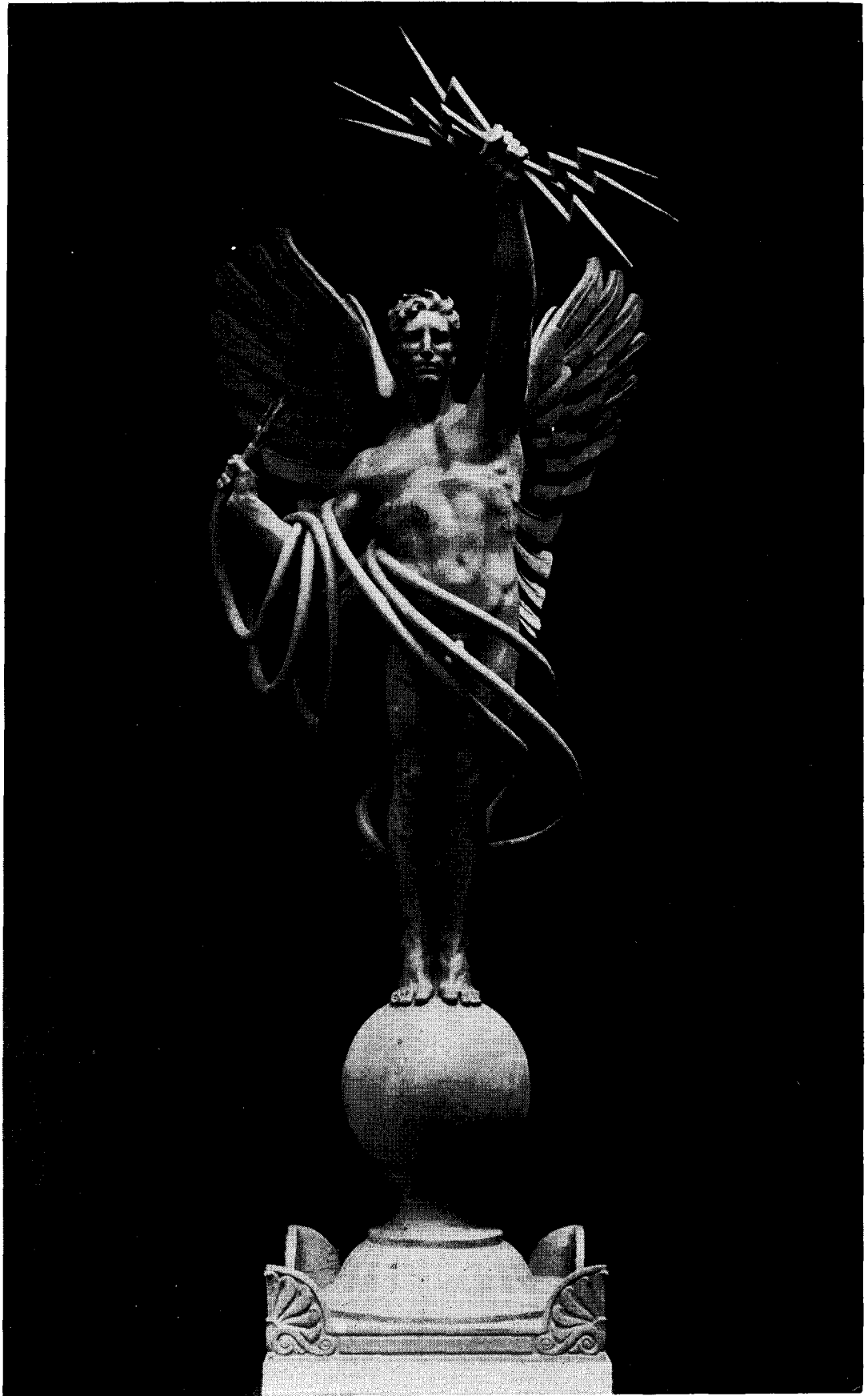


**Principles of Electricity**  
applied to  
**Telephone and Telegraph Work**

1953 Edition





*Spirit of Communication*

# **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**

**January, 1953**

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## PREFACE

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**T**HIS book has appeared in a number of editions over the past three decades, the latest prior edition being dated 1938. Its purpose has remained the same throughout this period—namely, to serve as a basic text on elementary electrical principles for use in the technical training of employees in the Long Lines Department of the American Telephone and Telegraph Company.

Although truly fundamental principles do not change with time, the rapid growth and development of the communications art over the years has necessitated the incorporation of numerous new or changed illustrative applications in each new edition. This, of course, remains true for the present volume. In addition, the accelerated extension in recent years of radio and very high frequency techniques has made it desirable to include a great deal of new material in the fields of electronics and electromagnetic radiation. This has required the introduction of certain basic concepts and principles not dealt with in earlier editions, as well as numerous examples to illustrate their applications in practice.

The variety and scope of the subjects covered in the book make it necessarily rather voluminous even though some material included in previous editions has been omitted in this. Every effort has nevertheless been made to treat each subject taken up as briefly as is consistent with a reasonably adequate presentation of the related theory and fields of application.

It is important for the reader to recognize that the book is not, and is not intended to be, a comprehensive treatise embracing the entire field of electrical communication. Its subject is electrical theory. Such descriptions of communication equipment and circuits as are included were selected primarily to illustrate practical applications and many of them have been somewhat arbitrarily chosen. The objective has been only to cover the essential general principles of basic electrical theory and to illustrate each principle briefly by one or more of its significant applications in the communications field.

The use of higher mathematics is avoided entirely in this book, and even the more elementary branches are employed 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. In addition, it has been thought desirable, in the Chapters dealing with the solution of alternating-current networks and with wire transmission theory, to make some use of simple Vector Notation. This may involve the introduction of certain mathematical concepts not familiar to all readers, but it is believed that the great simplification that may be effected with this convenient mathematical tool will more than justify any additional study time that the reader may find needed to master its practical use.

A general knowledge of elementary Physics and Mechanics is also assumed. The first Chapter of the text, however, reviews very briefly some of the fundamental physical principles that are particularly applicable to various subjects discussed in later Chapters.

A word of caution is perhaps needed regarding the use of the circuit drawings, tables, and other statistical data included at various points in the text. The circuit drawings are presented primarily as a means of illustrating the principles under discussion. Although they may be reasonably representative of actual practice, they may or may not conform in detail with any situation familiar to the reader. 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 data issued in current formal instructions.

C. F. MYERS,  
*Supervisor of Instruction,*  
L. S. CROSBY,  
*General Personnel Supervisor.*

32 Avenue of the Americas,  
New York, N. Y.,  
January, 1953.

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**Principles of Electricity**  
applied to  
**Telephone and Telegraph Work**



# CHAPTER 1

## BASIC PHYSICAL CONCEPTS

### 1.1 Introduction

Electricity is an agency of remarkable flexibility, with almost unlimited applications. By means of electricity, energy available at one place, in the form, for example, of burning coal or falling water may be transferred to another place perhaps many miles away, and utilized there in any convenient form, such as heat, light, or mechanical work. Electricity may be used to change the physical or chemical state of substances, and it may be employed to control processes, to check quality, and even to perform mathematical computations. Through the electron, basic unit of electricity, we may probe the physical and biological worlds, and study hidden secrets of the universe.

The power engineer is able, by applying his knowledge of the behavior of electricity, to accomplish the transformation and transmission of large amounts of energy that may be needed for many and various purposes; and the generation, flow and delivery of this energy also can be controlled electrically. The communications engineer, by similarly applying his knowledge of the same basic principles, is able to transmit intelligence in a variety of forms over great distances and without loss of meanings. The source may be a human voice, a musical instrument, printed page or picture, and it can be reproduced in like form, completely recognizable and intelligible, wherever needed.

Since this book is prepared primarily for telephone people, we shall discuss these basic principles with reference, naturally, to applications in the communications field. This will mean, in general, that we will be interested mostly in small values of electrical energy, and in the transmission of such energy over relatively great distances. This interest will embrace both direct current and alternating current concepts and in the underlying alternating current phenomena, the scope will cover a wide range of frequencies, including energy flowing over wires and energy radiated through space.

There are various possible approaches to beginning the study of electricity. In this book, we

shall adhere to the conventional, classical approach, beginning with the study of direct current and direct-current circuits, and proceeding from there to alternating current theory, with particular emphasis in the later Chapters on electronics and high-frequency phenomena.

Before beginning the study of electric circuits and the electrical transmission of energy, however, we shall consider briefly, in this Chapter, some of the more fundamental concepts of the physics of electricity.

### 1.2 Electric Charge

The earliest recorded recognition of electrical phenomena dates back several thousand years when it was discovered that rubbing a piece of amber with woolen cloth would cause it to attract other light bodies, such as bits of paper. The amber had thus acquired a property by virtue of which it exerted a certain force on other materials in its vicinity. This property became known as an electric charge. Then it was discovered that a similar phenomenon occurred when glass was rubbed with silk. Subsequent experimentation demonstrated, however, that the charge on the glass was opposite in sign from the charge on the amber. To distinguish them, the former was labeled "positive charge" and the latter "negative charge". In both cases the magnitude of the charge depended upon the extent of the surface rubbed and the intensity of the friction.

It was also found that two like bodies, such as two pieces of glass or amber, when so treated would exert a repelling force on each other. In other words, positive charges repel each other and negative charges repel each other. On the other hand, a force of attraction was found to exist between a positively charged substance and a negatively charged substance. These attracting and repelling forces although relatively small are millions of times larger than the force of attraction exerted by gravity between similar uncharged bodies.

The "practical" electrical unit for measurement of these electric charges is called the coulomb in honor of the man who, late in the eigh-



teenth century, first measured the extent of the forces of attraction and repulsion between charged bodies. He found that the forces (1) vary directly as the product of the charges and (2) inversely as the square of the distance between the charged bodies.

### 1.3 The Electron

The electron represents 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  $1.6 \times 10^{-9}$  coulombs and the mass  $9 \times 10^{-28}$  grams, or somewhat less than 1/2000 of the mass of a hydrogen atom. All negative charges consist of integral numbers of these unit electron charges.

All physical substances are made up of electrons, particles called protons which carry a positive charge of the same magnitude as that carried by an electron, and uncharged particles called neutrons having approximately the same mass as protons. Protons and neutrons have been determined to occupy the same amount of space as an electron but have about 2000 times the mass of the electron. It is sometimes convenient to think of the neutron as a locked-up combination of one proton and one electron, with the opposite charges counteracting each other and thus making the particle electrically neutral.

The atom of any chemical element is made up of a combination of neutrons, protons, and electrons and the number and arrangement of these particles determines what chemical element the atom represents. The atom of the simplest and lightest chemical element, hydrogen, has a nucleus consisting of a single proton around which a single electron revolves in certain fixed orbits. Heavier elements have nuclei consisting of a specific number of protons and a specific number of neutrons around which revolve, in certain fixed orbits, a number of electrons equal to the number of protons. The number of electrons revolving around the nucleus gives the **atomic number** of the element and the sum of the number of protons and neutrons in the nucleus gives the **atomic weight**. Thus the second lightest element in the **Periodic Table of Elements**, helium, is made up of

two protons, two neutrons and two electrons as indicated in Figure 1-1.

The heaviest natural element, uranium, in its most common form consists of 92 protons, 92 electrons and 146 neutrons. Elements may have atoms deviating to some extent from the common form in that the number of neutrons contained in the nucleus, and hence the atomic weight, is slightly greater or less than the normal. Such deviations are called "isotopes". It should be noted, however, that the number of electrons and protons remain constant as any change in these would represent a different element.

Electrons are attracted toward the atom nucleus and are repelled by one another with tremendous forces relative to their mass—forces enormously 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 one or more of the electrons farthest out from the atom nucleus is less strongly attached 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, 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. Thus, as previously discussed, when a glass rod is rubbed with silk the contact of the two different substances results in the breaking away of electrons from surface atoms in the glass and their attachment to surface atoms in the silk. The glass is then positively charged as a result of the subtraction of electrons, and the silk is negatively charged because of the addition of electrons.

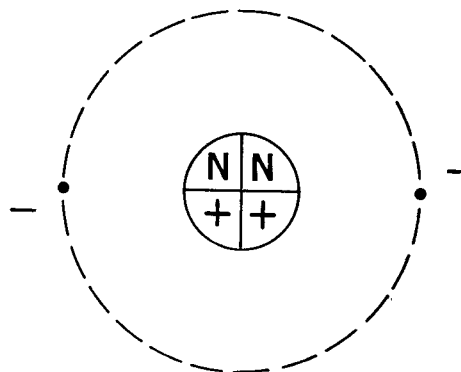


FIG. 1-1 HELIUM ATOM

In the case of materials known as electrical conductors—usually metals such as silver, copper, aluminum and others—the outer-most electron of each atom is so loosely attached that it is practically free to move about through the metallic structure under the influence of small external forces. In a good conductor, such as copper, the total number of such free electrons is enormously great—a cubic centimeter of copper contains  $8 \times 10^{22}$  atoms and a comparable number of free electrons.

When an external electric driving force is applied to such a metallic conductor, these free electrons move through the conductor in a continuous stream. This represents a flow of electricity through the conductor and the rate of such flow—that is, the number of electrons or, what is to say the same thing, the quantity of electricity that passes a given point in unit time—is electric current. For example, the practical unit of current, the ampere, represents the passage, at a given point in a conductor, of one coulomb of electricity or  $6.24 \times 10^{18}$  electrons per second.

It should be observed, that in this conduction process the movement of any single electron is anything but uniform. The individual electron may, after an exceedingly slight movement attach itself to an atom, thus releasing another electron to move forward—again in a more or less random manner. The total number of electrons involved is so extremely great, however, that the net movement of electric charge along the conductor appears to be a uniform process despite the random movement of the individual electrons.

In substances known as insulators, the outer electrons are attached much more strongly to the atoms. There are relatively few free electrons, and it is accordingly much more difficult to cause a current of electricity to flow in such materials.

#### 1.4 The Electric Field

Experiment shows that the presence of an electric charge at a point in space can be detected at points distant from the point where the charge is located. This means that the electric charge exerts an influence or force in the space surrounding it. To explain this, it is customary to say that there is an **electric field** surrounding the charge, which is always present whenever a charge exists and which may properly be considered as an integral part of the charge itself. The “strength” of this field at any point is directly proportional to

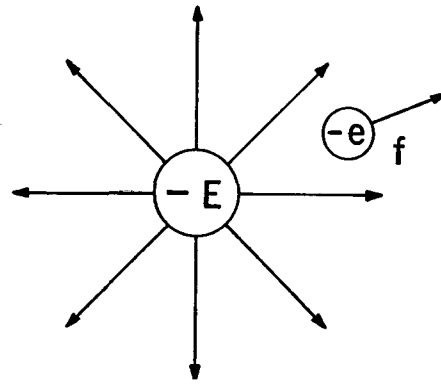


FIG. 1-2 ELECTRIC FIELD

the strength of the charge and inversely proportional to the square of the distance from the point to the location of the charge.

The strength of an electric field at any given point is called its **intensity** and is measured in terms of an electric intensity vector having both magnitude and direction as illustrated in Figure 1-2. If a small negative electric charge  $e$  is brought up to the neighborhood of a large fixed negative charge  $E$  it will be found to be repelled by the fixed charge with a force that can be represented by the length of the vector  $f$  and in the direction indicated by the arrow.

The electric field of a single electron, or a corresponding positive charge, is of course so weak as to be hardly measurable but, as pointed out earlier, charges may easily be developed of such magnitude that the intensity of their fields is great enough to exert powerful forces in their immediate vicinity.

Any relatively small test charge brought into such an electric field will, if free, move toward or away from the fixed charge causing the field, depending on the relative polarity of the test charge. The path of the moving test charge will of course be determined by the direction of the forces acting on it at each instant and it is usual to designate such a path as a **line of electric force or induction**. The configuration of an electric field—or of the electric forces within the field—may then be illustrated conveniently by groups of these lines extending between charges of opposite sign as shown in Figure 1-3. The number and direction of the lines passing through any small area of the field indicates the intensity of the field in that area; and the total number of lines emanating from a charge source may be considered as a measure of the strength of the charge. Lines of electric force need not be thought of as having

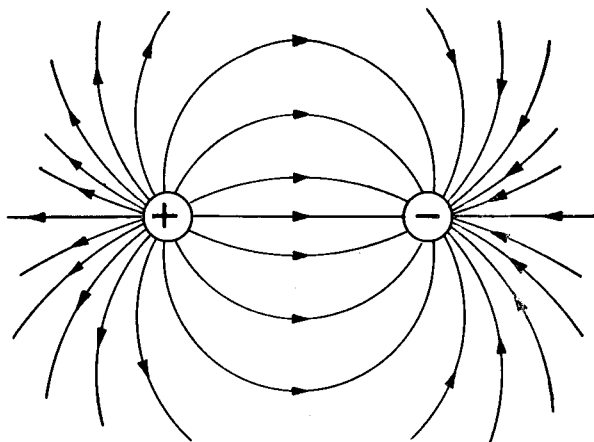


FIG. 1-3 FIELD BETWEEN OPPOSITE CHARGES

physical reality but the concept is so convenient that the line of electric force or induction is generally used as a basic unit of measurement. The strength of a field at any point may then be stated as the number of lines per square centimeter, where each line is considered as extending between unit charges of opposite sign.

### 1.5 The Magnetic Field

Also of long standing in the classical theory of electricity is the concept of the magnetic field. This field, as discussed more extensively in Chapter 4, represents the forces to be found in the vicinity of magnets. Like the electric field, its intensity at any point is directly proportional to the strength of the magnetic poles to which the field is due, and inversely proportional to the square of the distance from the poles. This field is similarly pictured by **lines of magnetic force or induction** and its intensity is measured in terms of number of lines of magnetic force per square centimeter. Unlike the electric field, however, magnetic lines of force are not thought of as terminating but as always forming closed loops.

A magnetic field is not only present in the space within and surrounding any magnet but, as is pointed out in Chapter 4, such a field may be detected in the space surrounding any conductor in which electric current is flowing. This phenomenon seems to require the introduction of a new concept, since there is no apparent identity between a static field produced by a magnet and a field associated with current flow in a conductor

and having nothing to do with any magnets or magnetic materials. Yet the two fields appear to have precisely the same characteristics.

Modern electrical theory resolves this problem by considering the magnetic field as a result or accompaniment of a moving electric field. Thus, current flow in a conductor is a movement of electrons, which are tiny electric charges. These charges have accompanying electric fields which of course must move as the electrons move. Extending this idea further, the magnetic field of a magnet is ascribed to the spinning movements of electrons in the magnetic material. From this view, the magnetic field may even be thought of as merely an aspect of the moving electric field. Nevertheless, the concept of the magnetic field as such, and of lines of magnetic force, continues to be very useful in the practical analysis of many electrical phenomena, and it need not be discarded.

### 1.6 Electro-Magnetic Fields

Under the field concepts discussed above, either an electric field or a magnetic field may exist alone only under static conditions. Under dynamic conditions, where either type of field is changing or moving, the total field must be electro-magnetic—i.e., both an electric field and a magnetic field must exist simultaneously. This is a necessary result of the fact that the moving electric field is always accompanied by a magnetic field, and vice versa. In dealing with electro-dynamic phenomena, accordingly, it is reasonable in many situations to consider only a single composite field which may be designated electro-magnetic. There are often distinct advantages, however, in viewing certain dynamic phenomena on the basis of coexistent electric and magnetic fields, each with its force lines always at right angles to those of the other.

In our study of direct currents and direct-current circuits in the immediately following Chapters, we shall not be much concerned with fields other than static magnetic fields. As we go into the study of high-frequency alternating currents and the behavior of various electronic devices, however, we shall find that dynamic electric and magnetic fields will assume a place of increasing importance—particularly in connection with radio transmission and other transmission at high and super-high frequencies.

