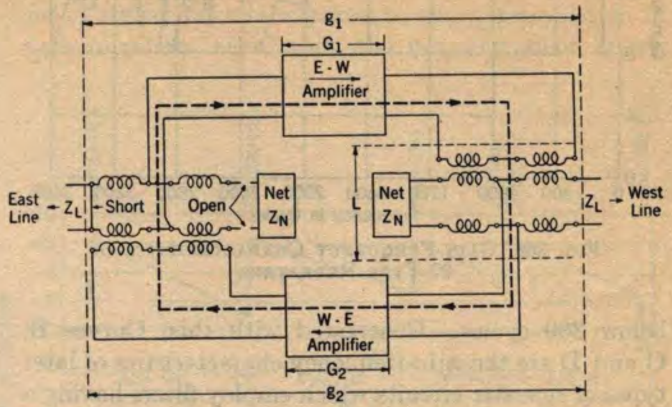


Fig. 369. Principle of Repeater Balance Test

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

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

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

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

$$g_1 + g_2 + 13 = 7 + R.L. \tag{169}$$

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

$$R.L. = g_1 + g_2 + 6 \tag{170}$$

178. Singing Points

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

$$S.P. = g_1 + g_2 + 6 \tag{171}$$

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

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

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

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

179. Balancing Networks

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

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

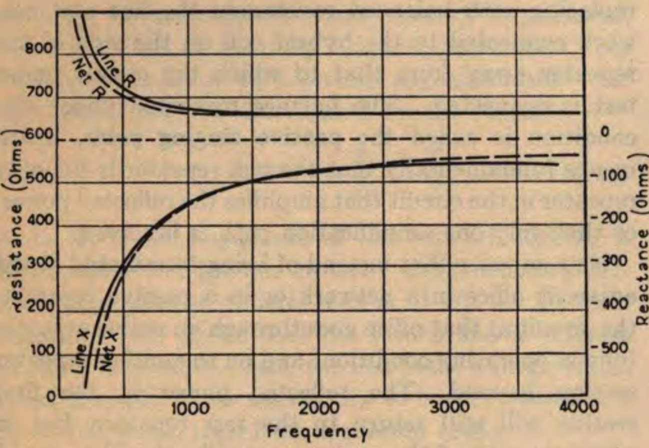


Fig. 370. Impedance Characteristics of Open Wire Circuit and Its Balancing Network

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

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

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

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

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

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

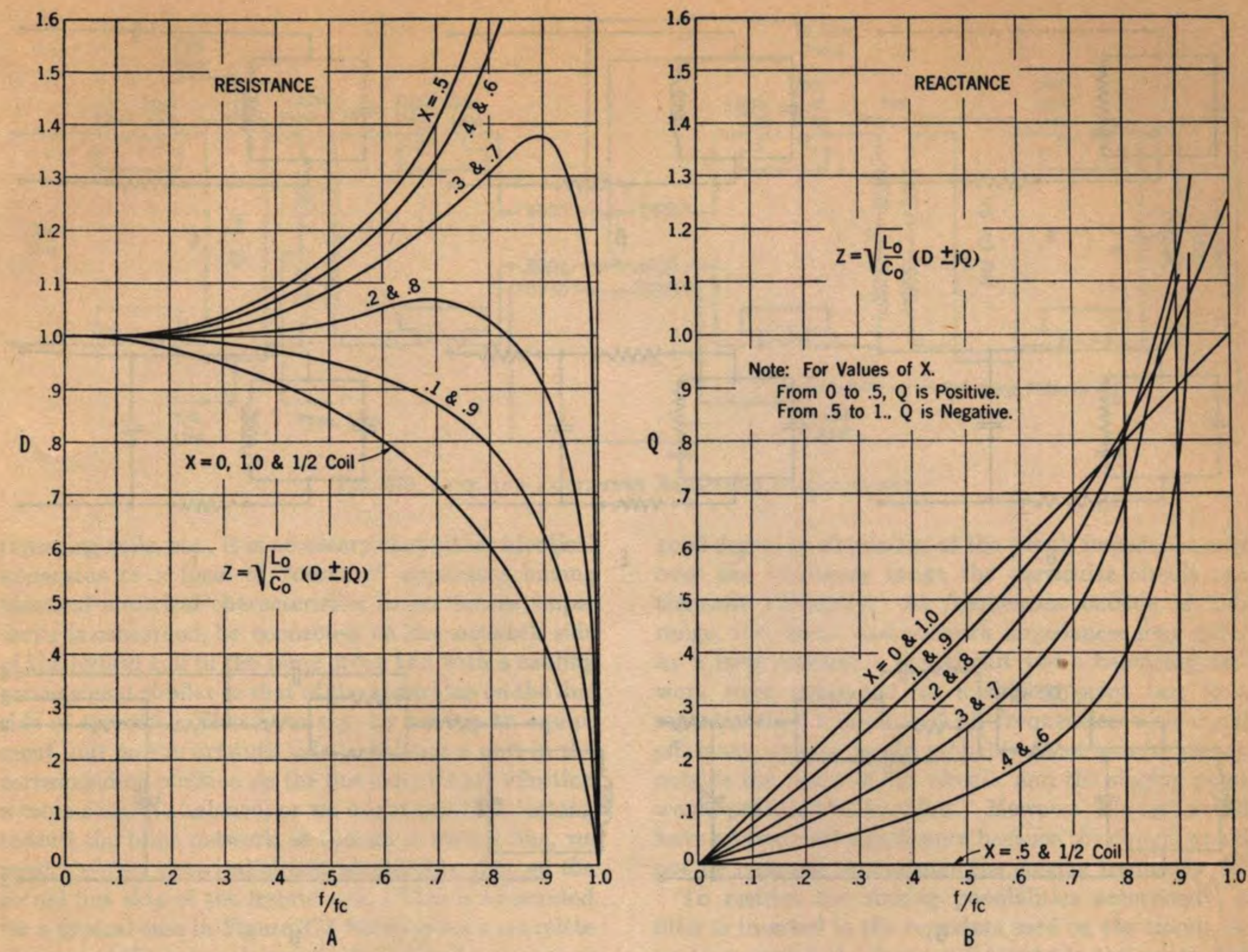


Fig. 371. Impedance Components of Ideal Loaded Circuit at Various End Sections

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

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

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

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

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

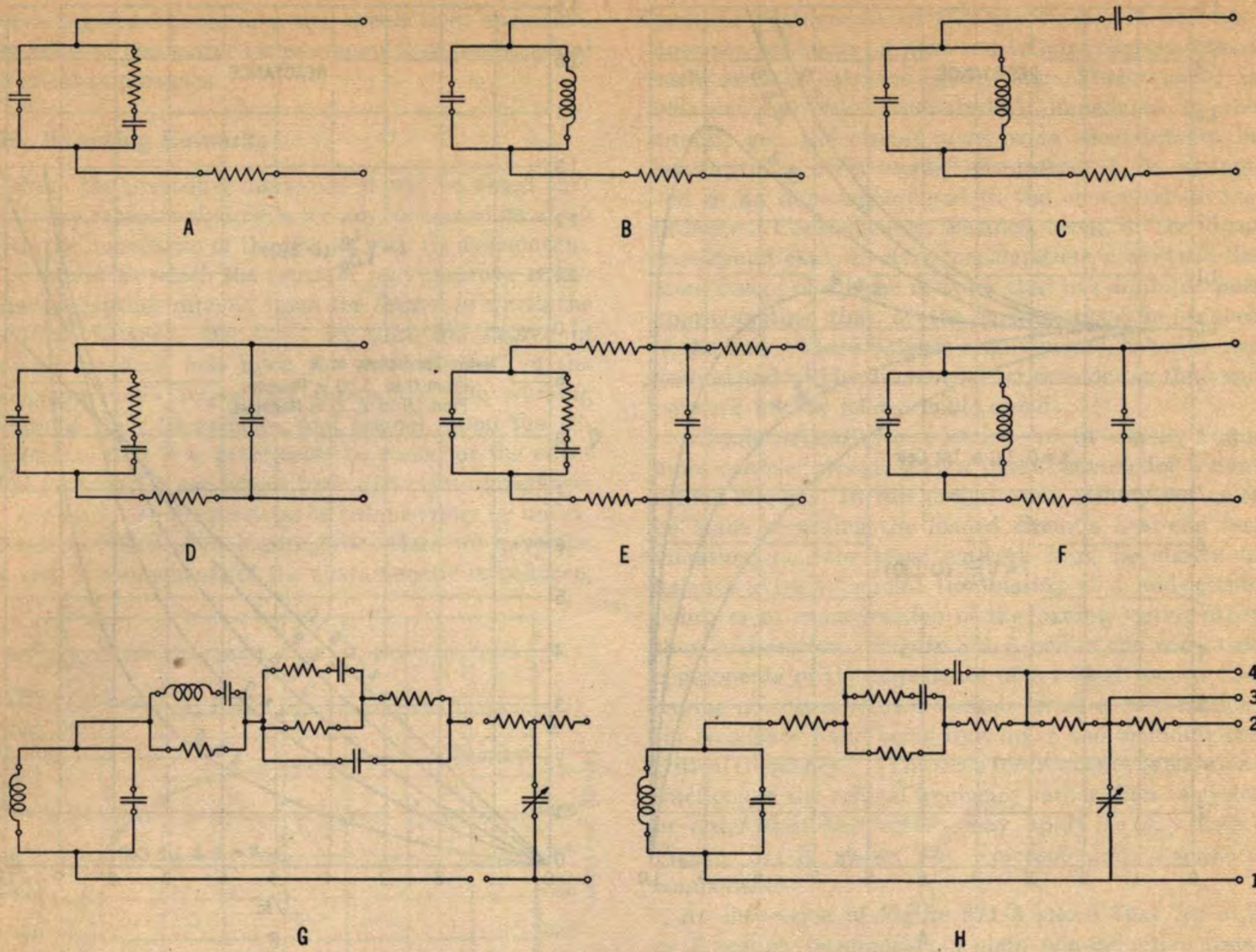


FIG. 372. STANDARD BALANCING NETWORK CIRCUITS

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

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

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

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

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

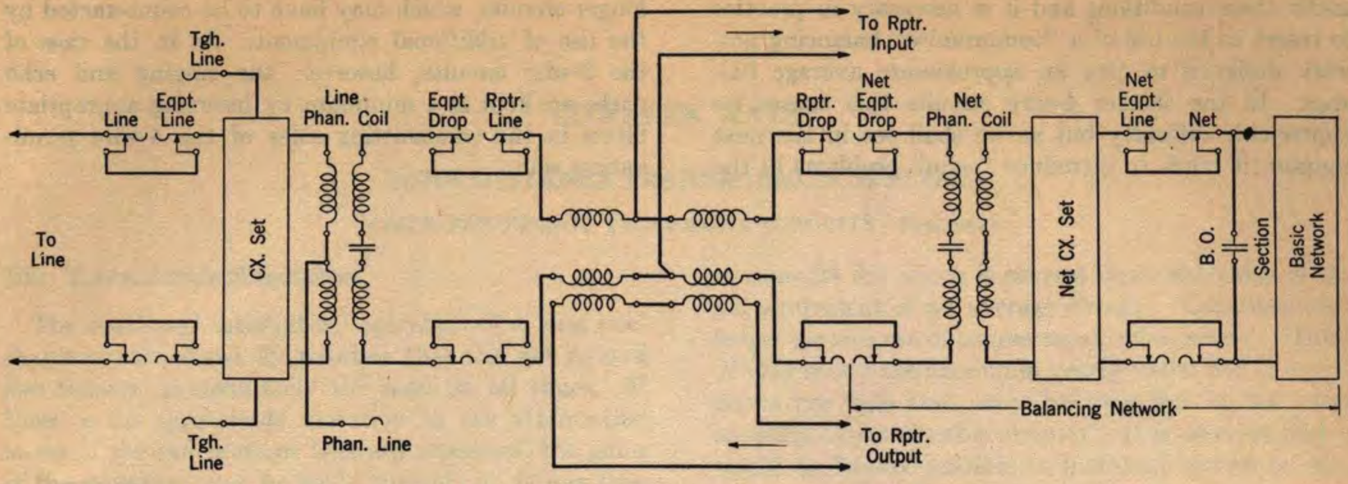


FIG. 373. LINE AND EQUIPMENT BALANCING ARRANGEMENT

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

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

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

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

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

CHAPTER XXIX

LONG DISTANCE TRANSMISSION SYSTEMS

VOICE-FREQUENCY TELEPHONE CIRCUITS—*Continued*

180. Transmission Regulation

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

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

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

TABLE XVIII

Transmission Equivalents in Decibels per Mile of 16 AWG & 19 AWG Cable Circuits at 55° F. Showing Yearly Variations in Equivalents on Account of Temperature Changes

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

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

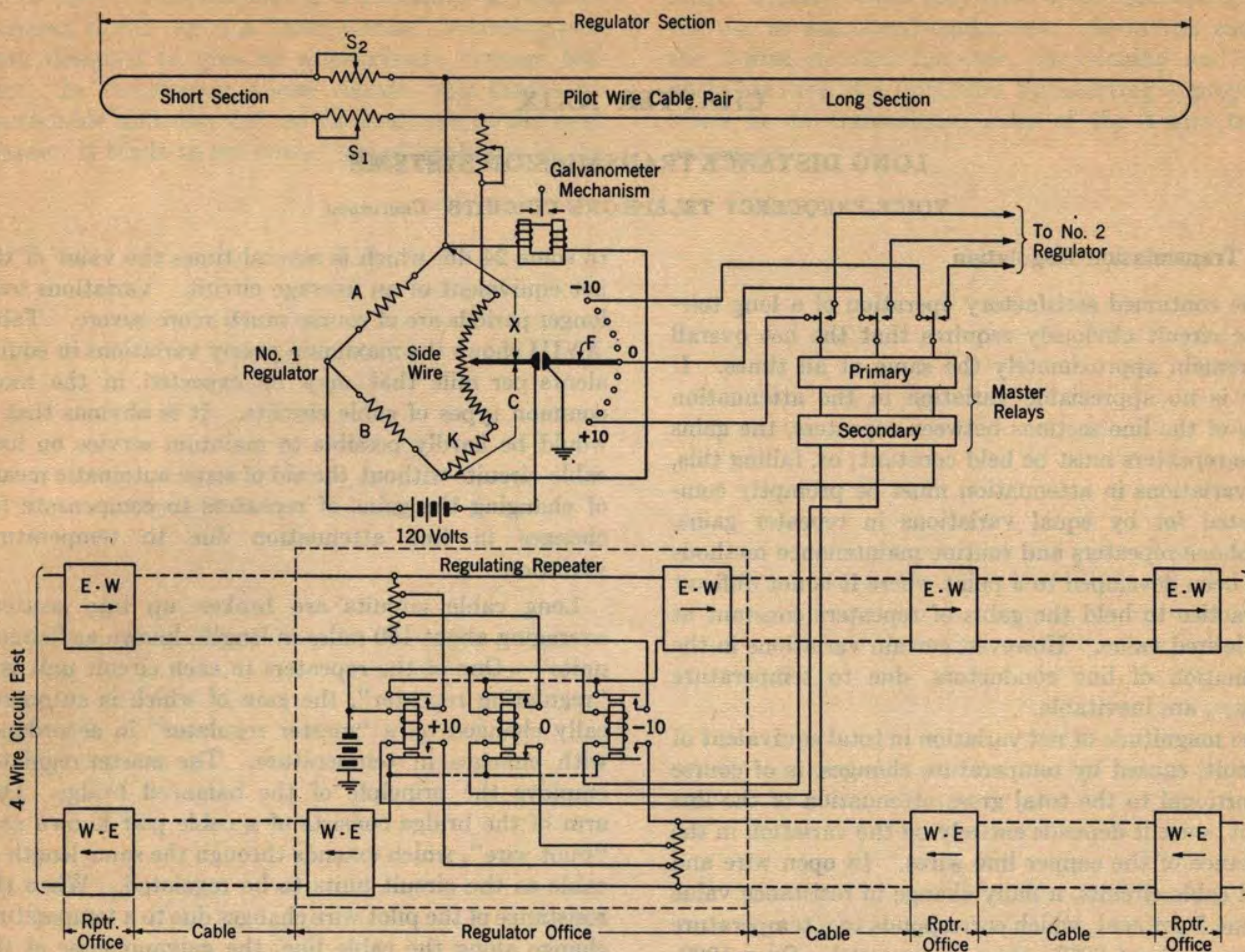


FIG. 374. PILOT WIRE TRANSMISSION REGULATOR CIRCUIT

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

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

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

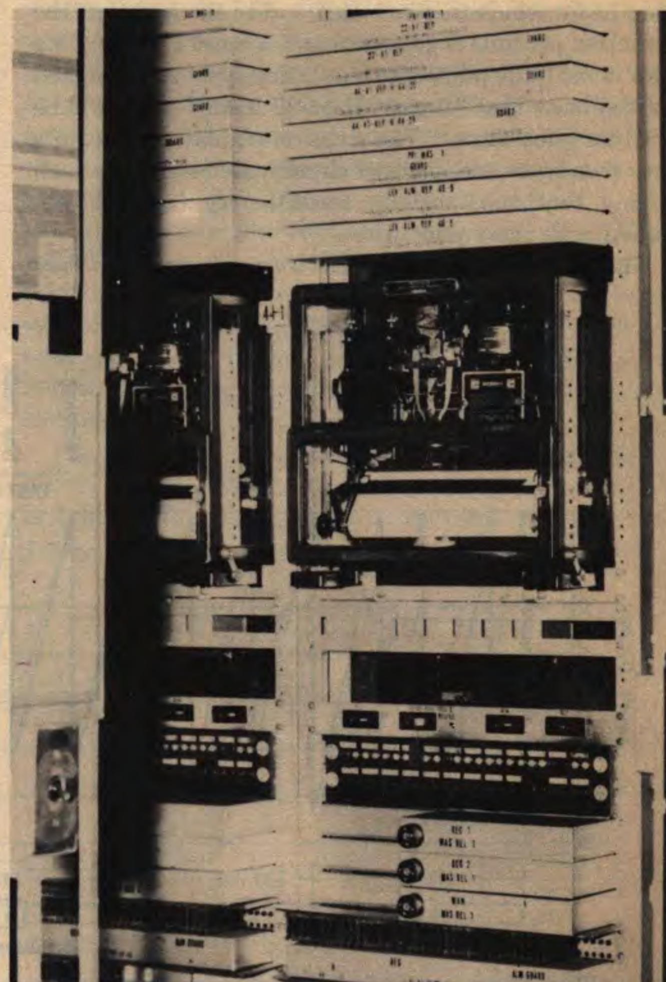
The regulating networks of the repeaters and the equalizing arrangements are designed on the basis of 55°F. being the average cable temperature. Consequently, the regulator is normally adjusted so that the slide wire will be at its mid-position (step 0) when the cable is at this temperature.

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

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

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

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



Transmission Regulator Equipment for Voice-Frequency Cable Circuits

regulating repeaters. Each of the different groups of secondary relays then controls directly the relays of all of the like regulating repeaters with which it is associated.

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

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

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

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

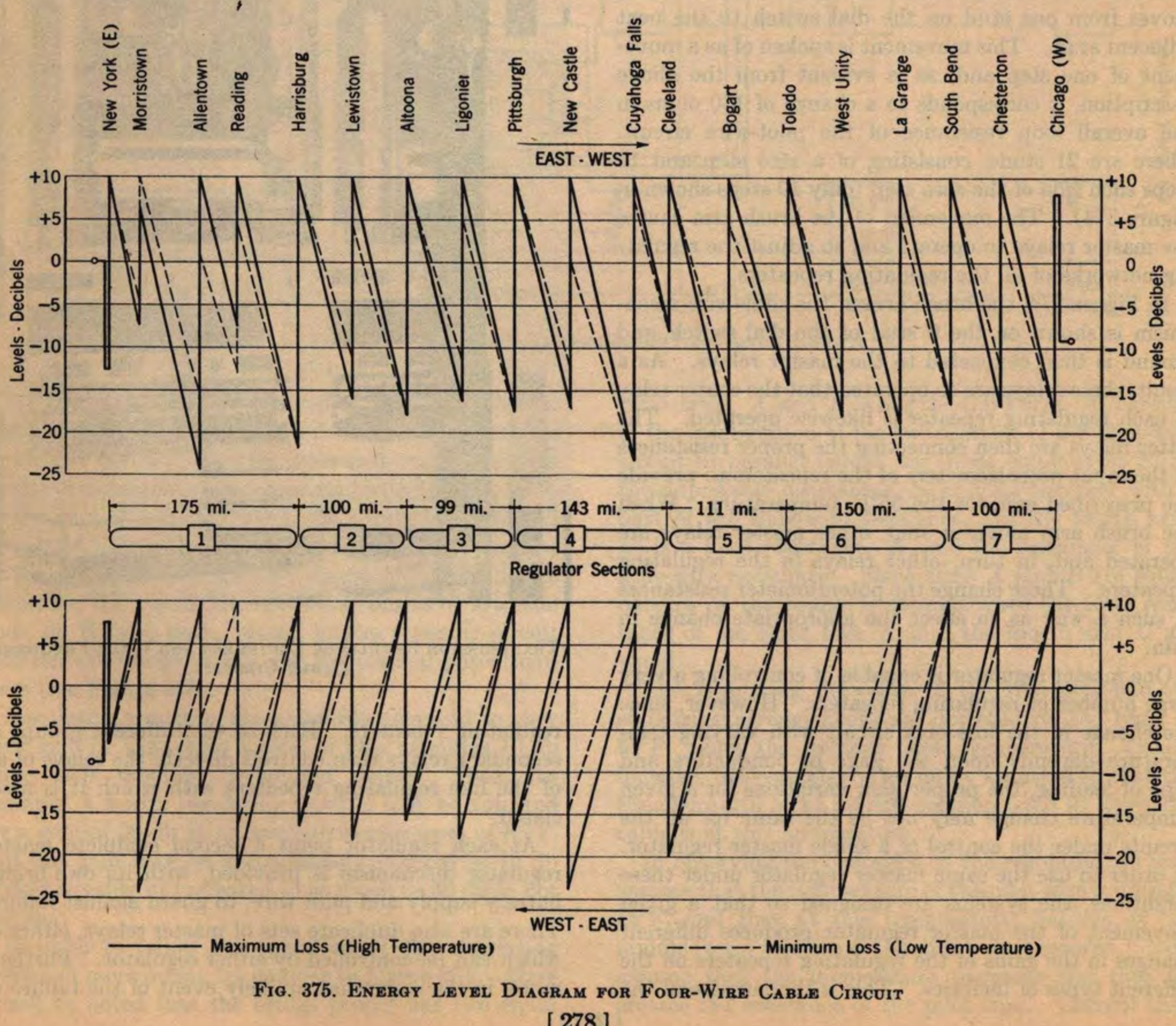


FIG. 375. ENERGY LEVEL DIAGRAM FOR FOUR-WIRE CABLE CIRCUIT

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

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

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

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

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

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

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

181. Echo Control

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

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

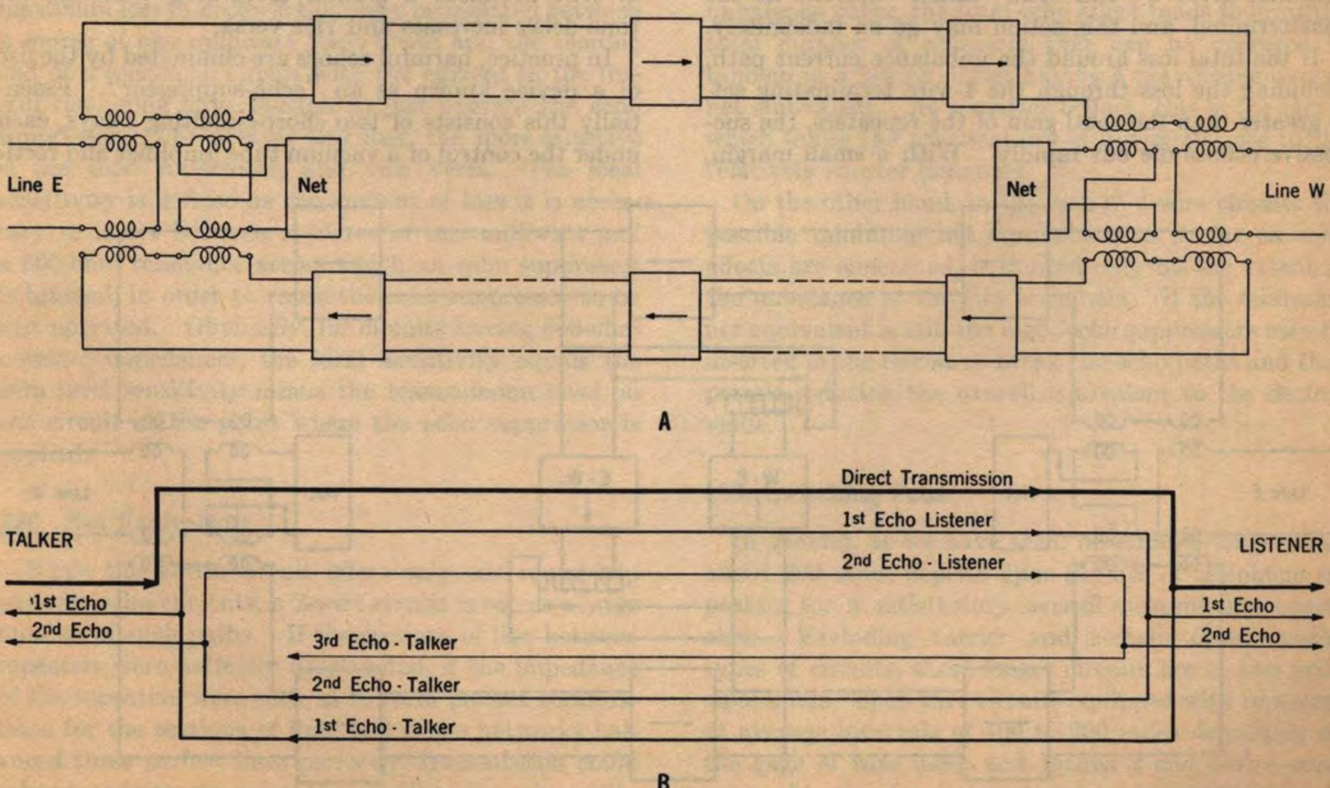


FIG. 376. ECHO PATHS IN FOUR-WIRE CIRCUIT

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

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

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

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

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

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

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

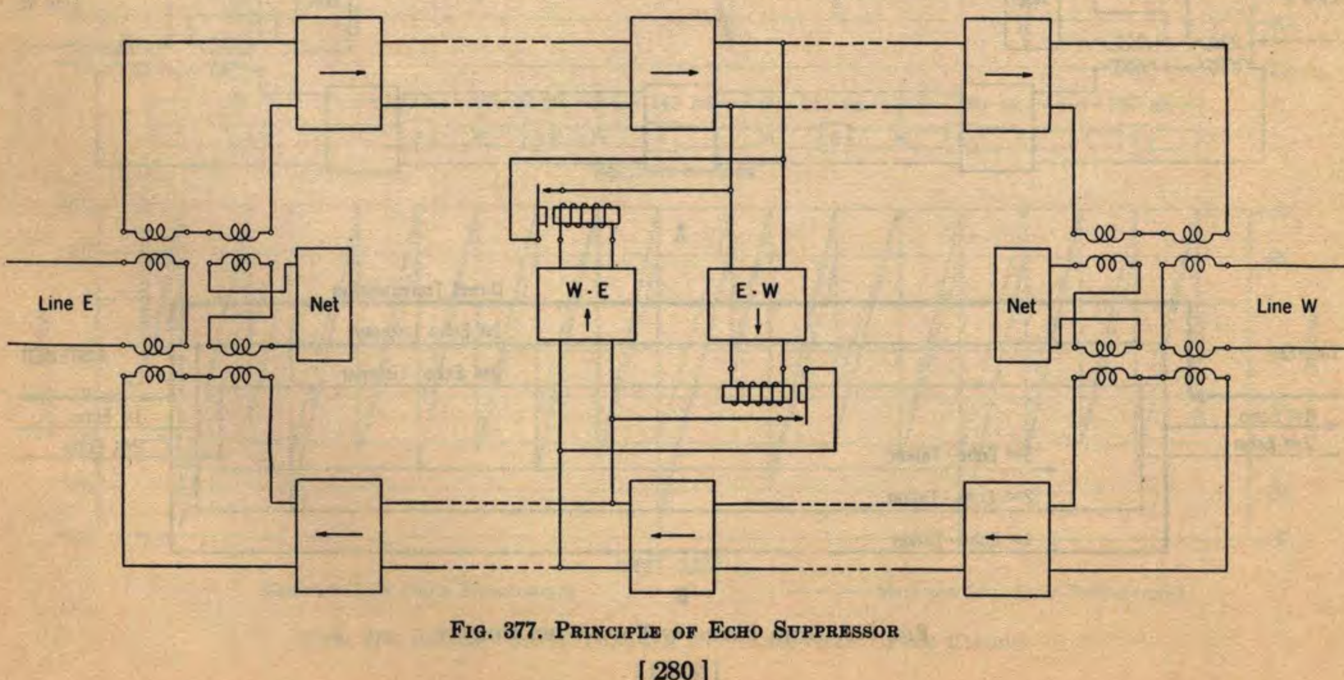


FIG. 377. PRINCIPLE OF ECHO SUPPRESSOR

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

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

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

182. Net Equivalents

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

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

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

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

183. Switching Pads

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

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

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

184. Signaling Systems

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

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

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

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

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

CHAPTER XXX

LONG DISTANCE TRANSMISSION SYSTEMS

CARRIER TELEPHONE AND TELEGRAPH CIRCUITS

185. Types of Carrier Systems

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

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

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

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

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

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

AUXILIARY CABLE CARRIER REPEATER STATION

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

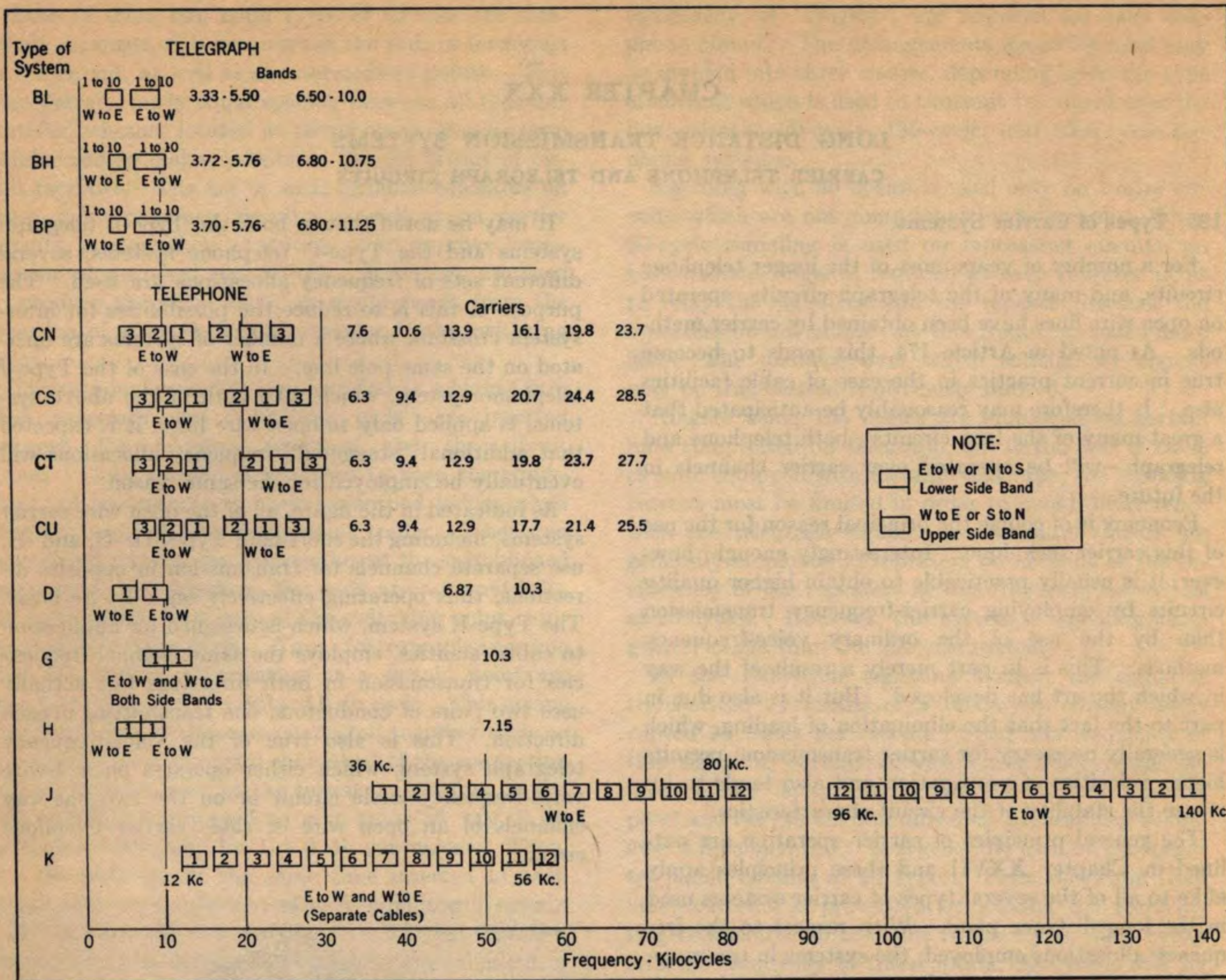


Fig. 378. Channel Frequency Allocations for Telephone and Telegraph Carrier Systems