## CHAPTER 2

### DIRECT CURRENT AND DIRECT-CURRENT CIRCUITS

#### 2.1 Direct Current Flow

In discussing electrons in the preceding Chapter it was pointed out that the movement of a stream of electrons through a conductor constitutes an electric current. Where such a flow of electrons is continuous in one direction it is called a **direct current**. It has long been customary to begin the practical study of electricity with an analysis of the behavior of such direct currents, primarily because the phenomena involved are relatively easy to deal with. We shall conform to this well-established study procedure in the beginning of this book although we shall later have to consider in some detail the behavior of currents which reverse their direction of flow at periodic intervals. These latter are called **alternating currents**.

#### 2.2 Electric Pressure or Electromotive Force

In order to establish a flow of current through a conductor it is necessary that there be connected in the circuit a source of electric pressure which sets up what is known as an **electromotive force**. When this condition is met the flow of electricity through a circuit is analogous in many respects to the flow of water through a closed system of pipes. Figure 2-1 shows a simple electric circuit consisting of a battery connected to a resistor  $ab$ . Figure 2-2 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 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 electric circuit, the battery supplies the electric 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 resistor.

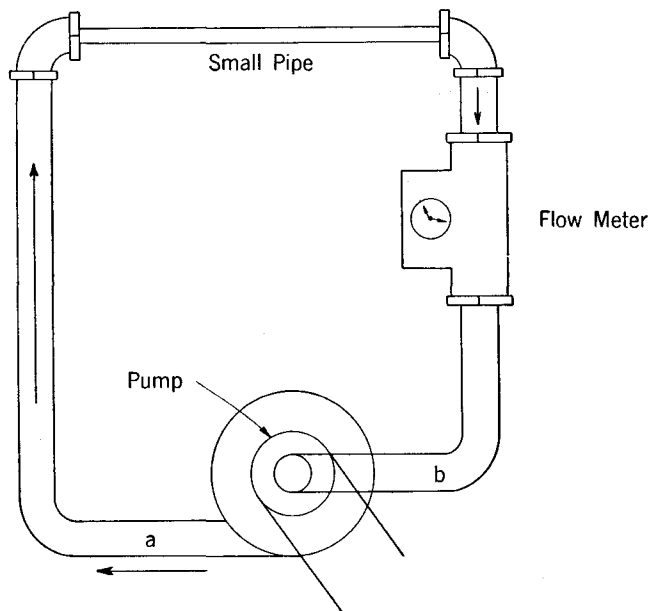


FIG. 2-2 WATER CIRCULATING SYSTEM ANALOGOUS TO SIMPLE ELECTRIC CIRCUIT

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 electric circuit, on the other hand, is measured in terms of a unit called the **volt**.

Since a source of electromotive force is by definition capable of exerting electric pressure on any external circuit connected to its terminals, it is customary to say that the "electric potential" of the positive terminal is higher than that of the negative terminal. The difference is the electromotive force of the battery or other source. In a closed circuit, such as shown in Figure 2-3, the

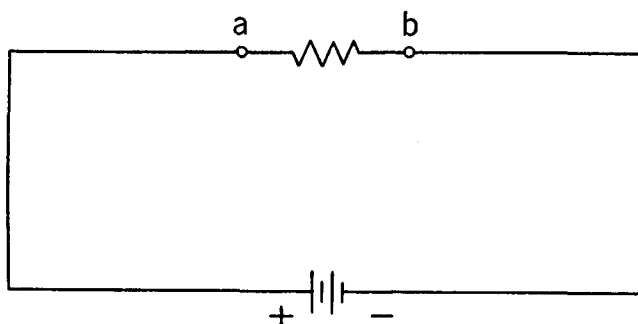


FIG. 2-1 SIMPLE ELECTRIC CIRCUIT

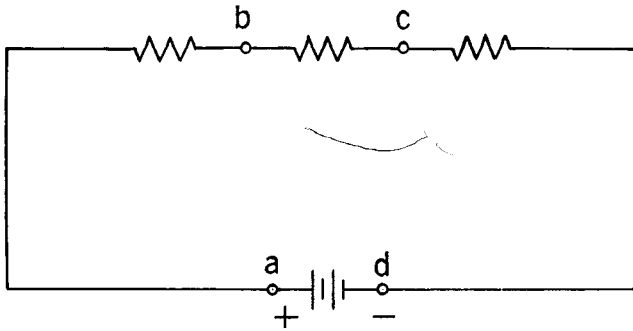


FIGURE 2-3

potential at point *b* will be lower than that at point *a*, that at *c* will be lower than at point *b* and so on around the circuit. Thus we say that there is a **potential drop** from *a* to *b* that may be measured in volts, the magnitude of which depends on the resistance of the conductor between *a* and *b*. Similarly there is another potential drop between *b* and *c*, and *c* and *d*.

### 2.3 Resistance

In Figure 2-2, if the small pipe is made longer the flow of water will be decreased although the pump maintains a 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 electric resistance in many respects. The resistance offered to a flow of water in a pipe is due primarily to friction between the moving molecules of water and the inner walls of the pipe. Resistance to the flow of electric current in a conductor is considered to be due to the multitudinous interactions or collisions between the moving electrons and the atoms of the conducting material. The practical unit of electric resistance is called the **ohm**, which by an act of the U. S. Congress, approved July 21, 1950, is defined as being equal to one thousand million units ( $10^9$ ) of resistance of the centimeter-gram-second (cgs) system of electromagnetic units.

**Note:** As defined in standard textbooks on physics, the cgs electromagnetic units are derived from the fundamental mechanical units of **length, mass and time** according to the basic principles of electromagnetism, with the value of the permeability of free space arbitrarily chosen as **unity**.

### 2.4 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. The practical unit of electric current is the **ampere** which is defined by law as being equal to **one-tenth of the unit of current of the centimeter-gram-second system of electromagnetic units**.

### 2.5 The Volt

The **volt** is the practical unit of electric pressure. It is **the electromotive force that, steadily applied to a conductor whose resistance is one ohm, will produce a current of one ampere**.

### 2.6 Open Circuits

The electric 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 electric circuits. Figure 2-4 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.

### 2.7 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 conven-

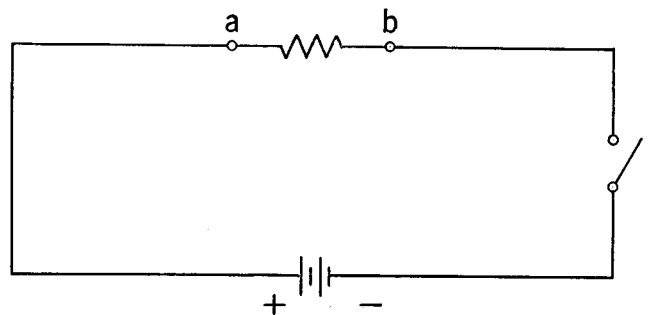


FIGURE 2-4

tions. Thus, Figure 2-5 illustrates an actual door bell circuit and Figure 2-6 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 2-7 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 equations. 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.

### 2.8 Ohm's Law

A German physicist named George Simon Ohm was the first to discover the relationship between current, electromotive force and resistance. 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.

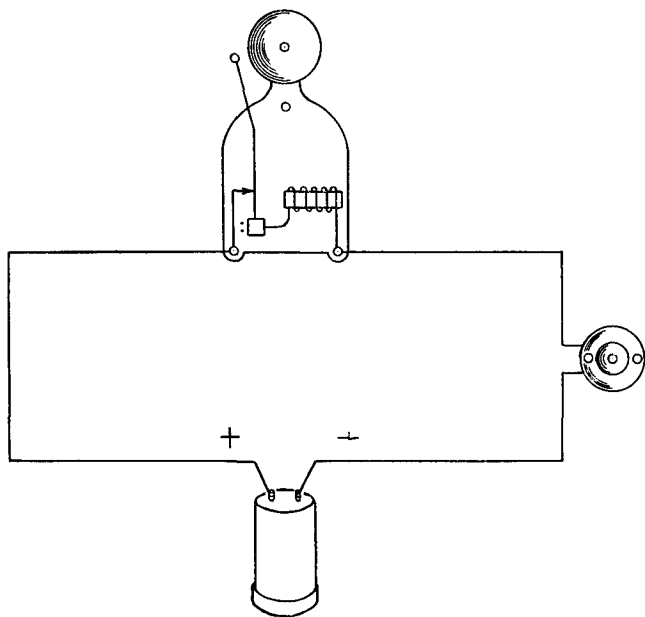


FIG. 2-5 WIRING OF DOOR-BELL CIRCUIT

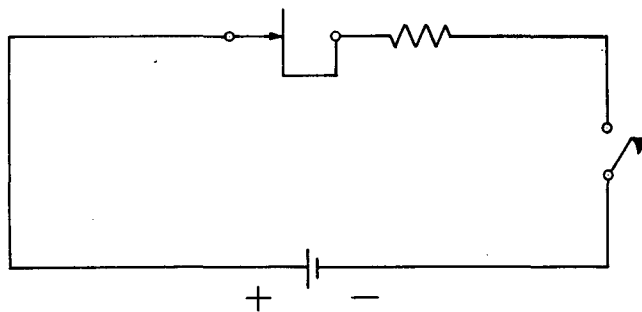


FIG. 2-6 CONVENTIONAL DRAWING FOR CIRCUIT OF FIG. 2-5

This law, mathematically expressed, is as follows:

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

If in the above expression we substitute the proper symbols for current, electromotive force and resistance we have the following equation:

$$I = \frac{E}{R} \quad (2:1)$$

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 2-8 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?

**Solution:**  $E = 24$

$R = 112$

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

### 2.9 Other Ways of Expressing Ohm's Law

Equation (2:1) 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—

$$E = IR \quad (2:2)$$

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



Crossed



Connected  
WIRES



Grounded



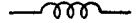
Single Cell



24 - Volt  
BATTERIES



RESISTOR



INDUCTOR



Fixed



Variable  
CAPACITOR



VOLTMETER



AMMETER



Switchboard  
LAMPS



Resistance  
LAMPS



FUSE



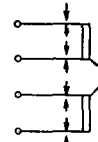
TELEPHONE RECEIVERS



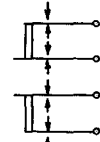
TELEPHONE  
TRANSMITTER



RINGER



Locking Lever - Type



Non-locking Lever - Type  
KEYS



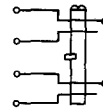
Regular Type  
JACKS



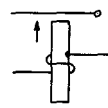
Twin Type



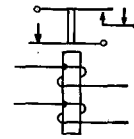
2 - Conductor  
PLUGS



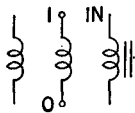
Twin Type



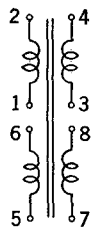
Single wound  
with make contact  
RELAY



Double wound with break  
& make - before - break contact  
RELAY

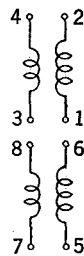


Retardation

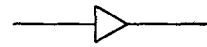


Repeating

COILS



TRANSFORMER



AMPLIFIER



ANTENNA



METALLIC RECTIFIER

FIG. 2-7 CONVENTIONS COMMONLY USED IN TELEPHONE CIRCUIT DRAWINGS

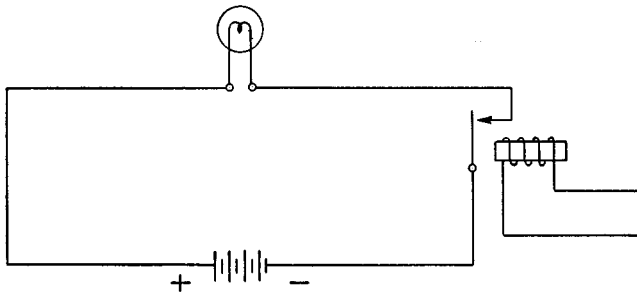


FIGURE 2-8

**Example:** In Figure 2-6 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:**  $R = 4$

$$I = .2$$

$$E = IR = .2 \times 4 = .8 \text{ volt, ans.}$$

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—

$$R = \frac{E}{I} \quad (2:3)$$

**Example:** What is the resistance connected between the points *a* and *b* in Figure 2-1 if the voltage of the battery is 1.3 volts and the current is .5 ampere?

TABLE I

SYMBOLS USED IN ELECTRICAL WORK

- I* Current in Amperes
- E* Electromotive Force in Volts
- R* Resistance in Ohms
- P* Power in Watts
- Q* Quantity in Coulombs
- V* Potential Drop in Volts
- G* Conductance in Mhos (is reciprocal of *R*)
- T* Time in seconds
- C* Capacitance in Farads
- L* Inductance in Henrys
- f* Frequency in Cycles per Second
- X* Reactance in Ohms
- Z* Impedance in Ohms
- Y* Admittance in Mhos

TABLE II

RESISTIVITY OF VARIOUS METAL CONDUCTORS AT 20° C.

MATERIAL	RESISTIVITY (MICROHM-CM)
Silver	1.62
Copper (Annealed)	1.72
Copper (Hard-drawn)	1.77
Chromium	2.7
Aluminum	2.82
Tungsten (Annealed)	4.37
Molybdenum	5.7
Zinc	5.92
Brass	7.5
Cadmium	7.60
Iron (Pure)	10.
Platinum	10.0
Nickel (Commercial)	11.0
Phosphor Bronze	11.5
Tin	11.5
Tantalum	15.5
Lead	22.0
German Silver	33.8
Mercury	95.8

**Solution:**  $E = 1.3$  volts

$I = .5$  ampere

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

2.10 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 2-2. The electrical instrument used for measuring the electric pressure of a source of electromotive force, or the potential difference between any two points in a circuit, is called the **voltmeter**.

Figure 2-9 shows a voltmeter being used to measure the voltage of a dry cell on an open circuit. Figure 2-10 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 resistors in series. If the voltmeter is connected across the points *a* and *b* as shown in Figure 2-11, which represents the

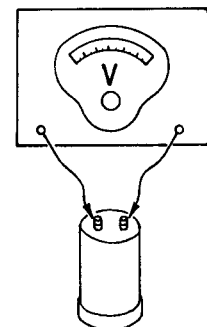


FIGURE 2-9



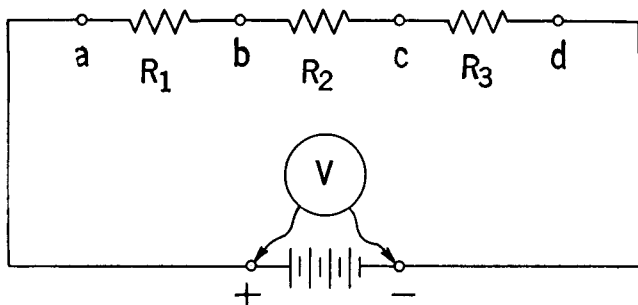


FIG. 2-10 VOLTAGE OF BATTERY ON CLOSED CIRCUIT

same circuit as Figure 2-10, its reading will be lower than when connected across the battery. Moreover if the voltmeter is connected across the resistors *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 closed circuit. If in Figure 2-11 the value of the resistance from *a* to *d* and 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 the circuit shown by Figure 2-11 is 15 ohms, the voltage of the electromotive force on closed circuit is 10 volts, the potential drop across  $R_1$  is 3 volts; what is the resistance of  $R_1$ ?

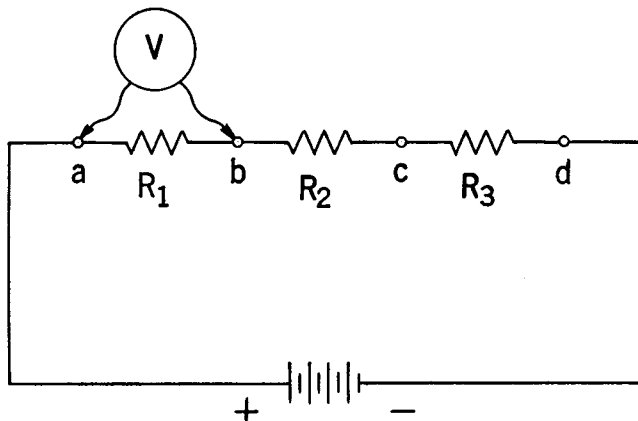


FIG. 2-11 VOLTAGE DROP ACROSS ONE RESISTANCE OF CLOSED CIRCUIT

**Solution:** For entire circuit:

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

For the part of the circuit in question—

$$E = 3 \text{ volts}$$

Therefore,

$$I = .67 \text{ ampere}$$

$$E = 3 \text{ volts}$$

$$R_1 = \frac{E}{I} = \frac{3}{.67} = 4.5 \text{ ohms, ans.}$$

## 2.11 Internal Resistance

If a dry cell, as shown in Figure 2-9, is placed in a closed circuit like that of Figure 2-1 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, 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 2-12, which represents the open circuit voltage plus a series resistor 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

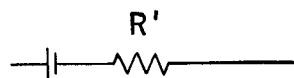


FIGURE 2-12

the telephone central office storage batteries are used almost exclusively because the internal resistance is negligible for most direct current considerations.

TABLE III  
ELECTRICAL PROPERTIES OF COPPER CONDUCTORS STANDARDIZED BY LONG LINES DEPARTMENT

CONDUCTORS	NO.	SIZE		WEIGHT	RESISTANCE	
		GAGE	DIAMETER IN INCHES	LBS. PER WIRE MILE	*OHMS PER LOOP MILE	OHMS PER 1000 FEET (SINGLEWIRE)
Open Wire	8	B. W. G.	.165	435	4.10	.389
	10	N. B. S. G.	.128	264	6.82	.645
	12	N. B. S. G.	.104	174	10.3	.975
Cable (side circuits of standard quadded cable)	10	A. W. G.	.102	168	10.1	.955
	13	A. W. G.	.072	82.6	20.5	1.99
	16	A. W. G.	.051	41.2	40.2	3.81
	19	A. W. G.	.036	20.5	84.1	7.95

\* The resistance values for open wire circuits are for 20° C or 68° F. Add 2/10 of 1% per degree Fahrenheit for higher temperatures. The resistance values for cable circuits are for 12.80° C or 55° F.

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.

### 2.12 Electric Power

In the simple circuits we have thus far considered we have only dealt with resistance, electromotive force, and electric 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. **In the electric 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 the power expended in a circuit having an electromotive force of one volt and a current of one ampere.

The following equation should be considered second only to Ohm's Law in importance:

$$P = EI \quad (2:4)$$

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

$$P = I^2R \quad (2:5)$$

This latter equation is apparent from Ohm's Law, which states that  $E = IR$  and we may, therefore, substitute  $IR$  for  $E$  in equation (2:4), which gives us  $I^2R$ .

**Example:** In Figure 2-10, what is the power expended in the resistor 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.

### 2.13 Properties of Electric Conductors

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" 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. Materials which are commonly used as insulators include: glass, mica, polystyrene, rubber, ebonite, cotton, silk, paper, bakelite, asbestos, porcelain and resins. 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 uniform conductor varies directly with its length.

b. The resistance of any uniform conductor varies inversely with its cross-sectional area.

Here we have the analogy to the water pipe previously mentioned but fortunately the electric conductors have more exact laws governing their electric 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 III shows the principal standard gages of wire used by the Long Lines Department and their resistance values.

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

- a. Four sizes of cable conductors are standard for most uses in 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 and #19.
- 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. #10 is the nearest A. W. Gage to 104 (#12 N.B.S.G.) but is slightly smaller.

## CHAPTER 3

### THE SOLUTION OF D-C NETWORKS

#### 3.1 Series Circuits

A simple circuit may contain any number of resistances. Figure 3-1 shows such a circuit with two resistors which, when connected as shown, are said to be **in series**. Figure 3-2 shows another circuit with the same resistors connected **in parallel**. Any number may be so connected in either case.

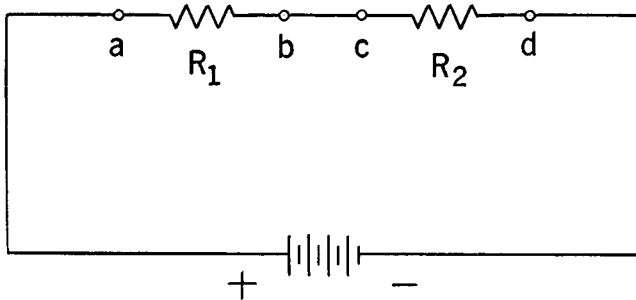


FIG. 3-1 SERIES CIRCUIT

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. It follows then that **the total resistance of a series circuit is equal to the sum of all the individual resistances**. Thus, in Figure 3-1 the total resistance is the sum of the two resistances; namely, that connected between *a* and *b* or  $R_1$ , and that connected between *c* and *d* or  $R_2$ .

$$R = R_1 + R_2 \quad (3:1)$$

#### 3.2 Parallel Circuits

If we apply Ohm's Law to either of the two parallel resistors of Figure 3-2, we shall find that the current in either case must be equal to the potential measured across the particular resistor divided by its value in ohms; and for this particular circuit, the potential measured across either resistor is the emf of the battery. The battery is in reality supplying two currents, one through the resistor *ab* and the other through the resistor *cd*. These two currents are united and flow together

in the conductors connecting the poles of the battery with the junctions of the two resistors. For any circuit having two resistors connected in parallel, the current supplied to the combination must therefore be greater than the current supplied to either of the resistors. If we think of the combination of resistors in Figure 3-2 as equivalent to a single resistor that might be substituted in their stead, it is accordingly clear that the combined value in ohms of two resistors in parallel must be less than that of either resistor taken singly.

We may make calculations for determining the current in a parallel circuit such as is shown by Figure 3-2, but these are more complicated than for a simple series circuit having more than one resistance, such as is shown in Figure 3-1. The solution of a parallel circuit is accomplished with the aid of Kirchoff's Laws in addition to Ohm's Law.

#### 3.3 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 3-2. If  $I$  is the current being supplied by the battery to the combination of the two resistors

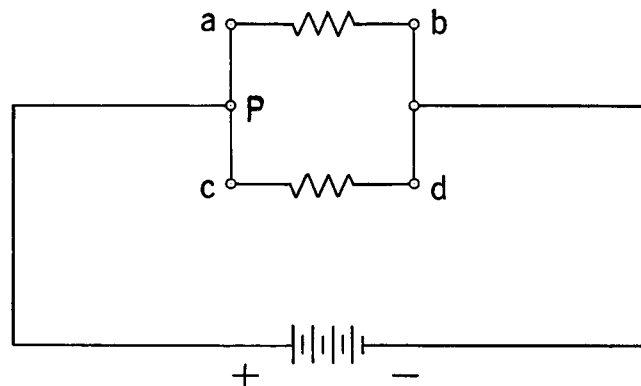


FIG. 3-2 TWO RESISTORS IN PARALLEL

in parallel, and  $I_1$  and  $I_2$  are the respective currents through the two parallel resistors, then

$$I = I_1 + I_2 \quad (3:2)$$

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

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

But

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

Therefore,

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

In this latter equation, the  $E$ 's can be cancelled and the equation written—

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2}} \quad (3:3)$$

and if we simplify this compound fraction by simple algebra—

$$R = \frac{R_1 R_2}{R_1 + R_2} \quad (3:4)$$

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

**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 3-3 shows a circuit having three resistors in parallel. An equation similar to (3:3) 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 the value of  $ef$ . But for problems involving more than two resistors in parallel, it is usually simpler to use the conductance method.

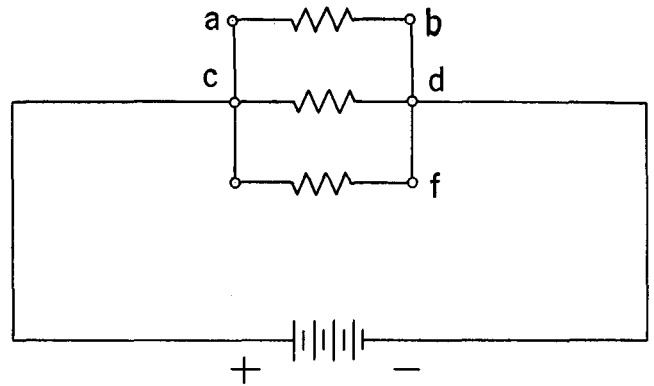


FIG. 3-3 THREE RESISTORS IN PARALLEL

### 3.4 Conductance

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

$$G = \frac{1}{R} \quad (3:5)$$

For a combination of resistors in parallel, such as is shown by Figure 3-3, the conductance of the combination is equal to the sum of the individual conductances, or

$$G = G_1 + G_2 + G_3 \quad (3:6)$$

In a circuit having a number of resistors 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.

**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:**

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

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

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

$$G = G_1 + G_2 + G_3 = .061 + .032 + .010 = .103$$

$$R = \frac{1}{G} = \frac{1}{.103} = 9.7 \text{ ohms, ans.}$$

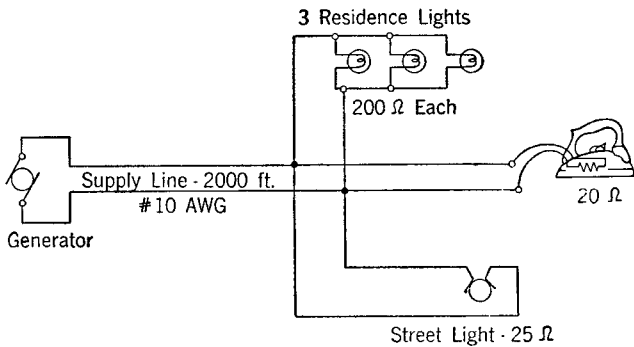


FIG. 3-4 SMALL ELECTRIC POWER SYSTEM

### 3.5 Direct-Current Networks

Several resistors may be connected in such a manner as to form very complicated networks. In practice many circuits are of this type. For example, Figure 3-4 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 3-5, and can further simplify this network as shown by Figure 3-6. 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 3-7 represents an A-board local cord circuit connected to a local switching trunk having 1/2 mile of 19-gage cable. The equivalent network is shown by Figure 3-8.

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 emf.

**Example:** What is the value of the current through each winding of the B-59 relay in Figure 3-7?

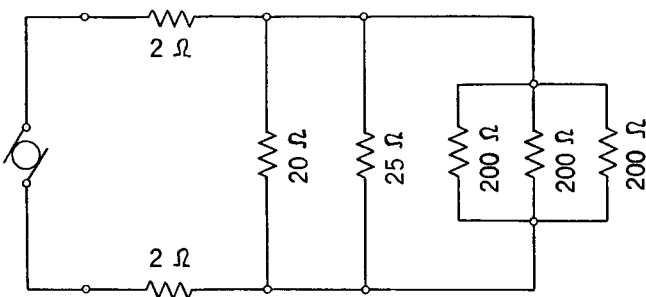


FIG. 3-5 CONVENTION FOR CIRCUIT OF FIG. 3-4

**Solution:** We must first find the total current through both windings and have:

$$I = \frac{E}{R} \text{ where } E \text{ is 24 volts and}$$

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

$$\text{Then } I = \frac{24}{166.3} = .14432 \text{ ampere.}$$

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

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

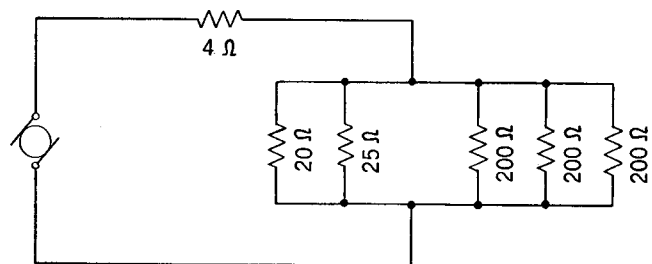


FIGURE 3-6

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

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

and

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

### 3.6 Kirchoff's Second Law

When current flows through a resistor there is always a difference in potential between the ends of the resistor, 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

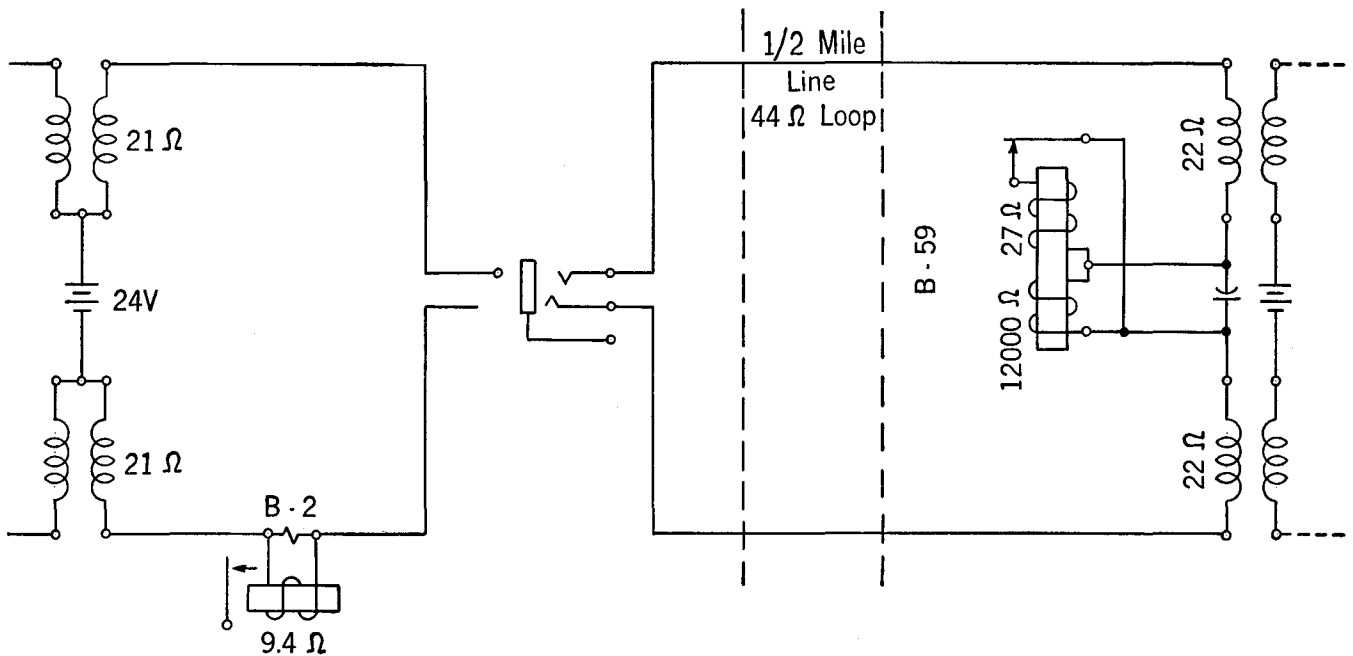


FIG. 3-7 CORD CIRCUIT AND SWITCHING TRUNK OF TELEPHONE CONNECTION

of the current and the resistance. This  $IR$  drop acts in the opposite direction to, or opposes, the emf which drives the current through the resistor.

In a closed circuit, such as is shown in Figure 3-9, the sum of the  $IR$  drops across the three resistors must be equal to the impressed emf. Thus if the drop across the resistor  $R_1$ , 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—

$$E = V_1 + V_2 + V_3 \quad (3:7)$$

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 emf's and the potential drops is equal to zero.

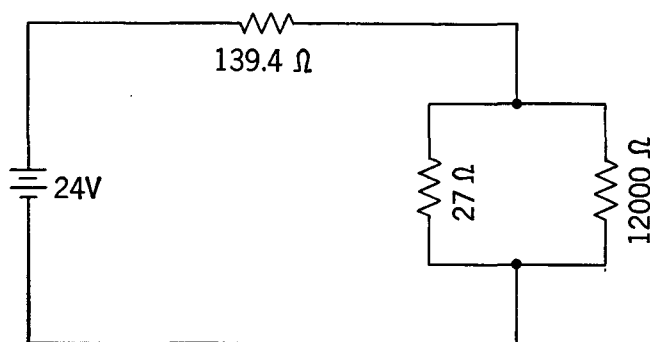


FIG. 3-8 SIMPLIFIED CONVENTION FOR CIRCUIT OF FIG. 3-7

In the case of Figure 3-9, Kirchoff's Second Law may be written as follows:

$$E - IR_1 - IR_2 - IR_3 = 0 \quad (3:8)$$

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 application 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 the current flow, the opposite sign must be given to the  $IR$  drops. 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 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 emf'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 emf'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 minus sign, it merely means that the actual direction of flow is opposite to the direction which was assumed.

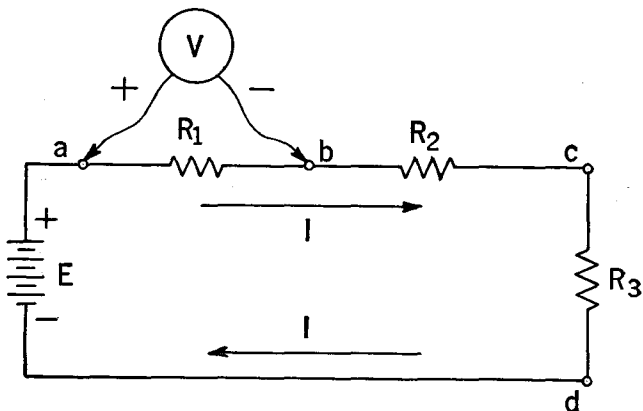


FIGURE 3-9

**Example:** Find the current values in each branch of Figure 3-10, 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  $R_2$  and  $R_4$  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 3-10 that is shown by Figure 3-11, we may write in accordance with Kirchoff's Second Law:

$$E - R_1(I_2 + I_3) - R_3I_3 = 0 \quad (a)$$

and for the closed loop shown by Figure 3-12

$$E - R_1(I_2 + I_3) - R_2I_2 - R_4I_2 = 0 \quad (b)$$

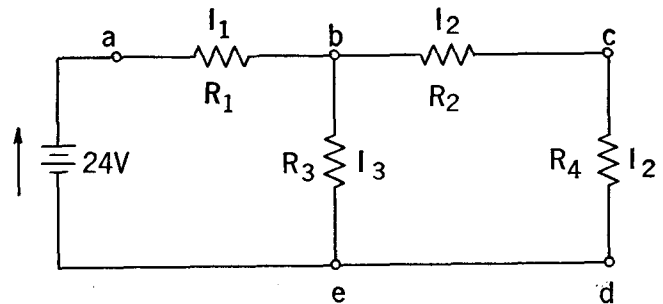


FIG. 3-10 SERIES-PARALLEL CIRCUIT

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:

$$24 - 5(I_2 + I_3) - 15I_3 = 0 \quad (a)$$

$$24 - 5(I_2 + I_3) - 10I_2 - 20I_2 = 0 \quad (b)$$

Simplifying, these equations become—

$$24 - 5I_2 - 20I_3 = 0 \quad (a)$$

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

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

$$168 - 35I_2 - 140I_3 = 0 \quad (a)$$

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

$$\hline 144 \qquad - 135I_3 = 0$$

or

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

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

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

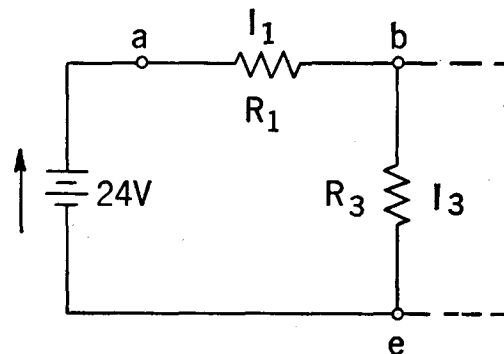


FIGURE 3-11



## CHAPTER 4

### MAGNETS AND MAGNETIC CIRCUITS

#### 4.1 Nature of Magnetism

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 also discovered at an early date that a bar or needle-shaped piece of this ore, when suspended so as to turn freely about a central axis, would turn so that one end always pointed to the north. For centuries, before it was discovered that there was any relationship between magnetism and electricity, this unique property was put to practical use in the mariner's compass—as it still is.

Investigation through the years showed that this magnetic property could be induced by one means or another in iron and steel, and to a lesser degree in nickel and cobalt, as well as in alloys of these metals. Not until early in the nineteenth century was it learned that the magnetic property could be artificially given to these metals by means of an electric current.

Magnets, as we know them today, are 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 **remalloy** (iron-cobalt-molybdenum), and cores for electro-magnets are ordinarily made of **soft iron** or of iron alloys such as **permalloy** (iron-nickel), **supermalloy** (molybdenum-permalloy), **perminvar** (cobalt-iron-nickel), or **permendur** (iron-cobalt).

#### 4.2 Permanent Magnets

Figure 4-1 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

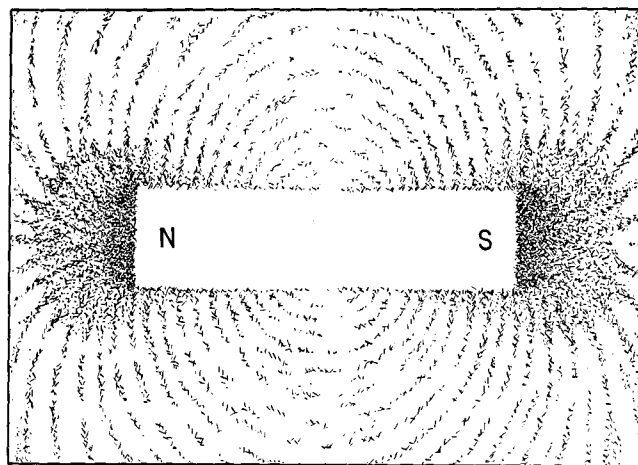


FIGURE 4-2

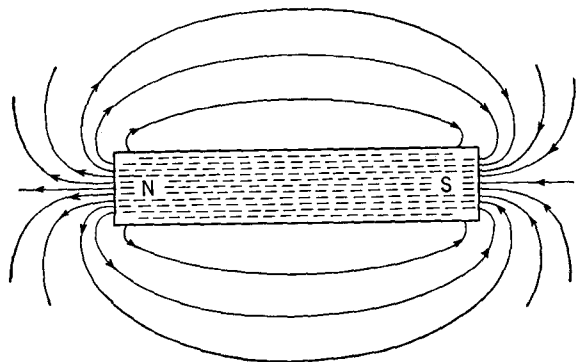


FIG. 4-1 MAGNETIC FIELD AROUND BAR MAGNET

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 have a magnetic field, which is represented by the curved lines in Figure 4-1. 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  $\phi$ . The flux per unit area is known as the **flux density** and is designated by  $B$ .

The lines of magnetic induction are thought of

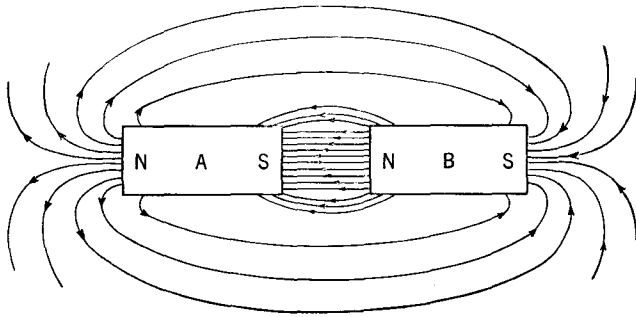


FIG. 4-3 MAGNETIC FIELDS AIDING

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 in Figure 4-1.