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

186. "Low-Frequency" Carrier Systems

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

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

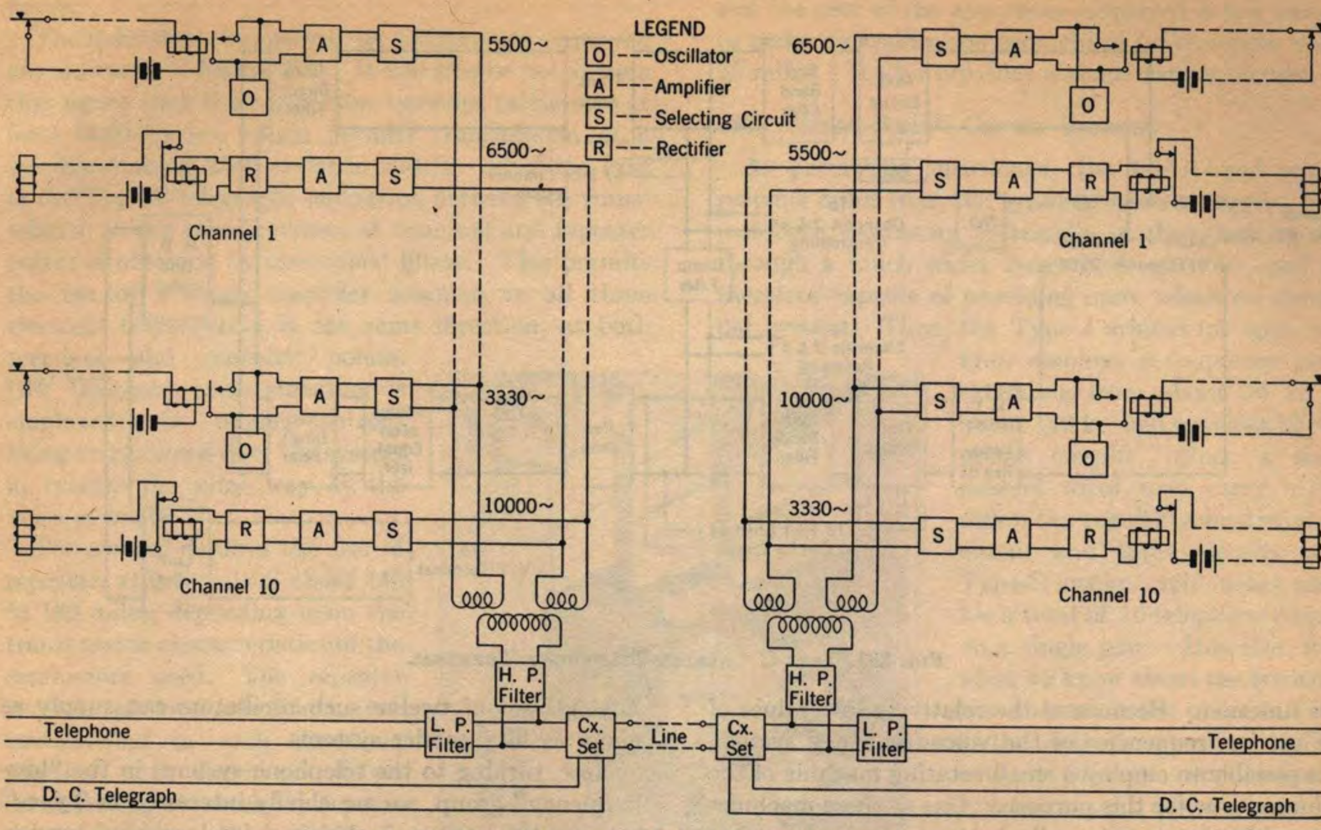


FIG. 379. HIGH-FREQUENCY CARRIER TELEGRAPH SYSTEM

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

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

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

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

VOICE-FREQUENCY CARRIER TELEGRAPH PANELS

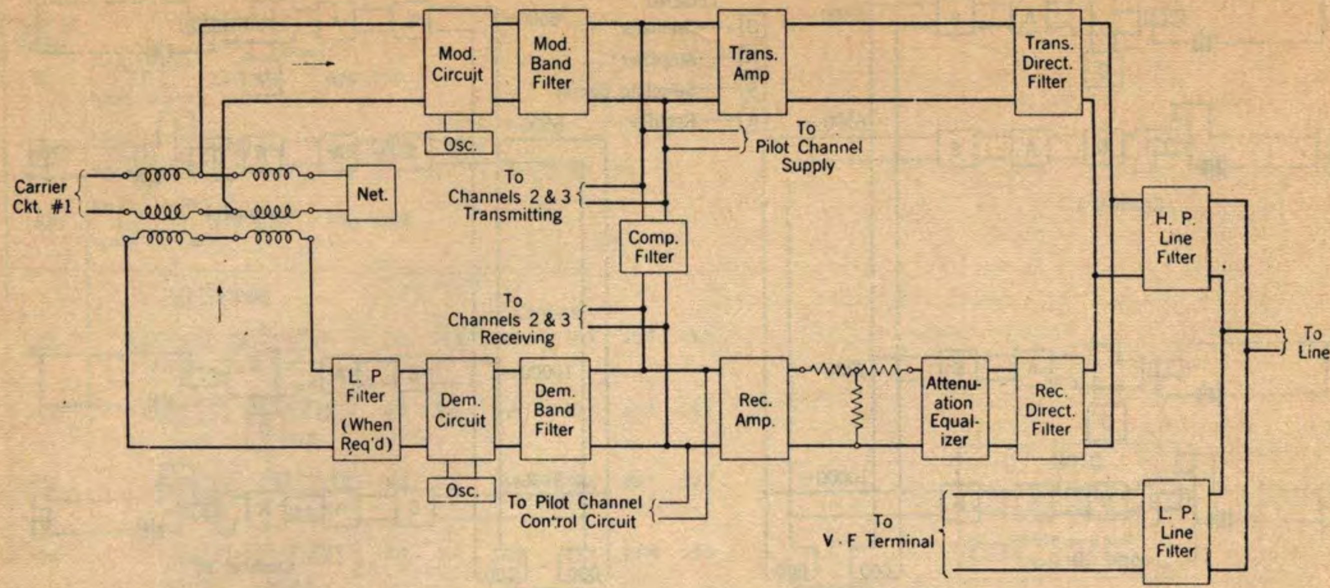


Fig. 380. Type-C Carrier Telephone Terminal

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

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

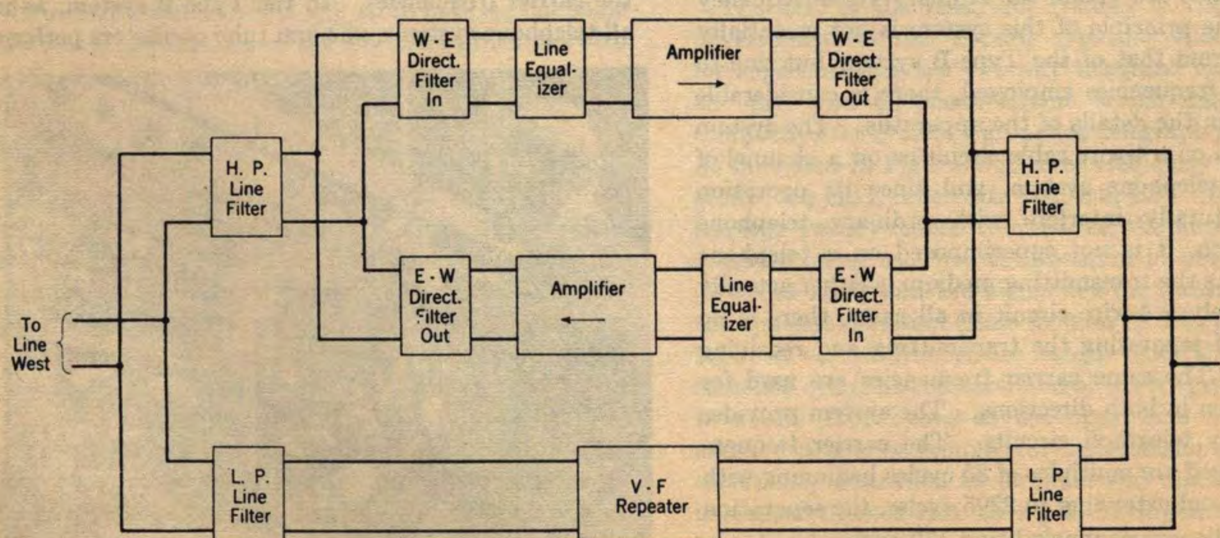


Fig. 381. Type-C Carrier Repeater Arrangement

tain satisfactory frequency synchronization at all times.

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

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

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

TERMINAL EQUIPMENT OF TYPE-C CARRIER SYSTEM

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

187. "Broad-Band" Carrier Systems

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

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

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

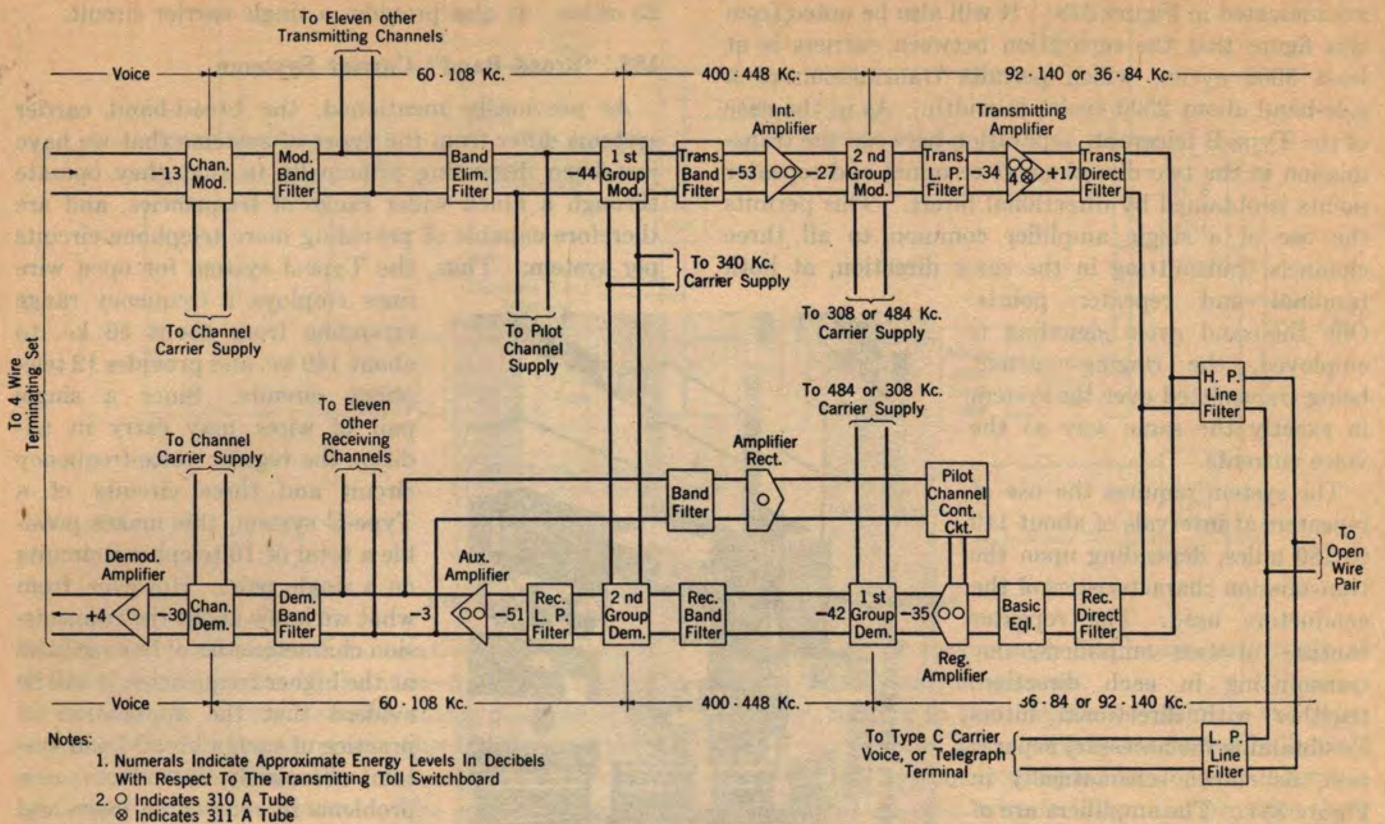


Fig. 382. Type-J Carrier Telephone Terminal

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

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

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

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

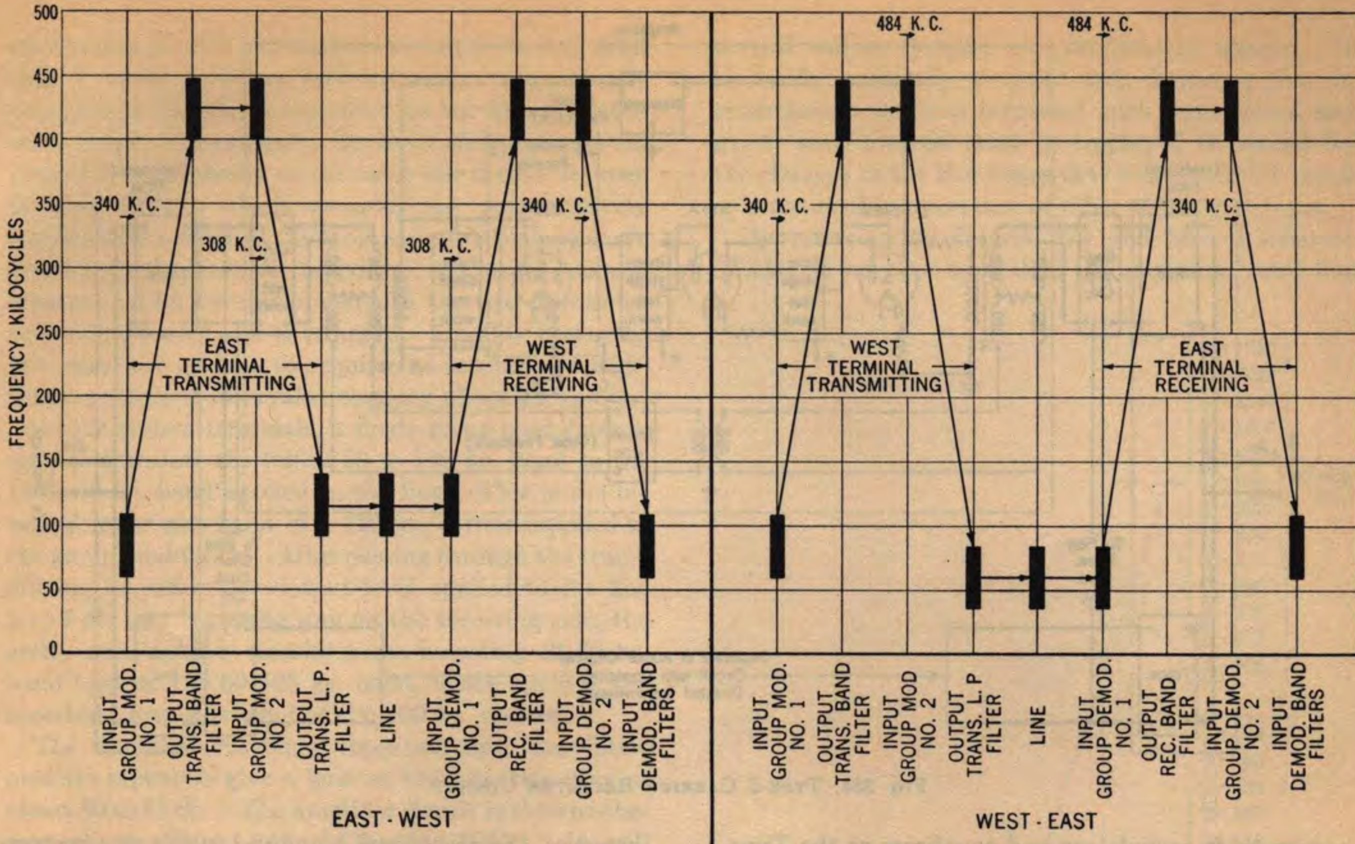


FIG. 383. FREQUENCY TRANSLATIONS IN TYPE-J CARRIER SYSTEMS

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

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

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

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

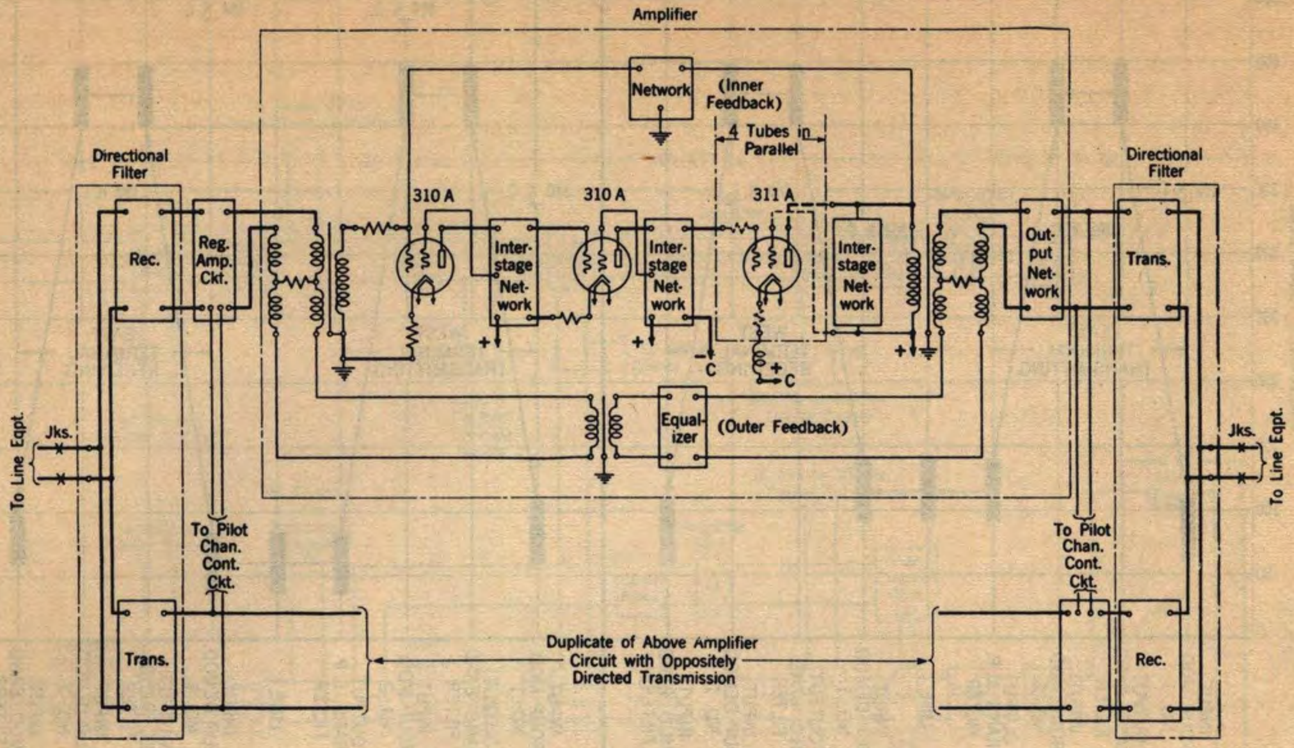


FIG. 384. TYPE-J CARRIER REPEATER CIRCUIT

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

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

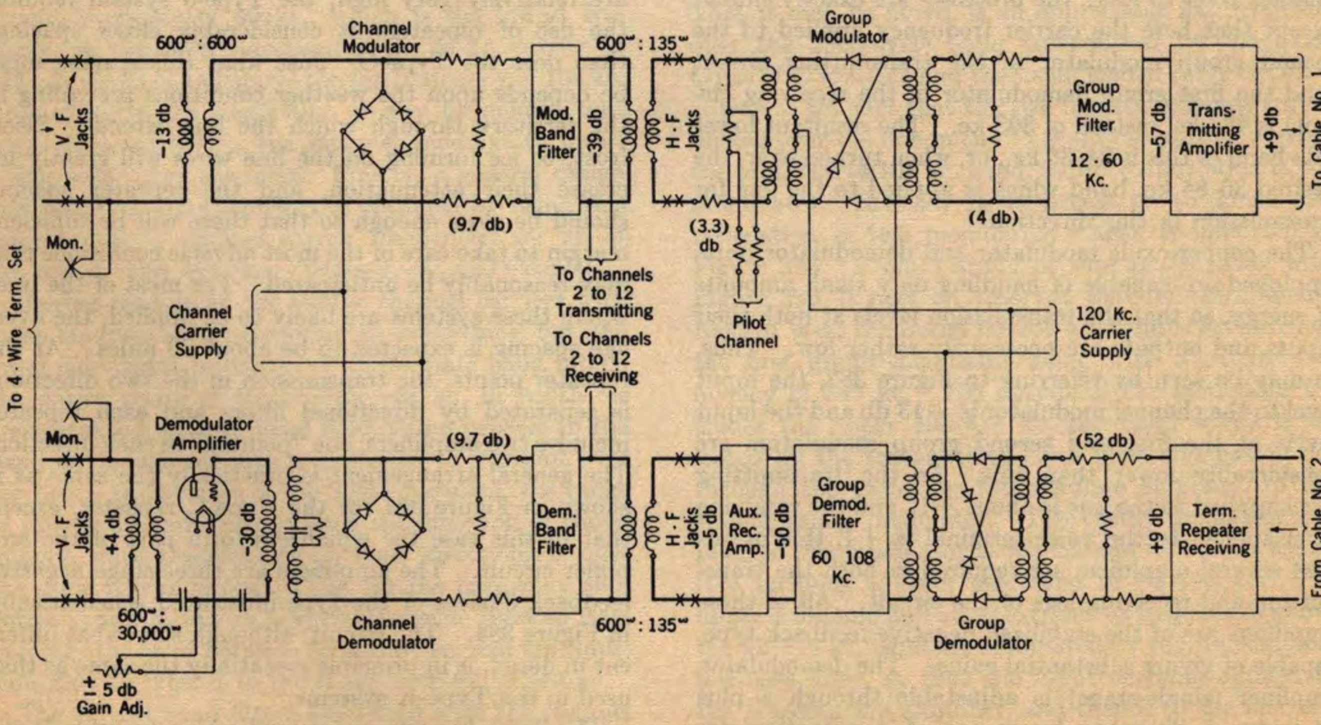


FIG. 385. TYPE-K CARRIER TELEPHONE TERMINAL

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

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

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

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

188. Transmission Regulation in Open Wire Carrier Systems

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

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

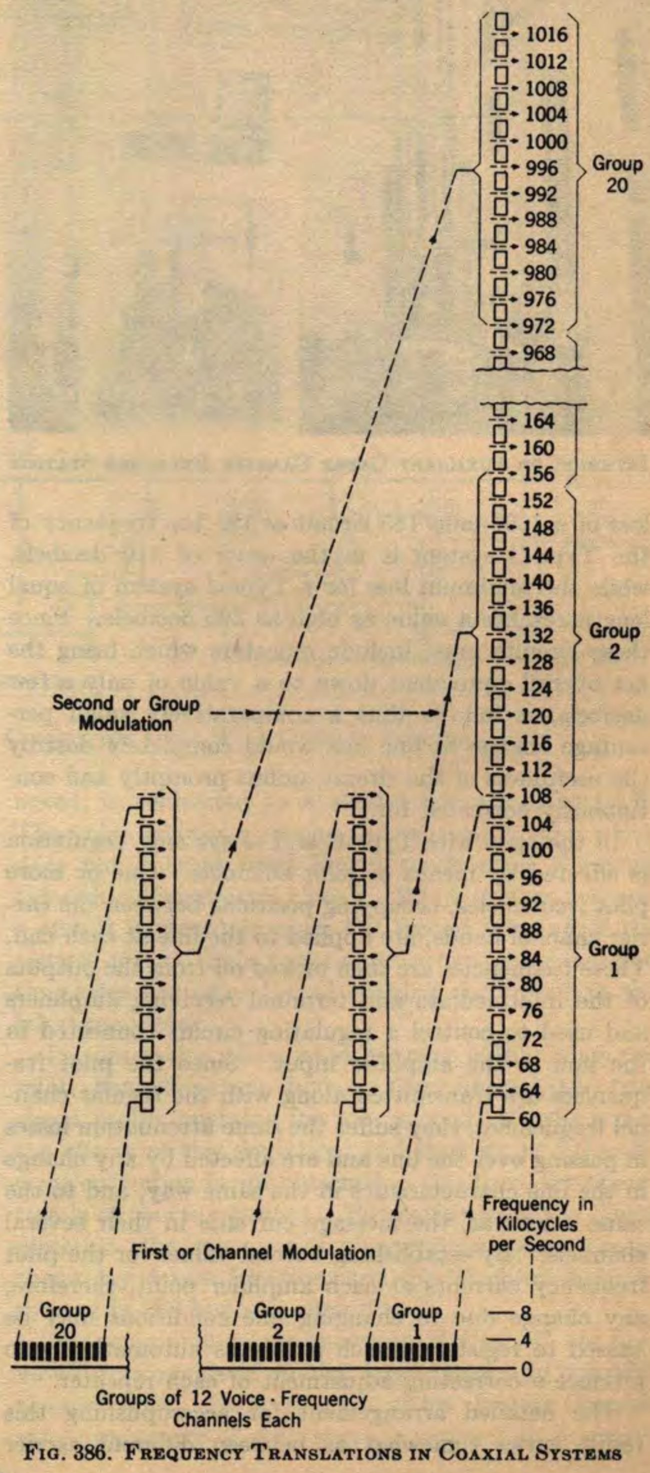
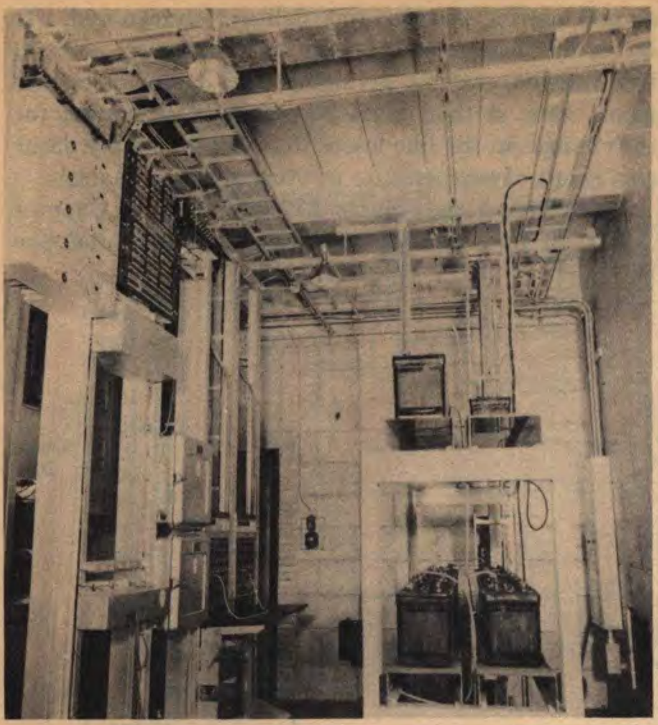


FIG. 386. FREQUENCY TRANSLATIONS IN COAXIAL SYSTEMS



INTERIOR OF AUXILIARY CABLE CARRIER REPEATER STATION

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

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

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

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

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

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

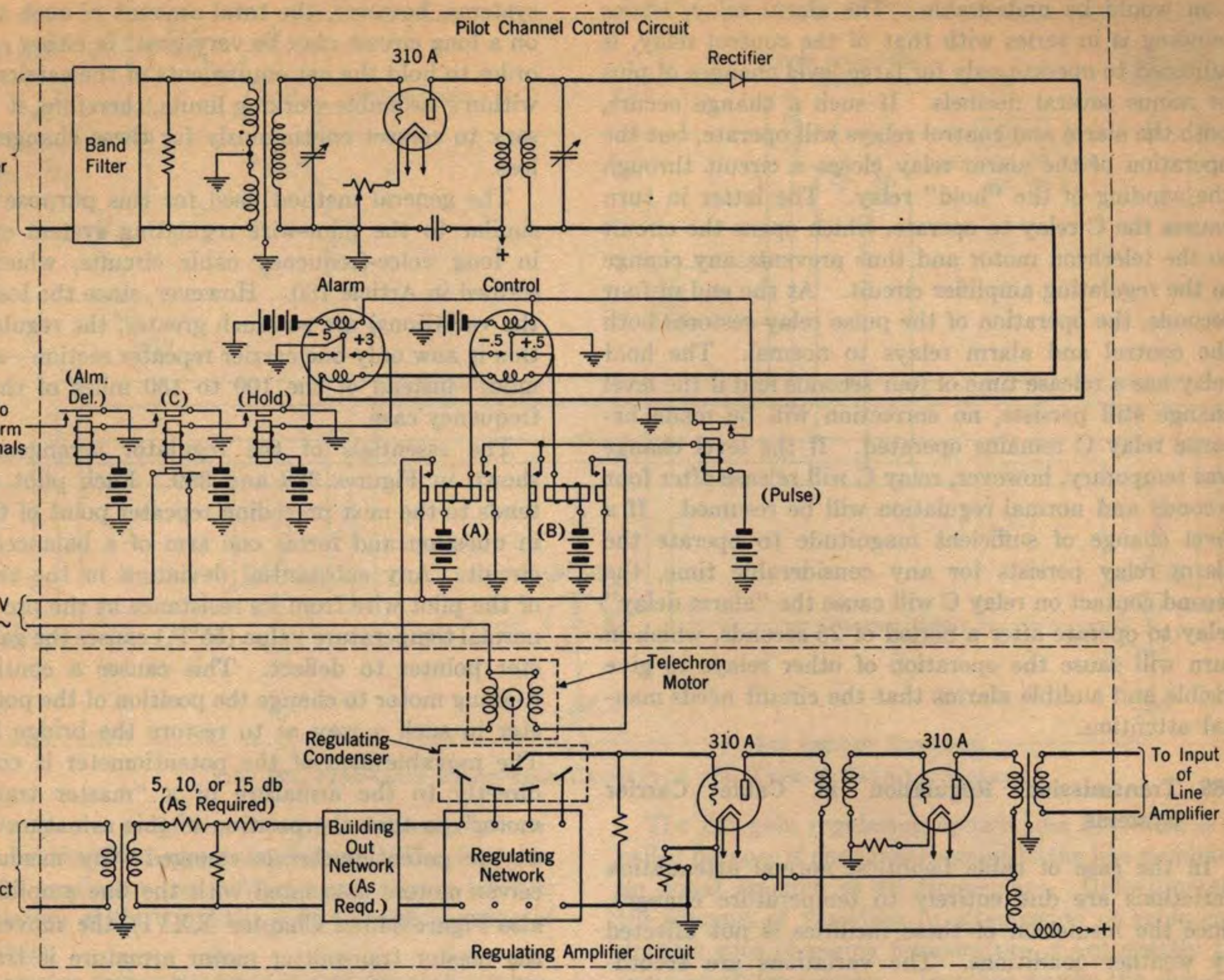


Fig. 387. Regulating Amplifier and Control Circuit for Open Wire Carrier Systems

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

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

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

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

189. Transmission Regulation in Cable Carrier Systems

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

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

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

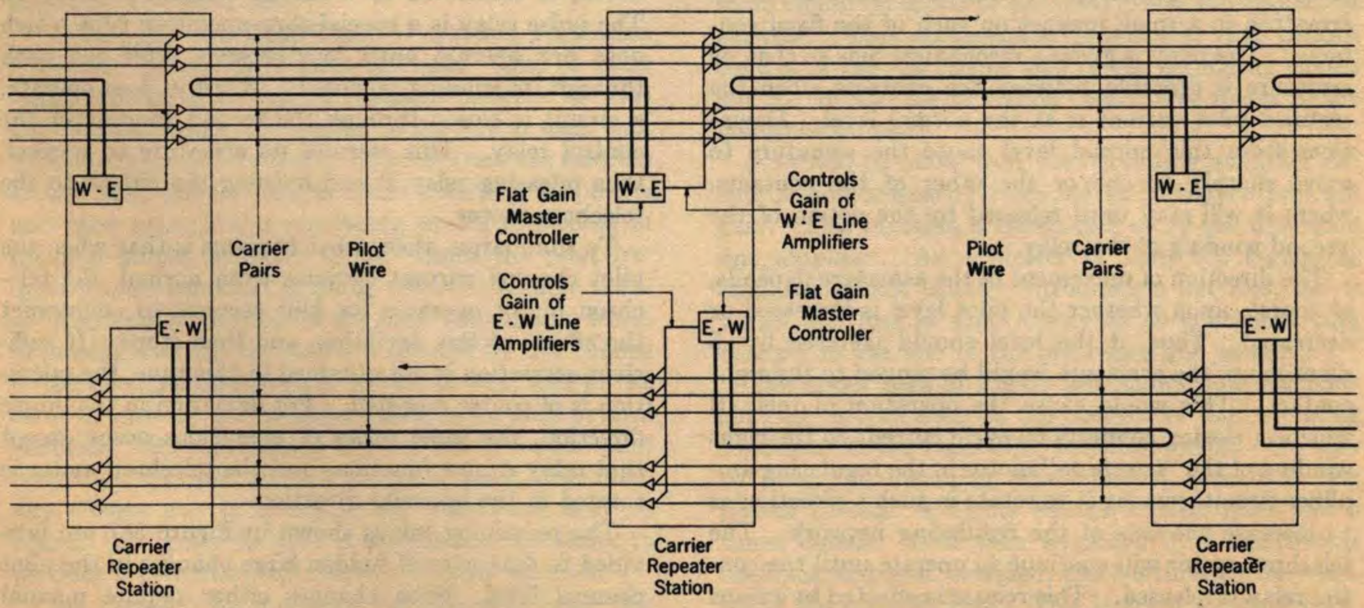


Fig. 388. Arrangement of Pilot Wires in Type-K Carrier System for Flat-Gain Regulation

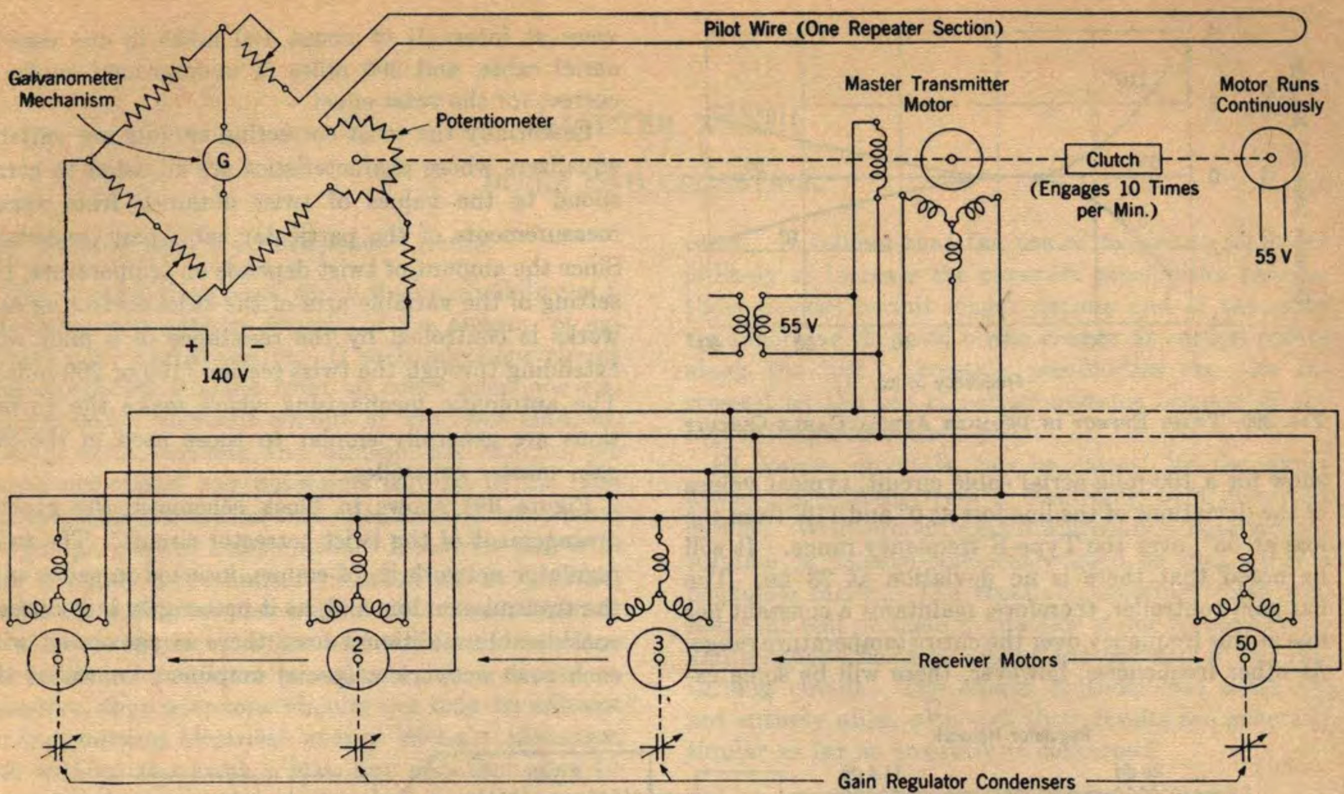


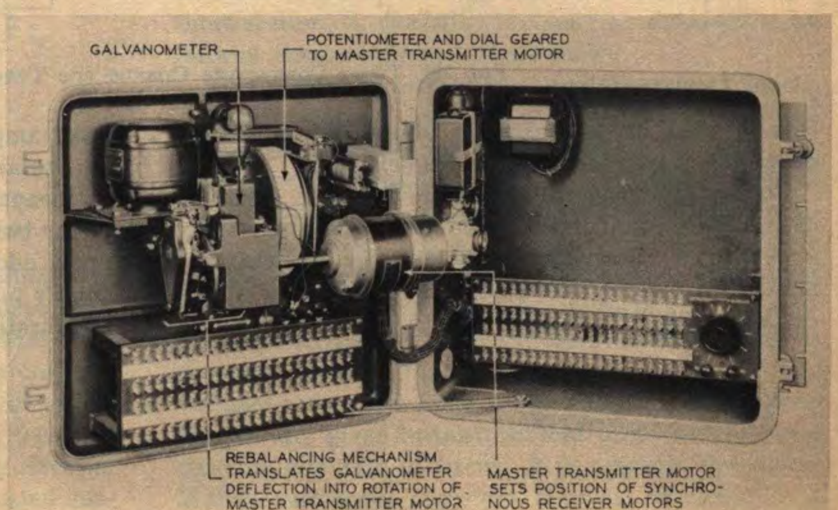
FIG. 389. FLAT-GAIN PILOT-WIRE REGULATOR FOR TYPE-K CARRIER SYSTEMS

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

The master transmitter motor and the receiver motors which it controls are merely an electrical means of gearing the bridge potentiometer shaft to the shafts of the flat gain regulator condensers in the line amplifiers. They are small synchronous motors whose armatures are normally stationary and are all held in exactly the same position when a common source of 60-cycle alternating current is applied to their field windings. However, if the armature of one motor (in this case the master transmitter motor) is moved by an external source, the armatures of all the other motors are moved at the same time by exactly the same amount, just as if all the armatures were connected directly together by a perfect mechanical gearing system.

A single master transmitter motor is capable of controlling 50 receiver motors and their associated regulator condensers. This means, of course, that one flat gain master controller is sufficient to take care of 50 carrier systems (600 circuits) in the same cable.

The flat gain regulating system just described is so called because it changes the gains of the line amplifiers an equal amount at all frequencies. Unfortunately, the amount of variation in attenuation of cable conductors with changing temperature is not exactly the same at all frequencies. In other words, the shape of the attenuation-frequency curve for cable circuits is usually slightly different at different temperatures. This effect is known as "twist". An idea of its magnitude may be had from the curves of Figure 390, which



TYPE-K CARRIER SYSTEM FLAT GAIN MASTER CONTROLLER

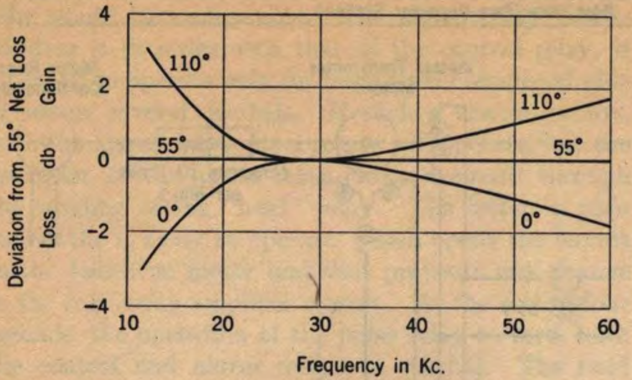


FIG. 390. TWIST EFFECT IN 100-MILE AERIAL CABLE CIRCUIT

show for a 100-mile aerial cable circuit, typical values of the deviations of the line loss at 0° and 110° from the loss at 55°, over the Type-K frequency range. It will be noted that there is no deviation at 28 kc. The flat gain controller, therefore, maintains a constant net loss at this frequency over the entire temperature range. At other frequencies, however, there will be some excess of gain at temperatures above 55° and some deficiency of gain at temperatures below 55°.

Compared in magnitude with the flat 28 kc. variation of loss with temperature, the twist variations may appear to be practically negligible. Nevertheless, they are too large to be allowed to accumulate over a long carrier system. It is necessary to employ special devices at intervals of about 100 miles in the case of aerial cable, and 200 miles in underground cable, to correct for the twist effect.

Essentially the twist correcting circuits are variable equalizers whose characteristics are adjusted to correspond to the values of twist obtained from actual measurements of the particular cable pair concerned. Since the amount of twist depends on temperature, the setting of the variable arm of the twist correcting networks is controlled by the resistance of a pilot wire extending through the twist section (100 or 200 miles). The automatic mechanisms which make the corrections are generally similar to those used in the flat gain master controller.

Figure 391 shows in block schematic the general arrangement of the twist corrector circuit. The twist regulator network is, of course, inserted in series with the transmission line and, as it necessarily introduces a considerable additional loss, there is associated with each such network a special amplifier, known as the twist amplifier, which has a fixed gain of 60 decibels. The total loss of the twist regulator circuit is adjusted by means of pads to be approximately equal to the gain of the twist amplifier so that in net effect the arrangement adds neither loss nor gain to the circuit, but simply provides the relatively small amount of special equalization required to overcome the twist effect.

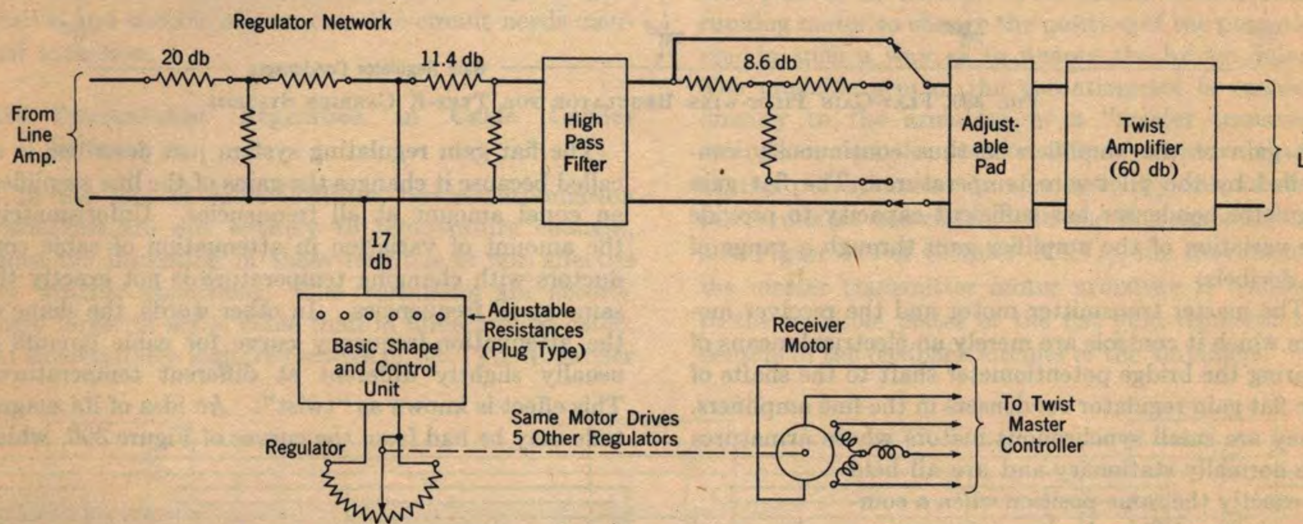


FIG. 391. TWIST CORRECTING CIRCUIT FOR TYPE-K CARRIER SYSTEM

CHAPTER XXXI

NOISE AND CROSSTALK

190. Induced Effects in Telephone Circuits

One of the factors upon which the intelligibility of a telephone conversation depends is the absence of excessive noise and crosstalk. If each telephone circuit was completely isolated from all other telephone circuits or other electrical circuits of whatever kind, including earth currents and atmospheric charges, we would not expect any potentials to exist in the telephone circuit other than those deliberately introduced for the purpose of transmission. However, this is in fact a purely hypothetical situation as, in practice, nearly every long telephone circuit is in close proximity to other telephone circuits, and sometimes to other electrical circuits such as power lines. It is necessary, therefore, that telephone circuits not only be efficient in transmitting electrical energy without distortion and without too great a loss, but also that they be protected against induced electrical currents coming from adjacent telephone circuits or from other electrical circuits.

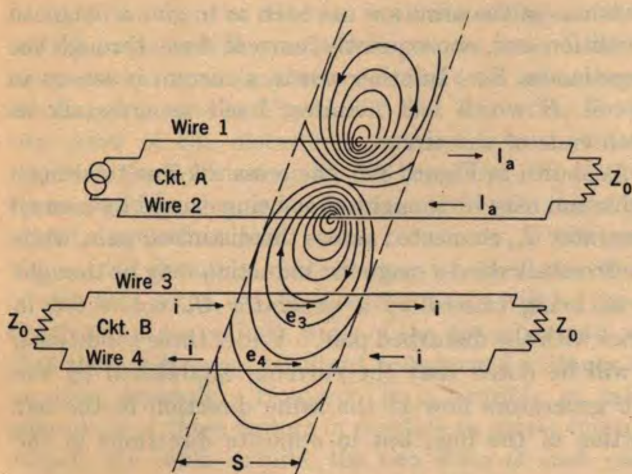


Figure 392

As a matter of fact, any two long paralleling telephone circuits that are not "balanced" against each other by means of transpositions, or otherwise, will crosstalk to such a degree as seriously to interfere with their practical use. Furthermore, because crosstalk is largely an inductive effect, its magnitude tends to increase with (1) the length of the paralleling circuits, (2) the strength (energy level) of the transmitted currents, and (3) the frequency of the transmitted currents. It follows that the use of telephone repeaters is likely to increase the crosstalk possibilities because these devices permit longer circuits and at the same time increase the level of the energy at certain points along the line. Crosstalk possibilities are also increased by the use of carrier systems because of the higher frequencies employed.

Of course, crosstalk can be caused by the direct leakage of current from a disturbing to a disturbed circuit. With properly maintained lines, however, insulation is usually sufficiently good to make this a negligible factor. The crosstalk coupling which presents the real problem in practice is due to the electric and magnetic fields set up by the currents in the disturbing circuit. The effects of these two fields are not entirely alike, although their results are generally similar as far as crosstalk is concerned.

191. Causes of Crosstalk

The effect of the magnetic field of one circuit on a second paralleling circuit is called "magnetic induction". Similarly, the effect on the second circuit of the electric field of the first circuit is called "electric" or "electrostatic induction". How magnetic induction causes crosstalk can be seen by referring to Figure 392. This shows the four wires of two telephone circuits running parallel to each other in the usual flat configuration employed on standard open wire lines. If we assume that an alternating voltage is applied to the left end of the disturbing circuit, *A*, which is made up of wires 1 and 2, the magnetic field existing about a short section, *S*, of this circuit may be represented by lines of magnetic induction as shown.

At a particular instant in the alternating cycle, the current in wires 1 and 2 may be represented by the equal and opposite vectors designated I_a. As I_a increases or decreases in value, the associated lines of magnetic induction will cut wires 3 and 4 of the paralleling telephone circuit, *B*. But with the relative spacing of the wires shown in the diagram, more lines will cut wire 3 than cut wire 4. Accordingly, the voltage induced by the magnetic field in wire 3 will be somewhat greater than that induced in wire 4. The voltages induced in both wires are in the same direction at any given instant, so that they tend to make currents circulate in circuit *B* in opposite directions. If they were equal, therefore, their net effect would be zero. But in so far as the induced voltage e_3 exceeds the induced voltage e_4, there is an unbalanced voltage, $e_3 - e_4$, tending to make a current circulate in circuit B. If the circuit is terminated at both ends in its characteristic impedance, Z_0, the current resulting from this unbalanced voltage induced in a short section of the circuit may be written as—

$$i = \frac{e_3 - e_4}{2Z_0} \tag{172}$$

The summation of the circulating currents caused by the voltages induced in each small section of the line may result in an appreciable current through both terminals of the circuit, which is due entirely to magnetic induction.

It should be noted that although the current in the disturbing circuit was considered in the above as being transmitted from the left end to the right end of the circuit, the crosstalk current induced in the disturbed circuit appears at both ends of the circuit. The crosstalk appearing at the left end of the disturbed circuit is known as **near-end crosstalk** and that appearing at the right end is known as **far-end crosstalk.**

Now turning our attention to electric induction, Figure 392 may also be used to show the equipotential lines of the electric field established about circuit A under the same conditions as in the previous example. This electric field will set up potentials on the surfaces of wires 3 and 4 and, with the configuration and spacing of the wires shown in the drawing, these potentials will not be equal. The resultant difference in potential between wires 3 and 4 will tend to cause crosstalk currents to flow to both ends of circuit B.

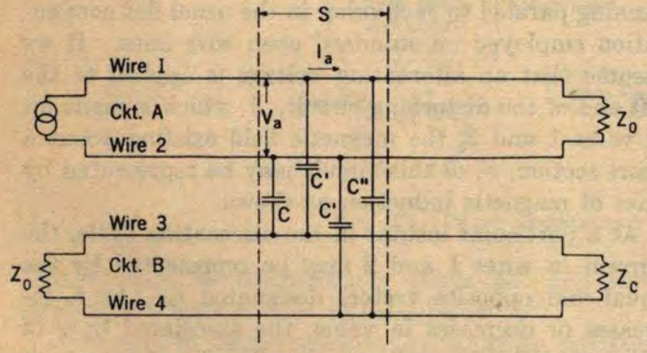


FIGURE 393

The crosstalk effect of electric induction may also be analyzed by consideration of the capacity relationships between the wires of the disturbing and disturbed circuits. Thus, referring to Figure 393, we know that in any unit length of the two circuits there is a definite capacity between wire 1 and wire 2 and between wire 3 and wire 4. Moreover, if the wires are equally spaced as shown, the separation between wires 2–3 is the same as that between wires 1–2 or 3–4,

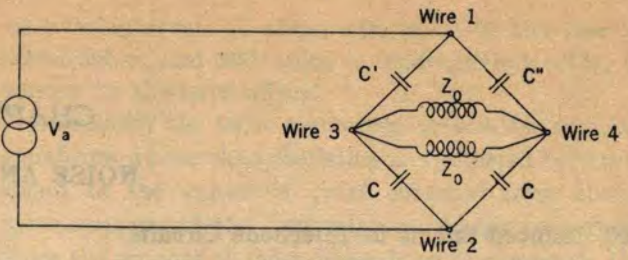


FIGURE 394

and there is therefore the same capacity between wires 2–3 as between the wires of either pair. This capacity is represented in the figure by the small condenser designated C. Similarly, the capacities between wire 1 and wire 3 and between wire 2 and wire 4 are designated by condensers whose capacity, C', is less than C because the separation between these wires is greater. There remains also the still smaller capacity between wire 1 and wire 4, which is indicated by the condenser, C''. Now if we assume an alternating current flowing in circuit A, there will be a difference of potential between wires 1 and 2, which will tend to cause small currents to flow through the various condenser connections to the wires of circuit B. The net effect can best be analyzed by redrawing the diagram of Figure 393 in the form of a Wheatstone bridge network as shown in Figure 394. A study of the capacity values of the arms of this bridge shows that the impedances of the arms are not such as to give a balanced condition and, consequently, current flows through the impedances Z_0. In other words, a current is set up in circuit B which will manifest itself as crosstalk at both ends of the circuit.

As shown in Figure 395, the crosstalk due to electric induction may be thought of as being caused by a small generator G_e, connected across the disturbed pair, while the crosstalk due to magnetic induction may be thought of as being caused by a generator G_m connected in series with the disturbed pair. Under these conditions, it will be noted that the currents established by the two generators flow in the same direction in the left portion of the line, but in opposite directions in the

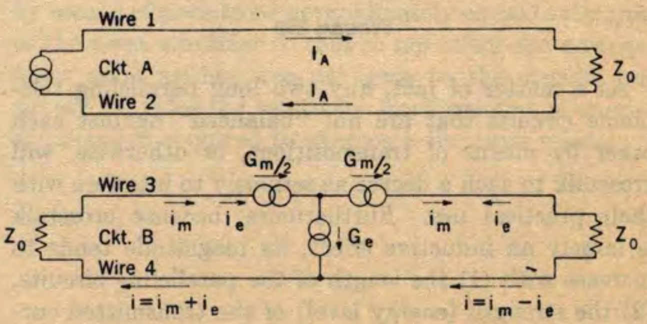


FIGURE 395

right portion. In other words, the crosstalk effects of magnetic and electric induction are additive in the case of near-end crosstalk, but opposed to each other in the case of far-end crosstalk.

192. Principles of Crosstalk Reduction

There are a number of possible ways of eliminating, or at least substantially reducing, the crosstalk induction discussed in the preceding article. One possibility is to arrange the paralleling wires in such a configuration that the effect of the field of one pair will be the same at both wires of the other pair, thus leaving no residual difference to cause currents in the second pair. Two possible ways to effect such a non-inductive configuration are shown in Figure 396. For a number of reasons, however, such wire configurations are not usually practicable.

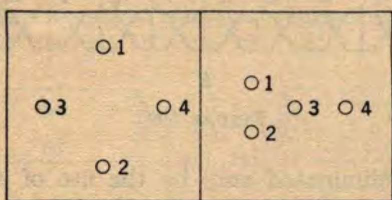


FIGURE 396

Another partial solution is to reduce the separation between the wires of either or both disturbing and disturbed pairs and, if practicable, at the same time to increase the separation between the two pairs themselves. A glance at Figure 392 will show that if the two wires of the disturbing pair are spaced closely together, the fields set up by the two wires will occupy approximately the same position and will therefore tend to neutralize each other. Similarly, if the two wires of the disturbed pair are close together, the effect of any field setup by the disturbing pair will be practically the same on both wires of the disturbed pair, so that there will be no resultant unbalanced voltages to produce crosstalk. However, it is possible to take advantage of these factors in practice to only a limited extent. In cable circuits, the two wires of each pair are close together, but so are the pairs themselves; in open wire circuits, there is considerable separation between pairs, but the two wires of a pair cannot be placed close together.

A third alternative is the use of **transpositions.** The principle involved here can be understood by referring to Figure 397, which shows the same four wires as were indicated in Figure 392. In this case, however, the relative pin positions of wires 1 and 2 are interchanged by means of a transposition in the middle of the short section, S. Under these conditions, it will be evident that when equal and opposite currents are flowing in the

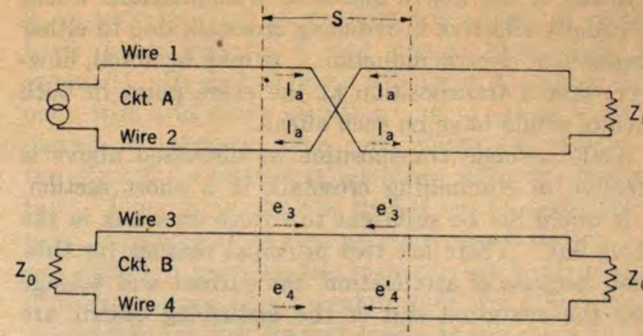


FIGURE 397

two wires of circuit A, the voltages induced in wires 3 and 4 will be in opposite directions on the two sides of the point where circuit A is transposed. Thus, as indicated in the figure, while e_3 is larger than e_4 and e_3' is larger than e_4', e_3 is exactly equal and opposite to e_3' and e_4 is exactly equal and opposite to e_4'. There is therefore no net voltage induced in either wire 3 or wire 4 and, consequently, no crosstalk from circuit A.

The same net effect would be obtained by inserting the transposition in the disturbed circuit B and leaving the wires of the disturbing circuit running straight through, as shown in Figure 398. In this case, the voltage induced in the wire nearer wire number 2 is broken into two equal parts represented by the vectors e_3 and e_3'. Similarly, the voltage induced in the wire farther away from wire number 2 is broken into two parts, e_4 and e_4'. But with the transposition as shown, voltage e_3 combines with voltage e_4' and voltage e_4 combines with voltage e_3'. The induced or crosstalk current in the section, therefore, is—

$$i = \frac{(e_3 + e_4') - (e_4 + e_3')}{2Z_0} \qquad (173)$$

But with the transposition in the center of the section as shown, it is obvious that—

$$e_3 + e_4' = e_4 + e_3'$$

The numerator of Equation (173) is therefore equal to zero and there is no resultant crosstalk.

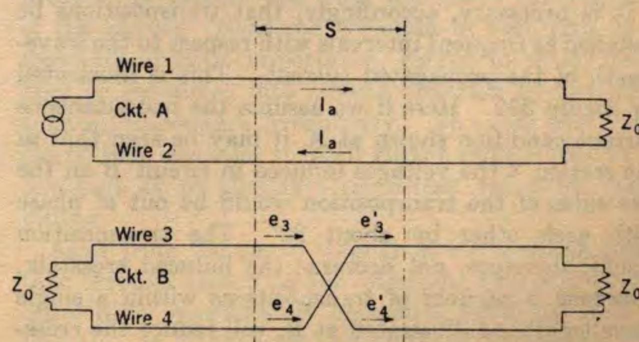


FIGURE 398

Either of the above discussed transpositions would be equally effective in reducing crosstalk due to either magnetic or electric induction. It may be noted, however, that a transposition at the same point in both circuits would have no such effect.

While a single transposition as discussed above is effective in eliminating crosstalk in a short section, *S*, it would not be sufficient to reduce crosstalk in the whole line. There are two principal reasons for this. First, because of attenuation, the current and voltage near the energized end of the disturbing circuit are many times as great as the current and voltage near the other end of the circuit. It could not be expected, therefore, that the induced crosstalk on the energized side of the transposition would be neutralized by the weaker crosstalk induced on the other side of the transposition. As a matter of fact, even in a short section, the transposition will not completely eliminate **near-end** crosstalk because the currents coming back from the far side of the transposition are necessarily attenuated somewhat more than the currents coming back from the near side of the transposition. On the other hand, the transposition is completely effective in the case of **far-end** crosstalk because the slightly higher currents induced on the energized side of the transposition are attenuated more in reaching the far end of the circuit than are the currents induced on the far side of the transposition.

The second reason why a single transposition is not effective in reducing crosstalk to the desired minimum is the phase change of the transmitted currents. In a long circuit, several wave-lengths may be included in the propagation of a voice current from one end to the other. Since crosstalk is an induced effect, its instantaneous value in any small section *S* depends upon the position of *S* with respect to the cycle of current in the disturbing circuit. If *S* is so located that the current or voltage in it has a maximum value, either positive or negative, we cannot expect the crosstalk induced here to be neutralized by the crosstalk in some other similar section, which is located at a point in the line where the voltage or current has a value nearly zero at the same instant.

It is necessary, accordingly, that transpositions be installed at frequent intervals with respect to the wave-length of the propagated current. This is illustrated by Figure 399. Here if we assume the instantaneous current condition shown at A, it may be seen that in the section *S* the voltages induced in circuit *B* on the two sides of the transposition would be out of phase with each other by about 90° The transposition would, therefore, not decrease the induced crosstalk. **However,** a number of transpositions within a single wave-length, as illustrated at B, will reduce the crosstalk to practical limits, although the crosstalk could be

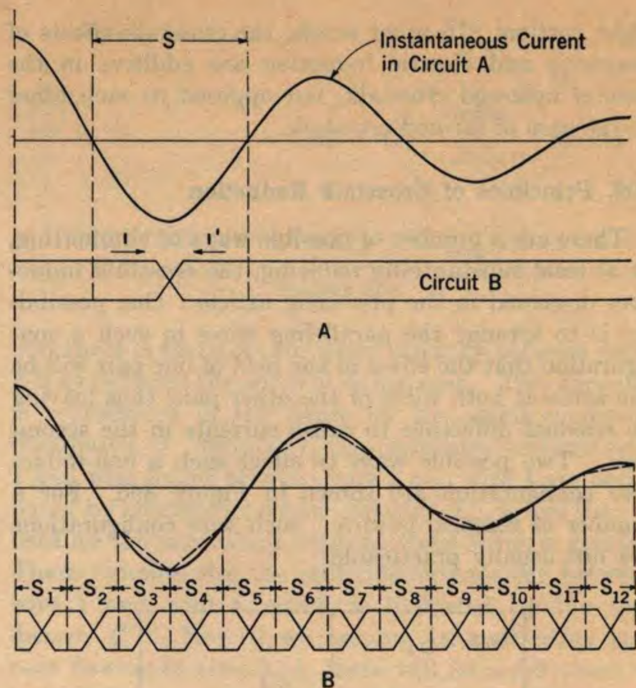


FIGURE 399

completely eliminated only by the use of an infinite number of transpositions. The dotted curve would then become identical with the solid curve.

For voice-frequency transmission, where the frequencies are relatively low and the wave-lengths correspondingly long, it is not difficult to obtain a sufficiently good approximation to this condition. Where high-frequency carrier systems are used, on the other hand, the wave-lengths are so short as to require very closely spaced transpositions. In open wire lines, spacings as close as every second pole are used in practice where the wires are carrying frequencies up to 140 kc. The degree of effectiveness of such closely spaced transpositions is of course dependent upon accurate spacing of the poles themselves, and is finally limited in practice by economic rather than theoretical factors.

There is another fundamental consideration in connection with crosstalk that is of the first order of importance. In what has been said above regarding transpositions, it has been tacitly assumed that the four wires which we were considering were of the same gage and material—and particularly, that the two wires of each pair were electrically identical. The latter is of course ordinarily the case in practice. However, there is always a possibility in open wire lines of the series resistance, or the insulation resistance, of one wire of a pair being slightly different from that of its mate due to imperfect joints, defective insulators, etc. When this occurs, the assumption made in connection with Figures 397 and 398 that the currents flowing in the two wires of the disturbing pair, or the induced currents in the two wires of the disturbed pair, were exactly equal and opposite is no longer true.

Thus, an unbalanced series resistance in wire 1 of Figure 397 would have the effect of reducing the current in wire 1 as compared to the current in wire 2 in an amount depending upon the value of the resistance. Under these circumstances, it is clear that the opposing currents induced in wires 3 and 4 would no longer be exactly equal, and there would therefore be some crosstalk despite the transposition. A resistance or leakage unbalanced in the 3–4 pair would cause a similar result. In this case, any resistance unbalance would cause one of the induced currents to be smaller than the other, with the result that some crosstalk current would flow through the terminal. It is extremely important, accordingly, that the two wires of every talking pair be so constructed and maintained as always to have identical electrical characteristics.

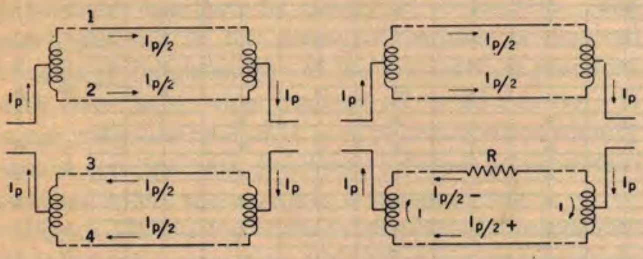


Figure 400

Resistance unbalance is particularly important when two pairs are used to create a phantom circuit. Here, under ideal conditions, exactly half of the phantom circuit current flows in each of the four wires, as shown in Figure 400. The phantom currents in wires 1 and 2 are then equal and in the same direction, and they therefore cause no current to flow through the terminal stations connected to these wires. The same is true of the side circuit made up of wires 3 and 4. A bad joint, or resistance unbalance of any other kind, in any one of the four wires will reduce the current in that wire somewhat. As a result, the phantom currents in the two wires of the pair concerned will no longer be equal and an unbalance current will flow through the side circuit terminal. In other words, the phantom circuit will crosstalk into the side circuit, or vice versa. The effect of resistance unbalance in this situation is ordinarily much more serious than its effect on crosstalk between two side circuits or two non-phantomed circuits.

193. Crosstalk Reduction Practices

In considering practical methods for keeping the crosstalk in long toll circuits at a reasonable minimum, it is desirable first to consider the effects of certain basic design features of long circuits with respect to crosstalk. In general, these will apply equally to both open wire and cable facilities, and at either voice or carrier frequencies. One such important feature is the effect of the location of telephone repeaters on crosstalk. Thus, it is obvious that if two circuits are in close proximity at a point near a repeater station, and one circuit is carrying the high current levels coming from the output of a repeater while the other circuit is carrying the low current levels approaching the input of a repeater, the tendency of the first circuit to interfere with the second circuit is very great. The very small percentage of the current in the first circuit which may be induced into the second circuit will be amplified by the repeater on that circuit along with, and to the same degree as, the normal transmission. The best practical remedy for this condition, of course, is to avoid such situations by keeping circuits carrying high level energy away from low level circuits as much as possible. Where such physical separation between circuits is not feasible, differences in energy level between adjacent circuits can frequently be minimized by proper adjustment of repeater gains when the circuit is designed.

Another basic element of circuit design is that in most of the longer voice-frequency cable circuits and in all carrier circuits, the effect of near-end crosstalk is minimized by the use of separate paths for transmission in the two directions. In cable circuits, the wires carrying the transmission in the two directions are physically separated as much as possible by placing them in different layers or segments of the cable; or, in the special case of cable carrier circuits, in different cables. An equally effective separation is obtained in open wire carrier circuits by using entirely different bands of frequencies for transmission in the two directions.

Furthermore, any near-end crosstalk occurring in spite of these physical separations is returned on the disturbed circuit to the output of an amplifier. Since the amplifier is a one-way device, the crosstalk can proceed no farther and does not reach the terminal of the circuit. Near-end crosstalk in such circuits is therefore of little importance, except in so far as it may be converted into far-end crosstalk by reflection from an impedance irregularity. To avoid this latter effect, it is essential that all circuit impedances be so matched as to eliminate important reflection possibilities.

Aside from the above techniques for avoiding crosstalk through circuit design methods, practical procedures differ considerably depending upon the type of facility. It is desirable, accordingly, to analyze the problem separately for open wire and cable facilities.

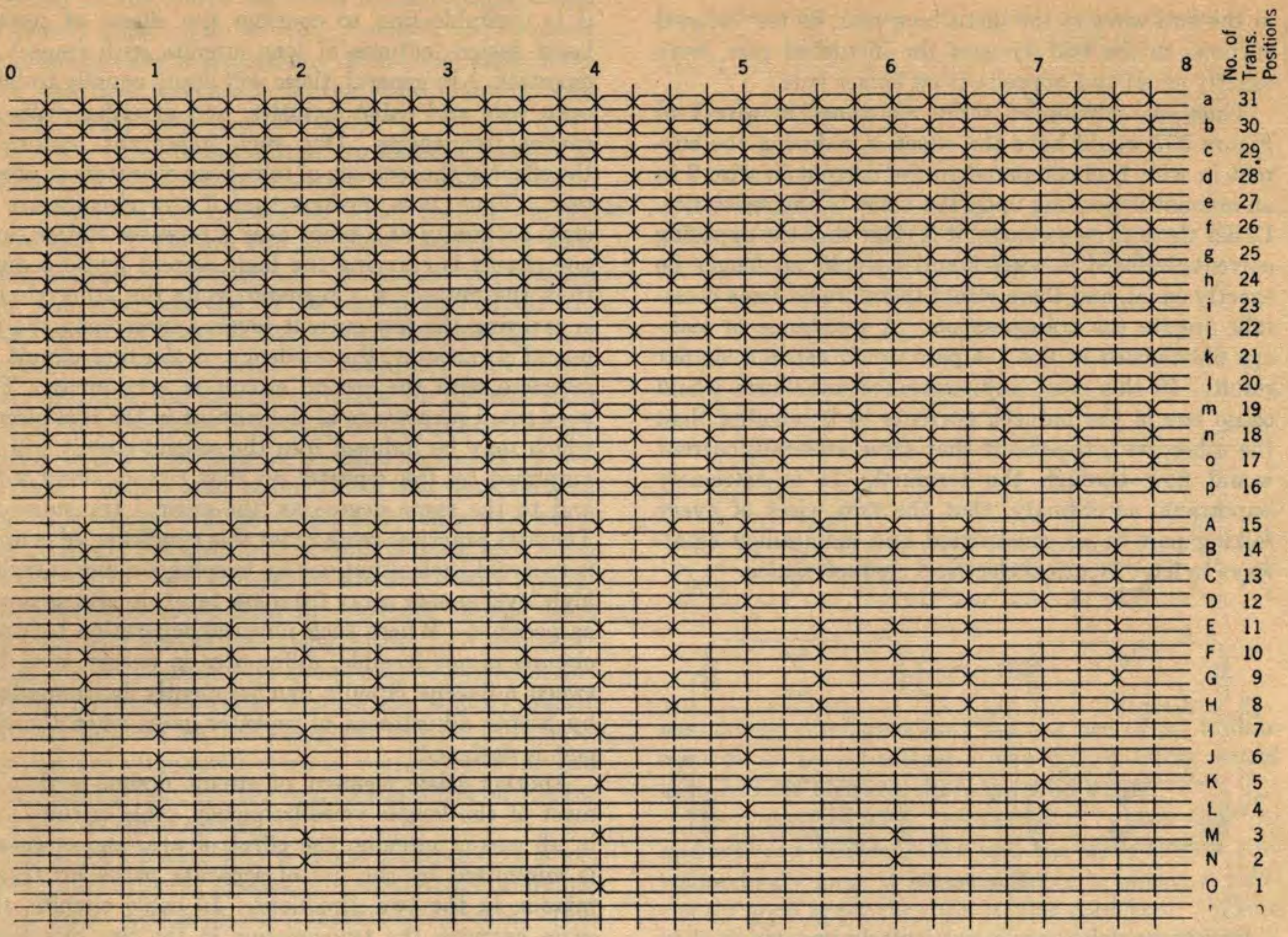


FIG. 401. STANDARD TRANSPOSITION CODE

In the case of open wire lines, crosstalk reduction depends upon three principal factors—namely, wire configuration on the poles, transpositions, and resistance balance. Resistance balance is primarily a question of maintenance and ordinarily presents no great difficulty. The use of high-frequency carrier systems, with their much greater crosstalk possibilities, has led to the development of new configurations of open wire lines in which the wires of individual pairs are spaced closer together and the pairs are spaced farther apart. One standard pole-head configuration of this kind is illustrated in Figure 276 of Chapter XXII, where it may be noted that the separation of the wires of each pair is 8 inches and the horizontal separation on the crossarm between any two wires of different pairs is at least 26 inches. The vertical separation between crossarms is 36 inches. Six inch spacing between the wires of a pair has also been used in a limited number of cases, but the danger of such closely spaced wires swinging together in the spans tends to restrict this practice to situations where weather conditions are particularly favorable.