Lines of magnetic induction are always closed loops. A vivid graphical demonstration, not only of the presence 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 4-2 shows how the filings arrange themselves under such a condition.

If a second magnet is placed at the end of the bar magnet shown in Figure 4-1, the magnetic field will become either like that shown in Figure 4-3 or that shown in Figure 4-4. 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 4-3, merely changing ends of one magnet will give the effect in Figure 4-4. 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. When 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.)

If the strength of the magnet in Figure 4-1 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 pole of another magnet located at any point in the magnetic field will depend upon the **intensity** of the field at that point. This field intensity is represented by the symbol  $H$ .

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 unit intensity. Thus in air the field intensity  $H$  and the flux density  $B$  have the same numerical value.

In Figure 4-1 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 4-5. 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 4-1. The strength of the field between the two poles of a horseshoe magnet is more intense than that of a straight magnet of equal strength. Thus, we not only shorten each line represented by a closed loop but, in so doing, create more lines. This gives us an analogy to the electric circuit, which we may call a **magnetic circuit**. In the electric 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 in-

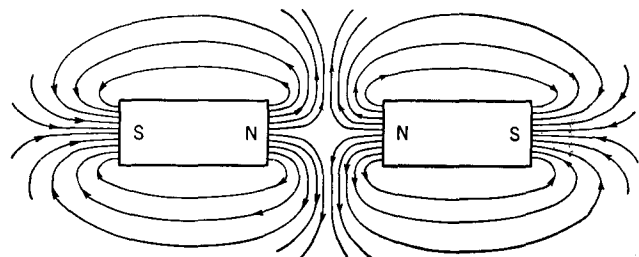


FIG. 4-4 MAGNETIC FIELDS OPPOSING

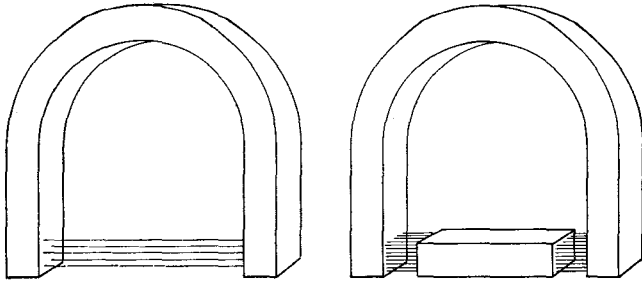


FIGURE 4-5

crease 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 electric circuit by substituting a conductor of lower resistance for one of higher resistance.

### 4.3 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  $\phi$ , 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—

$$B = \frac{\phi}{A} \quad (4:1)$$

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

Since the lines of magnetic induction are increased by the insertion of some material other than air in the magnetic field, it follows that the flux density depends upon the materials of the completed magnetic circuit and the strength of the magnet. This corresponds to an electric circuit wherein the current density in any given cross-section of conductor depends upon the resistance of the closed circuit and the electromotive force applied. There is, therefore, a property of a magnetic circuit which is analogous to the resis-

tance of an electric 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**. 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 circuit**. Expressed mathematically, this may be written—

$$\phi = \frac{M}{R} \quad (4:2)$$

where the symbol for flux is  $\phi$ , for magnetomotive force,  $M$ , and for reluctance,  $R$ . This may be compared to Ohm's Law as expressed by equation (2:1).

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

While we see that in many respects the magnetic circuit is analogous to the electric 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 electric circuit submerged in water. When the continuity of the metal conductors forming such an electric circuit is broken, the circuit will be completed through the liquid across its gap. Though

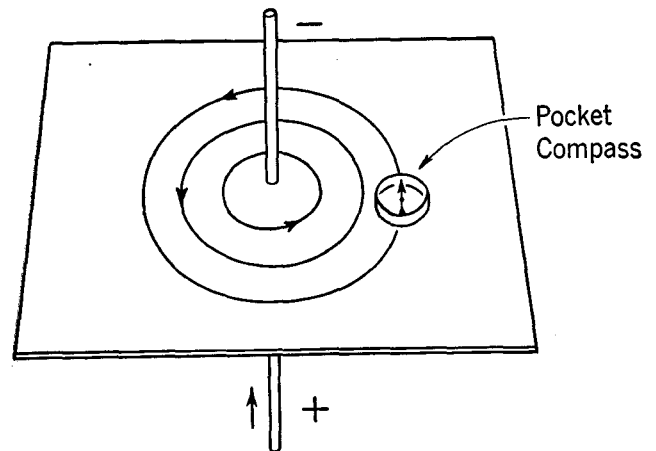


FIG. 4-6 MAGNETIC FIELD AROUND CURRENT-CARRYING STRAIGHT CONDUCTOR

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 in 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.

#### 4.4 Electromagnets

If a conductor carrying an electric current pierces a cardboard as shown in Figure 4-6, 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 4-7. Through such observations as these, we learn that **wherever an electric current is flowing there is also present a magnetic field, and the loops formed by the encircling lines of magnetic induction are always in a plane perpendicular to the conductor.**

If in either Figure 4-6 or 4-7 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 relationship. 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 4-6.

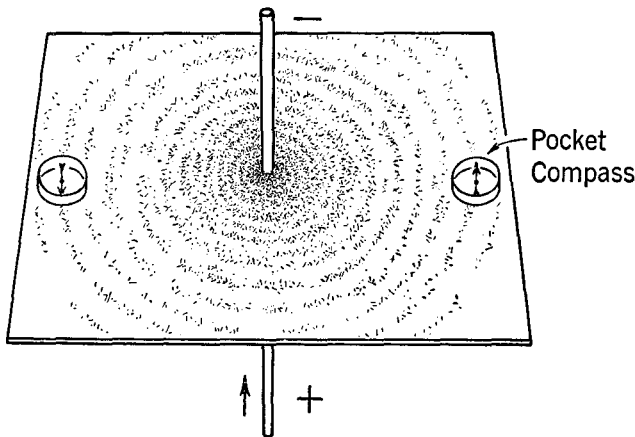


FIGURE 4-7

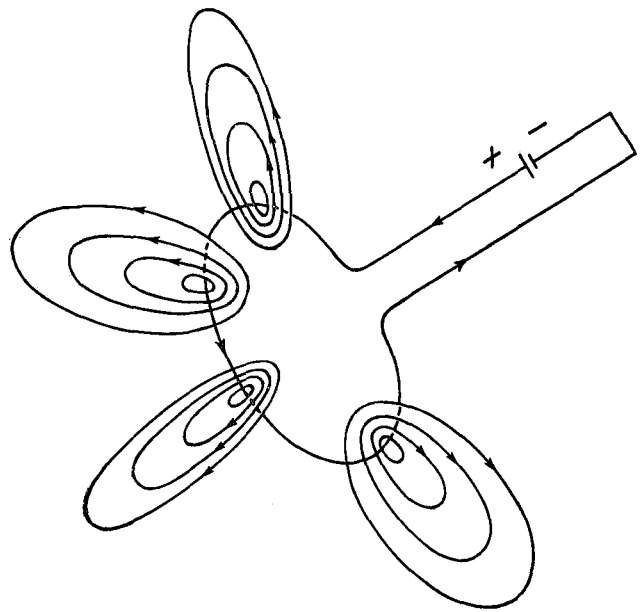


FIG. 4-8 MAGNETIC FIELD AROUND CURRENT-CARRYING LOOP

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 electric 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 4-8. 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 4-9 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 electric 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 4-10, 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

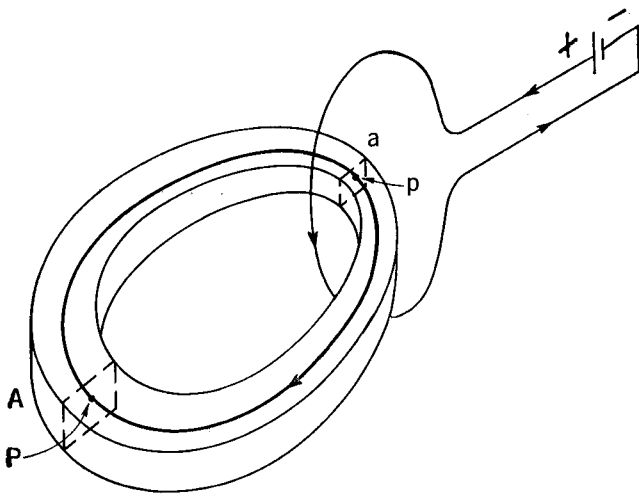


FIGURE 4-9

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

Comparing Figure 4-10 with Figure 4-1, we find that the current in the coil of wire creates a magnetic field similar to that of the bar magnet. In Figure 4-6 the relationship between direction of current flow and direction of lines of induction was shown by arrows. We use this same relationship in Figure 4-8 and going one step further, we may determine the north and south poles of the magnet formed by the coil shown in Figure 4-10. A simple way to remember the relationship for any winding is illustrated by Figure 4-11. Here if we 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 4-5 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 4-10 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 4-12, we have the customary electromagnet which is capable of exerting considerable force.

#### 4.5 Relation Between Current and Field Intensity

If we increase the current strength in the winding shown by Figure 4-10, we will find that the intensity of the magnetic field is increased pro-

portionately. 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 electric current for any given set of conditions.

A winding such as that shown in Figure 4-10 is called a **solenoid**. If such a solenoid is very long as compared to its diameter, the field intensity in the air on the inside of the solenoid is directly proportional to the product of the number of turns and the current, and inversely proportional to the length of the solenoid. Expressed mathematically—

$$H = k \frac{NI}{l} \quad (4:3)$$

where  $N$  is the total number of turns,  $I$  the current,  $l$  the length of the solenoid, and  $k$  is a constant depending on the units in which the other factors are expressed. In the rationalized mks (meter-kilogram-second) system of units,  $k$  is equal to 1,  $I$  is in amperes,  $l$  is in meters and  $H$  is then **ampere-turns per meter**.

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 unit length of circuit**. Its analogy in the electric 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 4-10 in ampere-turns, as the field intensity times the length of the solenoid—thus:

$$M = H \times l = kNI \quad (4:4)$$

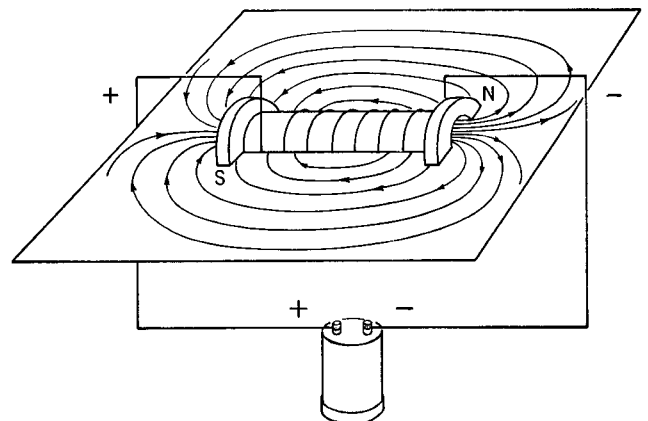


FIG. 4-10 MAGNETIC FIELD AROUND AIR-CORE SOLENOID

#### 4.6 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 4-10, the number of lines of induction will be greatly increased. This means that the flux density, or the number of lines per unit area 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. Because of the lowered 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.

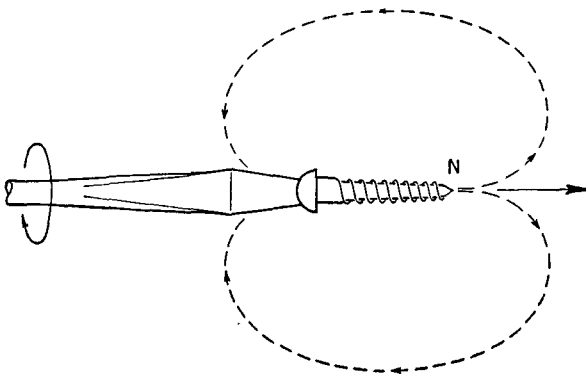


FIGURE 4-11

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 electric circuit. This term is known as **permeability** and is represented by the Greek letter  $\mu$ . 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:

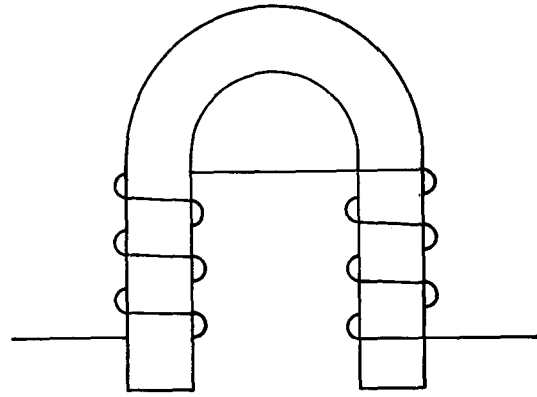


FIG. 4-12 HORSESHOE ELECTRO-MAGNET

$$\frac{\phi}{A} = H \times \mu \quad (4:5)$$

or

$$B = H \times \mu \quad (4:6)$$

where  $B$  is the conventional symbol for flux density.

This equation may also be written in other forms:

$$\mu = \frac{B}{H} \quad (4:7)$$

or

$$H = \frac{B}{\mu} \quad (4:8)$$

#### 4.7 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 electric conductor. In the electric circuit, the resistance or conductivity remains very nearly fixed for any degree of current strength, unless there is some change in

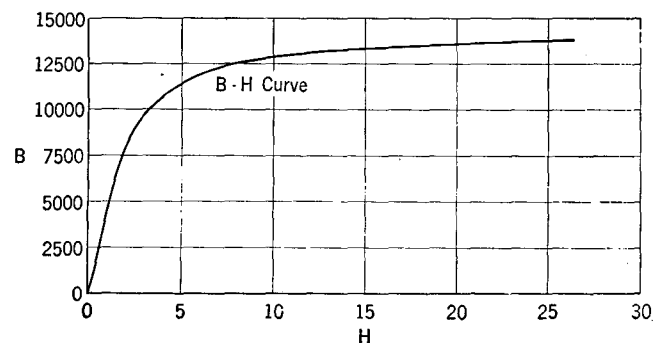


FIG. 4-13 B-H CURVE FOR IRON

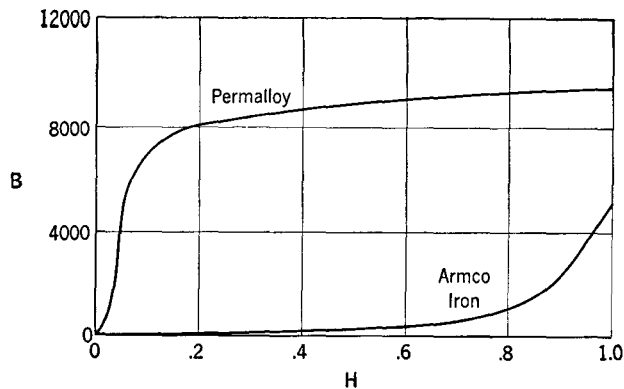


FIG. 4-14 B-H CURVE FOR PERMALLOY

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 unit area 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.

#### 4.8 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, and also

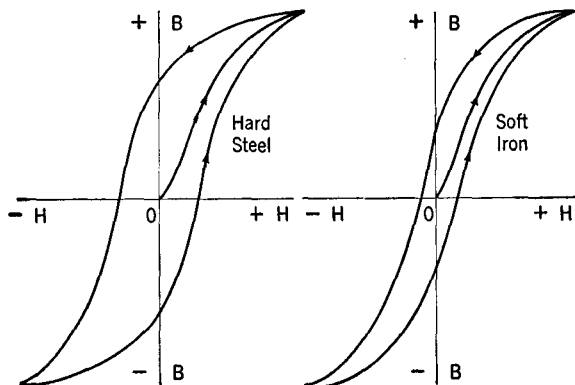


FIG. 4-15 HYSTERESIS LOOPS FOR SOFT IRON AND STEEL

to all of the magnetic alloys. Table II shows the resistance of electric conductors compared with copper. A similar table could be compiled for electric 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 4-13 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 magnetic material has some such curve. A

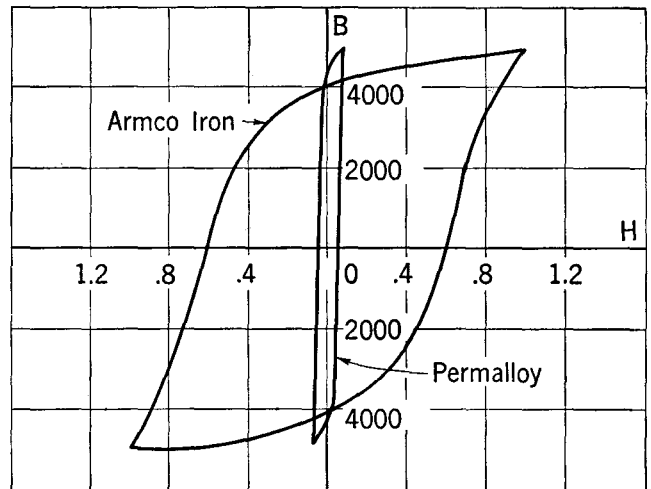


FIG. 4-16 HYSTERESIS LOOPS FOR MAGNETIC IRON AND PERMALLOY

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

4-14 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.

#### 4.9 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 4-15. This is called the **hysteresis loop**. Here it is seen that after the iron

has once reached the 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 4-15 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 4-16 where the hysteresis loop for permalloy and iron are compared.



## CHAPTER 5

### ELECTRICAL MEASUREMENTS IN DIRECT-CURRENT CIRCUITS

#### 5.1 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. Basic instruments of this kind include 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 electric (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.

In direct-current measurements, both ammeters and voltmeters are ordinarily galvanometer types

of instruments. The ammeter is used for measuring the flow of current in a circuit. For this reason it must be inserted directly in the path of the current as was done with the water flowmeter shown in Figure 2-2 (Chapter 2). It can readily be seen that if the meter is to measure accurately the current flow in the circuit, it must not disturb to any appreciable extent the current or voltage relations in the circuit to be measured. To insure this, the resistance of the meter is made extremely low. Thus, when an ammeter is inserted in a circuit as in Figure 5-1, the current flow measured by the meter is equal to

$$\frac{E}{R_1 + R_2 + R_3 + r}$$

where  $r$  is the resistance of the meter. Since  $r$  is relatively insignificant, this value is not appreciably different from the current flow without the meter, which is equal to

$$\frac{E}{R_1 + R_2 + R_3}$$

For the same reason the voltage drops across the resistors,  $IR_1$ ,  $IR_2$  and  $IR_3$ , the sum of which must equal  $E$ , remain effectively undisturbed.

The voltmeter is used to measure voltage or potential drop. It is connected across a battery or a resistor across which exists a potential drop. This is also illustrated in Figure 5-1. Since the current or voltage relations in the circuit are also to remain undisturbed, the voltmeter must not draw an appreciable current from the circuit

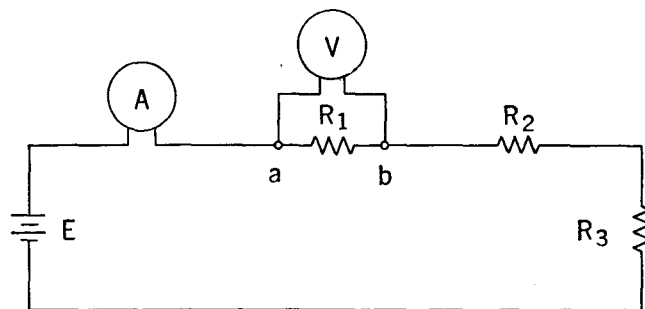


FIG. 5-1 VOLTMETER-AMMETER RESISTANCE MEASUREMENT

across which it is connected. For this reason the voltmeter, unlike the ammeter, must have an extremely high resistance.