The basic principle of transpositions was outlined in the preceding article. It was noted there that a large number of transpositions was needed in any long section of line to reduce crosstalk to the desired extent. In the entire discussion, moreover, only two pairs were considered. In practice an open wire line usually carries many more wires than this, and obviously there are crosstalk possibilities between any two pairs on such a line. These possibilities are greater between the pairs that are adjacent to each other, but all of the other possibilities are sufficiently large that they must be taken into consideration in designing a transposition system for the line. A practical system must also guard against crosstalk between side and phantom circuits and between the phantoms themselves, when such circuits are used.

There is still another extremely important factor which has not been considered up to this time. This is the possibility of crosstalk from one circuit to another via a third circuit. In a line carrying many circuits, there are a large number of these tertiary circuits via which crosstalk might be carried from any one pair to any other pair. Even the hypothetical line that we considered in the first place, carrying only four wires, has two such tertiary circuits. These are the phantom circuit, made up of the two wires of one pair transmitting in one direction and the two wires of the other pair transmitting in the opposite direction; and the "ghost" circuit, made up of the four wires acting as one side of a circuit, with a ground return. (Note that these circuits exist as tertiary crosstalk paths regardless of whether a working phantom circuit is actually applied to the four wires.) Needless to say, the presence of these tertiary circuits in a line complicates the problem of designing effective transposition systems. So much so, indeed, that no attempt can be made here to analyze this problem in detail.

Transposition systems for open wire lines are designed for unit lengths ranging from a few hundred feet to some six or eight miles. The purpose of the design is to approach as closely as possible to a complete crosstalk balance in each such unit section. Any number of sections can then be connected in tandem. The non-uniformity in the length of sections is required because of discontinuities in the line, such as junctions with other lines, wires dropped off or added, etc. It is naturally desirable that such points of discontinuity should coincide with junctions between transposition sections, where the crosstalk is balanced out.

Figure 401 illustrates the fundamental transposition designs used in a section with 32 or less transposition poles. These fundamental types are frequently extended to include 64, 128, or even as many as 256 transposition poles per section, on lines to which high-frequency carrier systems are applied.

Physically, there are two standard methods for effecting transpositions between wires on pole lines. These, known as "point type" and "drop bracket" transpositions, are shown respectively in Figures 402 and 403. The former (point type) is widely used on lines carrying carrier systems because it does not change the configuration of the wires in the adjacent spans, as does the drop bracket type. Where very high frequencies are used, as in the Type-J carrier system, this becomes extremely important. In fact, the sensitivity of these carrier systems to crosstalk is so great that every possible effort has to be made to avoid even slight deviations in the amount of sag of the wires in the spans between poles.

Turning now to cable, the most striking feature of this type of facility with respect to crosstalk is that the conductors are crowded closely together. This is particularly true of the two wires of each circuit pair, which are separated by only thin coatings of paper insulation. As we have already seen, this close spacing of the two wires of a pair in which equal and opposite currents are flowing tends to minimize the external

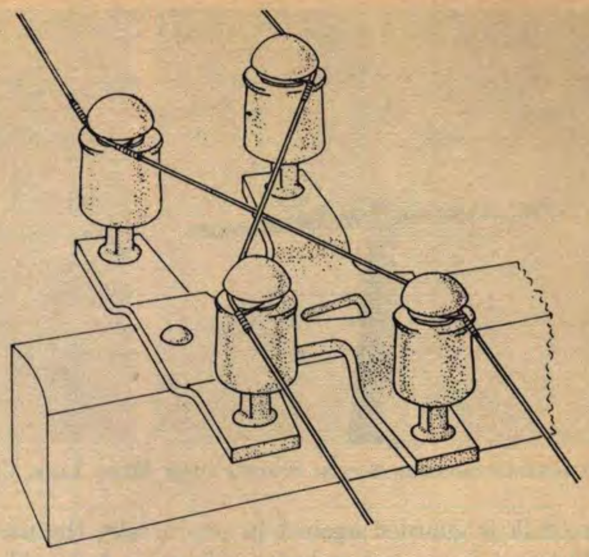


Fig. 402. Point-Type Transposition

effect of the electromagnetic field of the pair. Moreover, in the process of manufacture, the cable conductors are very thoroughly transposed by twisting the two wires of each pair together, by twisting the two pairs of each group of four wires together to form quads, and by spiralling the quads in opposite directions about the cable core. Cables are also so manufactured and installed that their conductors are practically free from series resistance unbalances or insulation leakages. On the other hand, the close spacing of many circuits within the cable sheath, as well as their proximity to the sheath itself, offsets the above advantages to a considerable extent.

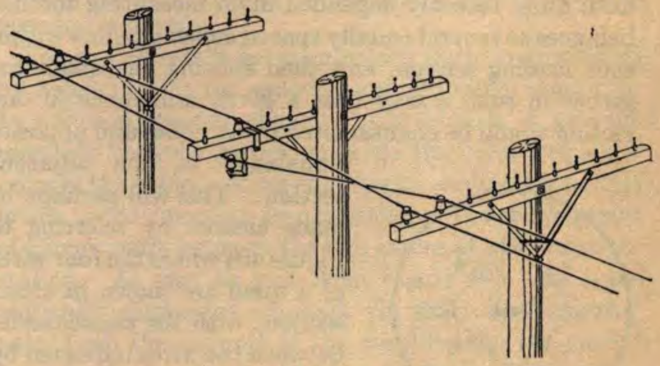
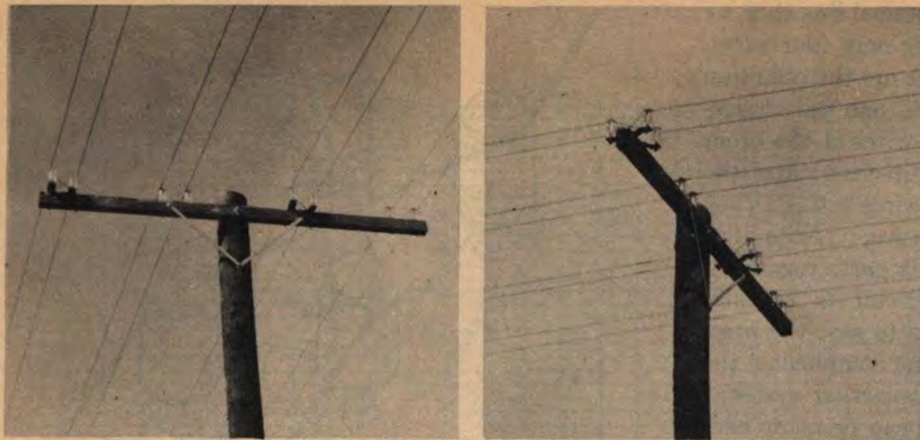


Fig. 403. Drop-Bracket Transposition

At voice frequencies, magnetic induction (inductive coupling) between circuits in a cable is normally so small as to be of relatively little importance in creating crosstalk. The same cannot be said of electric induction (capacity coupling). Despite the most careful manufacturing methods, the capacity unbalances between cable conductors usually remain large enough to cause objectionable crosstalk in long circuits. This



Transpositions in 8-inch Spaced Open Wire Line Using Point-Type Fixtures

crosstalk is guarded against in practice by the use of additional balancing techniques when a toll cable is installed.

Voice-frequency crosstalk between circuits in different quads of a cable can be reduced to a satisfactory minimum at that time by splicing the successive lengths of cable in a more or less random manner such that no two quads are adjacent for more than a small part of their total length. This technique of course has no effect upon the crosstalk between circuits in the same quad. To reduce this crosstalk, it is necessary to measure the capacity unbalances of each quad at the time of installation and then to correct such unbalances as are found large enough to be likely to cause serious crosstalk.

There are two principal methods of effecting this latter correction. The method most extensively used until quite recently depended upon measuring the unbalances at several equally spaced splicing points within each loading section, and then splicing the quads together in such a way that a given unbalance in one section would be counteracted by an equal and opposite unbalance in the adjacent section. This will perhaps be made clearer by referring to Figure 404 where the four wires of a quad are shown in cross-section, with the capacitances between the wires indicated by small condensers. The wires marked 1 and 2 form one pair of the quad, and the wires marked 3 and 4 the other. (The capacities between the pairs themselves are not shown because they have no effect on crosstalk.) The ideal condition in such a quad is that the values of all four capacities A, B, C, and D shall be equal, and that capacity E shall equal capacity F and capacity G shall equal capacity H. In this case there is no unbalance within the quad and no crosstalk. However, if it is found, for example, that capacity A in a certain quad of one section of the cable is too low, this quad can be spliced to a quad in an adjacent section of equal length in which capacity A is too high by an approximately equal amount. The net unbalance of the connected quad over the two sections will thus be made to approach zero.

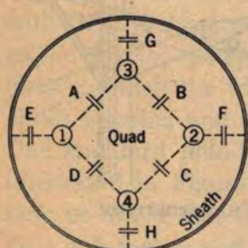
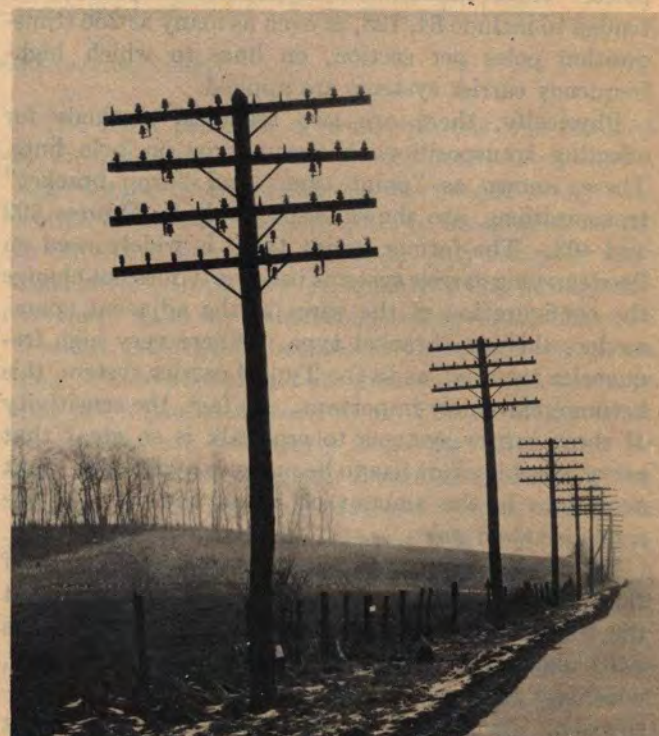


Figure 404

In current practice, it is usually found more economical to counteract the unbalances in part by connecting small balancing condensers into the circuits at one or two points in each loading section. This, combined with a limited number of "test splices" as above, effects the net result desired with greater accuracy, and reduces the number of capacity unbalance tests that have to be made when a cable is installed. These balancing condensers consist of short lengths of two parallel insulated fine-gage wires wound helically around a non-conducting core. Two terminals of this tiny condenser are connected across the two line conductors whose capacity it is desired to increase, and the other ends of the wires can be cut off at whatever point is necessary to give the condenser the precise value of capacity required. A large num-



Transpositions in 12-inch Spaced Open Wire Line Using Drop-Bracket Fixtures

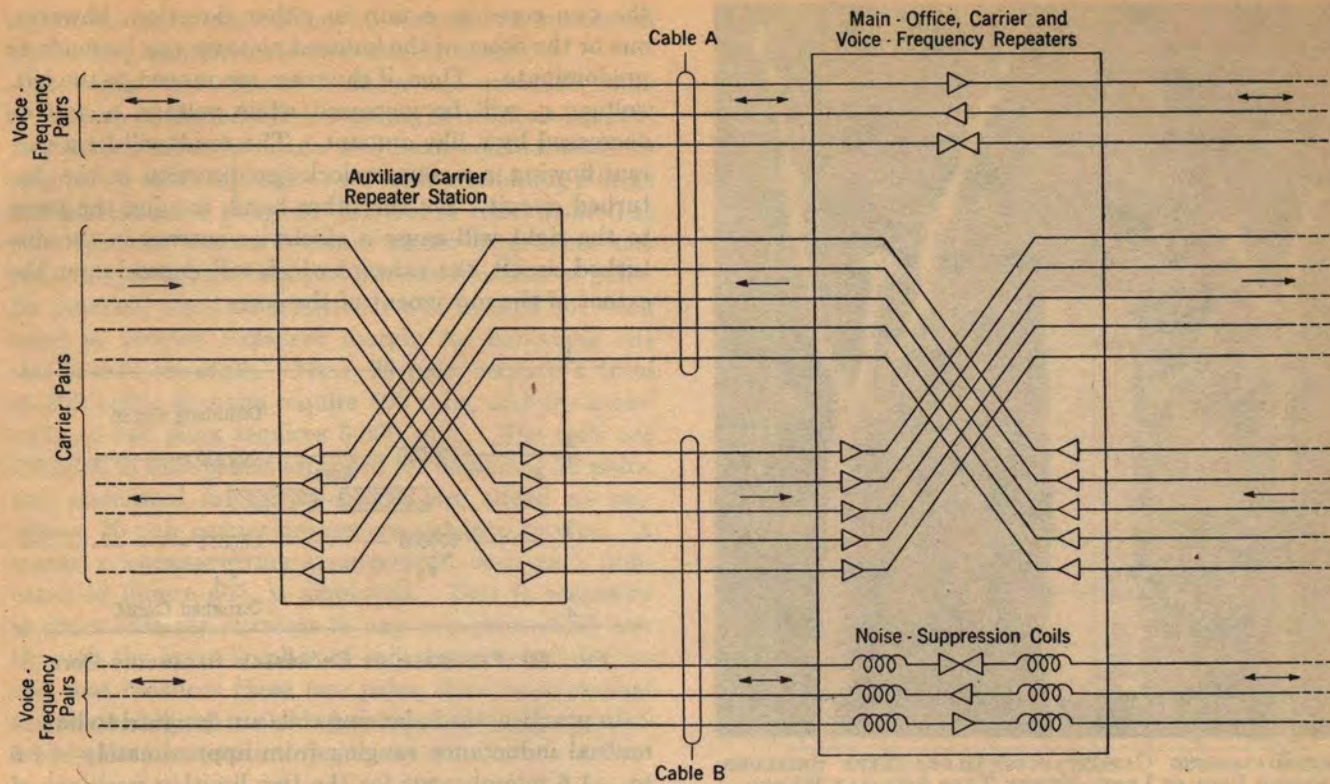


Fig. 405. Arrangements for Reducing Noise and Crosstalk in Type-K Carrier Systems

ber of these condensers can be included within the lead sleeve at a splicing point. In certain cases, where the cable conductors are to be used for 4-wire circuits, it is practicable to balance the capacities for a whole repeater section by adding condensers of this type at one end.

The capacity balancing methods outlined above have been found adequate in practice for keeping crosstalk to a tolerable minimum in **voice-frequency** cable circuits. When **carrier** systems are applied to cable circuits, the crosstalk problem becomes much more severe. In this case, while capacity coupling is still of consequence, inductive coupling becomes much more important as a cause of crosstalk. In fact, at the highest frequencies of the Type-K carrier systems, it predominates over capacity coupling as a cause of crosstalk in the ratio of about 3 to 1. Accordingly, additional crosstalk reduction measures must be applied to cable conductors used for carrier systems.

The crosstalk possibilities at these high frequencies are so great, in fact, that a number of basic changes in circuit design are required. In the first place, the carrier pairs are used for carrier transmission only. Next, the transmitting paths in the two directions are kept entirely separated by using separate cables for transmission East to West and West to East (or a special cable, with a shield between pairs transmitting in opposite directions). The circuits in the two directions are likewise kept separated within the terminal offices and repeater stations, and shielded office wiring is used in all cases. This means that the energy levels of the carrier currents are approximately the same in all physically adjacent conductors, and that near-end crosstalk possibilities are completely eliminated (assuming that reflection effects have been properly guarded against).

Far-end crosstalk between carrier pairs is minimized by balancing out the capacity and inductive couplings. In addition, special precautions are taken to prevent crosstalk between carrier pairs via the voice-frequency pairs in the cable. Most effective in accomplishing this latter, is the complete transposition of the entire group of carrier pairs between the two cables at each repeater station. As may be seen from Figure 405, this automatically eliminates crosstalk via the voice-frequency pairs from the outputs of the amplifiers in the carrier pairs to the inputs of amplifiers in other carrier pairs. Carrier filters or noise suppression coils are also inserted in the voice-frequency pairs at voice-frequency repeater stations, and certain other points, to discourage the transmission of induced currents of carrier frequencies over the voice-frequency conductors.

The methods of balancing out capacity coupling between the carrier pairs themselves are essentially the same as were discussed above in connection with voice-frequency transmission. In balancing out crosstalk

CABLE BALANCING CAPACITANCES—OLDER TYPE SHIELDED TWISTED PAIR AT LEFT—NEWER TYPE SPIRALLY WOUND PAIR AT RIGHT—UNIT USED AT END OF REPEATER SECTION IN CENTER

due to inductive coupling, entirely new methods have had to be devised. The fundamental problem is to balance every carrier pair against every other carrier pair in the same cable, in each repeater section. The method used depends upon counteracting the crosstalk currents with equal currents flowing in the opposite direction. Thus, if in a given disturbed circuit a crosstalk current is flowing in a clockwise direction, it is desired to set up an equal current in the circuit flowing in a counterclockwise direction.

This result can be effected by means of tiny transformers connected between each carrier pair and every other carrier pair. However, since it is necessary to control the magnitude of the artificially induced currents and also to cause them to flow in either direction, depending upon the direction of the crosstalk current, the transformers must be designed so that the coupling between circuits can be adjusted and so that they can be poled in either direction. The method used to obtain this result is indicated schematically in Figure 406. Here, it may be noted that there are really two separate transformers, one having a reversed winding in the disturbing circuit so that a current, I, flowing in the disturbing circuit will induce oppositely poled voltages in the disturbed circuit. If the cores of the two transformers are centered as shown in the drawing, the induced voltages will be exactly equal and the net effect on the disturbed circuit will be nil. By moving the two cores as a unit in either direction, however, one or the other of the induced voltages can be made to predominate. Thus, if the cores are moved to the left, voltage e_1 will be increased while voltage e_2 will be decreased by a like amount. The result will be a current flowing in a counterclockwise direction in the disturbed circuit. On the other hand, moving the cores to the right will cause a clockwise current in the disturbed circuit, the value of which will depend upon the extent of the movement of the cores.

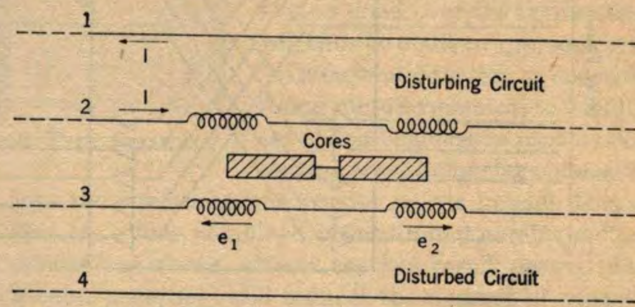
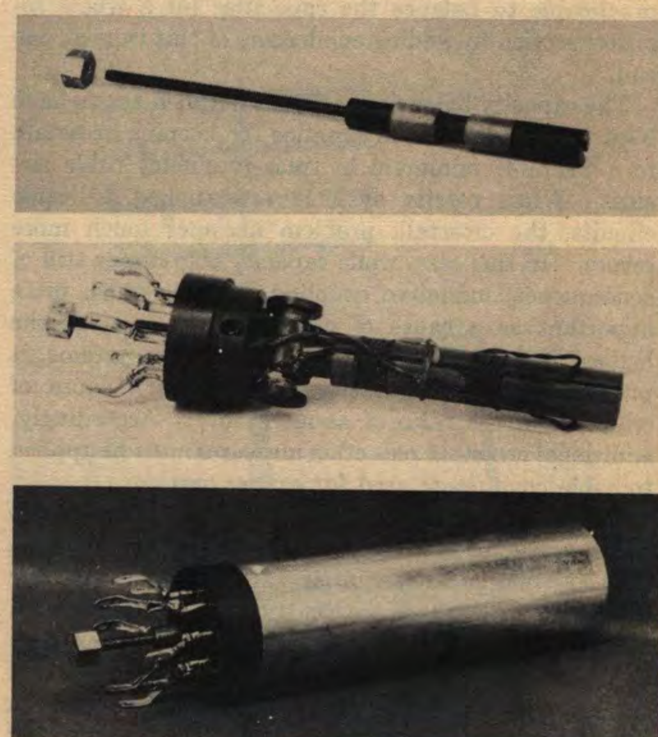


FIG. 406. PRINCIPLE OF CROSSTALK BALANCING COIL

In practice, the balancing coils are designed to have a mutual inductance ranging from approximately $+1.6$ to -1.6 microhenrys for the two limiting positions of the cores. The coils are mounted in small cylindrical containers about $4\frac{1}{2}$ inches long and $1\frac{3}{8}$ inches in diameter and arranged for rack mounting. The position of the coil cores is controlled by a screw by means of



CONSTRUCTION OF CROSSTALK BALANCING COIL which the core can be moved through its maximum travel of ½ inch in about 16 complete turns.

In using these coils to balance out crosstalk, measurements of the inductive coupling between each pair of conductors must be made and each coil adjusted to counteract this coupling. In a cable containing a large number of carrier pairs, the number of coils required at each repeater station becomes rather large since one coil is required for every possible combination of pairs. In practice, also, an additional coil is used for each quad to provide sufficient margin for balancing out side-to-side crosstalk. Thus, 20 pairs require a total of 200 coils, 40 pairs require 800 coils, and the maximum of 100 pairs requires 5,000 coils. The coils are installed in unit panels arranged for balancing 20 pairs, and additional intergroup panels are added as successive 20 pair carrier groups are put into service. A special crisscross wiring arrangement, such as is indicated in Figure 407, is employed. This is necessary in order that the currents in any two pairs shall flow through the same number of coils before reaching the coil that balances these two pairs, thus insuring that the phase shift up to the balancing coil will be approximately the same on both pairs.

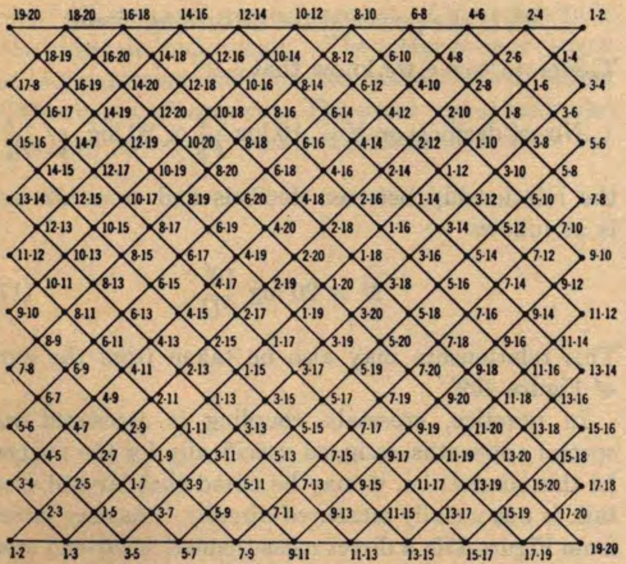


FIG. 407. METHOD OF CONNECTING CROSSTALK BALANCING COILS

194. Noise

Currents within the voice-frequency range, induced into a telephone circuit from electrical power circuits, are manifested to a listener on a disturbed telephone circuit as noise. In many cases crosstalk currents may also appear merely as noise. This is particularly true in the case of cable circuits where any crosstalk heard is likely to come simultaneously from a considerable number of other circuits, and appears to the listener on the disturbed circuit as a special form of noise, called "babble". In other words, it is just an unintelligible conglomeration of speech sounds coming from a large number of sources.

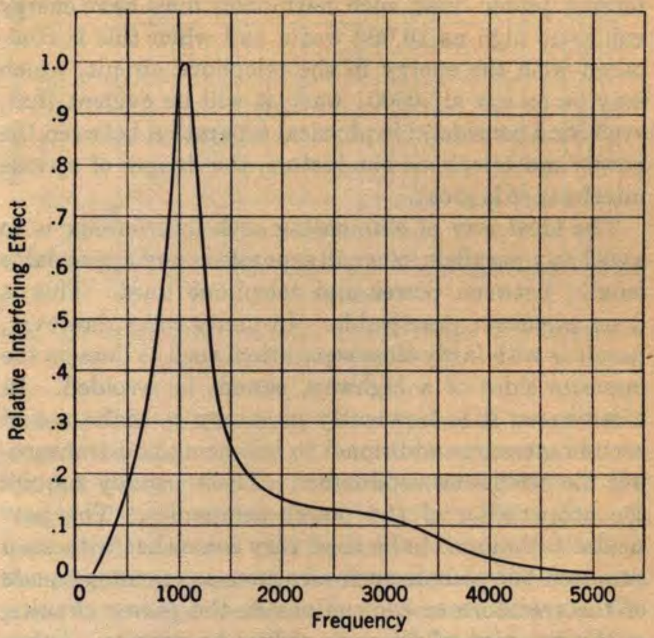


FIG. 408. RELATIVE INTERFERING EFFECT OF NOISE AT DIFFERENT FREQUENCIES

The disturbing effect of noise to a listener depends first, of course, upon its volume. It also depends upon the frequency of the noise currents. Figure 408 shows the results of tests that have been made to determine the relative disturbing effects of various noise frequencies. It will be noted that the disturbing effect peaks up rather sharply in the neighborhood of 1100 cycles. Where noise is of appreciable volume—particularly in the more sensitive frequency range—it is naturally annoying to the telephone user and may seriously reduce the intelligibility of conversation. It is accordingly necessary to keep the noise in working telephone circuits below those limits where its interfering effect on conversation will be important.

Since noise is essentially an induced effect like crosstalk, similar measures are used to counteract it. Careful resistance balancing of the telephone conductors, the use of transposition systems, and other measures taken to avoid crosstalk, are likewise effective in reducing noise. However, such measures alone may be inadequate to keep noise within the desired limits. This is a result of the fact that paralleling power lines are the principal source of noise, and the power carried over such lines is greater by tremendous percentages than that carried over any telephone circuit.

Of course, the usual fundamental frequency of power transmission is 60 cycles and this frequency is too low to interfere appreciably with telephone transmission. Unfortunately, however, the currents transmitted over the power line usually include several harmonics of the fundamental frequency, and these may lie well within the range of telephone frequencies. In high tension power lines, such harmonics may have energy values as high as 10,000 watts and when this is compared with the energy in the telephone circuit, which may be as low as .00001 watt, it will be evident that, even for a considerable physical separation between the power and telephone conductors, the danger of serious interference is great.

The ideal way of eliminating such interference is to avoid any parallels, of small separation and appreciable length, between power and telephone lines. This is done whenever practicable. In many cases, however, parallels with fairly close separation, such as lines on the opposite sides of a highway, cannot be avoided. In these cases, it is frequently necessary to make use of certain measures additional to balancing and transposing the telephone conductors. These usually require the cooperation of the power companies. The particular techniques to be used vary somewhat with each situation but include such measures as rearrangements of the transformer connections in the power circuits, or the insertion of filters to reduce harmonics. Other methods frequently employed include changes in the configuration of the power wires on their poles to make for better electrical balance, and transpositions of the power wires. Such power line transpositions have essentially the same effect in balancing out the magnetic fields as do transpositions in a disturbing telephone circuit.

195. Crosstalk and Noise Measurements

The ideal objective of the various methods discussed above for counteracting crosstalk and noise induction in telephone circuits is, of course, to eliminate their effects altogether. In practice this ideal is rarely attained. But certain practical limits are established, and every reasonable effort is made to keep the crosstalk and noise below these limits. In designing and maintaining circuits, therefore, it is necessary to be able to make definite quantitative measurements of both crosstalk and noise. As in any other kind of measurement, this requires the establishment of definite units.

Since crosstalk represents the transmission of telephone currents from one circuit to another, it can be measured directly in terms of the transmission loss between the two circuits in the same way as direct telephone transmission. This loss may be expressed either in terms of decibels or as "crosstalk units". The number of crosstalk units (CU) is equal, by definition, to 10^6 times the ratio of the current in the disturbed circuit to that in the disturbing circuit at the two points under consideration; or, if the circuit impedances are not identical, 10^6 times the square root of the power ratios. Expressed as an equation—

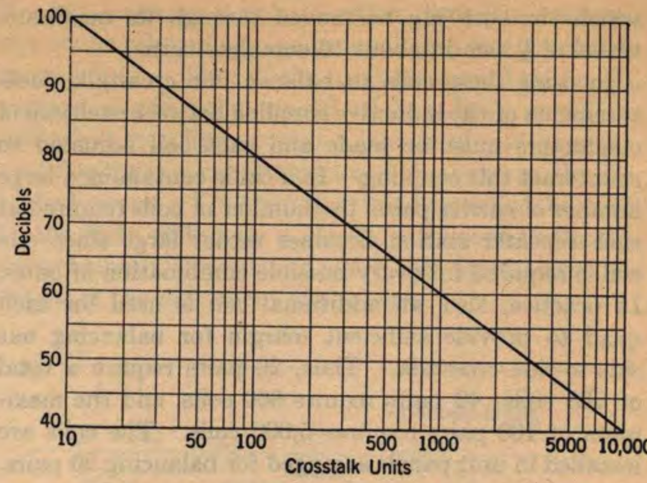


Fig. 409. Relation Between Decibels and Crosstalk Units

$$\text{No. of crosstalk units or } CU = 10^6 \times \sqrt{\frac{P_R}{P_S}} \qquad (174)$$

where P_R is the power in the disturbed circuit and

P_S is the power in the disturbing circuit.

Therefore, since we know that—

$$\text{No. of decibels or } N = 10 \log \frac{P_S}{P_R} \text{ or } 20 \log \sqrt{\frac{P_S}{P_R}}$$

the relationship between decibels and crosstalk units is as follows—

$$N = 20 \log \frac{10^6}{CU} \qquad (175)$$

This relationship may also be taken from the curve of Figure 409.

In practice, crosstalk coupling is measured with special apparatus designed specifically for the purpose. In this connection, it may be noted that far-end crosstalk is not usually measured directly. As may be seen from Figure 410, a direct measurement of far-end crosstalk would involve determining the ratio of the current I_a at one end of the disturbing circuit A to the current

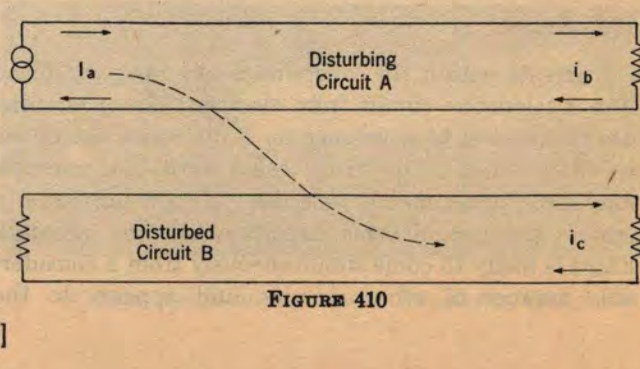


Figure 410

i_c at the other end of the disturbed circuit B. It is usually more convenient to measure the ratio between i_c and the current i_b at the far end of the disturbing circuit. This ratio is called the "measured far-end" crosstalk. If desired, it can be readily converted into true far-end crosstalk by multiplying by the current ratio $\frac{i_b}{I_a}$ of the current in the disturbing circuit—or, if the measured crosstalk is expressed in decibels, by simply adding the transmission loss of the disturbing circuit in decibels.

For measuring noise, a basic reference point has been selected, which is equal to 10^{-12} watts of 1000-cycle power. Noise may then be measured as the **number of decibels above this reference noise.** A unit known as the Noise Unit is also employed. Its value is such that the reference noise defined above is equal approximately to seven Noise Units. The relationship between Noise Units and decibels above reference noise is given approximately by the curve of Figure 411.