We have learned that the combined resistance of resistors in parallel is lower than that of the smallest resistor. It follows that a high resistor in parallel with a relatively small resistor will not appreciably change the total net resistance. This leads to a general rule for the use of voltmeters—**the resistance of a voltmeter must be very much higher than the resistance of the circuit across which the voltage drop is being measured.**

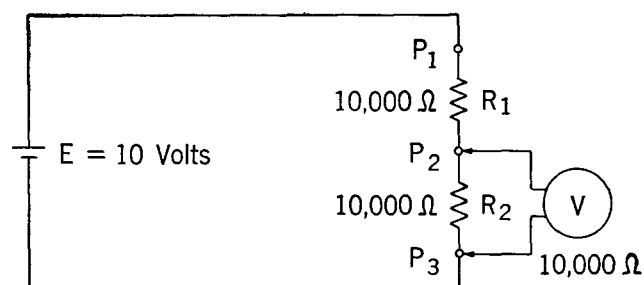


FIGURE 5-2

In order to show the error introduced by using a voltmeter whose resistance is comparable with that of the circuit element across which it is bridged, consider the circuit shown in Figure 5-2. If the voltmeter were not connected to the circuit, the current flowing in the circuit would be

$$I = \frac{E}{R_1 + R_2} = \frac{10}{10,000 + 10,000} = .0005 \text{ amps.}$$

The voltage drop either between points  $P_1$  and  $P_2$  or between  $P_2$  and  $P_3$  would be

$$IR_1 = IR_2 = .0005 \times 10,000 = 5 \text{ volts}$$

If, however, a voltmeter having a resistance of only 10,000 ohms is connected to  $P_2$  and  $P_3$  as shown, the total resistance between these points is the net parallel or combination resistance of the 10,000-ohm voltmeter and the 10,000-ohm resistor  $R_2$ ; that is, 5,000 ohms. Then the total current in the circuit is

$$I = \frac{10}{10,000 + 5,000} = .000667 \text{ amps.}$$

Consequently, the voltage drop between  $P_2$  and  $P_3$  now is

$$IR_2 = .000667 \times 5,000 = 3.33 \text{ volts.}$$

Therefore, the voltage read by the voltmeter is

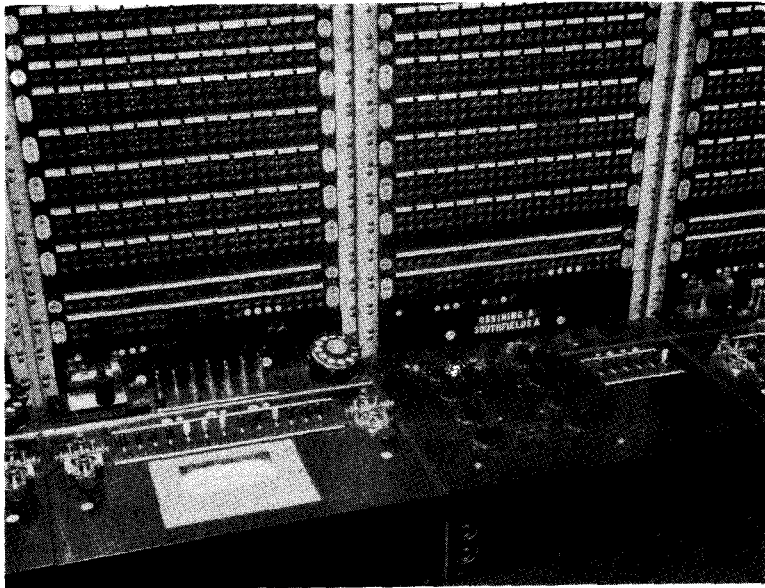
only 3-1/3 volts instead of the 5 volts which would actually exist between  $P_2$  and  $P_3$  if the voltmeter were not used, or if a voltmeter were used whose resistance was much greater than the 10,000-ohm resistor across which it is bridged.

To avoid error, or when it is desired to measure the voltage across very high-resistance circuits, an "electron-tube voltmeter" is often employed. This is essentially a voltmeter that is associated with an electron-tube amplifier, the amplifier making it possible to obtain a full scale reading on the meter without drawing an appreciable amount of power from the circuit to which the voltmeter is connected. In other words, the input resistance of such a voltmeter may be practically infinite.

If a voltmeter measures at any given instant the emf across any direct current 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 equation  $P = EI$ , 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 basic group. These are the Wheatstone bridge and the megger. The megger is a combination of a magneto source of relatively high electromotive force and a sensitive meter, calibrated to read values of very high resistances connected across its terminals. The Wheatstone bridge is a network of resistors which is used in connection with a galvanometer for measuring an electrical resistance by an accurate comparison method. A more detailed discussion on the use of this device follows later in this Chapter.

In using any electrical instrument it must be remembered that Ohm's Law is never failing and that it applies to every circuit branch. Thus, in the case of the ammeter, the current that will flow through it 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 ohms 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 poten-



TESTSHELF OF NO. 5 TESTBOARD SHOWING VOLTMETER AND WHEATSTONE BRIDGE TESTING ARRANGEMENT

tial, 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.**

## 5.2 Simple Resistance Measurements

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.

Two basic methods for measuring electrical resistances are discussed below:

### a. Voltmeter-Ammeter Method

Figure 5-1 shows a simple series circuit. Let us assume that it is desired to determine the value

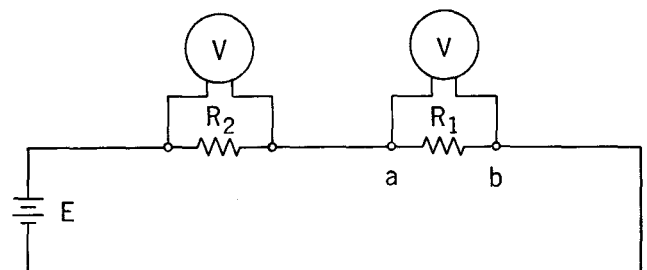


FIG. 5-3 DROP IN POTENTIAL RESISTANCE MEASUREMENT

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 emf 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 5-1, the voltmeter reading is 5 volts and the ammeter reading is .5 ampere; what is the value of resistance  $R_1$ ?

**Solution:**  $R_1 = \frac{V_1}{I_1} = \frac{5}{.5} = 10$  ohms, ans.

### b. Drop in Potential Method

If in Figure 5-3, it is desired to determine the value of the resistor  $R_1$ , the "drop in potential method" can be used if a second resistor  $R_2$  of known value is inserted in series and the voltage drops across both  $R_1$  and  $R_2$  are measured. Since the two resistors are in series, the same current is flowing through both and from Ohm's Law:

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

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

$$R_1 = R_2 \frac{V_1}{V_2} \quad (5:1)$$

**Example:** If in Figure 5-3 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?

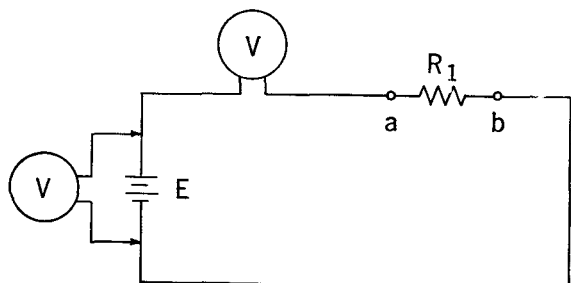


FIG. 5-4 INSULATION MEASUREMENT

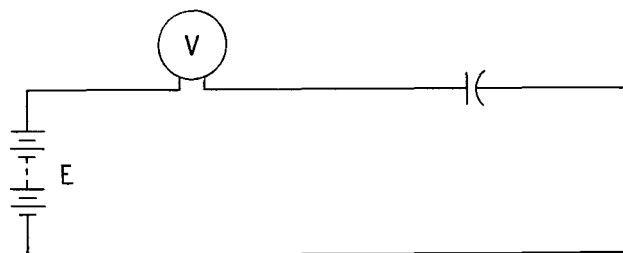


FIGURE 5-5

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

### 5.3 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 5-3 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 equation for this special case may then be written—

$$R_1 = R_2 \frac{E - V_2}{V_2} \quad (5:2)$$

where  $E$  is the emf of the battery.

If  $R_1$  is very high in value such as a "leak" due to poor insulation, it can be measured using equation (5:2) but instead of using a second known resistance, the voltmeter itself may be inserted in series with the battery and  $R_1$ , as shown in Figure 5-4. 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 resistor  $R_2$  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.

Figure 5-5 shows the drop of potential method with series voltmeter for measuring the insulation of a capacitor. Figure 5-6 shows a "leak" between two cable conductors and Figure 5-7 a "leak" between an open wire and ground, both being measured in the same manner.

For this application equation (5:2) is ordinarily written—

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

or

$$X = r \left( \frac{E}{V} - 1 \right) \quad (5:3)$$

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.

**Example:** The voltmeter shown in Figure 5-5 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 capacitor?

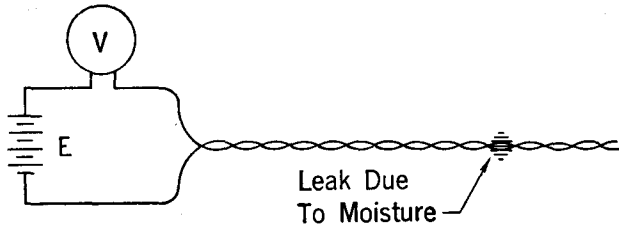


FIG. 5-6 METALLIC INSULATION TEST

**Solution:**

$$\begin{aligned} X &= r \left( \frac{E}{V} - 1 \right) \\ &= 100,000 \left( \frac{150}{8} - 1 \right) \\ &= 1,775,000 \text{ ohms, ans.} \end{aligned}$$

**Note:** Insulation resistance is usually expressed in megohms instead of ohms on account of its normally high value. One megohm equals one million ohms. Equation (5:3) 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.

#### 5.4 Theory of the Wheatstone Bridge

In practice the Wheatstone bridge provides a faster and more accurate means of making resistance measurements although the principles involved are not different from those of the potential drop method of measurement outlined in

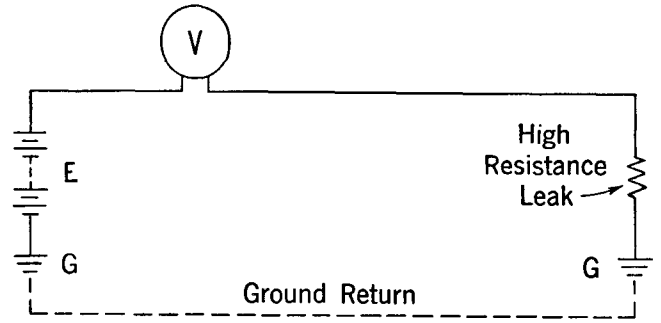


FIG. 5-7 TEST FOR INSULATION OF SINGLE WIRE

the preceding Article. Before reviewing specific Wheatstone bridge measurement techniques, however, it may be well to analyze briefly the broad general theory of this very widely used measuring instrument.

In Figure 5-8, the voltmeter has one terminal permanently connected to  $a$  and the other terminal may be moved along the resistor  $ab$ . The voltmeter reading will be zero when both terminals are at  $a$ , and will gradually increase as  $P$  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 the two points, or we may write:

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

where  $V'$  is the potential drop measured between  $a$  and any other point  $P'$ .

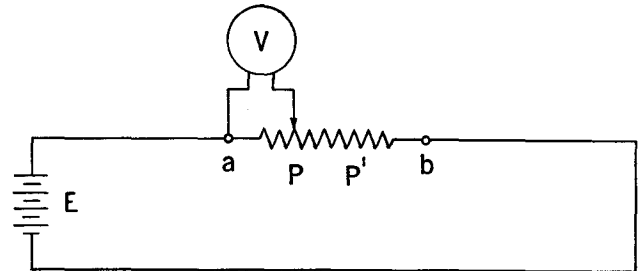


FIGURE 5-8

If instead of having one resistor as shown in Figure 5-8, we have two parallel resistors as shown in Figure 5-9, and one terminal of the voltmeter is moved along resistor  $ab$  while the other terminal is moved along resistor  $cd$ , 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

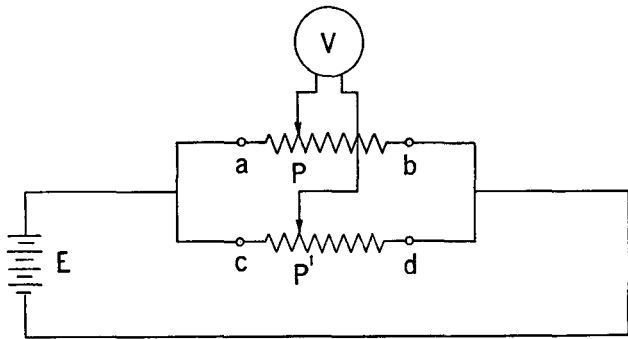


FIGURE 5-9

$P'$  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:

$$\frac{aP}{cP'} = \frac{Pb}{P'd} \quad (5:4)$$

In Figure 5-10 let us assume that the resistances 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 needle, we may write the same relation as was given by equation (5:4), namely:

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

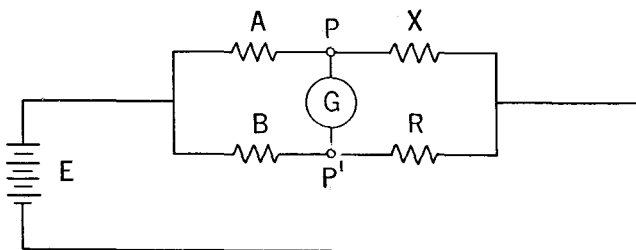


FIGURE 5-10

This equation can be expressed:

$$X = \frac{A}{B} R \quad (5:5)$$

which is the usual equation of the Wheatstone bridge.

Figure 5-11 illustrates the conventional method of showing the Wheatstone bridge. It is almost identical to the arrangement shown by Figure 5-10, but has the resistors 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 resistors  $A$  and  $B$  are called the **ratio arms** of the bridge and the resistor  $R$  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.

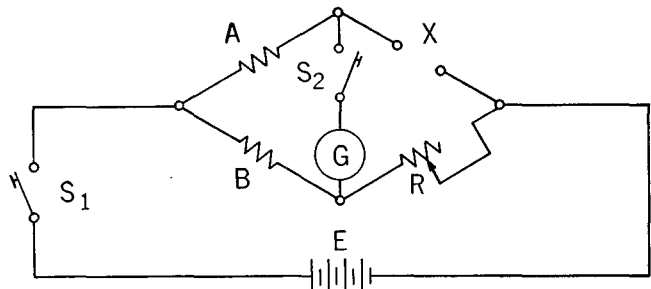


FIG. 5-11 STANDARD WHEATSTONE BRIDGE CONVENTION

Though Figure 5-11 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 equation (5:5) 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 5-11 the galvanometer and battery 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 three in particular that are used extensively in telephone and telegraph work. One is a small portable type bridge, which is used sometimes at local test desks. Another is Wheatstone bridge KS-3011 which is standard for permanent mounting on primary toll testboards and the third is Wheat-

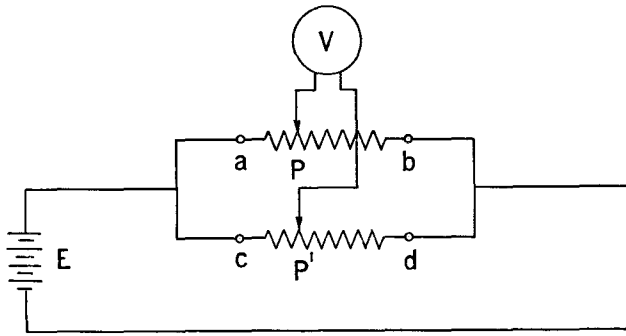


FIGURE 5-9

$P'$  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:

$$\frac{aP}{cP'} = \frac{Pb}{P'd} \quad (5:4)$$

In Figure 5-10 let us assume that the resistances 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 needle, we may write the same relation as was given by equation (5:4), namely:

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

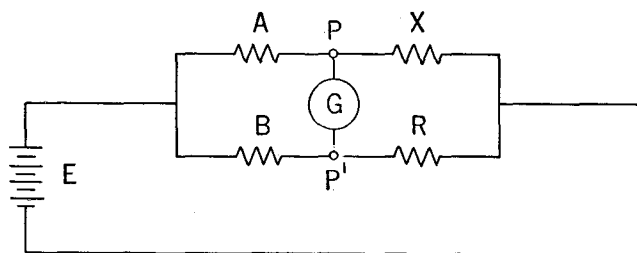


FIGURE 5-10

This equation can be expressed:

$$X = \frac{A}{B} R \quad (5:5)$$

which is the usual equation of the Wheatstone bridge.

Figure 5-11 illustrates the conventional method of showing the Wheatstone bridge. It is almost identical to the arrangement shown by Figure 5-10, but has the resistors 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 resistors A and B are called the **ratio arms** of the bridge and the resistor 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.

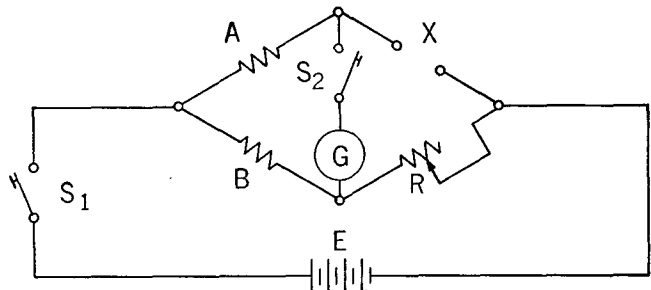
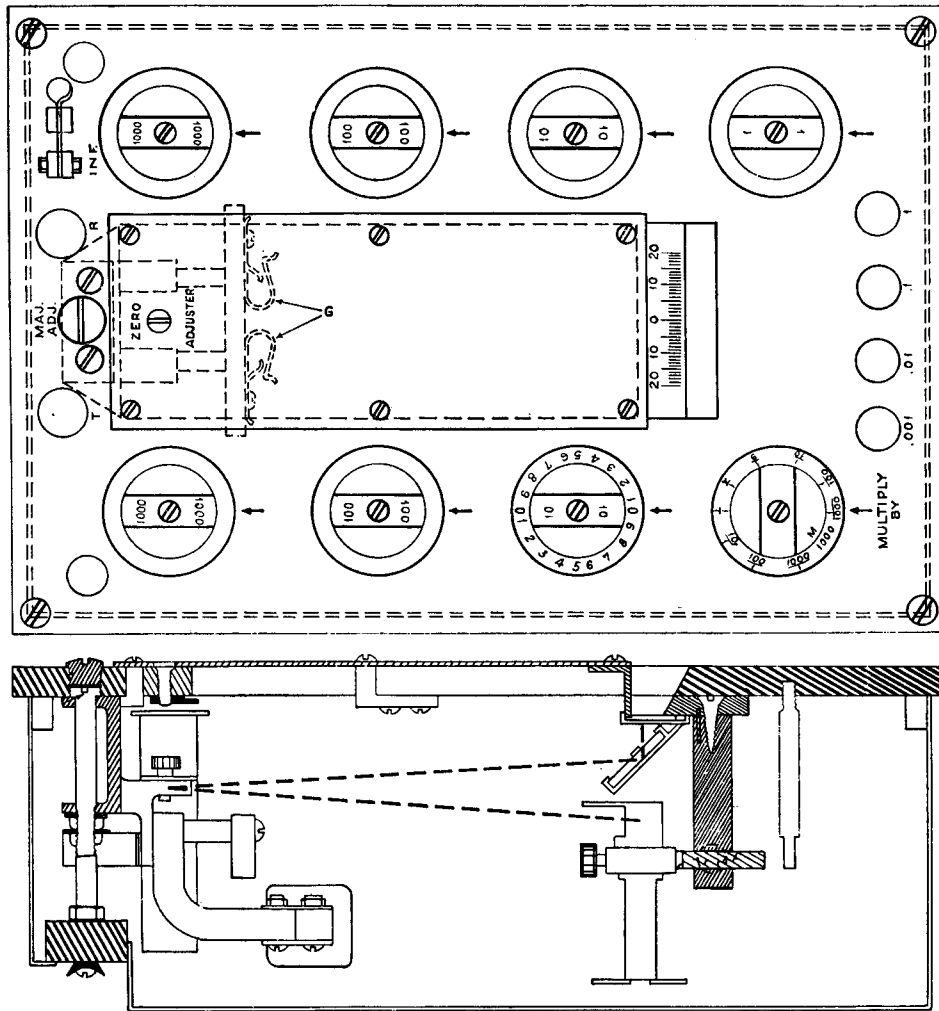


FIG. 5-11 STANDARD WHEATSTONE BRIDGE CONVENTION

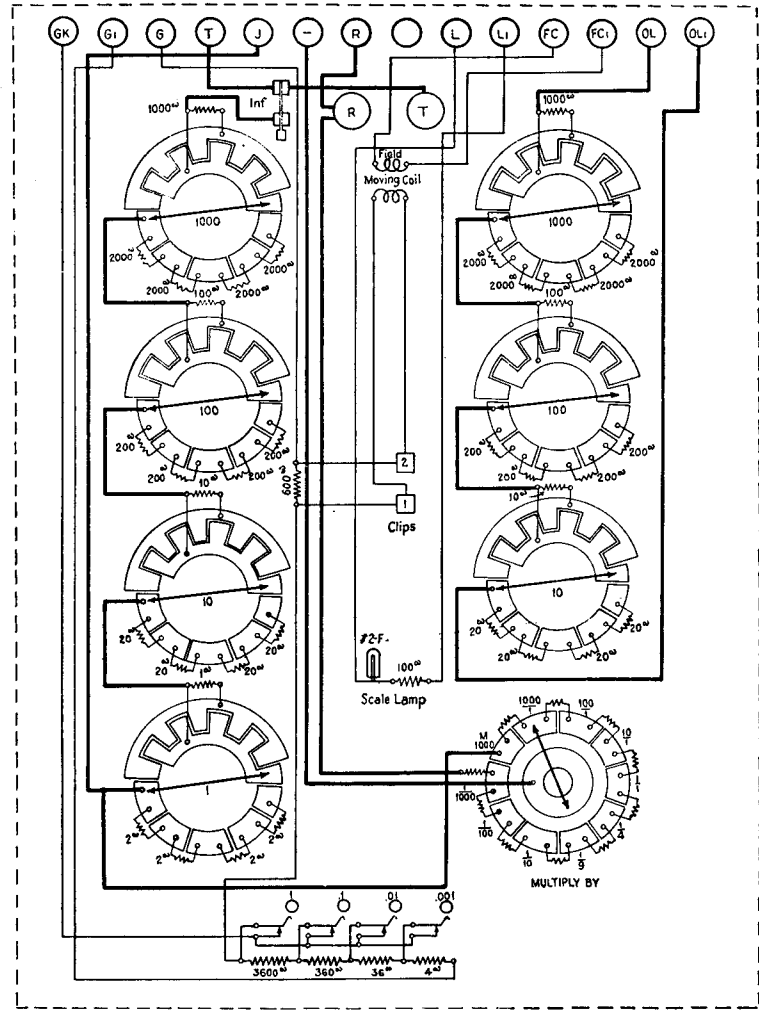
Though Figure 5-11 shows the resistance branch R as variable and the arms A and B as fixed, a balance could also be obtained by changing the ratio A/B in equation (5:5) instead of varying the value of R.

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 5-11 the galvanometer and battery 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 three in particular that are used extensively in telephone and telegraph work. One is a small portable type bridge, which is used sometimes at local test desks. Another is Wheatstone bridge KS-3011 which is standard for permanent mounting on primary toll testboards and the third is Wheat-



Dial and Galvanometer Arrangement



Wiring Diagram

FIG. 5-12 WHEATSTONE BRIDGE PER KS-3011



stone bridge KS-5588 used as part of a mobile "teawagon" type unit.

The dial and circuit arrangements for KS-3011 are shown by Figure 5-12. As indicated this bridge employs a reflection type galvanometer having a tiny mirror attached to the moving element. This permits the detection of smaller current values than is possible with smaller bridges where the indicating needle is attached to the moving element of the galvanometer.

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

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 5.7. The rheostat arm has a total resistance of 9999 ohms and is adjustable by means of four dials in steps of one ohm. The resistances in the rheostat arm are accurate to 1/10 of 1 per cent and in the ratio arms to 1/20 of 1 per cent. Three additional dials are provided for use in connection with open location tests.

### 5.5 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 5-13 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 determined from the resistance measurement and the values given in Table III, or by comparison with the measured resistance of another loop of known length.

**Example:** In Figure 5-13 the cable conductors are #19 AWG, 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?

**Solution:** Let  $d$  = distance in miles

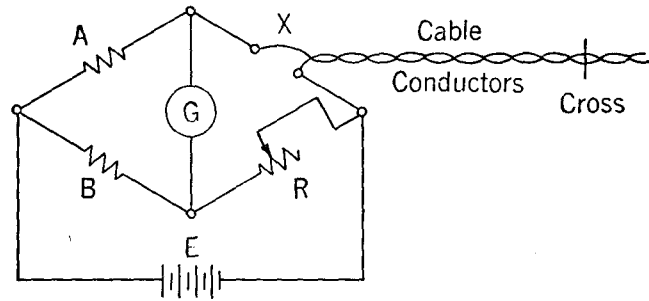


FIG. 5-13 LOOP RESISTANCE MEASUREMENT TO CROSS

Loop resistance of cable per mile from Table III is 84.1 ohms

$$X = 55 \text{ ohms}$$

$$d = 55 \div 84.1 = .654 \text{ mile, ans.}$$

If the cross shown in Figure 5-13 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 .654 mile from the office. It is therefore necessary in using the loop method to make two measurements to determine accurately 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 .654 mile away, is some point between .654 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 loca-

tion is the mean, or point half way between the two. Of course, in practice it may not be 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 5-14. This permits testing in both directions from the same office. If the exact length of the good pair is not known, it can be determined by making a measurement with the distant end crossed.

**Note:** As pointed out in the next Article, the quickest and most accurate method of locating a cross in practice is by the use of a Varley measurement. But the theory underlying the foregoing should be thoroughly mastered before taking up the later type of test.

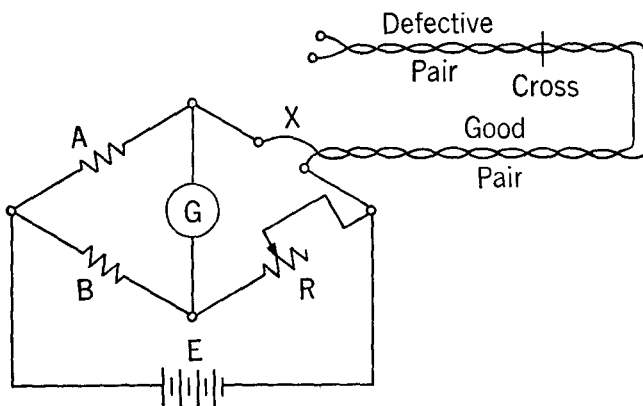


FIGURE 5-14

**Example:** In Figure 5-14 the good cable pair when short-circuited at the distant end has a loop resistance 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 AWG?

**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:

$$d_1 = \frac{37}{84.1} = .440 \text{ mile}$$

The length of the good pair is:

$$l = \frac{63}{84.1} = .749 \text{ mile}$$

The imaginary distance from the distant end is:

$$l - d_2 = \frac{108}{84.1} - .749 = .535 \text{ mile}$$

Then the actual cross is at point half way between .440 mile from the near office and .535 mile from the distant office or  $.749 - .535 = .214$  mile from the near office. The actual location is therefore

$$\frac{.440 + .214}{2} = .327 \text{ mile, ans.}$$

The resistance of the cross caused an error in the single measurement location of  $.440 - .327 = .113$  mile of cable pair. This expressed as resistance is  $.113 \times 84.1 = 9.5$  ohms, ans.

An important 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 all four wires are crossed at B.

**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:**

$$W_1 + W_2 = 90 \quad (a)$$

$$W_2 + W_3 = 90 \quad (b)$$

$$W_1 + W_3 = 88 \quad (c)$$

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

$$W_1 - W_3 = 0 \quad (d)$$

Adding (c) and (d) we have

$$2W_1 = 88, \text{ or } W_1 = 44 \text{ ohms}$$

Substituting in (a) we have

$$44 + W_2 = 90 \text{ ohms or } W_2 = 46 \text{ ohms}$$

Substituting in (c) we have

$$44 + W_3 = 88 \text{ ohms or } W_3 = 44 \text{ ohms}$$

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

$$44 + W_4 = 88 \text{ or } W_4 = 44 \text{ ohms}$$

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

### 5.6 Varley Loop Tests

The Wheatstone bridge may be used to locate a defect due to a grounded 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 5-15 shows the theory of an ordinary Varley loop test for a ground.

In comparing this figure with Figure 5-11, 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 5-11 becomes in Figure 5-15 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 5-15 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 III 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 5-15 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 grounded Varley measurement.

Thus, referring again to Figure 5-15, it will be seen that when a Varley balance is obtained, with the bridge ratio arms equal—

$$R + d = l + x \quad (a)$$

or

$$d = (l + x) - R \quad (b)$$

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

$$L = l + x + d \quad (c)$$

from which

$$d = L - (l + x) \quad (d)$$

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

$$2d = L - R \quad (e)$$

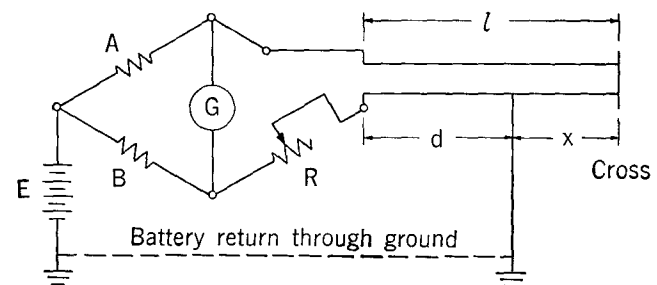


FIG. 5-15 GROUNDED VARLEY TEST

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 precisely the unit resistance value of the two wires, we still do not know the exact 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 as  $k$ , and the total length of the line as  $D$ , then—

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

from which

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

or applying (e) above

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

**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 the distance of the ground from the measuring end?

**Solution:**

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

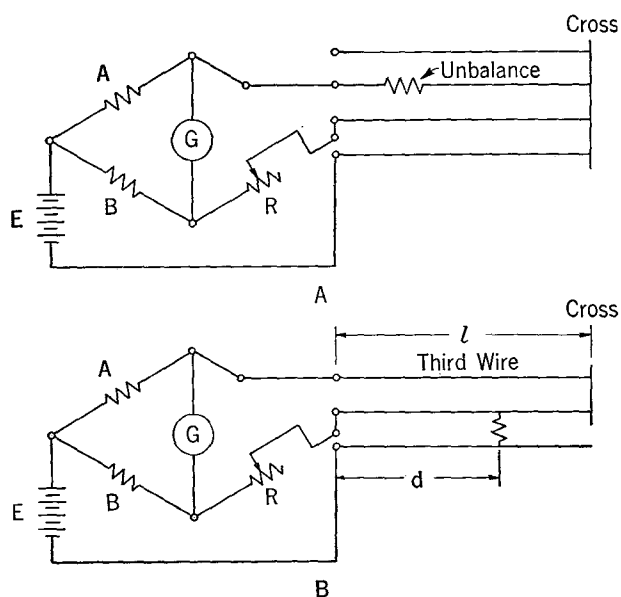


FIG. 5-16 METALLIC VARLEY TESTS

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 5-16 (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 5-16 (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

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

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.

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.

In practical line testing work, it is customary to use all three of the types of measurements we have been discussing for locating a cross or ground—i.e.: a grounded Varley measurement, a metallic Varley measurement, and a loop resistance measurement. This method, commonly called **the 3-Varley method** will give accurate results regardless of whether the several line wires used (including the defective wire) have the same overall resistance, regardless of the resistance of the cross or ground connection, and regardless of the insulation resistance between the wires and between the wires and ground.

Figure 5-17 illustrates the arrangement for making such measurements. In this case, where a grounded fault is to be located, two good wires, which may or may not have the same resistance, and the faulty wire are connected together at the distant office. A metallic Varley measurement which will indicate the difference in resistance between one good wire and the faulty wire, a grounded Varley measurement which gives the resistance from the home office to the fault, and a loop measurement giving the total resistance between the two offices are made.

The loop measurement is taken by connecting the switch arm *s* to the  $V_1$  position (Figure 5-17) and adjusting *R* until the bridge is balanced. This reading will be designated as  $V_1$ .

The grounded Varley measurement is made by switching to the  $V_2$  position and adjusting for balance. This reading is designated as  $V_2$ .

A metallic Varley measurement is made by switching to the  $V_3$  position. This reading is  $V_3$ .

If the *A* and *B* arms are set at equal values for

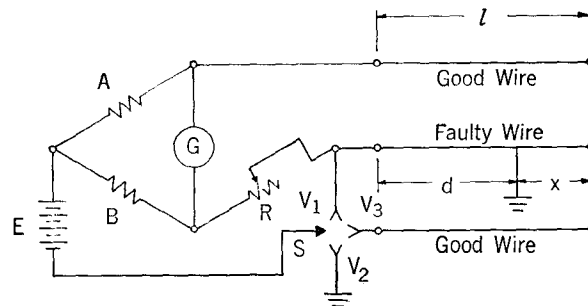


FIG. 5-17 3-VARLEY TEST FOR GROUND

all three measurements, we can see by inspection of Figure 5-17 that

$$V_1 = l + x + d$$

$$V_2 = l + x - d$$

$$V_3 = l - x - d$$

Now we may note that

$$V_1 - V_3 = 2d + 2x$$

which is twice the resistance of the faulty wire. Similarly

$$V_1 - V_2 = 2d$$

which is twice the resistance from the home office to the fault location. Then calling the distance to the fault *k* and the total length of the wire *D* as in (h) preceding, it is evident that

$$k = \frac{V_1 - V_2}{V_1 - V_3} \times D$$

It is particularly important to note in connection with this 3-Varley measuring method that the measured values in all cases include only values of the resistance of the faulty wire itself. The specific resistance values of other wires used in making the measurements are immaterial since they do not appear in the final equations.

**Example:** Assume that in Figure 5-17, a toll circuit is 42 miles long and that the following 3-Varley measurements are obtained:

$$V_1 = 3970$$

$$V_2 = 1700$$

$$V_3 = 20$$

What is the distance from the measuring office to the fault?

**Solution:**

$$\begin{aligned} k &= \frac{V_1 - V_2}{V_1 - V_3} \times D \\ &= \frac{3970 - 1700}{3970 - 20} \times 42 = 24.15 \text{ miles} \end{aligned}$$

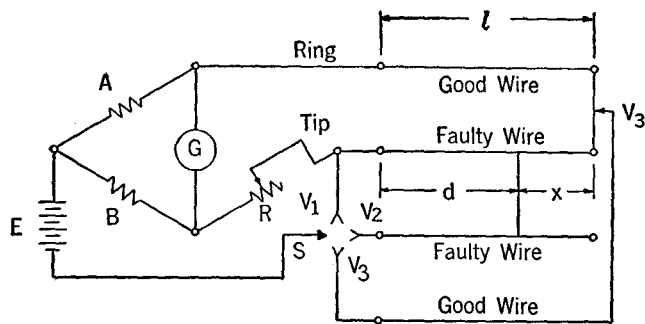


FIG. 5-18 3-VARLEY TEST FOR CROSS

The 3-Varley method is equally applicable for locating faults due to crosses where the measuring set-up would be as shown in Figure 5-18 and the calculations for determining the location would be the same as in the case of the ground fault discussed above.

To simplify the above discussion on fault location we have in each case assumed that the ratio arms  $A$  and  $B$  of the bridge could be set at equal values. Where the values of the resistances being measured are such that it is desirable to use unequal values of the ratio arms, the indicated  $R$  values must of course be qualified by the ratio  $\frac{A}{A+B}$  to obtain the correct measurement. When unequal values of the ratio arms are necessary the experienced tester will prefer to use the 1/9 ratio because the value of  $\frac{A}{A+B}$  then becomes a convenient 1/10 and it is only necessary to point off one decimal place to obtain the correct result.

### 5.7 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 5-19. 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.

**Example:** In Figure 5-19, 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

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

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

The advantage of the Murray test in locating a fault such as discussed above lies in the fact that the test does not require the use of a third wire as would be necessary in the Varley method. Except in certain special conditions involving rural lines, one pair service cables, etc. the Murray test is rarely used in telephone practice for locating grounds or crosses. The Murray type connec-

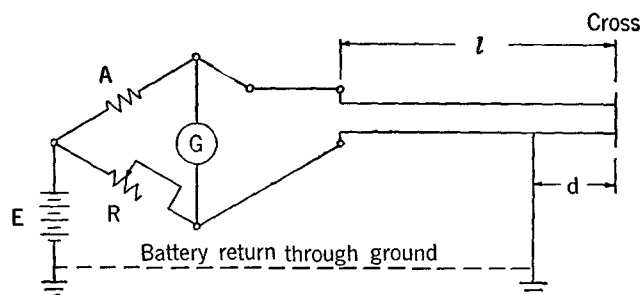


FIG. 5-19 MURRAY TEST

tion is commonly used, however, for locating opens. But since the wires here are open, it is obvious that no ordinary d-c measurement can be made. Instead, a low frequency a-c is generated by means of an "interrupter" which reverses the battery voltage 8 times per second and simultaneously reverses the polarity of the galvanometer connections. The bridge when balanced then compares the "capacitance" of the good wire to its far end with that of the defective wire to the point where it is open.

As we shall see in later Chapters, the capacitance value of the wire is directly proportional to its length and the ratio given by the bridge reading will therefore be equal to the ratio of the length of the good wire to the length of the defective wire to the fault.

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 many toll cables, parts of each section are aerial and other parts are underground. There is normally a considerable temperature difference between aerial and underground facilities and a corresponding difference in the wires contained in the cable sheaths. In such wires, it is necessary to apply correction factors to the measurements taken in order to locate a fault accurately. 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 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.

## CHAPTER 6

### THE DIRECT-CURRENT DYNAMO-ELECTRIC MACHINE

#### 6.1 Induced Electromotive Force

Chapter 4 describes how lines of magnetic induction exist around any wire in which there is an electric 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 6-1 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 4-11 in Chapter 4 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 6-1(D) and (E).