Noise measuring meters are now in general use by means of which the noise on a circuit can be read directly from a scale calibrated in decibels above reference noise (see Article 207). These same meters are also used to measure crosstalk volume in the same terms. This gives a general measure of the magnitude of the crosstalk on a circuit. The meter is so arranged that either measurement can be made without taking the circuit out of service. The noise meter may also be used to measure crosstalk coupling in decibels but

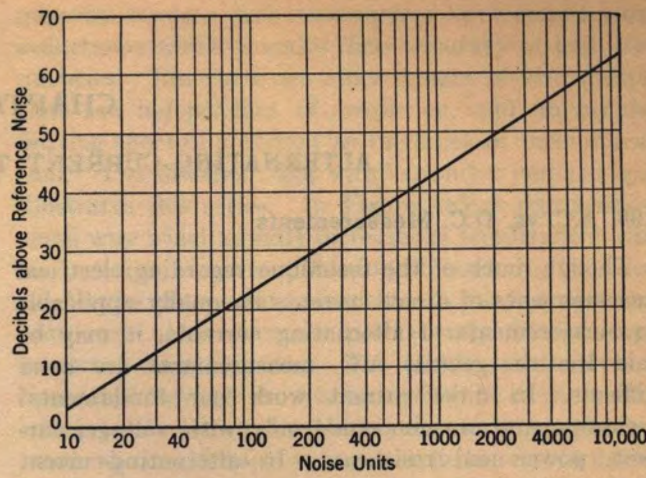


FIG. 411. RELATION BETWEEN NOISE UNITS AND DECIBELS ABOVE REFERENCE NOISE

such a measurement requires removing the two circuits in question from service. The power source applied to the disturbing circuit is preferably an oscillator of the "warbler" type, as described in Article 208. Then if the output volume of this oscillator (which is applied to the disturbing circuit) is first measured with the noise meter in terms of decibels above reference noise and the crosstalk coming from this same source is measured on the disturbed circuit, the difference between the two measurements gives the crosstalk coupling in decibels directly. This, of course, can be readily converted to crosstalk units if desired.

CHAPTER XXXII

ALTERNATING-CURRENT TESTS AND MEASUREMENTS

196. A.C. vs. D.C. Measurements

Though much of the technique regarding electrical measurements of direct currents is equally applicable to measurements of alternating currents, it may be said that in general A.C. measurements are more difficult. In direct-current work our fundamental measurements are concerned only with voltage, current, power and resistance. In alternating-current work, while voltage, current, and power are still the fundamental quantities, their interrelationship is no longer simple but involves considerations of phase, frequency, etc. Again, the measurement of alternating voltages and currents must presuppose some standard wave shape and make some supposition as to the basis of measuring a quantity which is ever varying, i.e., we may measure an instantaneous value, an effective value or a maximum value. Furthermore, in dealing with wave shapes other than sine waves, we must effect some analysis into a fundamental sine wave and harmonics of this fundamental, in order to analyze the conditions correctly (see Appendix IV). These new conditions are responsible for complications incidental to the measurement of the quantities which correspond to those we encounter in direct-current work, and introduce the necessity for more elaborate and painstaking methods for the complete analysis of A.C. phenomena. Moreover, it may be said that a degree of instability is inherent in certain of the properties met with in A.C. work, whereas in D.C. work this difficulty is not encountered. To illustrate, the direct-current resistance of a coil winding remains practically fixed, with the exception of minor changes in values due to temperature, while the alternating-current resistance of the coil may be less stable due to certain additional factors upon which it depends. These include the magnetic properties of the iron core and the physical relationship of the winding to the iron core, both having to do with certain power losses which in turn affect the resistance to alternating current.

It is not always the practice in alternating-current work to make measurements with the basic units, i.e., ampere, volt, watt, etc. We may employ as standards other units based either directly or indirectly upon the fundamental units. For instance, in telephone transmission work it is quite possible to determine the attenuation from the relationship—

$$\alpha = 2.303 \log \frac{I_s}{I_r}$$

by making current measurements, but this method is seldom used in practice. Instead, as we know, the decibel is used as a comparison standard and the measurement is expressed in decibels rather than as a numerical value of α.

On the other hand, the instruments designed to measure fundamental quantities are none the less important because apart from their field use, which may be limited in some cases, the same principles of operation employed in these devices are frequently employed in connection with other measuring apparatus. We shall, therefore, consider first of all the measurement of the fundamental alternating-current quantities.

197. A.C. Ammeters, Voltmeters, and Wattmeters

Indicating instruments, such as ammeters, voltmeters, and wattmeters for alternating-current measurements, are similar in appearance and in manipulation to direct-current instruments but have certain differences in design. A direct-current ammeter, such as was described in Chapter IV, if connected in series with an alternating E.M.F., would tend to indicate the instantaneous value of the current in the circuit. Now this value is constantly changing, and in the case of a 60-cycle power circuit, for example, the change is from zero to a maximum to zero 120 times a second, 60 such changes occurring while the current flow is in one direction, and the remaining 60 while the current flow is in the opposite direction. It is not possible for the needle to fluctuate so rapidly and consequently it would stand at zero, not responding to any value of current through the instrument. However, by substituting a coil winding for the permanent magnet of the D.C. instrument, we can obtain a definite deflection on the ammeter scale and this indication will depend upon the **effective** value of the current.

This, briefly, is the fundamental difference between D.C. and A.C. ammeters and voltmeters of this type. The coil winding and the moving element are connected in series so that whenever the current reverses in one, there is a similar reversal in the other. Consequently, the reaction between the magnetic field of the coil and the current in the moving element is always such as to turn the moving element in the same direction. This is illustrated by Figure 412. Here a movable coil is suspended within, and by means of a spiral spring is held perpendicular to, a stationary coil. The magnetic

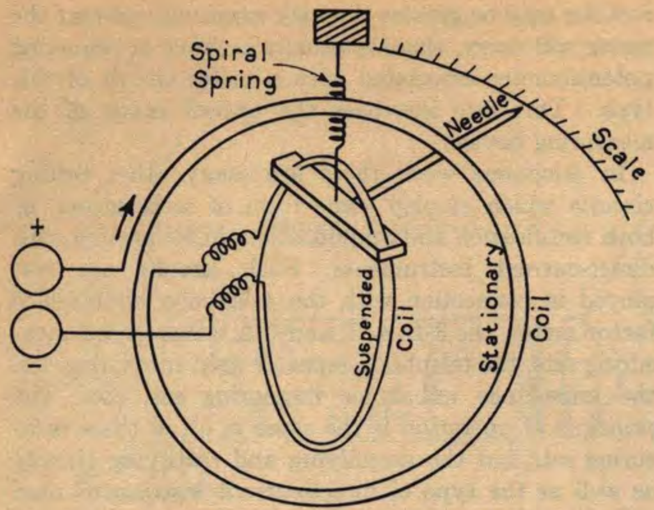


FIG. 412. DYNAMOMETER TYPE A.C. METER

field of the movable coil tends to align itself with that of the stationary coil, and the direction of rotation of the movable coil is the same regardless of the direction of the current through the two coils in series. If the instrument is a voltmeter it must, of course, have a high resistance; if an ammeter, it must have a low resistance, which can be secured by the use of a shunt.

In alternating-current work, unless the angle of lead or lag between the voltage and current is known, it is not possible to use an ammeter and a voltmeter for the measurement of power, inasmuch as the power equation is—

$$P = EI \cos \theta$$

instead of

$$P = EI$$

The power, therefore, cannot be determined by simply multiplying together the measured voltage and current values; a wattmeter must be used for accuracy. The A.C. wattmeter likewise employs two coils, but in connecting such an instrument in the circuit, one coil is connected in series so that the current in it varies as the line current, while the other coil is connected across the circuit so the current in it is proportional to the voltage. Such an arrangement automatically takes care of any phase difference between voltage and current, and the indication of the wattmeter, therefore, depends upon the power in the circuit.

The ammeter, voltmeter, and wattmeter described in the foregoing are said to employ the "dynamometer" principle. The commercial types of these instruments for alternating-current work are usually designed for a single frequency, or at best a narrow band of frequencies. It is not possible, for example, to use the rotating coil mechanism designed for 60-cycle power circuits in connection with telephone current frequencies because such instruments have considerable inductance which impairs their accuracy at high frequencies. But there are other designs of instruments that are independent of frequency, and employ the heating effect of a current as the basis of their operation. The so-called "hot wire" ammeter perhaps best illustrates this series. In Figure 413, h represents a small wire which rapidly increases in temperature with an increase in current flowing through it. W_1 and W_2 are the instrument connections to this hot wire, and both ends of the wire are permanently fixed, though insulated from the case. The middle of the wire is connected through the insulating link L to the needle which is pivoted at P. As in other indicating instruments the needle has a spring attached to it, but unlike other indicating instruments, when the temperature of the wire is above normal this spring tends to make it stand at full scale reading. However, when the hot wire is at normal temperature, it is so constructed as to pull the needle to the zero position. As the current flows through the wire and increases its temperature, the wire expands and the needle is allowed to give a scale reading. The scale is so calibrated as to indicate

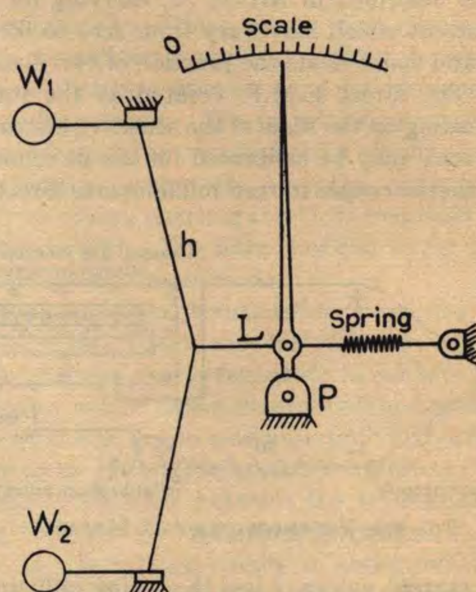


FIG. 413. HOT WIRE TYPE A.C. METER

the effective value of the current flowing through the wire. To take care of changes in atmospheric temperatures, the wire is so mounted on the case that the expansion between the mountings compensates for the effect of changes in temperature on the wire itself.

Although the hot wire type of instrument is independent of frequency, it has other practical limitations. It is not only sluggish in action, since time is required for the heating of the wire, but there is danger of burning out the instrument because any appreciable overload will produce a temperature great enough to melt the wire.

Neither the dynamometer nor the hot wire type of instrument is suitable for measuring extremely small alternating-current quantities such as are often encountered in communication work. The actual voice current, when transmitted over a telephone circuit, may vary from less than 10 milliamperes at the talking station to $\frac{1}{10}$th of one milliampere at the receiving station. For the high degree of sensitivity that is required for such measurements, amplifying and rectifying devices are often used in connection with direct-current meters in preference to the types of instruments we have discussed in the foregoing.

198. The Use of Rectifying and Amplifying Devices in Connection With Measuring Instruments

The simplest method of measuring high-frequency alternating currents, employing a direct-current meter, is to use a thermocouple connected to a sensitive millivoltmeter or microammeter. Such an arrangement is used in various transmission measuring sets and is illustrated in Figure 414. Here we have a thermocouple, as described in Article 52, carrying an alternating current which may vary from zero to 60 milliamperes and which heats the junction of two dissimilar metals. The direct E.M.F. created at the junction gives a reading on the scale of the sensitive instrument, and this scale may be calibrated for use in connection with the thermocouple to read milliamperes direct.

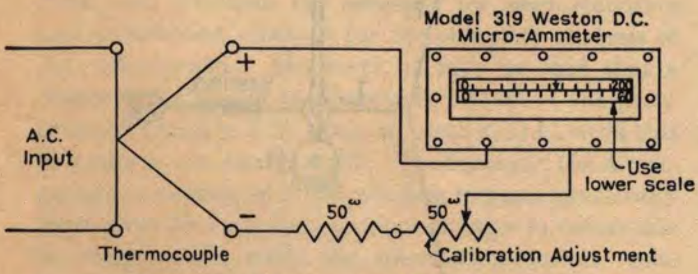


Fig. 414. Thermocouple A.C. Meter

Where current values of less than a few milliamperes are to be recorded, the same model or a similar but more sensitive model of instrument may be used in connection with an amplifying and rectifying circuit. Such an arrangement is commonly employed in transmission measuring circuits. In this case only the sensitive D.C. meter is used in the output circuit of the rectifying device, no thermocouple being necessary. The amplifying circuit, of course, amplifies the weak alternating voltages impressed on its input while the rectifying circuit converts these amplified voltages into direct current to be measured on the D.C. meter. To take care of the possibility that the output current of the rectifier may be greater than the maximum current the meter will carry, there is usually a shunt or adjusting potentiometer associated with a meter circuit of this type. This also increases the overall range of the measuring device.

In telephone work there are many other testing circuits which employ some form of rectification, or both rectification and amplification, in connection with direct-current instruments. Such circuits are employed in connection with the telephone interference factor meter, the 3-B, 4-B, and 6-A transmission measuring sets, the telephone repeater gain measuring set, the impedance unbalance measuring set, etc. The principle of operation is the same in all of these measuring sets, but the amplifying and rectifying circuits as well as the type of direct-current instrument used may vary in detail.

199. Transmission Measurements

Although the apparatus used in making transmission measurements is somewhat complex, this is due entirely to the difficulties inherent in the measuring of high-frequency currents of low value, and not in any way to the theory involved. The most obvious way of determining the attenuation of a circuit is to measure the currents sent and received at the two terminals, and thus learn the value of the current ratio, from which the attenuation may be determined. However, this method does not lend itself to rapid work and for routine measurements, transmission measuring sets reading directly in decibels are much more suitable. These sets are arranged either to make measurements directly, or to compare the loss of the circuit under test with a known calibrated loss. The former types are considered in the next article. The general arrangement of the comparison types for making transmission measurements is illustrated in Figure 415. The set is first calibrated by connecting a voltage to a fixed artificial line which causes a definite known loss. The entering current, after passing through this line, is amplified and rectified and passes through a potentiometer to a D.C. meter. The value of the applied voltage is then adjusted to such a value as to give any desired deflection of the D.C. meter, usually mid-scale. After calibrating, connections are changed so that the same voltage is applied to a variable artificial line in series with the circuit whose equivalent is to be determined. By cutting out sections of the artificial line the total loss in the circuit is made the same as that in the calibrating circuit, so that the D.C. meter gives the same deflection in both cases. The dials are arranged to read the loss in the "unknown" circuit directly. This is accomplished by constructing the variable artificial line to cause a maximum loss exactly equal to

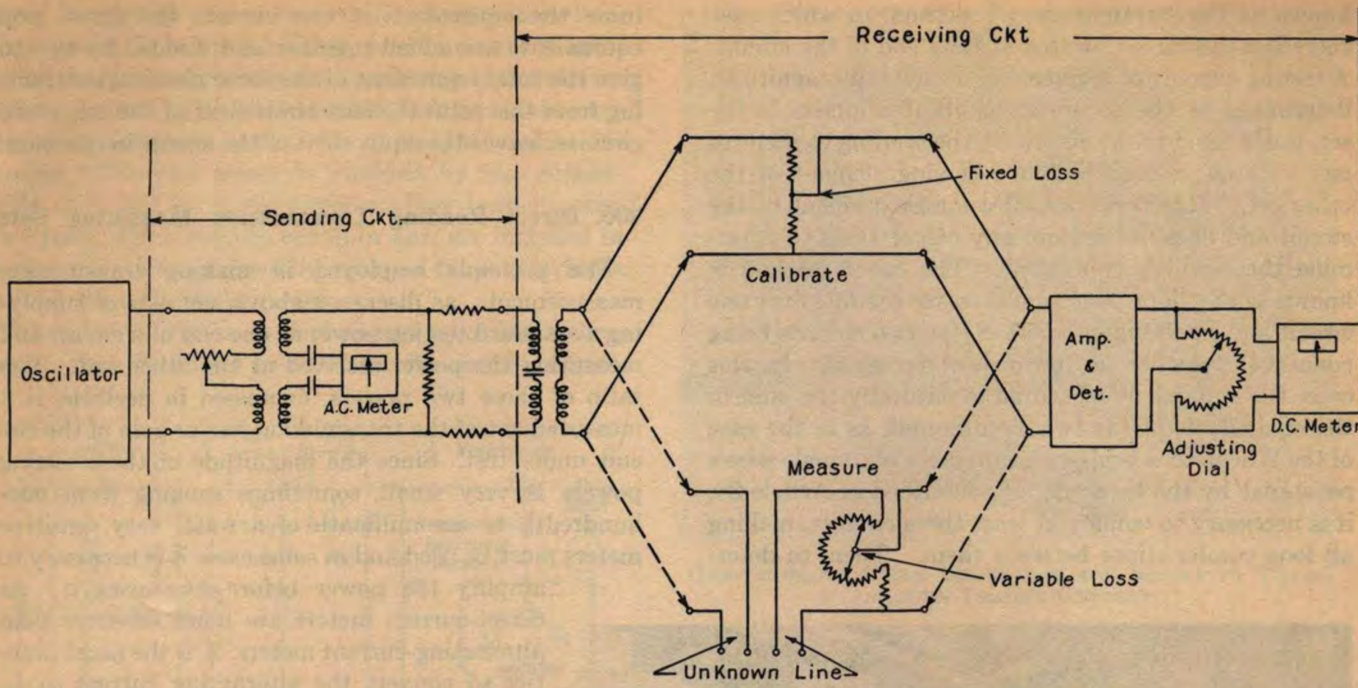


FIG. 415. PRINCIPLE OF TRANSMISSION MEASURING SET

the loss caused by the fixed artificial line. When the measuring dials are set at zero, the variable line will cause the maximum loss; with the dials set at 10 db, the total loss caused by the variable line will be the maximum minus 10, and so on.

Considering the component parts of the complete circuit, we have a generator or sending element, a load or receiving element, and a line connecting the two, which may be the calibrating standard or the circuit under test, depending upon the position of the switching key. In order that any reflection losses on the connection may be reduced, there are associated with the sending and receiving elements variable networks which may be adjusted to have impedance values closely approximating those of the circuits to which the elements are connected. Also, in order that line noises caused by inductive effects may not appreciably impair the accuracy of the readings, a filter which stops all currents except those between 850 and 1150 cycles is connected in the receiving element. It is obvious that noise will cause an error inasmuch as it increases the received energy and the ratio of energy received to energy sent will, therefore, be incorrect.

Since all artificial lines, networks, etc. in this apparatus are made up of resistances, it is possible to make measurements at any desired frequency in the voice band. A study of the equivalents of the majority of circuits measured at a single frequency and measured with a band of voice frequencies for comparison, has led to the standardization of 1000 cycles for tests of this nature. This particular single frequency gives for the various conditions, results most closely approximating the talking equivalents. This, however, does not mean that measurements are not sometimes made at other frequencies. The transmission measuring set is often employed in the study of a circuit's "quality" by making measurements over the entire voice-frequency band and charting the "loss-frequency" curves (see Article 139). The filter is of course cut out when such measurements are made.

Certain models of transmission measuring sets are adapted to permanent office installations. The testing circuits of these models terminate in cords or jacks for the sending and receiving connections, and are equipped with a reversing key to send and receive in either direction without interchanging the connections. Trunks are usually installed between the transmission measuring set and the testboard, switchboard, repeater bays, etc. A talking circuit is also provided at the measuring set, which may be used to talk over a circuit connection to the trunks or over a call circuit to a test operator in the toll operating room. Portable transmission measuring sets are also available which can be conveniently used for measuring the transmission equivalents of cord circuits, composite sets, switching trunks, and miscellaneous equipment units that affect the transmission features of a talking connection. These do not have the range of the permanently mounted sets but their principle of operation is not widely different.

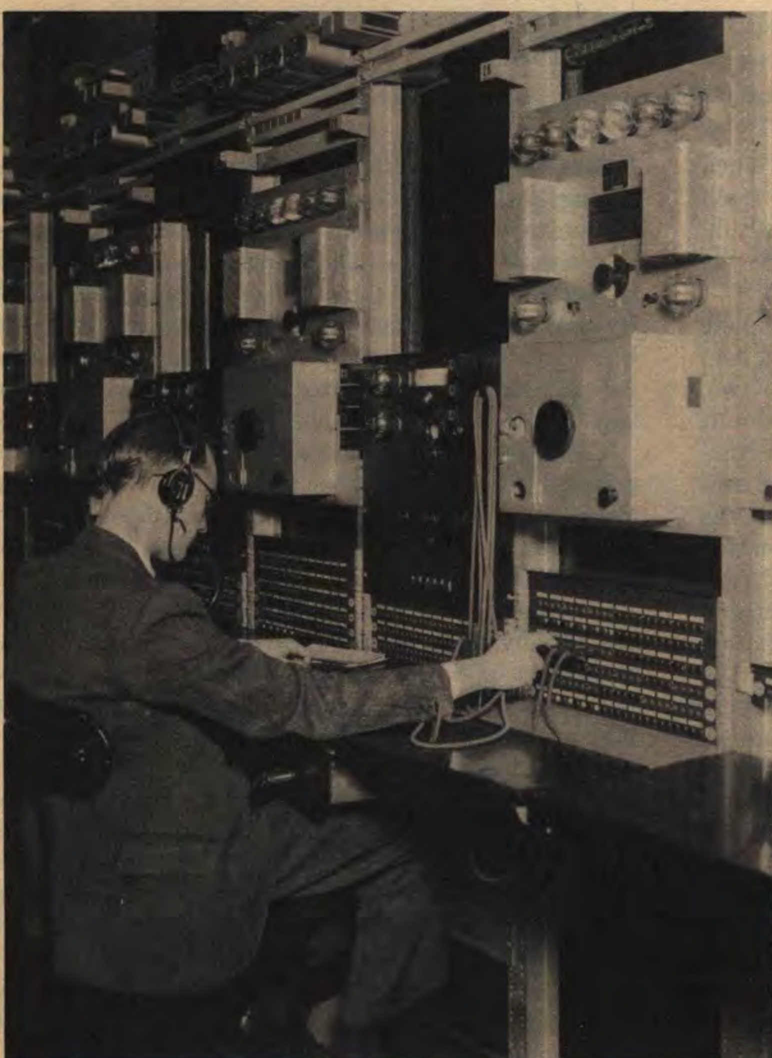
In making measurements with the types of sets discussed above two methods are employed. One is known as the "straight-away" method, in which case there is a testing set located at each end of the circuit. A testing current of a given frequency and magnitude, determined by the accurately calibrated meters in the set, is delivered to the circuit by the sending element of one set and received by the receiving element of the other set. This permits a direct measurement of the circuit and does not require any calculations to determine the circuit's equivalent. The other method is known as the "loop test" and involves sending over one circuit and receiving on another, the two circuits being connected together at the distant terminal. In this case, the equivalent measured is naturally the sum of the equivalents of the two circuits and, as in the case of the Wheatstone bridge measurement of a single wire's resistance by the loop method, described in Article 38, it is necessary to employ at least three circuits, making all loop combinations between them. Then, to determine the equivalent of one circuit, the three loop equivalents are added together and divided by two to give the total equivalent of the three circuits; subtracting from this total the loop equivalent of the other two circuits, leaves the equivalent of the circuit in question.

200. Direct Reading Transmission Measuring Sets

The principle employed in making transmission measurements, as discussed above, consists of supplying a standard testing power at one end of a circuit and measuring the power received at the other end. The ratio of these two powers, expressed in decibels, is a measurement of the transmission loss or gain of the circuit under test. Since the magnitude of these testing powers is very small, sometimes ranging from one-hundredth to one-millionth of a watt, very sensitive meters must be used and in some cases it is necessary to amplify the power before measuring it. As direct-current meters are more sensitive than alternating-current meters, it is the usual practice to convert the alternating current to direct current by means of vacuum tube or copper-oxide rectifiers and then measure with direct-current meters.

In accordance with these principles, the simplest transmission measuring arrangement would consist in connecting a known source of power to one end of the unknown circuit and a direct reading meter, calibrated in decibels, to the other end of the circuit. Until quite recently, such a simple arrangement has not been feasible, either because sufficiently sensitive meters were not available or because the amplifiers used for increasing the currents to the required values were not sufficiently stable. It has been necessary accordingly to employ the comparison method, described in the preceding article. This method obviates errors due to variations in the amplifier circuit because it consists in making two successive measurements, which merely compare the unknown circuit with a known loss, and does not measure absolute values.

The development of highly stable negative feedback amplifiers, copper-oxide rectifiers, and more sensitive direct-current meters has made possible direct reading measuring systems. Meters employing some of the newer alloy steels are now available, which will measure transmission losses of around 20 db below one milliwatt without an amplifier. Figure 416 is a simplified circuit of such a measuring set that accurately measures losses up to 20 db. In using this set, one milliwatt of power at 1000



Transmission Testboard Showing Variable Oscillators and Transmission Measuring Sets cycles is applied at one end of the circuit under test. This power is supplied by a small magneto generator designed especially for furnishing transmission testing power. The transmission testing set is connected at the other end of the circuit, where the incoming 1000-cycle power is rectified by the copper-oxide units and led to the direct reading decibel meter. Two pads, which may be cut in or out, are included in the input circuit to extend the range of the set to values beyond the scale of the meter.

Where transmission measurements involving values greater than 20 db are to be made, the simple copper-

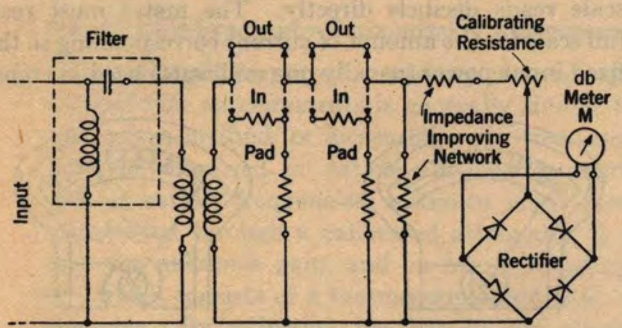
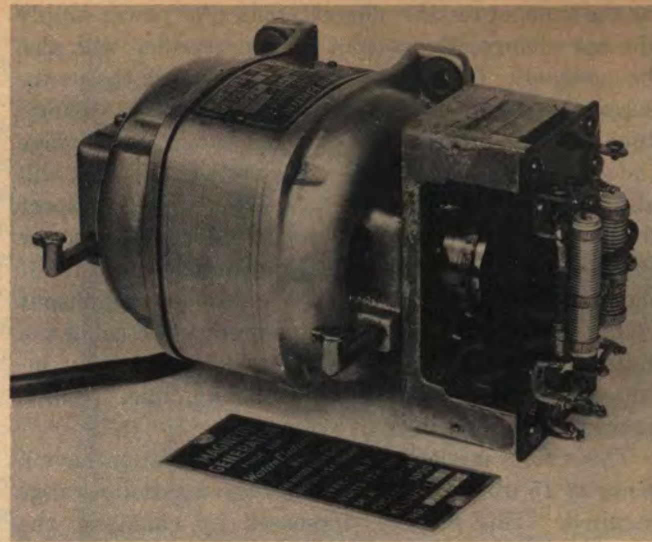


Fig. 416. Direct Reading Transmission Measuring Set

oxide rectifier and meter circuit illustrated in Figure 416 is inadequate. To meet this condition, a negative feedback amplifier is included in the measuring set. As we know, such an amplifier is very stable and may be operated over long periods without adjustments. The amplifier is shown in Figure 417. It consists of a high impedance input transformer, T, bridged across a 600-ohm terminating resistance (which may be removed when level measurements are made), two pentode tubes, a copper-oxide rectifier, R, and a meter, M, calibrated directly in decibels. The negative feed-



One-thousand Cycle Generator for Supplying Transmission Testing Current

back voltage is introduced into the control grid circuit of the first tube through resistances A, B, and C by connecting one of the rectifier terminals to contact P of the potentiometer. These resistances, together with resistance D, form the cathode drop resistance of the control grid circuit, and any potential applied across them affects the potential on the control grid of the first tube. With the potentiometer, P, contact as shown in Figure 417, the negative feedback is a maximum and the net overall amplification of the amplifier is therefore a minimum. Changing the potentiometer contact to a lower step gives less feedback and results in a greater net gain of the amplifier.

When a constant potential is applied to the input terminals of the amplifier, a constant voltage is also applied to the control grid of the first tube. As long

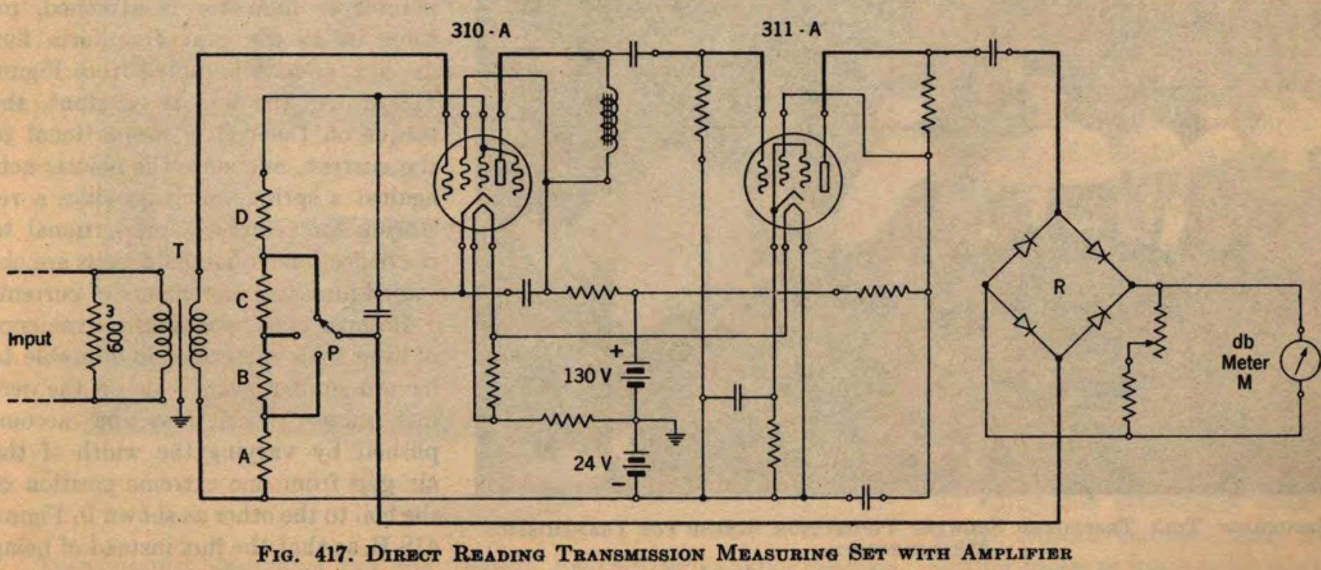


Fig. 417. Direct Reading Transmission Measuring Set with Amplifier

as the tube, or rectifier characteristics, or power supply do not change, the output of the rectifier will also be constant. If, however, due to any cause the power supply changes, or the characteristics of a vacuum tube or the rectifier changes in such a way as to change the output to a higher value, the voltage fed back will also be increased. Because it is negative with respect to the input, this reduces the input sufficiently to lower the net output voltage to its proper value again. If the change were in the reverse direction, the output voltage would first be lowered thereby decreasing the voltage fed back to the input, which effectively increases the input voltage and the amplifier returns again to its proper output voltage.

The meters used with this measuring system have a range of 15 db, which is less than the maximum range required. This range is increased by changing the amplifier gain in steps of 10 db. The resistances A, B and C in the feedback path are used for this purpose. In practice the contact arm of the potentiometer, P, may be controlled by relays which in turn are remotely controlled by keys, jacks, or dials at various points in the office. The db meters may be located where desired, without reference to the amplifier-rectifier. They may also be placed in lantern slide projectors to throw a greatly enlarged meter scale on a screen, which can be read from a distance of 50 feet or more. These arrangements are extremely flexible since a single amplifier-rectifier and one or more associated decibel meters may be used to make measurements from any one of several points in an office.

The amplifier-rectifier circuit is designed to give a constant gain through a wide range of frequencies. It may, therefore, be used, with an appropriate variable oscillator at the sending end, to measure the loss-frequency characteristics of any circuit through the voice-frequency range. It may also be used in connection with a recording device to measure the variations in the loss of a circuit over a period of time.

201. Decibel Meters

The decibel meter, associated with the amplifier-rectifier circuit discussed in the preceding article, is merely a special D.C. ammeter designed so that its scale reads decibels directly. The meter must read full scale for the amount of current corresponding to the fixed input power (usually one milliwatt) used in trans-

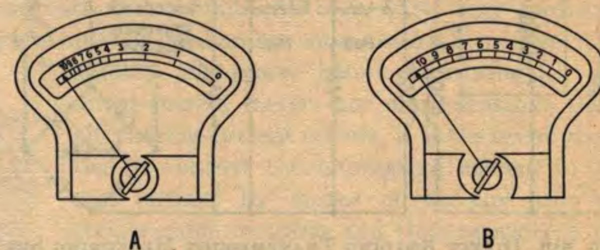
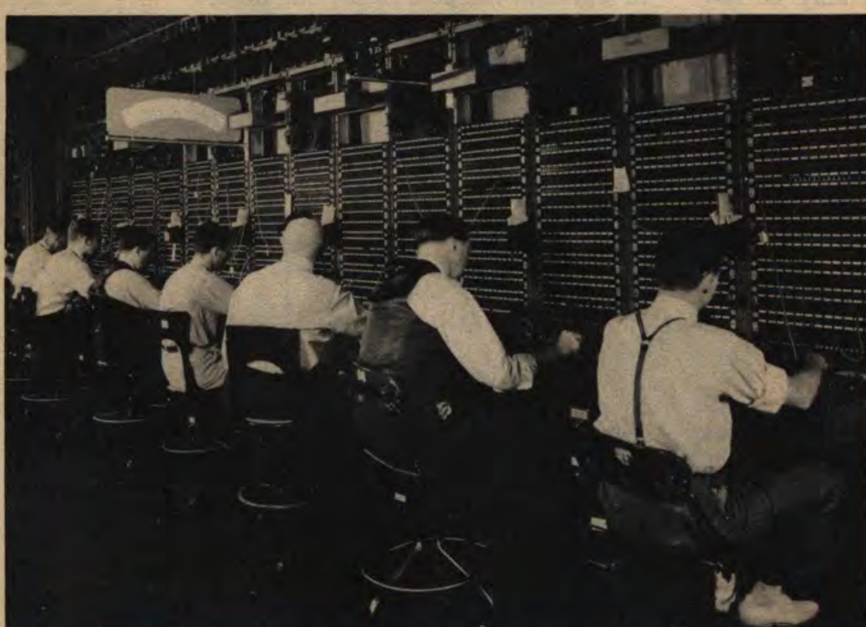


Fig. 418. Decibel Meter

mission measuring, and this full scale deflection of the meter is marked "zero" loss. If the meter is of the usual ammeter design, the points on the scale for subsequent units of db loss are not evenly spaced, but crowded together as illustrated in Figure 418-A, because the decibel is a logarithmic unit.