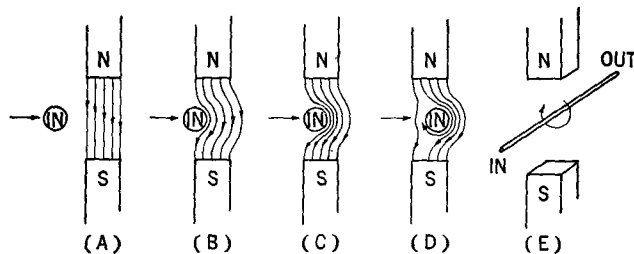


FIG. 6-1 WIRE MOVING THROUGH MAGNETIC FIELD

This rule, stated in another way, is called the **right-hand rule** for remembering the induced emf relation. It is illustrated in Figure 6-2. 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 emf, or the direction of current flow. If a galvanometer

is connected to the conductor, as in Figure 6-3, it will be found that the effect is more noticeable when the conductor is moved swiftly. From these

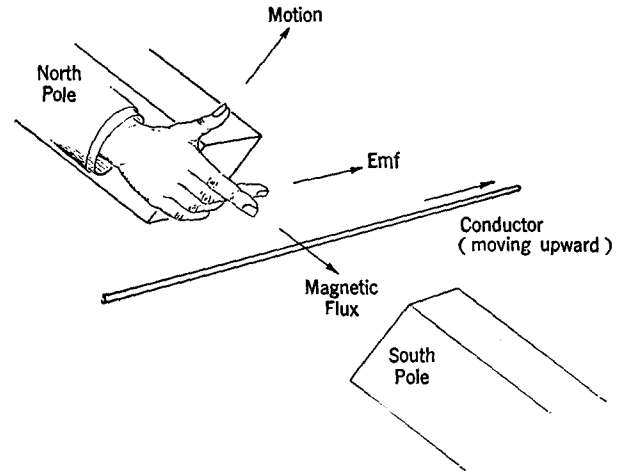


FIG. 6-2 RIGHT-HAND RULE

and other similar experiments we learn that the law for induced emf may be stated as follows:

**When any conductor is made to cut lines of magnetic induction there will be an emf induced in it, and the direction of the emf, 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 magnitude of the in-**

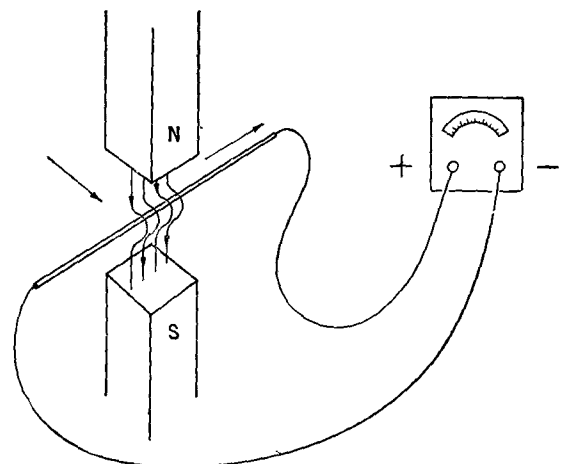


FIGURE 6-3



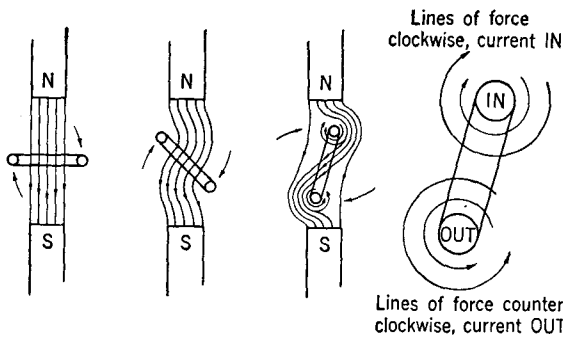


FIG. 6-4 LOOP ROTATING IN MAGNETIC FIELD

duced emf depends upon the rate of cutting magnetic lines, or the number of lines cut per second.

### 6.2 Emf 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 magnetic field between the poles of a magnet, as shown in Figure 6-4. 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 emf induced has a different direction in the two conductors.

But, because the loop is complete, these emf's will aid in causing a continuous current in the direction  $a-b-c-d$ , as shown by Figure 6-5(A). The values of these emf's at any given instant will depend upon the position of the loop. When the plane of the loop becomes perpendicular to the magnetic field as in Figure 6-5(B), each conductor will be moving parallel to the direction of the lines, the loop will be in a neutral position, and the generated emf will have decreased to zero. If the loop is then turned through an angle of  $90^\circ$  in the same direction (Figure 6-5(C)), it will again be cutting lines at the maximum rate, but the emf will be reversed with respect to the loop

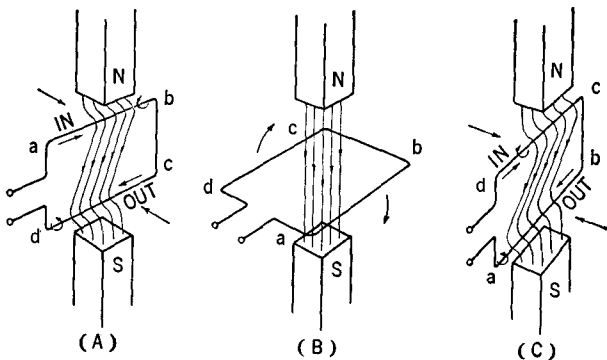


FIGURE 6-5

itself and the current will be in the direction  $d-c-b-a$ , or opposite to that in Figure 6-5(A).

With the loop revolving at constant speed, the emf 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 emf 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 emf may be determined by the horizontal motion of the loop per angular degree through which it turns.

Figure 6-6 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  $c$ , 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 direc-

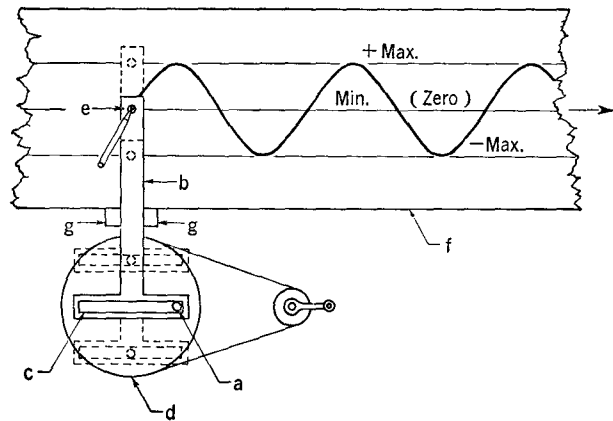


FIG. 6-6 MECHANISM FOR DRAWING SINE WAVE

tion, 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, 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

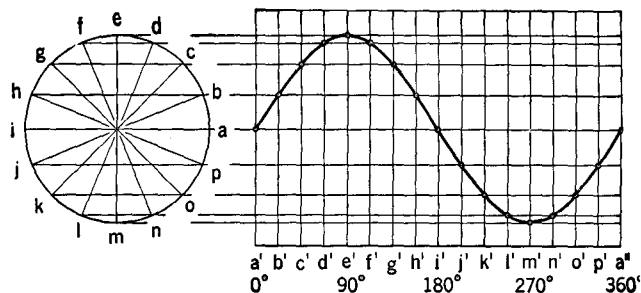


FIG. 6-7 GRAPHICAL CONSTRUCTION OF SINE WAVE

mid-point of its travel represents the emf induced in the conductor. This is apparent since the induced emf 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 emf 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 6-7, 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.

### 6.3 Principle of the Direct-Current Generator

The revolving loop or **armature** shown in Figure 6-5 may be connected to slip-rings, as shown in Figure 6-8(A). In this case the resulting emf between the two terminals or **brushes** will reverse in direction as the loop revolves, giving rise to an

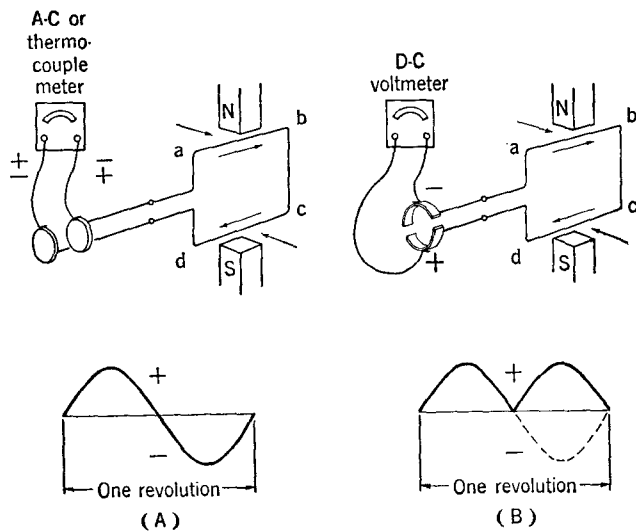


FIG. 6-8 SLIP-RINGS AND COMMUTATOR

**alternating** emf, one cycle of which is plotted in the Figure. If it is desired to produce a unidirectional emf, it is necessary to 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 6-8(B). This commutator reverses the connections to the armature leads just as the emf or current is reversed, changing the negative half-cycle to a positive pulsation. The resultant emf 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 6-9 in which two loops and four commutator segments are shown. The resultant emf is represented by the full lines at the right of the Figure. Comparing Figure 6-8 with Figure 6-9, it may be seen that an increase in the number of loops causes a smaller fluctuation in the armature emf. An armature wound with many turns therefore produces a practically continuous non-pulsating emf, causing a **direct current** to flow when the external circuit is closed.

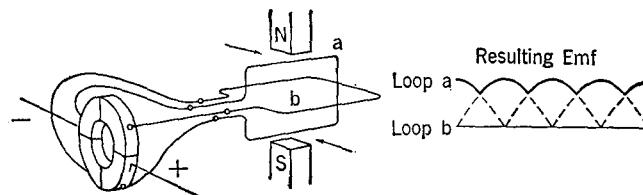


FIG. 6-9 EFFECT OF ADDITIONAL LOOPS

In Figures 6-5, 6-8 and 6-9, 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 6-10. 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 shown schematically in Figure 6-11. As the emf 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

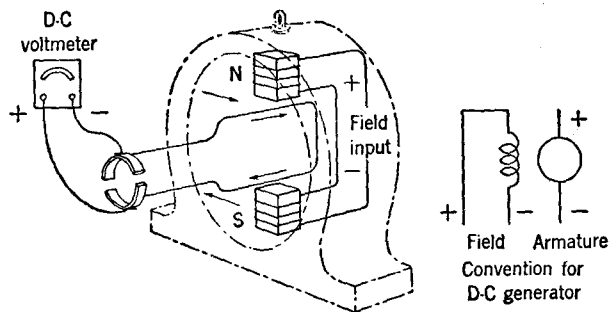


FIG. 6-10 SEPARATELY EXCITED D-C GENERATOR

the field current will cause a change in the armature emf. 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 emf. On the other hand, in the case of the series wound generator, the armature emf increases with the load current. The compound wound generator is designed to neutralize this change in armature emf 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 6-12 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 6-13. 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.

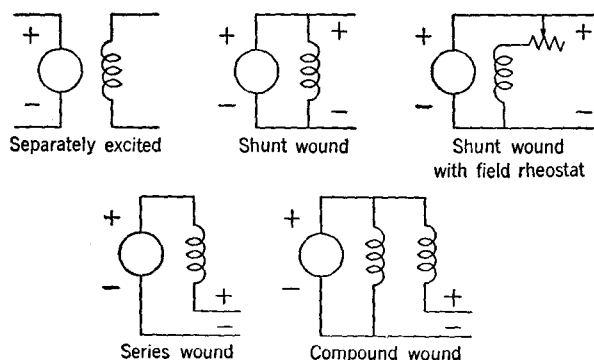


FIG. 6-11 CONVENTIONS FOR STANDARD TYPES OF D-C GENERATORS

## 6.4 D-C Generators for Supplying Central Office Power

Direct-current generators are widely used in telephone 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 certain relays and other apparatus; also various higher voltages, ranging up to a maximum of 130 volts, for electron tube plate supply and operation of telegraph circuits, etc. 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

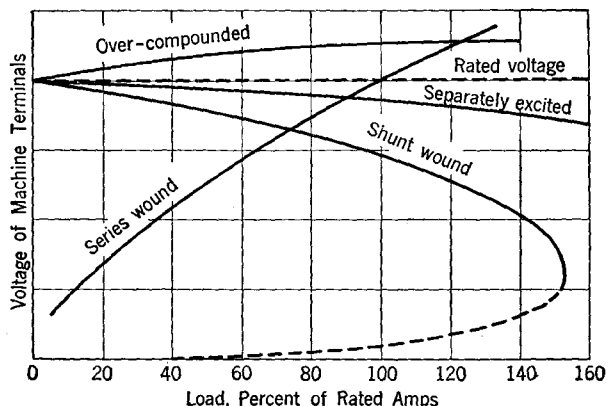


FIG. 6-12 LOAD-VOLTAGE CHARACTERISTICS OF STANDARD TYPES OF D-C GENERATORS

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 have an important filtering effect in reducing noise caused by the generators.

There are several arrangements of generators and batteries that have been used in the past to develop central office power. The 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.

In power plants where a battery, continuously floated across the line, is used, it is desirable for

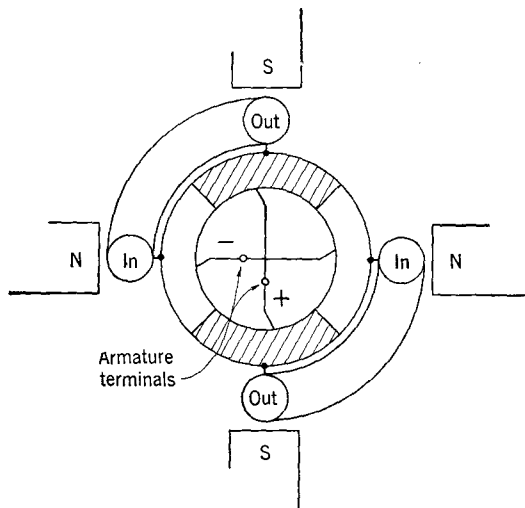


FIG. 6-13 FOUR-POLE GENERATOR

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.

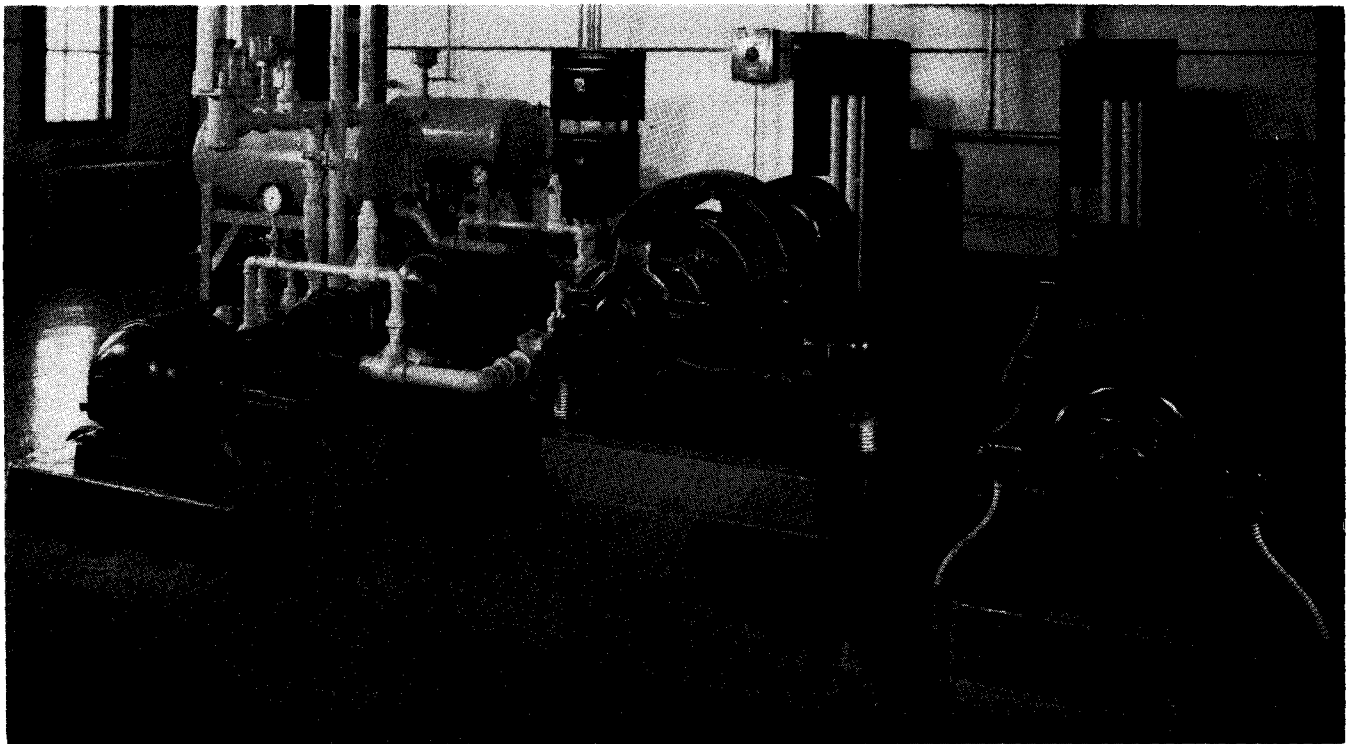
The required flat characteristic could be obtained from an ordinary compound wound generator of proper design; but such a machine is not safe to use because in case of failure of the outside power, the floating battery would run the

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.

The present practice is to use regulated voltage rectifiers for continuous float power plants where the load requirements are small, and shunt wound generators with automatic voltage regulation for power plants of larger capacity.