Of course, the D.C. ammeter is ordinarily designed to have equal spacing on its scale for equal increments of current. This is a result of the fact that the moving coil, to which the pointer or indicator is attached, rotates in an air gap of uniform flux density, as may be noted from Figure 418-A. As the flux is constant, the torque on the coil is proportional to the current, and since the pointer acts against a spring which provides a restoring force directly proportional to the deflection, equal deflections are obtained for equal increments of current.

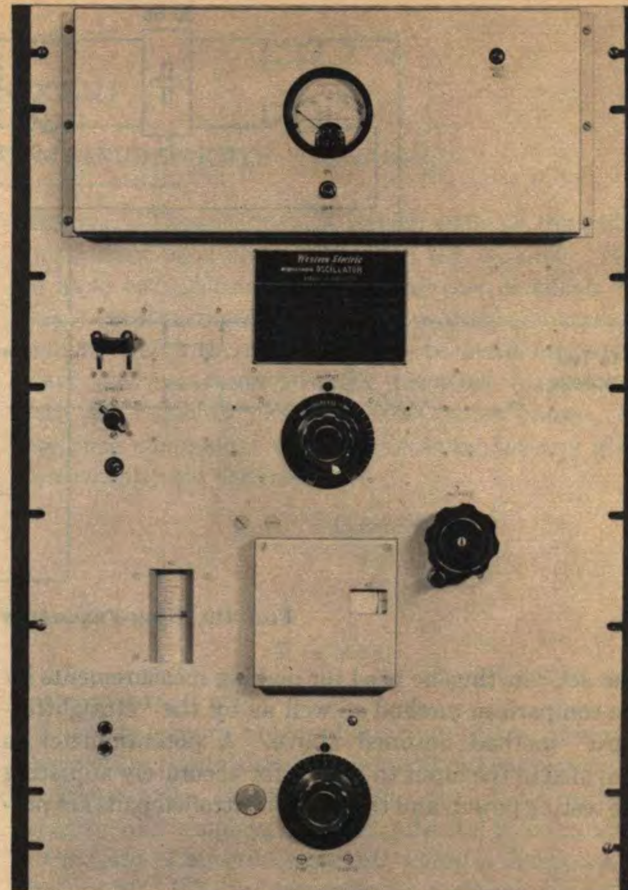
Because of the practical convenience of even scale meters, it is desirable to have a similar even scale on the decibel meter. This may be accomplished by varying the width of the air gap from one extreme position of the coil to the other as shown in Figure 418-B, so that the flux instead of being



Secondary Toll Testboard Showing Projection Meter for Transmission Measurements the same at all positions of the moving coil, increases logarithmically toward the position of greater loss. With this arrangement, the position for maximum current on the right-hand end of the scale is still marked zero db loss. But in order that the movement of the pointer along the scale shall be the same for each decibel, the air gap is so proportioned as to make the flux correspondingly larger toward the larger loss end, thus offsetting the decreasing increments of current. However, a meter of this type requires a higher operating current than the ordinary ammeter because the flux is less at maximum current.

202. Carrier-Frequency Transmission Measurements

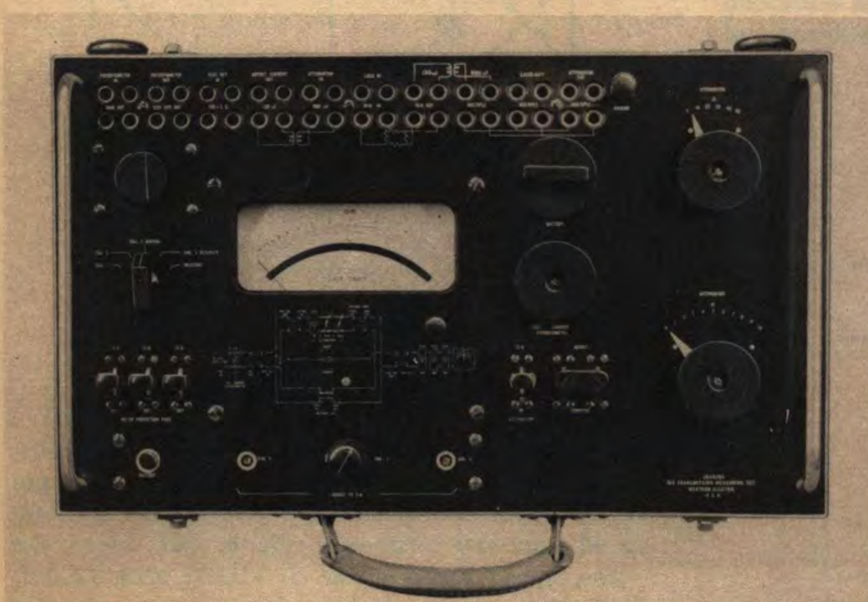
The principle employed in making carrier-frequency transmission measurements is generally similar to the comparison method for measuring voice-frequency circuits, as discussed in Article 199. In measuring a gain at carrier frequencies, a known input power is transmitted through a calibrated attenuator in series with the unknown gain, and on into a metering circuit which consists of a thermocouple and D.C. milliammeter. By adjusting the loss in the attenuator until the output of the unknown gain device is equal to the input power at the attenuator, the loss in the attenuator is made equal to the unknown gain. In the case of measuring a loss, the same basic principle is employed. The known input power is now transmitted through a calibrated attenuator, the unknown loss, and a fixed known gain. Then when the attenuator is adjusted so that the output of the known gain device is equal to the input power, the unknown loss is equal to the difference between the known gain and the calibrated loss.



High-Frequency Oscillator (50 to 150,000 Cycles) for Broad-Band Carrier Testing Work

A transmission measuring set using this principle was developed primarily for making carrier-frequency measurements on the Types-C, -J, and -K carrier telephone systems. It consists essentially of a thermocouple, milliammeter, attenuators, keys, jacks and repeating coils, arranged in such a manner as to facilitate the various types of tests which are required. Because of the wide range of frequencies involved heterodyne oscillators, capable of furnishing frequencies in the carrier range, are required for making measurements.

A simplified schematic of this measuring set is shown in Figure 419. Two paths, which are controlled by a switching key, are provided between the input to the set and the thermocouple and meter. One path contains an adjustable attenuator while both paths contain suitable jacks for inserting losses or gains to be measured.



Transmission Measuring Set for Testing Broad-Band Carrier Circuits

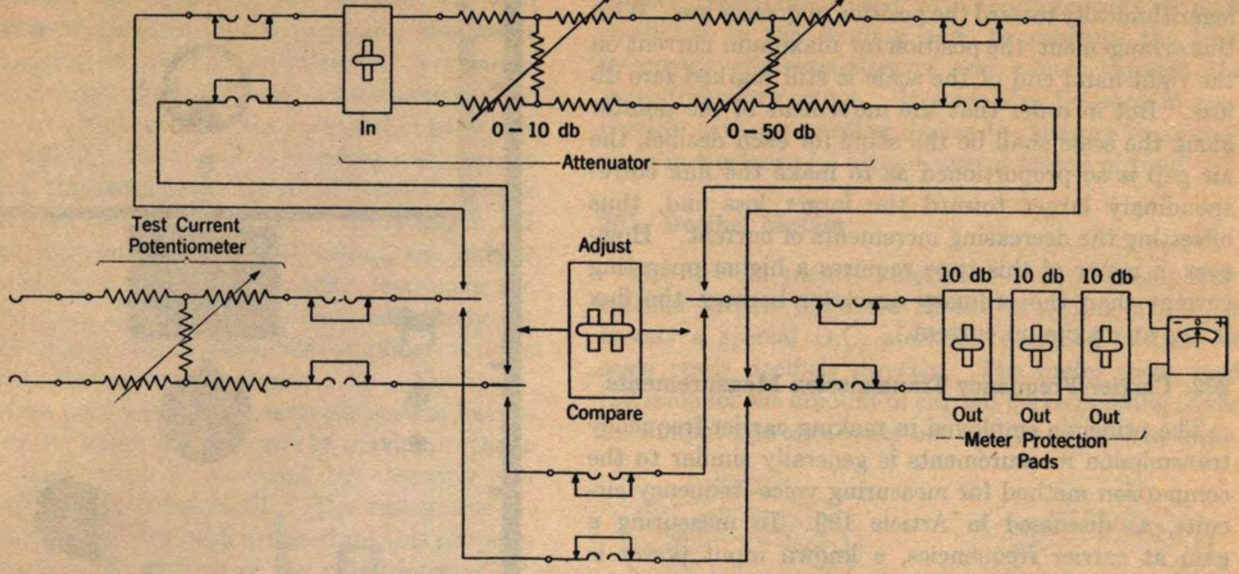


FIG. 419. HIGH-FREQUENCY TRANSMISSION MEASURING SET

The set can thus be used for making measurements by the comparison method as well as by the "straightforward" method outlined above. A potentiometer is provided in the input to the set for accurately adjusting the testing power, and three key-controlled pads are provided ahead of the meter for protection against burning out the sensitive thermocouple. The thermocouple and milliammeter are arranged to read with reference to a test power of one milliwatt. The set is calibrated on direct current from a dry cell battery contained in the set.

CHAPTER XXXIII

ALTERNATING-CURRENT TESTS AND MEASUREMENTS—(*Continued*)

203. The Impedance Bridge

In Chapter V we first encountered the Wheatstone bridge and learned the theory of "balance" which permits us to measure the D.C. resistance of any circuit. Later we saw that the principle of balance can also be applied to alternating currents and in Article 117 we discussed briefly an A.C. Wheatstone bridge circuit. In telephone work it is frequently necessary to make A.C. bridge measurements, and to this end we utilize the "impedance bridge" illustrated in Figure 420. It can be seen that this is merely a modification of the Wheatstone bridge, with an alternating source of E.M.F. substituted for the battery, a telephone receiver (where the frequencies are within the voice-current range) substituted for the galvanometer, and a variable impedance consisting of resistance and inductive or capacitive reactance, substituted for the adjustable resistance arm of the Wheatstone bridge.

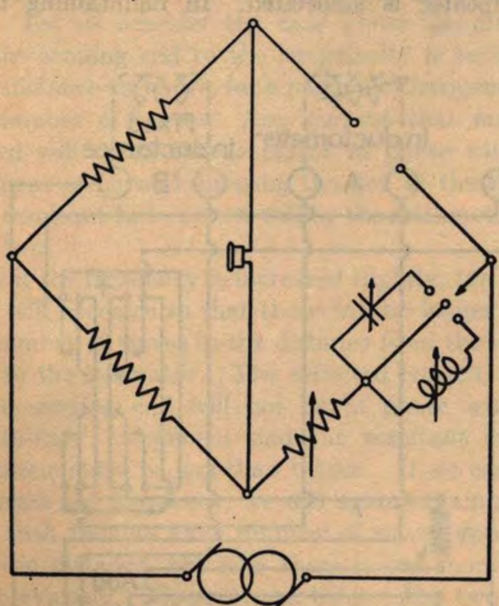


FIG. 420. PRINCIPLE OF IMPEDANCE BRIDGE

The similarity here needs no further discussion, but it will be noted that the construction of an adjustable impedance introduces the necessity of having some means of varying either or both the resistance and the reactance. The variable impedance arm must be adjustable with respect to two distinct components—resistance and reactance, and provision must be made to make the balancing reactance either positive or negative. Theoretically, a device such as that illustrated may be used in obtaining the balance. Here we have two distinct adjustments, one of which is a simple resistance and the other of which is an arrangement by which the reactance may be made any value, either positive (inductive) or negative (capacitive). With the capacity or inductance values known, the reactance component can be calculated for any given adjustment from the formula—

$$X = -\frac{1{,}000{,}000}{2\pi fC}$$

or

$$X = 2\pi fL$$

as the case may be.

There is a more practicable arrangement, however, which will accomplish the same result as that illustrated by Figure 420. It eliminates the use of the variable capacity and employs only an adjustable inductance in the form of an inductometer. This is illustrated in Figure 421. The inductometer is a device consisting of two similar coils, one of which may be rotated with respect to the other. Such an arrangement permits of a variation of inductance from almost zero, when the magnetic fields of the two coils oppose each other, to a maximum when the fields aid each other and the inductances add directly. The condenser is eliminated by introducing a transfer switch by which the in-

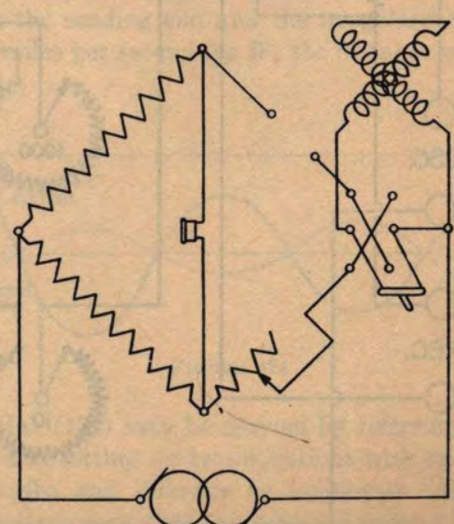


FIG. 421. IMPEDANCE BRIDGE WITH INDUCTOMETER

ductometer may be connected in series with the unknown impedance so as to neutralize a negative reactance component rather than balance it. Inasmuch as the inductometer is calibrated in terms of millihenrys or henrys, as the case may be, the reactance component must be calculated by the use of the formula—

$$X = 2\pi f L$$

This component will be negative when the switch is so thrown that the inductometer is in series with the unknown impedance, and positive when the switch is so thrown that the inductometer is in series with the adjustable resistance arm.

In the actual construction of an impedance bridge it is of course not possible to employ an inductometer having zero resistance. It is necessary, therefore, to have a compensating resistance which is similarly connected to switch contacts so as to always be in the opposite arm to the one in which the inductometer is connected. Figure 422 is a diagram of connections of the 1-B Impedance Bridge used in the telephone plant for measuring line impedances. Due to the fact that it is not possible to secure absolute zero inductance with the inductometer, a second fixed inductance is used, which may be switched from one arm to the other. Zero inductance can then be secured by throwing the fixed inductance on one side of the bridge and the inductometer on the opposite side, so that the fixed one neutralizes its value on the scale of the other. Since either inductance unit may be switched to either side of the bridge, it can be seen that the reactance values that can be measured range from zero to a value of $\pm 2\pi f\,(L_a + L_b)$, where L_a and L_b denote the values of the two inductance units. Like the inductometer, the fixed inductance also has a compensating resistance which is so wired to the transfer switch that it will always be in the proper arm of the bridge to balance the resistance of the coil.

A vacuum tube oscillator is used in connection with the bridge as the A.C. supply, and to permit measurements over the entire range of voice frequencies the oscillator has an adjustable resonant circuit which allows any desired frequency value within this range to be obtained. It is also equipped with a filter for eliminating the harmonics of the particular frequency used, thereby affording a sine wave testing current.

204. Line Impedance Measurements

In Chapter XXVIII we learned that the successful operation of 22-type telephone repeaters necessitates the use of a balancing network for each line with which the repeater is associated. In maintaining the re-

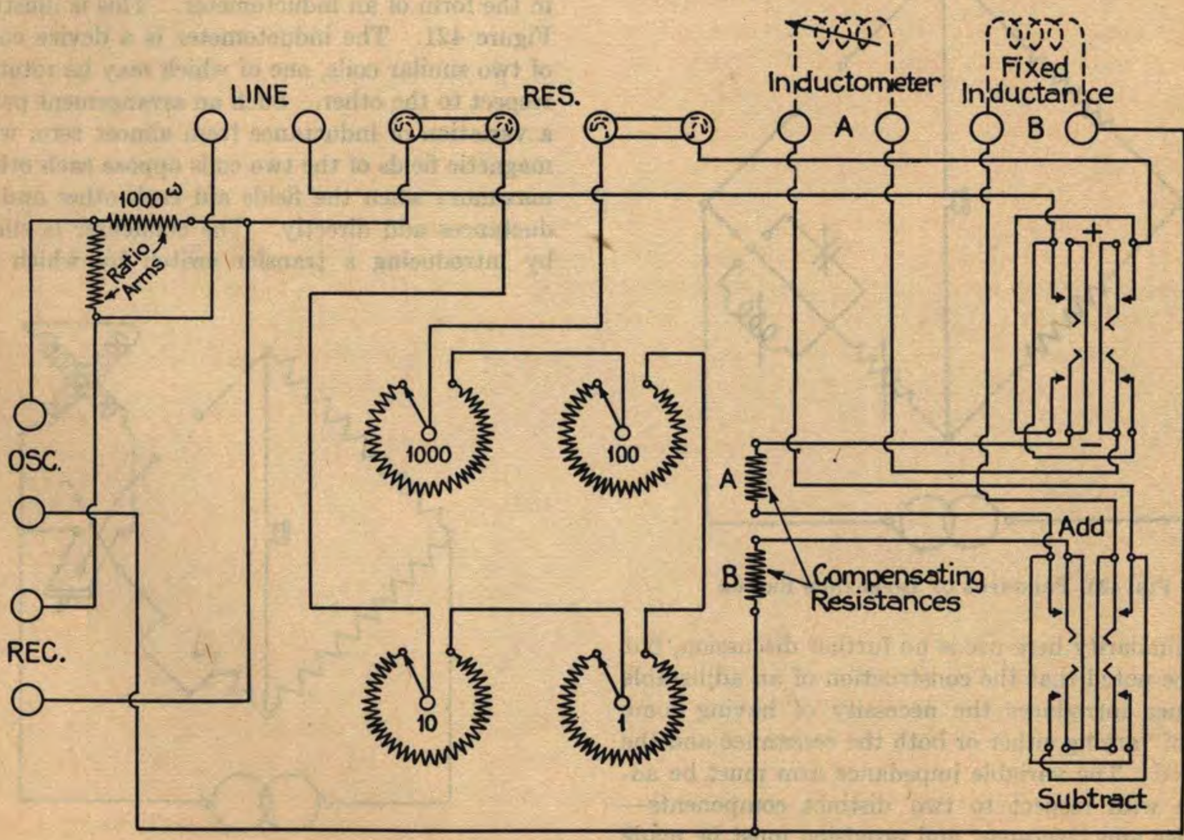


Fig. 422. Circuit of No. 1-B Line Impedance Bridge

quired degree of balance between the network and the line, measurements throughout the voice range of frequencies are sometimes required in order to check the impedance values and to locate any irregularities in the line circuit that may seriously affect the balance. Here an impedance irregularity means any condition throughout the length of the line that may cause a partial reflection of the electrical wave. From our earlier considerations of wave propagation we know that any wave will suffer both attenuation and phase displacement as it travels along a line, and this is, of course, also true of the reflected wave. Consequently, the magnitude of this wave and its phase position relative to the wave just leaving the generator, will change as it moves from the point of reflection to the sending end. Further, a little consideration will show that the phase of the reflected wave when it reached the generator will depend on the time it takes to travel from the irregularity to the generator, or what amounts to the same thing, on the distance from the irregularity to the generator.

For a clearer analysis of the effect of an irregularity on a line, let us resolve the current entering the line into two components, one the current that would enter the circuit if there were no irregularity, and the other the reflected current that is present due to the irregularity. Let us consider the case where the distance from the sending end to the irregularity is such that in this distance there are, for a particular frequency, an even number of waves. Any current that may be reflected will reach the generator in phase with the component of current entering the line at that point, so the resultant line current will be the arithmetic sum of the two.

Now if the frequency is increased slightly, the wave-length will decrease so that there will no longer be an even number of waves in the distance from the irregularity to the generator. The reflected current reaching the sending end will not be in phase with the "smooth-line" component and the resultant current will consequently be less than before. If we continue to increase the frequency, we will again obtain a condition such that an even number of waves appear in the given distance, but now there is one more wave than previously. However, as before, the two component currents, being in phase, combine arithmetically to give the resultant current. In other words, the sending end current will vary with frequency. Therefore, since $Z = E/I$, if we measure the impedance of the circuit, we will find periodic variations such as appear in the curve of Figure 423. This is a typical impedance curve obtained by making measurements with the impedance bridge on a circuit having an impedance irregularity. Here points, M, N and O correspond to frequencies where the distance to the ir-

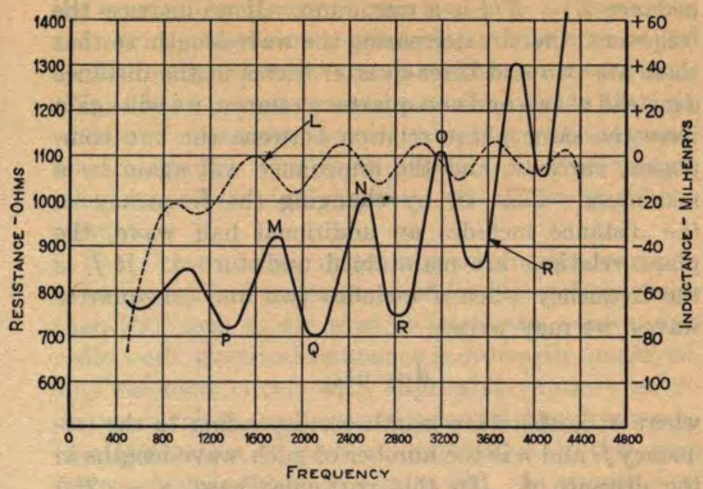


Fig. 423. Impedance Characteristic of Circuit with Single Irregularity

regularity is such that the reflected current is exactly opposite in phase to the outgoing current at the sending end, and points P, Q and R to frequencies where the reflected current is exactly in phase with the outgoing current at the sending end.

Such a curve permits a determination of the distance from the sending end to the irregularity by the use of the following formula:

$$d = \frac{W}{2(f_2 - f_1)} \qquad (176)$$

where d is distance to the irregularity, W is the average velocity of propagation of the telephone currents over the particular line in question, which may be taken from Tables XII and XIII, and $f_2 - f_1$ is the average frequency interval between the adjacent "humps" of the impedance curve. If, in the case of a loaded circuit, we employ the number of "loads per second" for the term W, d will give the number of loading points between the sending end and the irregularity. If we employ miles per second for W, the distance will be in miles, etc.

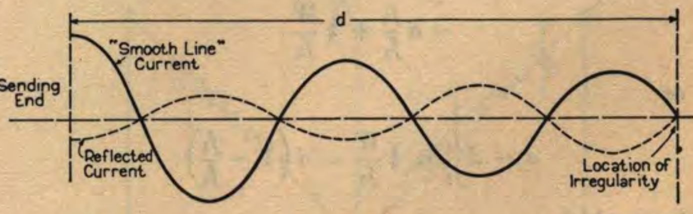


Figure 424

Equation (176) may be derived by referring to Figure 424 and setting up two equations with number of wave-lengths and distance as unknowns. Here the conditions are such that the reflected current is opposite in phase to the smooth-line component, so the impedance $Z = E/I$ is a maximum. If we increase the frequency, thereby decreasing the wave-length, so that there are two and three-quarter waves in the distance d instead of two and one-quarter as shown, we will again have the same phase relation between the two component currents, and the impedance will again be a maximum. That is, by changing the frequency so the distance includes an additional half wave, the phase relations are maintained undisturbed. If f_1 is the frequency when d includes two and one-quarter waves, we may write—

$$d = n\lambda_1$$

where λ_1 is the wave-length corresponding to the frequency f_1 and n is the number of such wave-lengths in the distance d. (In this particular case, $n = 2\frac{1}{4}$.) For the second case, where d includes an additional half wave-length, we may write—

$$d = (n + \tfrac{1}{2})\,\lambda_2$$

But

$$\lambda_1 = \frac{W}{f_1} \qquad \text{and} \qquad \lambda_2 = \frac{W}{f_2} \qquad \text{(from Equation 95).}$$

Substituting these values of λ_1 and λ_2 in the above equations, we have—

$$d = n\frac{W}{f_1}$$

$$d = (n + \tfrac{1}{2})\frac{W}{f_2}$$

Here we have two equations with two unknowns, namely d and n, and since we are not interested in n we will eliminate it; thus—

$$n = d\frac{f_1}{W}$$

Whence, we have—

$$d = \left(d\frac{f_1}{W} + \tfrac{1}{2}\right)\frac{W}{f_2}$$

$$= d\frac{f_1}{f_2} + \tfrac{1}{2}\frac{W}{f_2}$$

or

$$d - d\frac{f_1}{f_2} = \tfrac{1}{2}\frac{W}{f_2} = d\left(1 - \frac{f_1}{f_2}\right)$$

from which

$$d = \frac{W}{2f_2\left[1 - \dfrac{f_1}{f_2}\right]} = \frac{W}{2(f_2 - f_1)} \qquad (176)$$

To illustrate the use of Equation (176) let us consider a typical example.

Example: Calculate the distance to the irregularity on the 19-gage H-44 cable circuit whose impedance curve is shown in Figure 423.

Solution: On the R curve, we have "peaks" at 1000 and 3200 so that $(f_2 - f_1) = \frac{1}{3}(3200 - 1000) = \frac{1}{3} \times 2200 = 733$. Similarly from the L curve, $(f_2 - f_1) = \frac{1}{3}(3040 - 860) = 727$. Using the average value of $(f_2 - f_1)$ we have—

$$d = \frac{W}{2(f_2 - f_1)} = \frac{17640}{2 \times 730} = 12.1 \text{ loads, ans.}$$

As a matter of fact, the fault was located at the twelfth loading point.

It will be noted that the fault location secured in the above example is rather approximate. In modern practice considerably more accurate locations are desirable since clearing the trouble usually involves opening a cable splice. The standard instructions, covering fault location by means of impedance measurements, describe several refinements which may be applied to the basic technique described above in order to secure greater accuracy.

205. The Capacity Bridge

From the theory of the impedance bridge, it can be seen that it would be quite possible to measure capacity values with this device. However, since condensers are practically pure negative reactances it is more convenient to eliminate the variable resistance, thus simplifying the manipulation. Figure 425 shows the

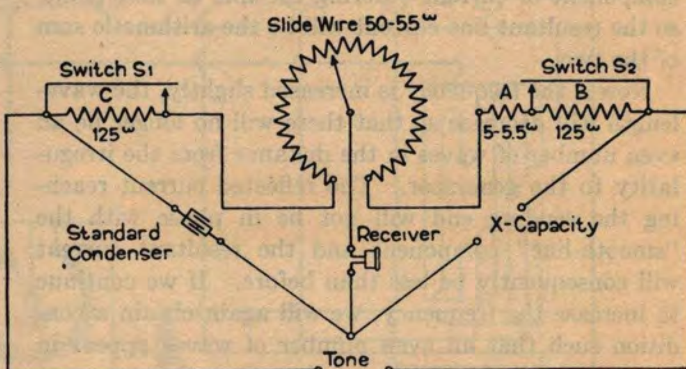


Fig. 425. Principle of the Capacity Bridge

impedance bridge when modified to measure capacities, in which form it is known as a "capacity bridge". Here, instead of having fixed ratio arms, we have variable arms and obtain balance by adjusting them until no sound is heard in the receiver. In this capacity bridge, the scale reading gives the ratio of the resistance arms, so the value of the unknown capacity is—

$$C_x = \text{Scale Reading} \times C_s$$

where C_s is the capacity value of the standard condenser.

206. Frequency Meters

As mentioned earlier, in alternating-current work we need to know the value of the frequency. There are several methods of determining this, the one employed in each case depending entirely upon the conditions. For frequency values lower than about 200 cycles, a simple vibrating reed device will give direct readings

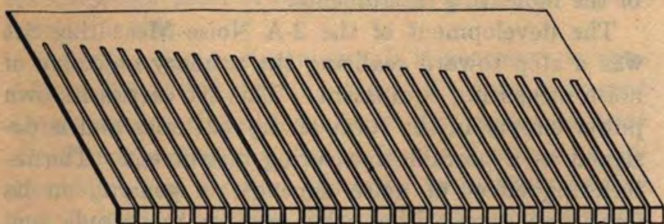


FIGURE 426

within a reasonable degree of accuracy. The theory of this instrument is illustrated by Figure 426. Here we have a comb shaped arrangement of metal reeds of varying lengths, each having a different natural period of vibration. A coil winding is connected to the circuit for which the frequency is to be determined and the core of this coil has long narrow pole pieces of the same length as the group of vibrating reeds. The reeds are

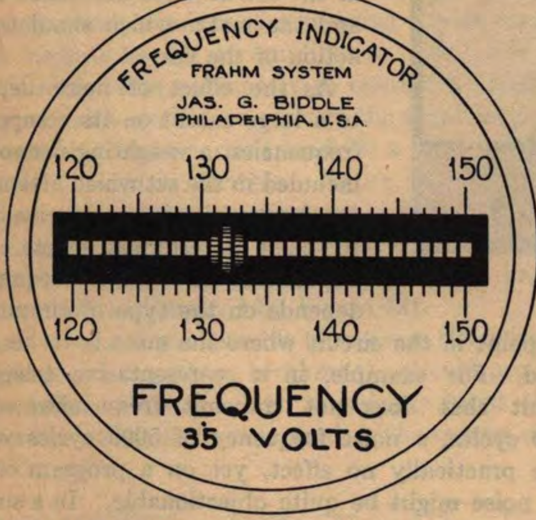


FIG. 427. VIBRATING REED FREQUENCY METER

inserted in the air gap between the pole pieces. There is a magnetic attraction, therefore, for each reed due to the alternating current in the coil, in the same way that there is magnetic attraction for the diaphragm of a telephone receiver. The particular reed which has a mechanical period of vibration corresponding to the current frequency will vibrate but other reeds will not respond to the magnetic pulses. The ends of the reeds are aligned under a scale as illustrated by Figure 427, and the frequency can be read by noting the long white line created by the reed vibrating with the greatest amplitude. This form of meter is employed in telephone offices for checking the 135-cycle ringing current.

The vibrating reed meter depends for its operation upon mechanical resonance, but for higher frequencies than 200, such as are used in telephone, carrier, and radio work, electrical resonance is ordinarily employed. Any resonant circuit with adjustable capacity or inductance values can be used for determining frequency. One illustration of such an application is the wave meter which is shown in Figure 428. In circuit A, if

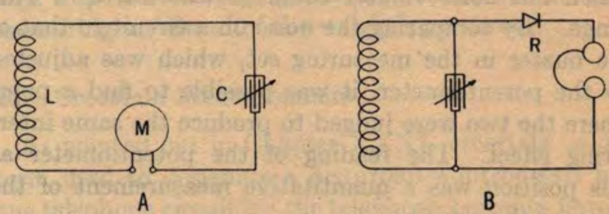


FIG. 428. WAVE METERS

an inductance L and a condenser C are so adjusted as to give resonance, the meter M will give a maximum reading. In circuit B, the resonant condition will obtain when maximum sound is heard in the telephone receivers which are connected in series with a rectifying device, R.

Instead of using the resonant circuit in the manner illustrated by Figure 428, it may be used in connection

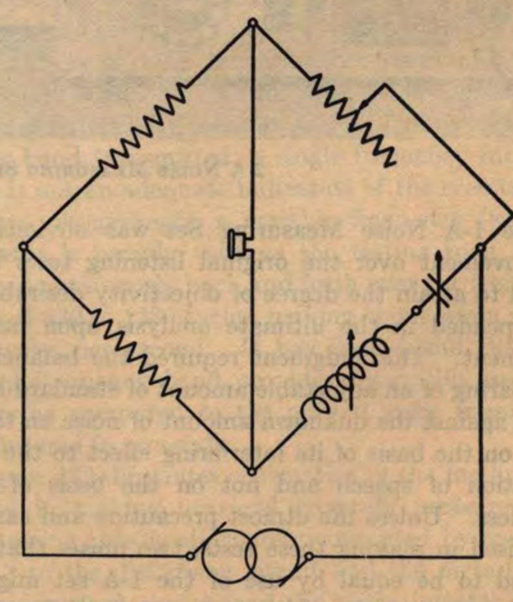
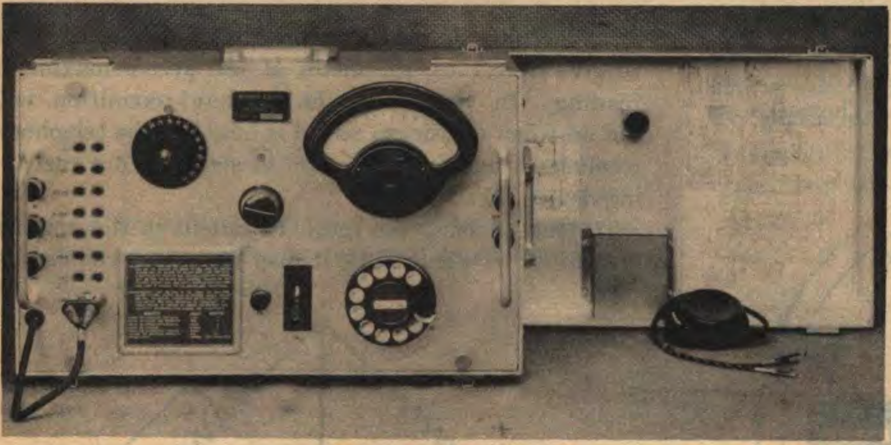


FIG. 429. IMPEDANCE BRIDGE ARRANGED FOR MEASURING FREQUENCY

with the impedance bridge, as illustrated by Figure 429. Here when the variable capacity and the variable inductance are resonant to the impressed frequency, it is possible to secure a bridge balance by means of the variable resistance arm alone. This scheme may be employed for calibrating vacuum tube oscillators.

207. Noise Measurements

From the early days of telephone service, some means of measuring the noise on telephone circuits has been in use. This at first consisted of merely listening on the circuit with a telephone receiver, which obviously resulted in a wide variety of ideas as to the amount of noise that was present. This led to the use of the 1-A Noise Measuring Set which consisted of a "buzzer" to produce a noise, and a potentiometer with which this noise volume could be varied over a wide range. By comparing the noise on a circuit to that of the buzzer in the measuring set, which was adjusted by the potentiometer, it was possible to find a point where the two were judged to produce the same interfering effect. The reading of the potentiometer at this position was a quantitative measurement of the noise on the circuit.



2-A Noise Measuring Set

The 1-A Noise Measuring Set was obviously an improvement over the original listening tests but it failed to attain the degree of objectivity desirable, for it depended in the ultimate analysis upon personal judgment. This judgment required the balancing or comparing of an adjustable amount of standard buzzer noise against the unknown amount of noise on the circuit, on the basis of its interfering effect to the interpretation of speech and not on the basis of equal loudness. Unless the utmost precaution and care was exercised in making these tests, two noises that were judged to be equal by use of the 1-A set might be found to have quite different interfering effects on a transmitted conversation.

These objections to the 1-A set were overcome to a large extent by the development of the 1-A Noise Amplifier. By using this amplifier in conjunction with a standard transmission measuring set, the ear was eliminated as the basis for determining the magnitude of the noise. This amplifier employed "frequency weighting" which was based on the information available as to the relative interfering effects of single frequency tones. While this amplifier was quite an improvement it fell short of being the ideal in other ways, such as portability and dynamic characteristics of the indicating instruments.

The development of the 2-A Noise Measuring Set was a step toward realizing the primary objective of noise measuring equipment. This set carries its own power supply in the form of dry batteries and is designed as a portable measuring instrument. The interfering effect of noise depends, in general, on its frequency composition, the relative magnitude and spacing of the various frequencies, the time duration of the noise, the type of circuit on which it occurs, and the person who is listening to it. These factors obviously complicate the design of a satisfactory noise measuring set. It is desirable to employ a meter to give the indication of noise; but before such a method can be satisfactorily used, the interfering effect of different noises on speech as heard by the ear must be determined, and then circuits must be incorporated in the set between the noise input and the meter which simulate the action of the ear.

As the effect of noise depends to a large extent on its component frequencies, a weighting network is included in the set which attenuates the frequencies in an inverse ratio to their interfering effects. The importance of various frequencies depends on the type of circuit and the point in the circuit where the noise is to be measured. For example, in a representative telephone circuit that does not transmit frequencies above 3000 cycles, a noise frequency of 5000 cycles would have practically no effect, yet on a program circuit this noise might be quite objectionable. In a similar manner, the subscriber's telephone subset which includes an induction coil and a condenser, affects the frequencies transmitted primarily in the low-frequency range. Therefore, the weighting that should be made in the measuring set would be different depending upon whether the measuring set was to be used at a subscriber's receiver or at a central office.

In the event the noise was of a single frequency, the weighting networks would insure that its measured effect would be the same as that determined by the average ear. The noise, however, usually comprises many frequencies, and the action of the ear must be considered in combining these different frequencies. The copper-oxide rectifier, which converts the amplified noise currents into direct current to operate the meter, is designed to combine the weighted noise currents in such a manner that the sum is approximately equal to the square root of the sum of the squares of all the weighted components. It has been found that the ear combines most ordinary noises in a very similar manner to this.

Following these adjustments, it is necessary to consider the effect of the time interval the noise is present on the circuit. It has been found that most telephone circuit noises must exist for at least 0.2 second to appreciably affect the ear. In order to incorporate this time factor in the noise measuring set, the meter has been given such dynamic characteristics that a noise duration of 0.2 second produces about the same deflection as a continuous noise of the same intensity. Noises of shorter duration produce proportionally smaller deflections.

The complete 2-A Noise Measuring Set includes (1) an input circuit which presents the proper impedance to the circuit to which the set is to be connected; (2) weighting networks for frequency discrimination so that for each type of measurements, the noises of different frequencies will be measured in accordance with their interfering effects; (3) a graduated potentiometer which provides a means of measuring a wide range of noise magnitudes; (4) an amplifier to raise the level of the noise currents sufficiently to operate a meter; (5) a copper-oxide rectifier to convert the amplifier output currents into direct current; and (6) a noise indicating meter calibrated in decibels. Figure 430 shows schematically this circuit arrangement. When measuring noise on ordinary types of telephone circuits at the central offices, the weighting network used has the characteristics illustrated by Figure 431.

As it is sometimes desirable to measure the noise on a telephone circuit without breaking the circuit, provision is made for bridging a high impedance input circuit across the circuit to be measured. The output of this high impedance bridging circuit connects to the noise measuring set ahead of the weighting network.

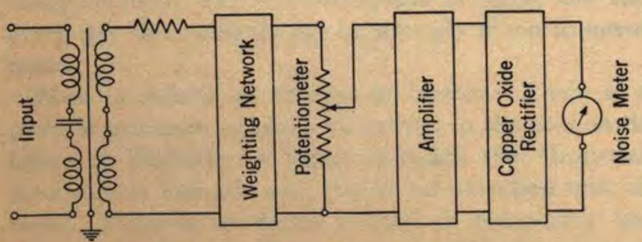


Fig. 430. Circuit of 2-A Noise Measuring Set

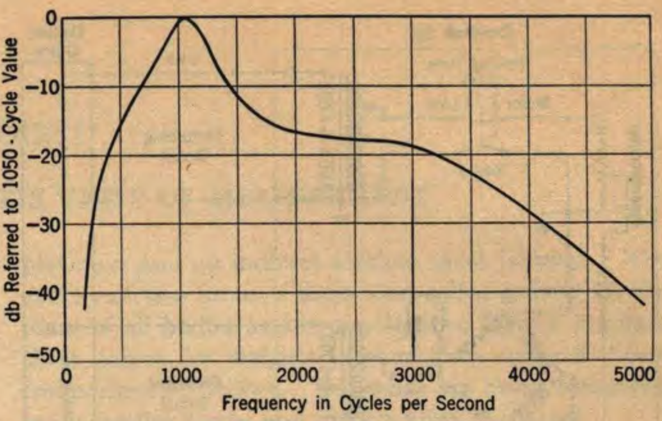


Fig. 431. Characteristic of Weighting Network of 2-A Noise Measuring Set

208. Crosstalk Measurements

As pointed out in Chapter XXXI, crosstalk is the term used to designate a disturbance introduced into one telephone circuit by the telephone currents flowing in another. This disturbance may be caused by inductance, capacitance, or resistance coupling between the two circuits, or by combinations of these three general types of coupling.

The amount of crosstalk induced from one circuit into another is, of course, determined by the degree of coupling between the two circuits. A measurement of this coupling in terms of loss is therefore a measurement of the crosstalk. If the loss is low, the crosstalk is obviously large, and vice versa. While this loss could be measured with a standard transmission measuring set, using a calibrated amplifier in the receiving side if the loss was beyond the measuring range of the set, the results would only be for single frequency values. As the crosstalk may occur at practically any frequency in the band transmitted, a single frequency measurement is not an adequate indication of the crosstalk intensity. Accordingly, a variable frequency (warbler) oscillator is usually used as the testing tone, which sweeps continuously back and forth over the frequency range of 830 to 1230 cycles, making about seven sweeping cycles per second. It has been found that this testing frequency band provides reasonably accurate results as compared to the normal voice frequencies encountered in crosstalk.

Figure 432 illustrates schematically the method employed in making near-end crosstalk measurements. The testing tone is alternately connected by means of Key 3 to the disturbing circuit and the metering (or loss measuring) circuit, and the potentiometer (crosstalk meter) in the measuring circuit is adjusted until

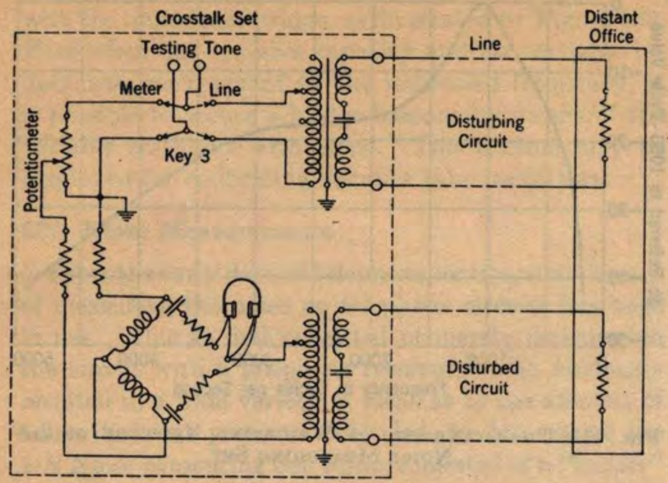


Fig. 432. Crosstalk Measuring Set Arranged for Near-end Crosstalk Measurement

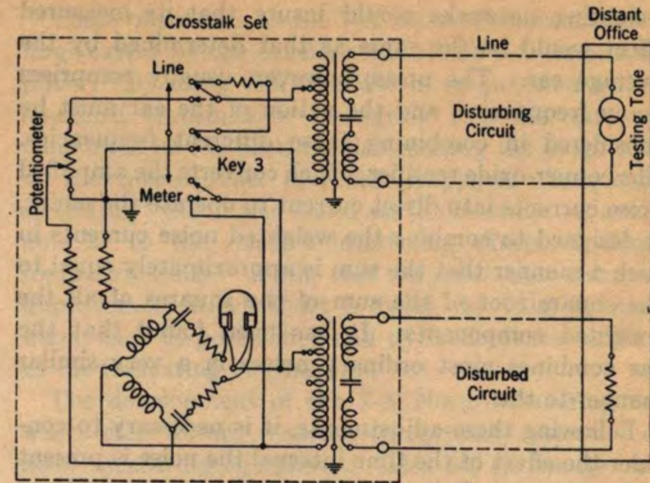


Fig. 433. Crosstalk Measuring Set Arranged for Far-end Crosstalk Measurement

the volume of the tone heard in the receivers is the same for both positions of the key. When the volume of the testing tone is the same over both paths, the loss injected by the potentiometer is obviously equal to the coupling loss between the two circuits. The setting of the potentiometer then gives a quantitative measurement of the near-end crosstalk from the disturbing to the disturbed circuit. By having this potentiometer calibrated directly in Crosstalk Units and Decibels, a direct reading is obtained.

It is important to note that in making near-end crosstalk measurements, any line noise on the disturbed circuit is present in the receivers when listening to the testing tone for both positions of Key 3. Therefore, such noise has no effect on the measurements.

In making far-end crosstalk measurements, the testing tone must be applied at the far-end of the disturbing circuit. The circuit arrangement used is illustrated in Figure 433. By operating Key 3, the disturbing circuit is connected to the calibrated potentiometer (crosstalk meter) where sufficient loss may be added to attenuate the testing tone to the same magnitude as that present in the receivers when Key 3 is operated to terminate the disturbing circuit. The loss measured by the potentiometer is then the measured far-end crosstalk. (As was pointed out in Article 195, it is necessary to add the line loss of the disturbing circuit to this value in order to get the true far-end crosstalk.)

In making crosstalk tests at frequencies above the voice band, some provision must be made in order that the crosstalk frequencies, which are above the audible range, may be heard in the regular telephone receiver. To accomplish this a detector-amplifier is usually employed, which functions as a heterodyne detector in that the presence of the high-frequency currents in its input is made known by an audible frequency in the receiver. This audible frequency is produced by beating the incoming high frequency with that of an oscillator included within the detector-amplifier itself. The arrangements illustrated in Figures 432 and 433 for near- and far-end crosstalk measurements, respectively, are used with the addition of a detector-amplifier in making crosstalk tests at these higher frequencies. The input test tone is supplied at a single frequency. A series of tests at various frequencies is normally made to determine the relative amount of crosstalk throughout the frequency band being considered.

APPENDIX I

PHYSICAL QUANTITIES AND THEIR UNITS OF MEASUREMENT

1. Mechanical Quantities

In addition to the general physical quantities, **force, work** and **energy** the essential ones belonging to the mechanical group are **time, length, area, volume, mass, density, speed, acceleration, and power.** The combined group consists of two classes, viz. (a) those that are fundamental, and (b) those that may be defined from their relations to the fundamental ones.

In dealing with any one physical quantity we need first to know its definition or exact nature and next the method whereby it may be measured. Its measurement will always require a comparison with some unit which is either a fixed standard or may be derived from some fixed standard or standards. In addition to this it may sometimes require the fixing of a positive or negative sign to the numerical size of the quantity.

2. The Three Fundamental Units

There are three units that are fundamental to all mechanical measurements. We may either use these directly to measure any quantity in which we are interested or to derive other units with which the quantity may be measured. These fundamental units, which we must preserve as standards and which must be remembered at all times because they cannot be produced mathematically, are **the unit of time, the unit of length and the unit of mass.**

We daily express periods of time by comparison with familiar units such as second, minute, hour, day and year and have vivid conceptions of their size or greatness, for example: 10 minutes, 2 hours, 3 years, etc. We likewise express distances by comparing them with simple units of length such as inch, foot, yard, mile, etc.

Though the fundamental units of time and length are too familiar to require discussion, the conception of the third fundamental unit or **the unit of mass** is easily confused with the conception of force, and the every day term **weight** may be wrongly taken to mean **mass:**

Mass is defined as amount of matter. There is a piece of platinum carefully preserved in the Standards Office at Westminster which is called the "Imperial Avoirdupois Pound", and this is our standard unit of mass. There is no direct method of comparing the mass of bodies of other material with this piece of platinum and an indirect method must be used. The earth exerts a force of attraction called **gravity** on the mass of all bodies and in practice we merely compare these forces (or **weights**) rather than make a direct comparison of masses. But what are being compared are in reality forces and not amounts of matter.

From the same units with which we express distances, that is, from the units of length, we may derive mathematically units for area or volume. For area, such units are the square inch, the square foot, the square mile, etc. and for volume they are the cubic inch, cubic foot, cubic yard, etc.

3. Density and Specific Gravity

Density is defined as the mass per unit volume of a substance. To measure it we employ the unit of volume (length $\times$ length $\times$ length) and the unit of mass, i.e., mass divided by volume. In practice we seldom use density in the absolute sense but use instead **specific gravity.** The specific gravity of any material is the ratio of the weight of a given volume of that material to the weight of an equal volume of water. For example, the specific gravity of cast iron, a cubic foot of which weighs approximately 500 lbs., is 7.7 since a cubic foot of water weighs approximately $62\frac{1}{2}$ lbs. Likewise the specific gravity of the electrolyte (diluted sulphuric acid) which is ordinarily used in storage batteries is about 1.200, which means that any unit of its volume will weigh 1.2 times as much as the same volume of water.

4. Velocity or Speed

We may express **speed** or **velocity** as the time rate of traveling **distance.** If any body in motion, such as a train, continues its motion without change for one unit of time, such as the hour, the distance is a quantitative measure of its speed. For example, a train's speed may be thirty miles per hour which means if it keeps moving for one hour with the same speed as at the instant observed, it will cover a distance of thirty miles. This does not mean, however, that the train will keep moving for one hour or that it will eventually move the distance of thirty miles; it merely means that if the conditions under which the train is moving at the instant observed are unchanged during a one hour period the train will traverse a distance of thirty miles.

5. Acceleration

If the train at any time should either acquire **more** speed or "slow down" it would be accelerated. **Acceleration** is defined as the time rate of changing speed. A train is positively accelerated when getting up to speed and negatively accelerated when slowing down. This quantity is an ideal example of a measurement requiring more than a mere numerical comparison. If in one minute's time the train should **increase** its speed one mile per minute, it would have one unit's **positive** acceleration or an acceleration of **plus** one mile—per minute—per minute; if in the same length of time it **decreased** its speed one mile per minute, it would have **negative** acceleration of one mile—per minute—per minute.

6. Force

Mass, such as the piece of platinum already mentioned, is drawn toward the center of the earth by the force of gravity. If it were free to fall, the force of gravity would give to it an acceleration of about 32 feet—per second—per second; that is, at the end of the first second from the time it started to fall it would have a velocity of 32 feet per second and at the end of the next second a velocity of 64 feet per second, etc. That influence which tends to set any body in motion or to change the direction or speed of any body already in motion is called **force.** The **unit** of **force** is that force which the influence of gravity exerts on the standard pound of mass when at sea level. It is called the **pound of force.**

When the forces acting upon a body in any one direction are equal to those acting in the opposite direction, the body is in equilibrium or at rest. When these forces become unbalanced, the body is set in motion and work is performed.

7. Work

Work is done when a force moves a body **in the direction of the force.** It is measured by the product of the force and the distance through which it acts. If a pound of force acts upon a body through a distance of one foot, one **foot-pound** of work is performed. If a six pound hammer is raised from the floor to a bench three feet in height, six times three, or eighteen foot-pounds of work is performed. If the vertical distance between the two floors of a building is 12 feet and a man weighing 150 pounds ascends from one floor to another he performs 12×150 or 1800 foot-pounds of work. Here it should be noted that the force acts through the vertical distance only and not through any horizontal distance he may travel while ascending flights of stairs. The distance must be measured parallel to the direction of the force.

Work in its scientific sense is measured by the result and not by effort or exertion. A man may attempt to move a weight until he is fatigued, but he performs no **mechanical work** unless he succeeds in elevating **the weight** against the force of gravity, changing its position in a given plane against the friction and inertia tending to hold it at rest, or overcoming any other resisting forces that may be acting upon it.

8. Power

In the same sense that speed is the time **rate** of traversing distance, **power** is the time **rate** of **doing work.** Its unit of measurement, therefore, not only requires the fundamental units of foot and pound for its derivation but must include time also. If a machine is capable of performing 33,000 foot-pounds of work, we have no conception of the size of the machine. It may be a small machine capable of performing this amount of work in fifteen days or a large machine capable of performing the work in three seconds, but a machine that can perform **33,000** foot-pounds of **work in one minute** is rated at **one horse power.**

9. Energy

Energy is ability to do work or is stored work and may therefore be expressed in the same units as work. A suspended weight or a fly wheel in motion are said to have stored mechanical energy by virtue of their ability to perform useful work.

Energy also exists in forms other than mechanical. A strong acid may have the ability to dissolve metals; a solution may have the ability to generate an electric current when in contact with two dissimilar metals. Food, when digested, permits man to perform useful tasks. These are examples of ability to perform work, but the **energy** is in a **chemical** and not a **mechanical state.**

Another very common form of chemical energy is that of coal which when burned gives off large quantities of heat. Heat is within itself a form of energy and may be made to perform work. It may be applied to a boiler in such manner as to generate steam under pressure and the steam may be used to operate an engine or turbine. The amount of **heat** required to raise the temperature of one pound of water one degree Fahrenheit is the **British Thermal Unit.**

The following have approximately equal energy values, viz.: 10,000,000 foot-pounds:

1. One pound of coal.
2. One day's supply of food for adult man of average muscular vigor.
3. Medium sized 24-volt storage battery fully charged.
4. Amount of water in average railroad water tank when elevated from ground level.

Energy may be converted from one form to another but **can never be produced or destroyed.** This does not mean, however, that one pound of coal could be made to pump sufficient water to fill the railroad water tank mentioned in the foregoing. The best known devices for converting heat or chemical energy into mechanical energy are very inefficient and only a small part of the coal's energy value can be made to produce useful work, the greater portion being lost by heat radiation or conduction rather than utilized by conversion.

10. The Metric System of Measurements

The system of measurements used by all civilized countries except Great Britain and the United States is called the **metric system.** It is based upon multiples of ten or upon the decimal plan, in a manner similar to that of our money system in this country. On account of its marked advantages over our awkward pound, foot, acre, quart, bushel, etc., it has been standardized for all scientific work, and consequently the electrical units are based upon this more convenient system. In the same manner that our money system greatly facilitates calculations by being based on the multiples of ten (namely, mil, cent, dime, dollar and eagle) the metric system provides units for measurement of all physical quantities which facilitate simple calculations involving these quantities.

A system of prefixes is used in the metric system to indicate multiples of ten or decimal parts. This is illustrated in Table I, which gives the metric units for measurements of length.

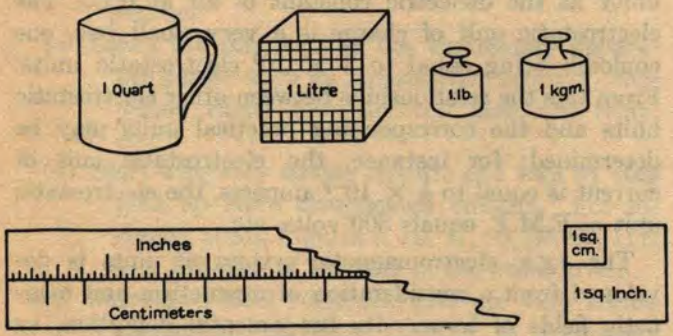


Fig. 1. Comparison of English and Metric Units

The measurement of volume and liquid capacity is based upon the litre, which is **a cube** with each dimension ten centimeters in **length.** It is, therefore, equal to 1000 cubic centimeters. It can be remembered as being a little larger than our quart.

The measurement of **mass** is based upon the **gram,** a very small unit and approximately equal to the mass of one cubic centimeter of water. It can be remembered as equal to one-fifth of our five-cent piece. In other words, a U. S. five-cent piece weighs 5 grams. The kilogram or 1000-gram weight is more commonly used. It can be best remembered as being a little larger than two pounds.

TABLE I

Metric Units of Length

Unit	Decimal
1 Millimeter	.001 meter
1 Centimeter	.01 meter
1 Decimeter	.1 meter
1 Meter	
1 Dekameter	10. meters
1 Hectometer	100. meters
1 Kilometer	1000. meters

There is only one system for measuring **time,** and the **second** has the same meaning in the scientific system as in our system.

Temperature is based upon the Centigrade scale instead of the Fahrenheit wherever the metric system is used. Zero degrees on the Centigrade scale corresponds to freezing point of water at sea-level atmospheric pressure (14.7 lbs. per square inch) or 32 degrees on the Fahrenheit scale; 100 degrees on the Centigrade scale is the temperature of boiling water and corresponds to 212 degrees on the Fahrenheit scale. All intermediate degrees are proportional, which means that one degree Fahrenheit is equal to five-ninths degree Centigrade.* Figure 2 illustrates the thermometer showing both scales.

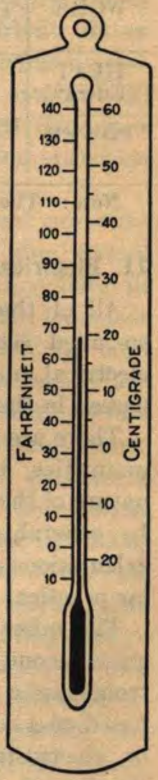


Fig. 2. Comparison of Fahrenheit and Centigrade Scales

Table II gives the conversion factors between the English and metric systems for the units of various physical quantities. Included in this table are several units that have not been discussed in the preceding pages, such as watt, dyne and joule, but the significance and use of such units is explained in the body of the text. Particularly important from our point of view are the relations between the watt or kilowatt and the horse power, and between the joule and the foot-pound.

* For converting temperatures from Fahrenheit to Centigrade the above relations may be expressed as a formula—

$$C^\circ = (F^\circ - 32)\ 5/9$$

and for converting Centigrade to Fahrenheit—

$$F^\circ = (C^\circ \times 9/5) + 32$$

TABLE II

Relation Between Metric and English Units

QUANTITY	CONVERSION FROM METRIC TO ENGLISH				CONVERSION FROM ENGLISH TO METRIC	
	Name	Abbreviation	Value	English equivalent	Name	Metric Equivalent
TIME	Second	sec.	(Same in both systems)		second	(Same in both systems)
LENGTH	centimeter meter kilometer	cm. m. km.	.01 meter — 1000 meters	.39 inch 1.094 yards .625 mile	inch foot yard mile	2.54 cm. .305 meter .915 meter 1.609 kilometers
MASS	gram kilogram	gm. kg.	— 1000 grams	.035 ounce (avoir.) 2.204 pounds	ounce pound	28.35 grams .454 kilogram
AREA	square-centimeter square-meter	sq. cm. sq. m.		.155 sq. in. 1.2 sq. yd.	square-inch	6.45 sq. cms.
VOLUME	cubic centimeter litre	cu. cm.	.001 litre 1000 cu. cms.	.061 cu. in. 1.06 quarts	cubic-inch cubic-foot gallon	16.39 cu. cms. 28.3 litres 3.8 litres
FORCE	dyne (seldom used) gram (of force) kilogram (of force)		.00102 gram 980 dynes 1000 grams	.035 ounce (of force) 2.20 pounds (of force)	pound (of force)	.454 kilogram (of force)
WORK	erg (seldom used) joule		1 dyne-centimeter 10,000,000 ergs	.738 ft. lb.	foot-pound	1.35 joules
HEAT (Energy)	gram-calorie large-calorie	gm.-cal.	gm. of water 1° C 1000 calories	3.087 ft.-lbs. 3.963 B.t.u.'s	British thermal unit	1055 joules 255 gram-calories
POWER	watt kilowatt	 kw.	joule per second 1000 watts	.74 ft.-lb. per sec. 1.34 horse power	horse power	746 watts or $\frac{3}{4}$ kw. approximately

Note:—The decimal fractions in the above table are not carried further than is essential for the applications made in this text.

11. Electrical Units

All of the practical units as well as some of the so-called scientific or absolute units for measuring electrical and magnetic quantities are defined and discussed in the body of the text.

There are three systems of units for measuring such quantities, the utility of each depending upon the nature of the calculations or measurements to be made. In general, except for certain kinds of theoretical calculations and laboratory experimentation, the familiar practical system is to be preferred.

The other two systems are the c.g.s. (centimeter-gram-second) electrostatic system and the c.g.s. electromagnetic system. The first of these systems is based on a consideration of electric charges and electric or electrostatic fields. As is well known electric charges or so-called "static electricity" may be produced on the surface of substances such as amber, rubber or glass by rubbing them vigorously with dry silk or fur. Such charges are accompanied by an electric or "static" field of force which, acting in the space around a charged body, causes it to attract non-charged bodies such as small bits of paper and to attract or repel other charged bodies, depending on whether the other bodies are charged with electricity of the opposite or the same sign. An electric field is always present wherever electricity is moving or tending to move. The electrostatic unit of charge or quantity of electricity is that possessed by each of two bodies which repel or attract each other with a force of one dyne when the bodies are one centimeter apart. This definition depends upon the arbitrary selection of unity as the dielectric constant of air at 0°C. The electrostatic unit of charge is a very small one, one coulomb being equal to 3×10^9 electrostatic units. From this the relationships between other electrostatic units and the corresponding practical units may be determined; for instance, the electrostatic unit of current is equal to $\frac{1}{3} \times 10^{-9}$ amperes, the electrostatic unit of E.M.F. equals 300 volts, etc.

The c.g.s. electromagnetic system of units is developed from a consideration of magnetism and magnetic fields of force. Its fundamental definition, as given in the text, is that of the unit of magnetic pole strength, i.e., unit pole strength is that possessed by each of two magnetic poles that repel or attract each other with a force of one dyne when placed one centimeter apart. This definition depends upon the arbitrary selection of unity as the permeability of air at 0°C. The c.g.s. electromagnetic unit of E.M.F. is equal to 10^{-8} volts, while the electromagnetic unit of current equals 10 amperes. Other relationships between electromagnetic and practical units may be derived from these.

APPENDIX II

CIRCUIT DIAGRAM READING

A casual glance at a complicated telephone circuit drawing often leads the student to think that circuit reading is like a strange language acquired only by the most difficult study. He is confronted with a maze of apparatus parts connected by a complex entanglement of wires, and he despairs at the thought of memorizing hundreds of such circuits, each different from the others. Fortunately, this impression is more or less of an illusion. Though circuit reading is somewhat of a "knack", it depends for the most part upon skill to be acquired by learning the underlying principles step by step. Furthermore, while the best telephone men may diligently study these principles and apply them in practice, they never attempt to memorize circuit drawings. The large number of complicated circuits that are essential to the proper handling of telephone service can never be mastered in this way.

Accordingly, the following hints are given not as a key to every telephone circuit, but as a review of the underlying principles and as a recommended procedure for those not familiar with toll central office circuits:

(a) **Learn the principles of current flow including Ohm's and Kirchoff's Laws.**

(b) **Memorize the conventional circuit symbols for the commonly used units of telephone apparatus.**

(c) **Learn the functions of the elementary apparatus parts**—such as transmitter, receiver, condenser, ringer, induction coil, jack, key, relay, etc.

(d) **Start with very simple circuits and step by step take up more complicated circuits.** For instance, study Figures 7, 10, 17, 20, 80, 81, 111, 112, 113, etc. of the text in the order given.

(e) **Remember that voice currents are very feeble alternating currents ranging in frequency somewhere between 200 and 3000 cycles, that ringing currents are alternating currents of 20, 135, or 1000 cycles, and that the majority of relays are operated with direct currents.** On most drawings, the direct path of voice currents is shown by heavy solid lines and the path for direct or ringing currents, when not over the same conductors as voice currents, is shown by light lines.

(f) **Bear in mind that direct currents will not flow through condensers and will flow through retardation coils and that alternating currents will effectively flow through condensers but are greatly reduced in their flow through retardation coils. Also, that direct current will not flow from one winding to another of an induction coil, repeating coil or transformer, and that alternating currents will flow, or to be more exact, are induced across from one winding to another of these coils.**

(g) **Learn first the purpose of operation, or in other words, that which the circuit is designed to accomplish.** For example, in Figure 113 of the text we might say that the circuits shown are designed to permit local telephone connections where one operator can establish all connections and the circuit design must permit common battery single position operation as follows:

Subscriber takes receiver off hook and subset circuit in conjunction with line circuit must give operator signal by lighting lamp in face of switchboard in front of her. Operator must be able to answer subscriber by connecting her telephone set to some one of her cord circuits and connecting this cord circuit to calling party's line. Line circuit must contain relay features which will extinguish lamp when operator makes this connection. Cord circuit must permit operator to make similar connections to called party's line and to ring called party. Ringing current must not reach ear of operator or calling party. Operator must have signal associated with cord circuit that will tell her when called party has answered and when both parties are through talking and hang up. The system must be so designed as to provide direct current over line to subscribers' telephones for transmitter supply. The circuits must be designed to give an efficient voice-current path from one subscriber to the other and from each subscriber to operator. Keys must be provided to permit operator to disconnect her head set from, or connect her head set to, cord circuit at will, thereby permitting her to handle connections with other cord circuits while two particular subscribers are talking.

We thus have the detailed performances for which the circuit must be designed. These are requirements for mechanical and electrical features that should be provided in a somewhat automatic manner. Knowing these, or in other words, knowing the operating procedure for establishing connections between two common battery subscribers, each part of the circuit will stand out as having a specific purpose and make circuit reading comparatively simple.

(h) **Get an approximate mental picture of the layout of the various apparatus parts.** For example, in Figure 113, we may think of the transmitter, receiver and hook switch in the telephone set located on a table or desk at the subscriber's office or residence; the condenser, induction coil and bell located in the bell-box at the subscriber's office or residence; the protector and relay equipment of the subscriber's line circuit located in the central office terminal room; the jack and lamp located in the face of the switchboard in front of the operator; the cord circuit equipment located somewhere in the switchboard position, with the plugs, keys and supervisory lamps of this circuit in the key-shelf where the supervisory lamps are visible and the plugs and keys are readily accessible.

(i) **Do not attempt to read wiring diagrams.** These are intended to assist the wireman in making the proper connections of cable pairs and other wires to the apparatus terminals and are not intended for circuit study. Make sure that the circuit drawing is a schematic or theory drawing and not a wiring diagram.

(j) **Learn thoroughly the operation of a few important telephone circuits pertaining to your particular branch of telephone work.**

(k) **If the circuit is a difficult and unusual one, ask someone to explain it to you in preference to following a written explanation of it.**

APPENDIX III

CONSTRUCTION AND USES OF CURVES

The use of graphical charts or curves is usually the most convenient and effective method of presenting data where two interdependent variables are involved. Such charts and curves are in most cases more readily understood than the corresponding mathematical equations and are not only easier to follow but take up less space than tables.

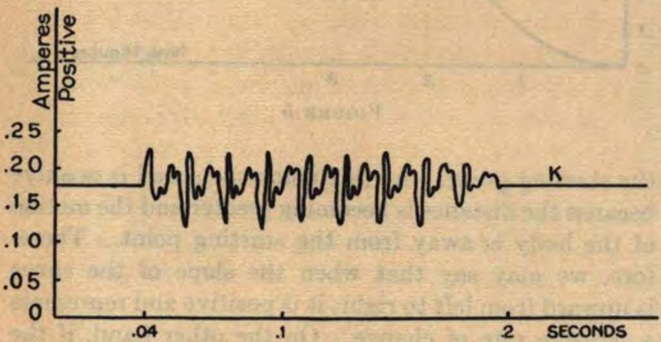


FIGURE 1

Any relationship between two variables, such for instance as $y = ax$, where y varies as x varies (or is said to be a "function of x") may be clearly pictured on a simple plane chart where one variable, say x, is plotted on a horizontal scale and the other variable, y, is plotted on a vertical scale. The two scales do not need to be alike, although each one must itself be uniform. Thus, Figure 1 shows the rather complex relationship between time and the current in the primary of a telephone subset induction coil when a certain vowel sound is spoken into the transmitter. Here we have a horizontal scale representing time, beginning at some instant designated as zero time and charted to the right in graduations of tenths of seconds and a vertical scale representing current values charted upward in graduations of .05 ampere. Thus curve K shows all

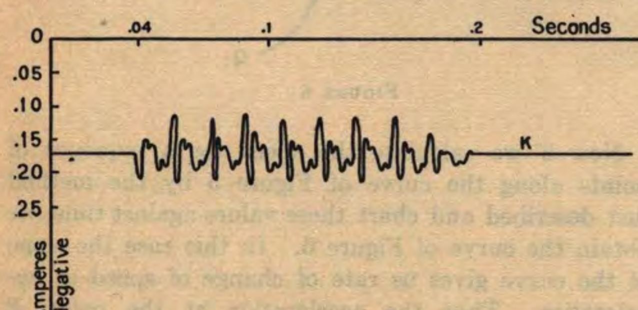


FIGURE 2

values of the current in the circuit for the interval of time considered and conveys a better and more complete idea of what is actually taking place in the circuit than could be expressed in words. During the interval of time between zero and .04 second, i.e., during the first .04 second considered, there is a steady current through the transmitter and induction coil primary of .18 ampere; but beginning at the instant represented by .04 second a vowel sound is spoken into the transmitter and its resistance is alternately lowered and increased causing fluctuations in the current value which are represented by the irregular portion of the curve between .04 and .2 second. The value of the current could of course be charted over any period of time but in this case we have sufficient values to show clearly the effect on the transmitter current of a certain spoken vowel.

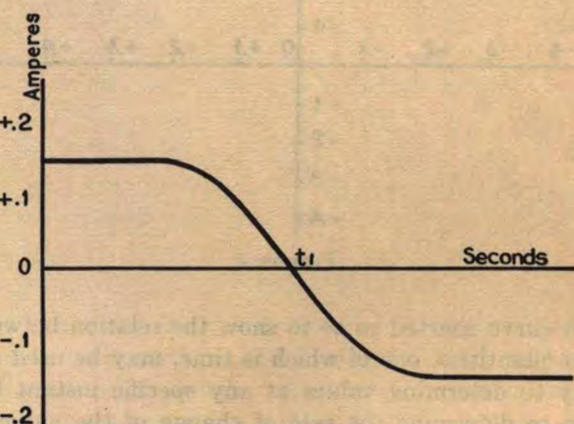


FIGURE 3

The chart of Figure 1 is adequate for showing the magnitude of a current varying with time. Frequently, however, cases are encountered where it is desirable to show not only the magnitude of the current but its direction as well. To take care of this a convention has been adopted that values plotted upward on the vertical scale shall be considered as positive and values plotted downward as negative. Similarly, values plotted to the right on the horizontal scale are positive and to the left, negative. Thus, if the battery causing the current in Figure 1, which we have considered positive, were reversed, we would have a current in the opposite or negative direction and our current-time diagram would be that of Figure 2. In the same way, to represent an alternating current we may use a current-time curve such as that shown by Figure 3, which beginning at zero time represents a sequence of positive current values decreasing to zero value after the time t_1, and followed by a sequence of negative current values (i.e., current in the opposite direction) increasing from zero at time t_1.

Obviously curves such as those illustrated and discussed above may be used for other purposes than charting the relation between current and time. They may quite as frequently be used to chart relations between voltage and time or, in fact, between any two quantities that are so related that one is a function of, or is dependent on the other. There are frequent occasions, also, for charting the relationship between two quantities both of which may be either positive or negative. For this purpose a diagram known as a "complete rectangular coordinate diagram" such as that illustrated by Figure 4 is used. The hysteresis curves shown by Figures 42 and 43 in Chapter III of the text are good examples of diagrams of this kind.

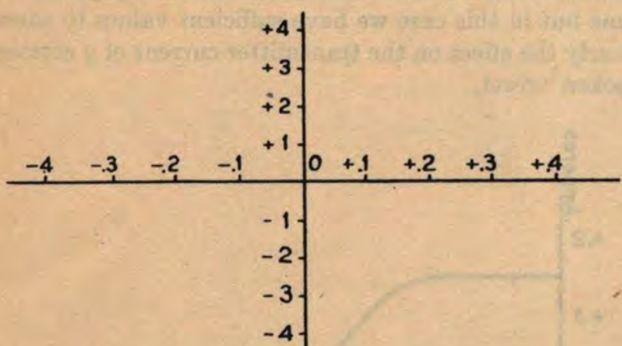


FIGURE 4

A curve charted so as to show the relation between two quantities, one of which is time, may be used not only to determine values at any specific instant but also to determine the **rate of change** of the value at any instant. The curve of Figure 5, for instance, which represents the relation between distance covered by some moving object and the time required to cover this distance, will tell us how far the object is from the starting point at any instant and also the speed or **rate of change of distance** at that instant. This speed or rate of change of distance is defined as the **slope** of the curve at the designated instant. Thus, the speed at which the moving object of Figure 5 is traveling .2 minute after it leaves the starting point may be determined by finding the slope of the curve at the point P, where a vertical line drawn upward from the .2 minute graduation on the time scale intersects the curve.

The slope may be determined by drawing through the point on the curve a straight line **tangent** to or having the same direction as the curve at the point and constructing on this straight line a right triangle of any convenient size but having one leg, as bc, vertical and the other, ab, horizontal; then the value of bc measured on the vertical scale of the chart divided by the value of ab measured on the horizontal scale is the slope of the curve at the point P. This quantity is distance divided by time and therefore is a measure of rate of change of distance or, in more useful terms, the **speed** at which the body is moving .2 minute after it leaves

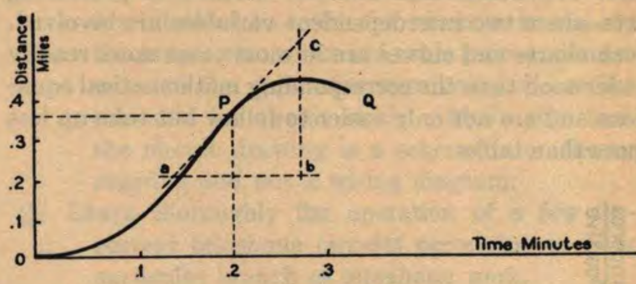


FIGURE 5

the starting point. At this point the speed is positive because the distance is becoming greater and the motion of the body is away from the starting point. Therefore, we may say that when the slope of the curve is upward from left to right, it is positive and represents a positive rate of change. On the other hand, if the body ceases to move away from the starting point and returns toward it, it may be said to have negative speed. Such a condition is illustrated by point Q of Figure 5 and here the slope of the curve is downward from left to right and is negative.

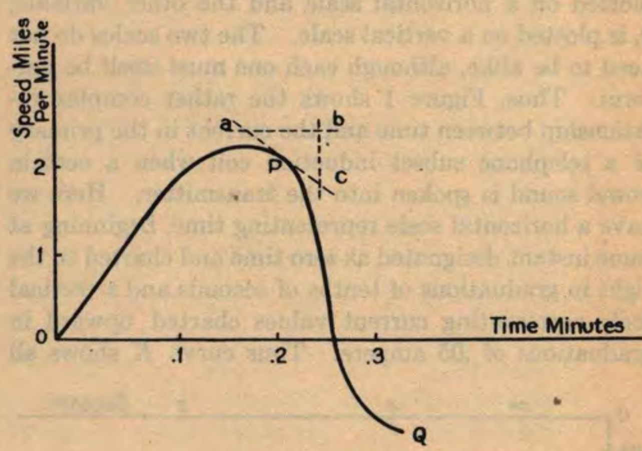


FIGURE 6

Now if we calculate the speed for a sequence of points along the curve of Figure 5 by the method just described and chart these values against time, we obtain the curve of Figure 6. In this case the slope of the curve gives us rate of change of speed or **acceleration.** Thus the acceleration at the point P of Figure 6, or .2 minute after the body starts to move, may be measured by drawing a tangent to the curve at the point, constructing a triangle as shown, and dividing the length of the line *bc* as projected on the speed scale by the length of the line *ab* as projected on the time scale. As before, we have positive acceleration meaning increasing speed, which is represented by positive slope, and negative acceleration meaning decreasing speed or "slowing down", which is represented by negative slope. At point *P* the acceleration is negative since the body is slowing down prior to reversing its direction of motion and returning toward the starting point.

APPENDIX IV

WAVE MOTION FREQUENCY SCALES

1. Vibratory Motion

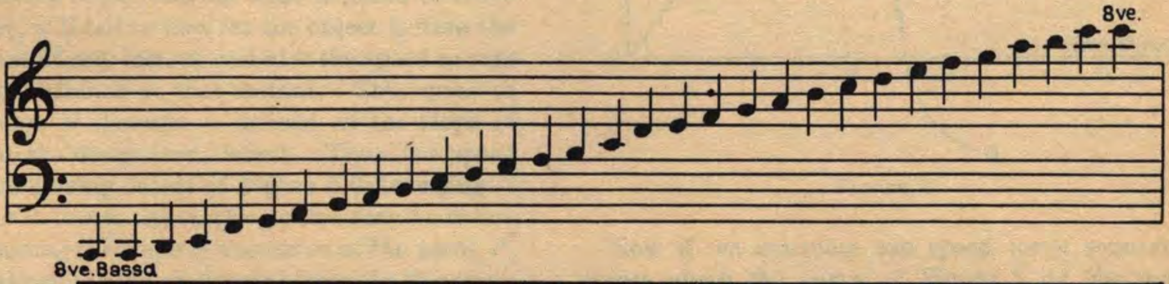
Manifestations of vibratory or wave motion are common in all nature and many forms take place about us continuously. Perhaps the most obvious form is the water wave which everyone has observed to be a form of vibratory motion. By means of condensations and rarefactions of the atmosphere, which are as truly a form of wave motion as the water wave, we experience the sensation called sound. By virtue of still another form of wave motion consisting of a vibration in the mysterious substance which fills all space, called "ether", we recognize the color of some brilliant object. To go further, if this vibration is of a certain particular frequency, the color registered by the retina of the eye may be red but if nearly twice this particular frequency, we are conscious of violet color instead of red. In an electrical conductor, we may think of an electrical current flowing first in one direction and then in the other, as another example of wave phenomena not altogether different from that of light, and although belonging to an entirely different category, similar in some respects to that of sound or even water waves.

Audible sound is defined as a disturbance in the atmosphere whereby a form of wave motion is propagated from some source with a velocity of approximately 1075 feet per second, the transmission being by means of alternate condensations and rarefactions of the atmosphere in cycles having a fundamental frequency ranging somewhere between 16 and 32,000 cycles per second. In Figure 1, we have a scale arranged by frequencies corresponding to periodic recurring octaves. It shows the musical staff notation, and the frequencies of the notes vary as numbers whose logarithms can be shown as a simple progressive scale. Figure 2 shows another sound scale as a simple table of frequencies again arranged by graduations corresponding to octaves or logarithms. Here, however, frequencies are shown by tens, hundreds and thousands instead of as notes of the musical scale. This figure also extends the frequencies to the absolute limits of human hearing or the so-called "limits of audibility". Vibrations coming within the limits of this scale are referred to as **audio-frequencies**, and an alternating electrical current having a frequency, for example, of 1000 cycles, would be carrying the equivalent of a monotonous pure tone lying somewhere between C″ and C‴ of the musical scale.

Figure 3 illustrates a scale of ether vibrations which transmit light, radiated heat, and the electromagnetic waves of wireless telephony and telegraphy. This scale is likewise arranged by graduations proportional to octaves. The number shown opposite each graduation corresponds to frequency, and the wave-lengths at the important division points are given in Angstrom units. As noted on the drawing, this is a very small unit equal to one ten-millionth of a millimeter. The division points are rather arbitrarily chosen in most cases and are intended only to group the various classifications in a general way. Except in the case of visible light, there is naturally no sharp dividing line between the different classes of radiation.

2. Alternating-Current Frequency Scales

All alternating-current frequencies that have any general application in practice, can be grouped and



	C_1	C	D	E	F	G	A	B	c	d	e	f	g	a	b
Relative Pitch	$\frac{1}{2}$	1	$\frac{9}{8}$	$\frac{5}{4}$	$\frac{4}{3}$	$\frac{3}{2}$	$\frac{5}{3}$	$\frac{15}{8}$	2	$\frac{9}{4}$	$\frac{5}{2}$	$\frac{8}{3}$	3	$\frac{10}{3}$	$\frac{15}{4}$
Frequencies	32	64	72	80	$85\frac{1}{3}$	96	$106\frac{1}{3}$	120	128	144	160	$170\frac{1}{3}$	192	$213\frac{1}{3}$	240

	c′	d′	e′	f′	g′	a′	b′	c″	d″	e″	f″	g″	a″	b″	c‴	c^{IV}
Relative Pitch	4	$\frac{9}{2}$	5	$\frac{16}{3}$	6	$\frac{20}{3}$	$\frac{15}{2}$	8	9	10	$\frac{32}{3}$	12	$\frac{40}{3}$	15	16	32
Frequencies	256	288	320	$341\frac{1}{3}$	384	$426\frac{2}{3}$	480	512	576	640	$682\frac{2}{3}$	768	$853\frac{1}{3}$	960	1024	2048

FIG. 1. MUSICAL SCALE

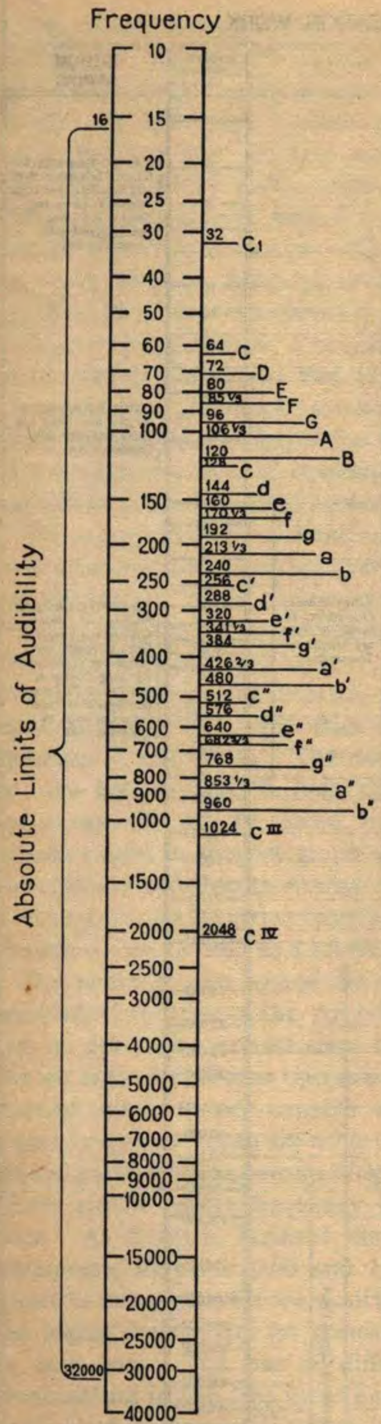


Fig. 2. Logarithmic Musical Scale

graphically pictured similarly to the foregoing scales for sound and light. Thus, we have represented in Figure 4 a chart of the alternating-current frequencies used in both power and communication work. This chart naturally includes many of the same frequency values shown in the foregoing scales, because frequencies identical in value with those of sound are commonly used in telephone circuits, and frequencies used in wireless telephony and telegraphy are also often used for various wire applications, such as carrier. A study of Figure 4 gives us the entire field of use of electricity in the alternating-current form, so ordered that we see at a glance the wide range of frequency values and their adaptability to different utilitarian applications.

This chart is divided into four columns, representing four different fields of use of electricity. The first column is headed "Power Work", the second "Telegraph", the third "Telephone", and the fourth "Telephone Signaling". Adjacent to each column, either on the right or left, is a scale of frequencies which is again arranged by octaves with the actual frequency values shown. In other words, it is a logarithmic rather than an arithmetic scale, which shows frequencies from 10 cycles upward.

In the "Power" column, we find a bracketed note opposite 25 cycles which covers a frequency often used in power work when alternating-current railways are involved and lighting is a secondary consideration. But 60 cycles is indicated as the standard frequency used most extensively in power work. Each standard has its advantages. Alternating-current machinery designed for 25 cycles need not be designed to operate at high speeds, and in some cases permits better power control than 60 cycles. On the other hand, 25 cycles is not desirable for lighting inasmuch as the ordinary lamp filament will sufficiently cool between positive and negative current peaks of the cycle (or in the $\frac{1}{50}$th second interval) to cause the light to "flicker". The 60-cycle system is ideally adapted to lighting and for most applications is well adapted to power work.

Near the bottom of this column is a note explaining that lightning frequencies range from a few hundred thousand cycles to the higher wireless frequencies. Of course, all exposed power and communication lines, as well as radio receiving and transmitting circuits, are subject to lightning hazards, and protection design is an application to circuits involving these frequencies.

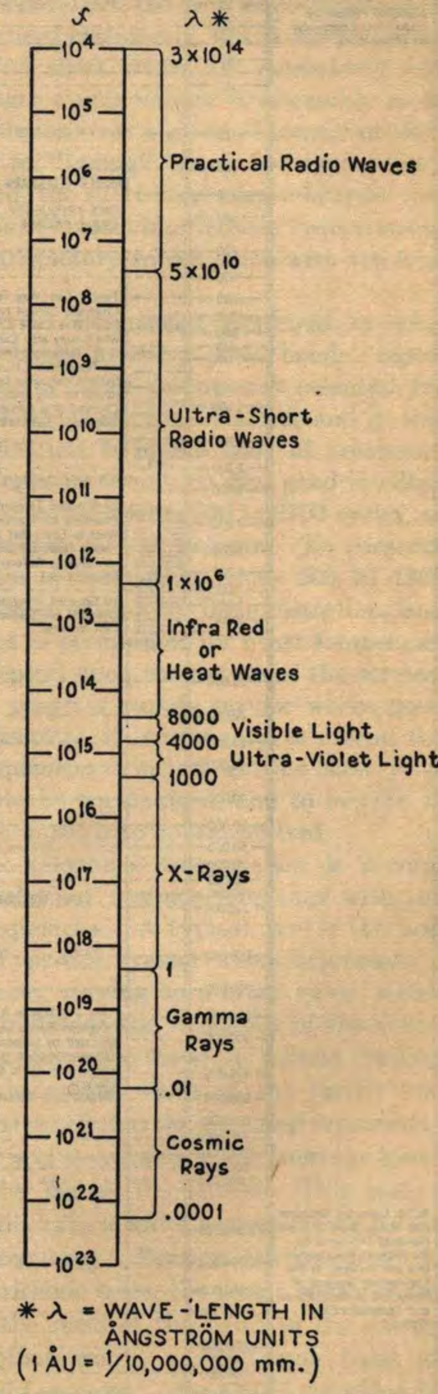
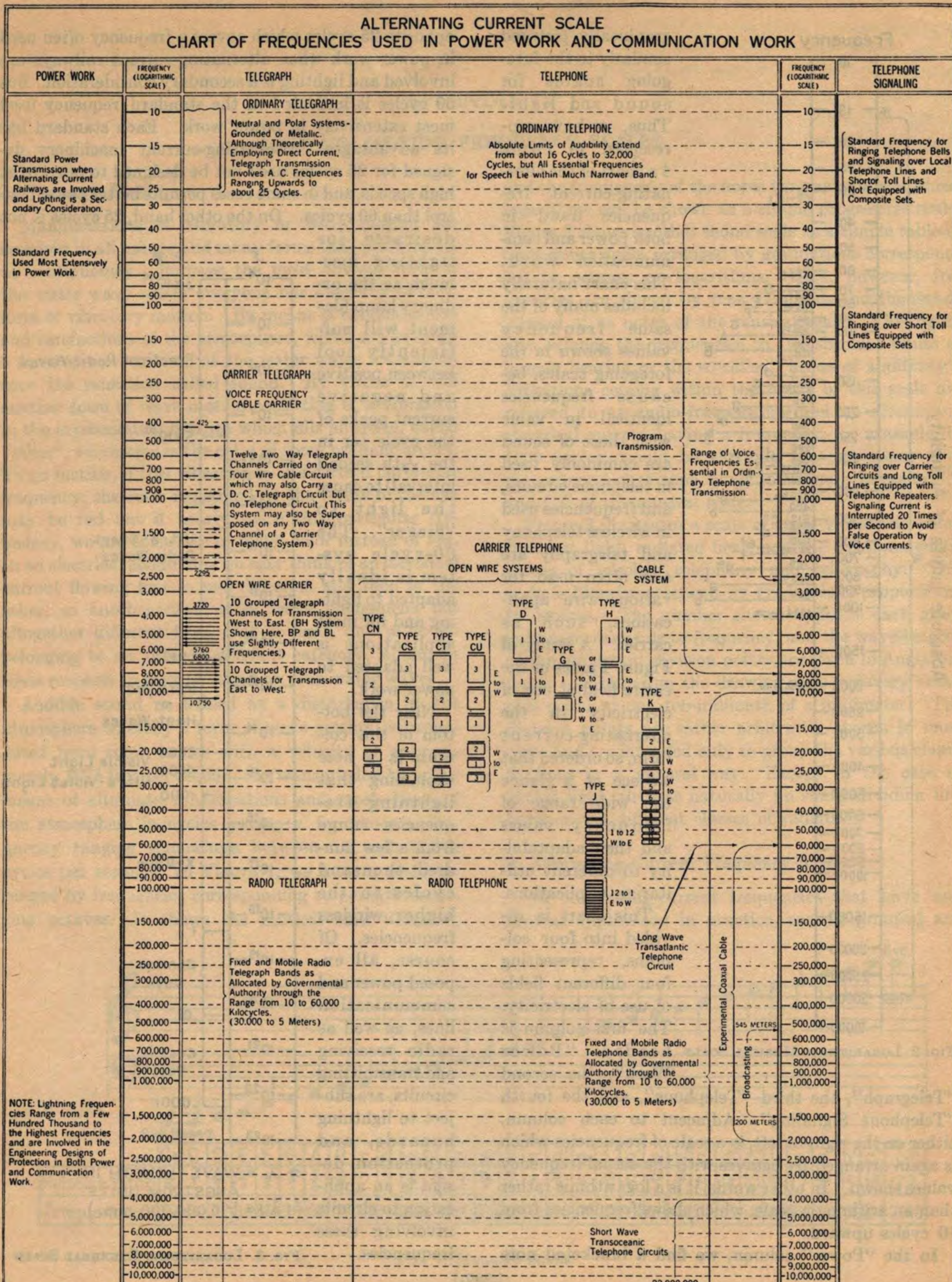


Fig. 3. Logarithmic Ethereal Scale


ALTERNATING CURRENT SCALE
CHART OF FREQUENCIES USED IN POWER WORK AND COMMUNICATION WORK
POWER WORK
FREQUENCY (LOGARITHMIC SCALE)
TELEGRAPH
TELEPHONE
FREQUENCY (LOGARITHMIC SCALE)
TELEPHONE SIGNALING
Standard Power Transmission when Alternating Current Railways are Involved and Lighting is a Secondary Consideration.
Standard Frequency Used Most Extensively in Power Work.
NOTE: Lightning Frequencies Range from a Few Hundred Thousand to the Highest Frequencies and are Involved in the Engineering Designs of Protection in Both Power and Communication Work.
ORDINARY TELEGRAPH
Neutral and Polar Systems-Grounded or Metallic. Although Theoretically Employing Direct Current, Telegraph Transmission Involves A. C. Frequencies Ranging Upwards to about 25 Cycles.
CARRIER TELEGRAPH
VOICE FREQUENCY CABLE CARRIER
Twelve Two Way Telegraph Channels Carried on One Four Wire Cable Circuit which may also Carry a D. C. Telegraph Circuit but no Telephone Circuit (This System may also be Superposed on any Two Way Channel of a Carrier Telephone System.)
OPEN WIRE CARRIER
10 Grouped Telegraph Channels for Transmission West to East. (BH System Shown Here, BP and BL use Slightly Different Frequencies.)
10 Grouped Telegraph Channels for Transmission East to West.
RADIO TELEGRAPH
Fixed and Mobile Radio Telegraph Bands as Allocated by Governmental Authority through the Range from 10 to 60,000 Kilocycles. (30,000 to 5 Meters)
ORDINARY TELEPHONE
Absolute Limits of Audibility Extend from about 16 Cycles to 32,000 Cycles, but All Essential Frequencies for Speech Lie within Much Narrower Band.
Program Transmission.
Range of Voice Frequencies Essential in Ordinary Telephone Transmission.
CARRIER TELEPHONE
OPEN WIRE SYSTEMS
CABLE SYSTEM
TYPE CN
TYPE CS
TYPE CT
TYPE CU
TYPE D
TYPE G
TYPE H
TYPE K
TYPE J
E to W
W to E
1 to 12 W to E
12 to 1 E to W
RADIO TELEPHONE
Long Wave Transatlantic Telephone Circuit
Experimental Coaxial Cable
Broadcasting
545 METERS
200 METERS
Fixed and Mobile Radio Telephone Bands as Allocated by Governmental Authority through the Range from 10 to 60,000 Kilocycles. (30,000 to 5 Meters)
Short Wave Transoceanic Telephone Circuits
22,000,000
Standard Frequency for Ringing Telephone Bells and Signaling over Local Telephone Lines and Shortor Toll Lines Not Equipped with Composite Sets.
Standard Frequency for Ringing over Short Toll Lines Equipped with Composite Sets
Standard Frequency for Ringing over Carrier Circuits and Long Toll Lines Equipped with Telephone Repeaters. Signaling Current is Interrupted 20 Times per Second to Avoid False Operation by Voice Currents.
10
15
20
25
30
40
50
60
70
80
90
100
150
200
250
300
400
500
600
700
800
900
1,000
1,500
2,000
2,500
3,000
4,000
5,000
6,000
7,000
8,000
9,000
10,000
15,000
20,000
25,000
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70,000
80,000
90,000
100,000
150,000
200,000
250,000
300,000
400,000
500,000
600,000
700,000
800,000
900,000
1,000,000
1,500,000
2,000,000
2,500,000
3,000,000
4,000,000
5,000,000
6,000,000
7,000,000
8,000,000
9,000,000
10,000,000

FIGURE 4

The second column of Figure 4 headed "Telegraph", covers four classes of telegraph transmission. First, the more common system designated "ordinary telegraph", including the familiar grounded neutral and polar systems and metallic cable systems; second, the voice-frequency carrier current telegraph system; third, the superposed carrier current telegraph; and fourth, the radio telegraph. The first application, or the more common telegraph system which makes and breaks a direct-current circuit is usually thought of as a direct-current system, but considering the rapidity of the makes and breaks and the wave-like nature of the telegraphic pulses, we have an approximation to an alternating-current transmission that we could describe as a telegraphic band of frequencies beginning at zero and shading into about 25 cycles.

The other applications make and break a steady flow of alternating current rather than direct current. The voice-frequency cable carrier system makes use of a series of frequencies beginning at 425 cycles and extending to 2295 cycles or higher. By this means twelve two-way telegraph channels are secured over a 4-wire cable circuit, with a separation in frequency between channels of 170 cycles. The use of the 4-wire circuit permits transmitting in both directions at the same frequency. It will be noted, however, that the frequencies used in this telegraph application are within the ordinary voice range so that a cable circuit used for a voice-frequency carrier telegraph system cannot at the same time be used as a telephone circuit.

Due to the greater cost of the wire facilities, it is not economical to operate the voice-frequency carrier system on open wire circuits since this would not permit the use of the facilities as telephone circuits. Open wire circuits, however, are capable of transmitting much higher frequencies than are cable circuits. This feature makes possible the superimposing on an open wire telephone circuit higher frequency carrier telegraph systems. As shown in Figure 4, these carrier systems use frequencies between 3300 and 11,250 cycles, ten frequencies being used for transmitting West to East and ten higher frequencies for transmitting East to West in each case. The use of different frequencies for transmitting in the two directions makes possible the application of the system on an ordinary 2-wire circuit.

The radio telegraph employs frequencies from about 10,000 cycles to more than 60,000,000 cycles. The wave-lengths corresponding to any radio frequency (or any ethereal vibration whatsoever) are calculated from the formula—

$$\lambda = \frac{300,000,000}{f}$$

where λ is the symbol for wave-length in meters and f is the frequency. This is evident from the known speed of ethereal waves, which is 300,000,000 meters per second. The number of wave-lengths in one second (or frequency) times the length of one wave must equal the distance traveled in one second, or 300,000,000 meters.

Wave-lengths are assigned for radio telegraph and radio telephone communication through Government or International Conference regulation and are accordingly subject to change from time to time. In the earlier days of the radio art, the long wave-lengths (low frequencies) were used exclusively, but in modern practice both long and short waves are extensively employed. The longer waves require considerably more power for transmission over a given distance but they are less subject to "fading" than the short waves. On the other hand, the short waves are relatively free from certain kinds of "static interference" which sometimes prevents satisfactory transmission with the long waves.

Alternating-current frequencies employed in **telephone transmission** are in every case "bands" representing the range of voice frequencies essential for intelligibility in a telephone conversation, and in this respect the application is unlike that of telegraph. The ordinary telephone circuit, to give good intelligibility, must employ a band from 200 to 2700 cycles, as shown by the bracket in this column. To preserve quality, the range of frequencies from 500 to 1800 should be transmitted with very little distortion, and the distortion that is permissible for other frequencies will, of course, depend upon the nature of the service. For example, in program supply service where good quality is very essential, it is necessary to extend the entire band of frequencies to include at least 3000 cycles if only speech is to be transmitted, and to include at least 5000 cycles if music is to be transmitted.

Carrier current telephone transmission is accomplished by "modulating" a single frequency with the band of voice frequencies. A typical case is the first channel of the Type-CN system which represents a 7600-cycle frequency varying to a lower value, which variation would represent the frequency of the voice. Thus, each carrier channel is shown as a band shading away from the designated value of the carrier frequency. It will be noted that the graphical representation of the bands is of decreasing width from the lower frequencies to the higher frequencies. This means that each band will have a width corresponding to the range of voice frequency. Because the frequency is plotted on a logarithmic scale, the scalar width of the band represents the ratio of the band to the carrier frequency. In other words, a 3000-cycle band at 30,000 cycles would represent only a 10% variation in the carrier frequency, while a 3000-cycle band with a

9000 carrier frequency would represent a 33⅓% variation.

Frequencies used for radio telephony extend through the same range as those employed in radio telegraphy. At the lower frequency end, they overlap the carrier frequencies and they extend upward far beyond the scale of the chart. Because they are subject to frequent change, no attempt is made to show the detailed allocations for various purposes. However, the ordinary (long wave) broadcast band is indicated as extending through the range between 550 and 1500 kc. The chart also shows the allocations for the short-wave transoceanic telephone circuits and for the single long-wave circuit between New York and London.

While **telephone signaling** is essentially a part of telephone communication service, it is, nevertheless, distinct from the transmission of the human voice. The fourth column of Figure 4 shows three telephone signaling systems. The standard ringing current for ringing telephone bells and signaling over local telephone lines, as well as over short toll lines not equipped with composite sets, has a comparatively low frequency of approximately 20 cycles, and is ordinarily referred to as "20-cycle ringing".

The 20-cycle current would naturally fall within the band of frequencies discussed in connection with "ordinary telegraph", which as has been explained, is essentially direct current, but shades into the alternating scale to about 25 cycles. Consequently, it would not be practicable to separate in the telephone office a 20-cycle ringing current from a telegraph current, and a ringing current of this frequency transmitted over composited telephone lines would interfere with the telegraph service. Conversely, telegraph currents would interfere with signaling. For this reason a second frequency, well outside the ordinary telegraph range of frequencies, is used for ringing over the shorter composited circuits. The standard frequency value assigned for this use is 135 cycles.

On long voice-frequency telephone circuits equipped with a number of telephone repeaters, there are advantages to be gained in eliminating the special apparatus which is required at each repeater point to relay the 135-cycle current around the telephone repeater circuit. This is necessary because the repeater set is ordinarily not designed to "pass" the 135-cycle frequency. Of course, the repeater is designed to amplify the band of frequencies representing the actual voice, and any ringing current within this range of frequencies would not only pass through the repeater but would at the same time be amplified. Such an arrangement would not require relaying equipment for the ringing current. This third ringing frequency is 1000 cycles interrupted 20 times per second, the latter feature being necessary in order that conversation over the circuit, which will ordinarily contain some 1000-cycle notes, will not operate the ringer. This type of signaling is also used on carrier telephone circuits for similar reasons.

In concluding the foregoing explanation of the various frequency values and frequency bands in general use, it should be remembered that this description is neither complete nor fundamentally descriptive of the distinction between various alternating-current applications. Though the frequency scale illustrates in one sense characteristic differences in the currents, it should be remembered that other characteristic differences such as energy values might illustrate an equally wide diversification. Furthermore, the use of a single frequency does not mean that no consideration must be given to the nature of other frequencies which may indirectly become involved in the same application. To illustrate, a 60-cycle, 3-phase power line when unbalanced has a residual current flow of 180 cycles, and the same power system with imperfect sine wave form may carry harmonics which give a band of frequencies extending through the voice-frequency range, resulting in serious inductive interference with communication circuits paralleling the power lines. Another case is radio telegraph transmission. While employing a single wave-length, this must have a sufficiently accurate sine wave form so that it will not radiate harmonics which would interfere with radio communications em-

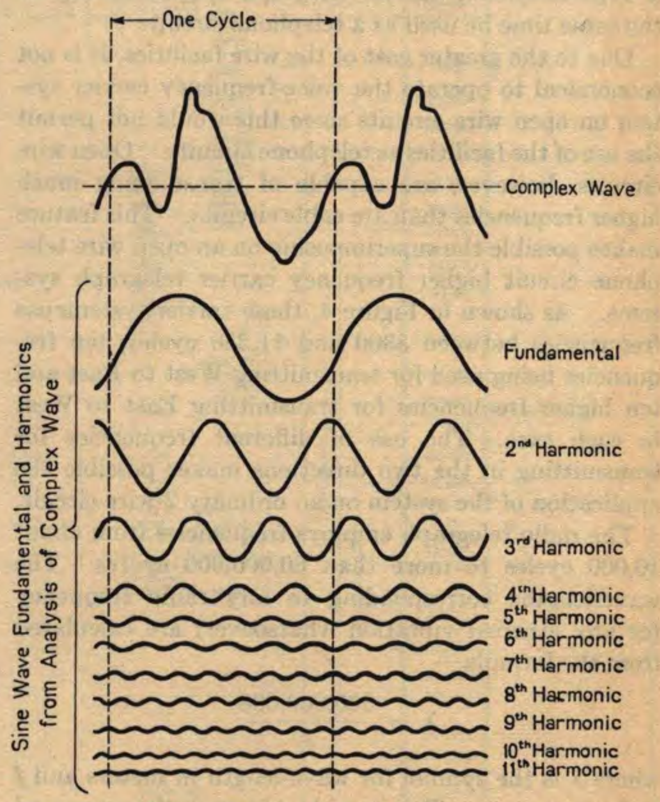

Fig. 5. Analysis of Sound Wave by Fourier's Theorem

ploying other wave-lengths. The application of alternating currents of a single frequency or a single band of frequencies, therefore, may and actually does in the majority of instances, involve the suppression of other frequencies as well as the use of the individual frequency.

3. Wave Analysis

It is known that the characteristic wave forms for vowel sounds are complex waves rather than sine waves of a designated frequency, and that alternating currents which represent the tones of the human voice are seldom, if ever, a simple sine wave. The same may be said for many other cases of alternating-current transmission. We would be hopelessly involved if we should attempt to deal with any current wave shape that might be encountered as distinct and apart from the sine wave, which is the basis of most alternating-current circuit calculations. Fortunately, for all practical cases, **any steady state alternating-current wave form, regardless of how irregular, may be considered as a series of sine wave currents being simultaneously transmitted.** That is to say, any irregular current wave shape can be analyzed or broken up into a series of sine wave shapes. This series can further be restricted to a fundamental frequency and multiples or harmonics of this fundamental frequency. For example, Figure 5 shows an irregular wave so analyzed, and by actually adding the successive and corresponding values of all the sine waves, we can chart the original and irregular wave shape.

The practical treatment of any irregular wave shape for electrical transmission, is then, divided into two steps:

a. **Analyze the wave shape into a fundamental sine wave and its harmonics.**

b. **Deal with each of these component sine waves individually.**

The actual mechanical analysis, such as that shown by Figure 5, is quite difficult, and while the analysis can be made from an "oscillogram", this is for the most part a combination of a laboratory process and involved mathematical calculations. For most purposes, however, we need be concerned with the concept only, rather than with the actual analysis.

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