### 6.5 Typical Central Office Power Plant

Figure 6-14 is a schematic of a typical power plant where the total load requirements are in the order of 100 to 4,000 amperes and both a generator and a rectifier are used as sources of supply. The battery is continuously floated in this type of plant and the charging unit voltage must, therefore, be maintained at a constant value. As indicated in the Figure, this is accomplished automatically by means of a motor-driven field rheostat associated with the shunt wound generator, and by an electronic regulating and control circuit associated with the rectifier. 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



POWER PLANT AT OVERSEAS TRANSMITTING STATION

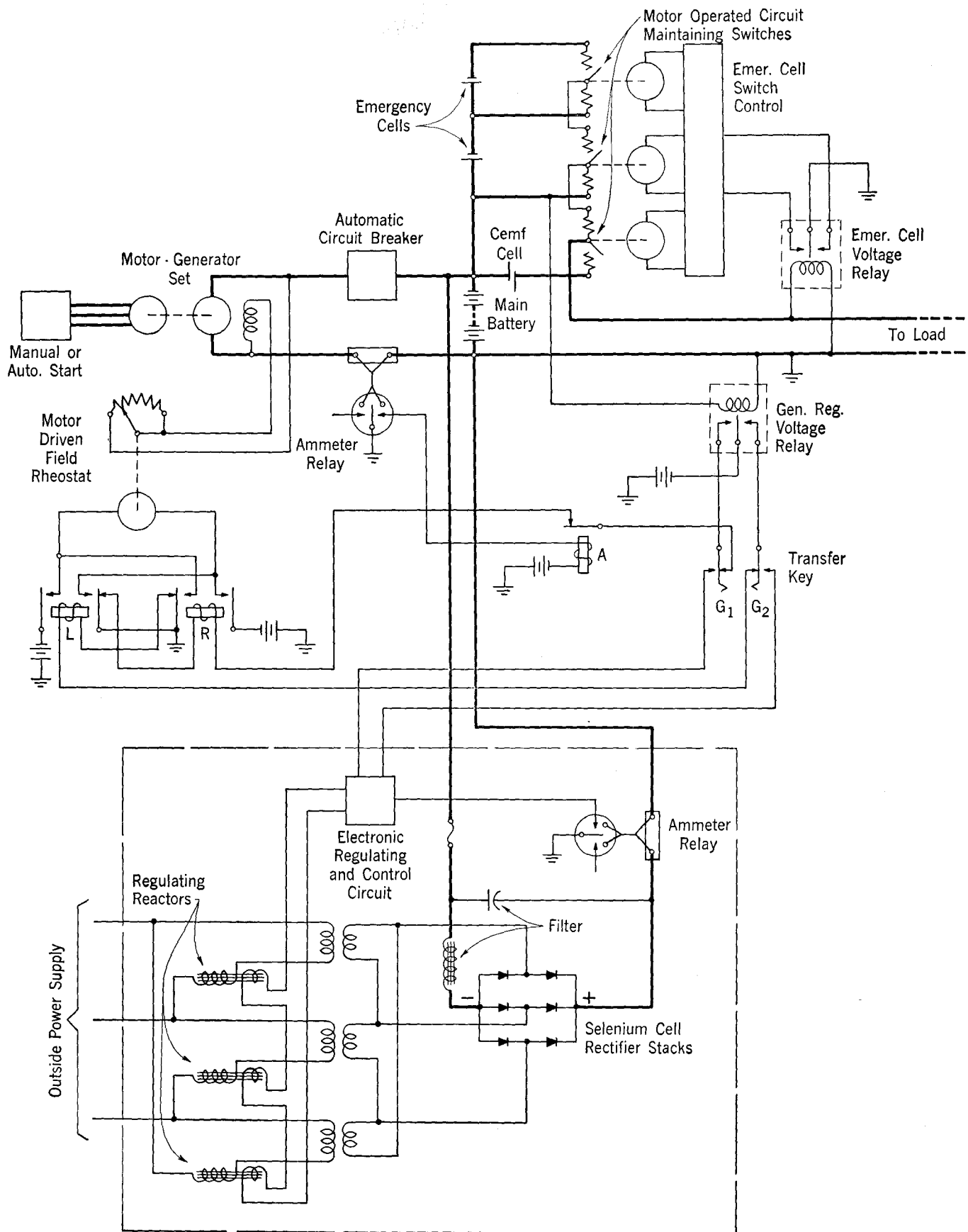
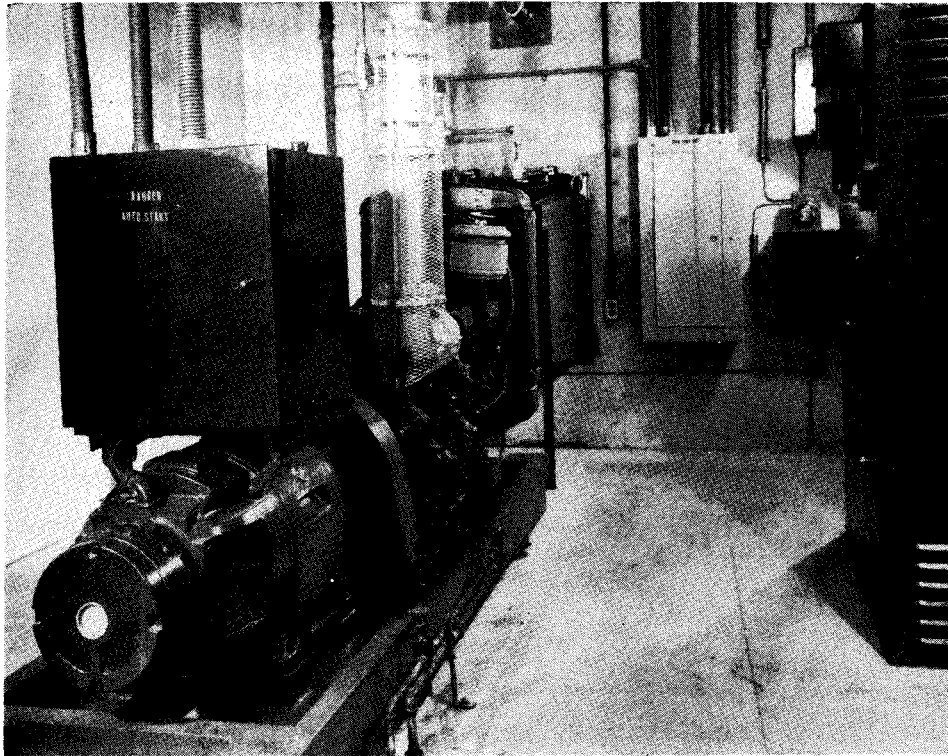


FIG. 6-14 TELEPHONE OFFICE POWER PLANT



AUXILIARY POWER PLANT AT RADIO RELAY STATION

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 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; or, in the case of the rectifier, causes the electronic control circuit to raise or lower the rectifier voltage as required. To avoid the possibility of overloading the charging unit, an ammeter relay is inserted in series with the line. When the unit is fully loaded, a contact on this relay closes causing relay *A* to operate and open the regulating voltage relay circuit. This prevents any further attempt on the part of the relay to increase the charging output.

The circuit includes two emergency cells which are connected to switches in such a way that one or both may be connected in series with the main battery. These cells are provided to take care of emergency conditions where the outside power supply fails and the charging units 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 cells are then automatically cut into the circuit by means of a voltage relay bridged

across the line, as shown.

Switches are provided for charging the emergency cells from the charging units in series with the main battery. However, since the emergency cells are not normally in use they are continuously supplied with a small trickle charge furnished by a copper-oxide rectifier (see Article 7.5) which normally maintains them in a fully charged condition. These switches and the copper-oxide rectifier are not shown in the drawing.

The main battery is, of course, kept in a continuously charged condition as long as the plant is operating normally.

When failure of the out-

side 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 charging unit above its normal value. But since the charging unit is connected directly to the load, an increase in its output voltage would also increase the load voltage. To avoid this, the circuit includes a **counter-emf cell** which is automatically inserted in series with the load circuit when the output voltage of the charging unit is increased above its normal value.

The cemf 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 substantially constant under wide variations of current. Physically, the cemf 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 7.3 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 6-14 shows one generator and one rectifier but additional charging units may be included

as required. To insure continuity of operation, a practical power plant always includes at least two charging units, and as many more may be added as are necessary to handle the maximum load. When the first unit becomes fully loaded the second is put into operation by throwing the transfer key shown on the drawing. Additional charging units when required are connected across the main leads to the battery but are manually controlled.

This power plant may be arranged so that the charging units 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 the needed units will be automatically started and connected to the line as the load increases and automatically disconnected and stopped as the load decreases.

## 6.6 Direct-Current Motors

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 6-8(B), 6-9 and 6-10, 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** where the left thumb represents direction of motion, the fore-finger 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 emf with a direction opposite to that of the applied emf is induced. This is called the **counter-electromotive force**, and the current in the armature is—

$$I = \frac{E_i - E_c}{R} \quad (6:1)$$

where  $E_i$  is the impressed emf,  $E_c$  is the counter emf, and  $R$  is the armature resistance.

Since there is low cemf 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 resistance 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".

## CHAPTER 7

### OTHER SOURCES OF DIRECT ELECTROMOTIVE FORCE

#### 7.1 Types of D-C Energy Sources

For an electric 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 emf; 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 capacitance relations.

If any device maintains an emf and sustains a current of electricity in a circuit, energy is supplied to the circuit. But the law of conservation of energy states that energy cannot be created or destroyed. Any source of emf 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 emf 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.

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 6. We shall at this time consider the various types of batteries and rectifiers, and the general battery requirements of telephone service.

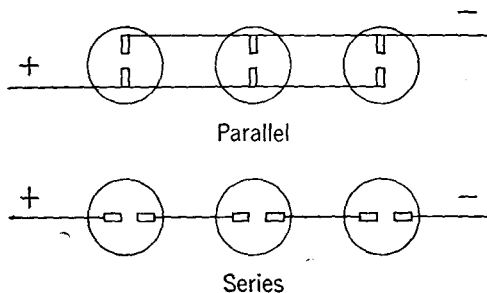


FIG. 7-1 CELLS IN SERIES AND PARALLEL

#### 7.2 Primary Batteries

Chemical batteries are divided into two classes, primary and secondary. A primary battery is one that generates an emf 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 emf 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.

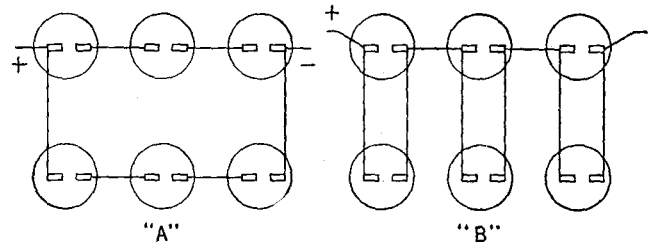


FIGURE 7-2

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 7-1, cells may be connected in parallel or in series, depending upon the value of the emf desired and the value of current to be sustained. If they are connected in series, the emf's are added, making the total emf of the battery the sum of the emf's of the individual cells. If they are connected in parallel, the emf of the battery is that of a single cell, but the current supplied to the circuit is divided between the several cells.

A battery may consist of groups of cells connected in parallel and these in turn connected in series, or vice versa. Figure 7-2 shows two methods of connecting six cells where the emf 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 are divided into two general groups called "wet" cells and "dry" cells. About the only wet cell used in the telephone plant is one 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, of course, not really dry but 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 7-3. The negative terminal consists of the zinc container and the positive terminal is a bar of carbon centered within it. The carbon is surrounded by a porous medium consisting of ground carbon and manganese dioxide, and this mixture is saturated with a solution of zinc chloride and ammonium chloride. 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 emf 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  $1\frac{1}{2}$  volts, which decreases with age, and has an internal resistance of .2 to .3

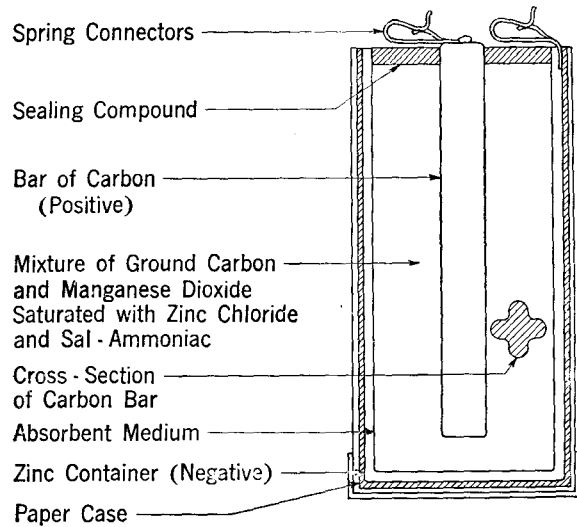


FIG. 7-3 BLUE BELL DRY CELL

ohms, 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,  $4\frac{1}{2}$ ,  $22\frac{1}{2}$  and 45 volts. The higher voltage blocks usually have intermediate taps giving various voltage values below the maximum.

### 7.3 Storage Batteries

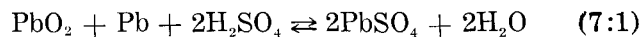
A chemical battery that is capable of storing electrical energy delivered to it from some other source, and delivering this energy to an electric

circuit at some later time, is called a storage battery. There are three principal types of storage batteries—the lead-acid, the Edison (iron-potassium hydroxide-nickel) and the nickel-cadmium. Because of its low internal resistance and more constant terminal voltage, the lead-acid type most nearly meets the exacting requirements for general telephone central office needs. (These requirements are discussed in Article 7.4 following.)

When the lead-acid cell is in a fully charged condition, the active constituents are a positive plate of lead peroxide ( $\text{PbO}_2$ ) and a negative plate of spongy lead ( $\text{Pb}$ ) in a dilute solution of sulphuric acid ( $\text{H}_2\text{SO}_4 + \text{H}_2\text{O}$ ). 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 ( $\text{SO}_4$ ) of the electrolyte, forming lead sulphate ( $\text{PbSO}_4$ ) and losing its negative charge; second, 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 ( $\text{SO}_4$ ) released by this action combines with hydrogen to form sulphuric acid. At the same time, hydrogen, traveling 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.

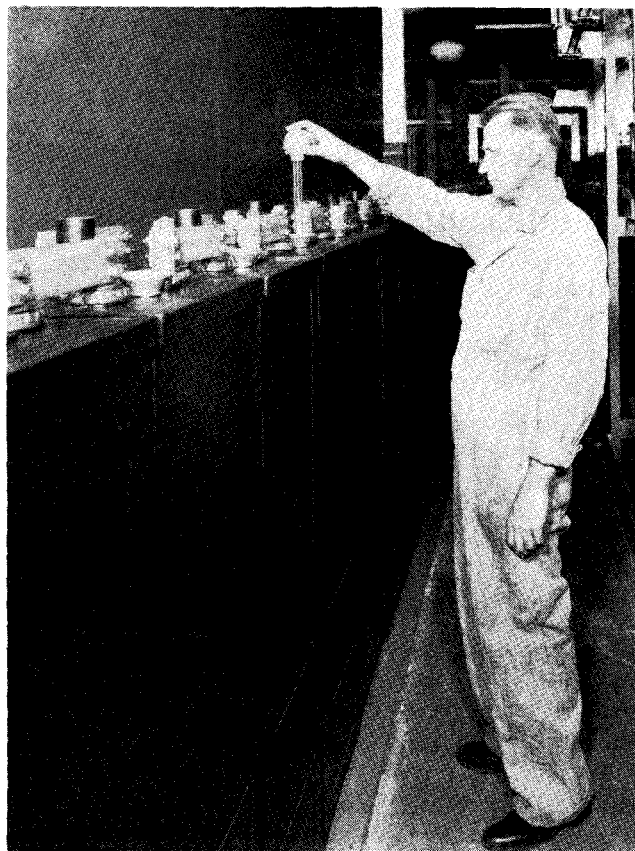
The following chemical equation may be used to explain the action of discharge when reading from left to right and the action of charge when reading from right to left:



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 sealed rubber containers. The containers are provided with vents to permit the release of gas and the plates are connected to terminals which project through the top. Larger cells are factory assembled in hard rubber tanks having sealed rubber covers as shown in the accompanying photograph.

In the practical operation of the lead-acid battery, we must be able to determine the state of charge or discharge at any time. It is not con-



ENCLOSED TANK TYPE STORAGE BATTERY

venient 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 electric charge held by the plates, resulting in a change in the emf 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.

The Edison storage battery has little application in telephone work. The more recent nickel-cadmium type of battery, however, is expected to become increasingly useful for such particular purposes as engine starting. It has discharge characteristics similar to the lead-acid battery but is not affected by short circuits and may be left in any state of discharge without detriment to the plates because the electrolyte (potassium hydroxide and distilled water) does not enter into the plate reactions but serves virtually only as a conducting medium. At high rates of discharge, as in engine starting, the voltage drops off less than for an equal capacity lead-acid battery. In consequence nickel-cadmium batteries for this service can be of somewhat less capacity.

#### 7.4 Power Plant Requirements in Telephone Offices

The telephone central office power plant must be not only reliable at all times but must meet other exacting requirements. Modern practice has led to the standardization of a common source of emf for the majority of the talking circuits, as well as for the operation of telephone and telegraph relays 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

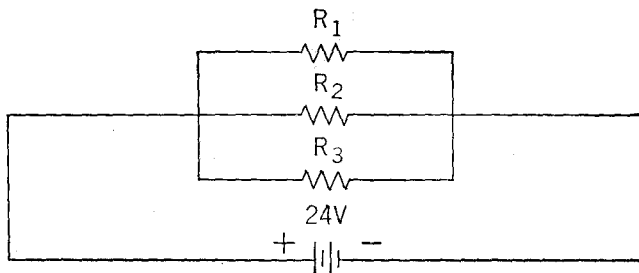


FIGURE 7-4

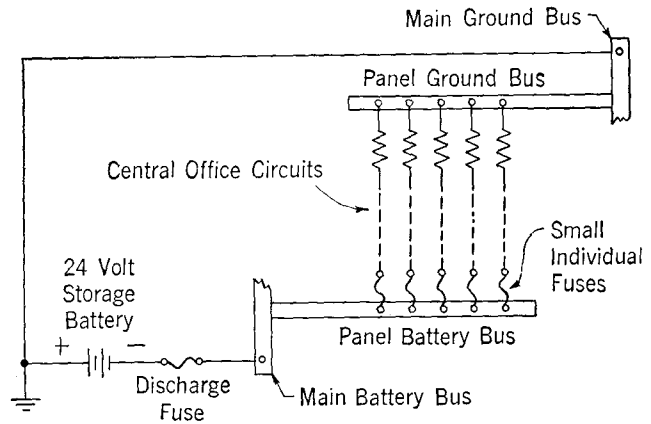


FIG. 7-5 STORAGE BATTERY LOAD CONNECTIONS

transmitter supply on long distance connections, 130-volt supply, both positive and negative, for telegraph repeater operation, and other voltages for telephone repeater and other electron tube 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**.

In our study of simple electric circuits, we have considered a single source of emf for each individual circuit. But we have learned that any number of resistances may be connected in parallel, as shown by Figure 7-4, 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 emf impressed on any single branch is the emf of the source and, theoretically, is independent of current flowing through other branches. This assumes, however, that the battery is a perfect source of emf without internal resistance.

Figure 7-5 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 7-4, excepting that as indicated in Figure 7-5, 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 capacitances 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 7-4, in which the current in any one resistance branch was seen to be independent of that in any other (provided the source of emf 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 7-6. Due to the resistance  $R_0$ , the current in one branch is no longer independent of that in other branches. Let us assign values as follows:

- $R_0 = 2$  ohms
- $R_1 = 5$  ohms
- $R_2 = 4$  ohms
- $R_3 = 3$  ohms
- $E = 24$  volts

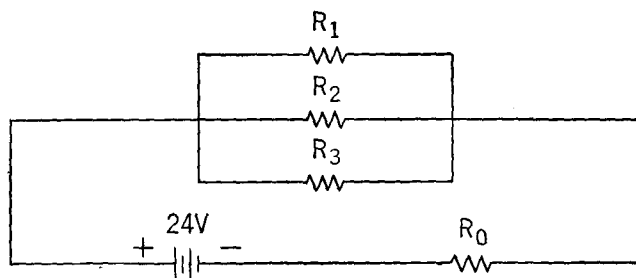


FIGURE 7-6

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 7-5, **unless the central office source of emf 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 emf has negligible internal resistance.

## 7.5 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. 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 as the older mercury arc tube, the newer mercury-vapor tube, and the Tungar tubes, as well as nearly all other types of electron tubes when properly connected. There are also the metallic or disc types of rectifiers of which the two most commonly used are the copper-oxide rectifier and the selenium rectifier.

The Tungar and mercury-vapor tube rectifiers

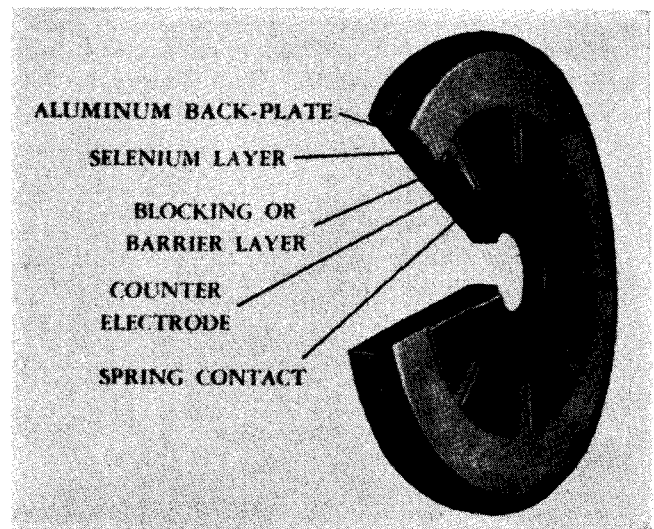
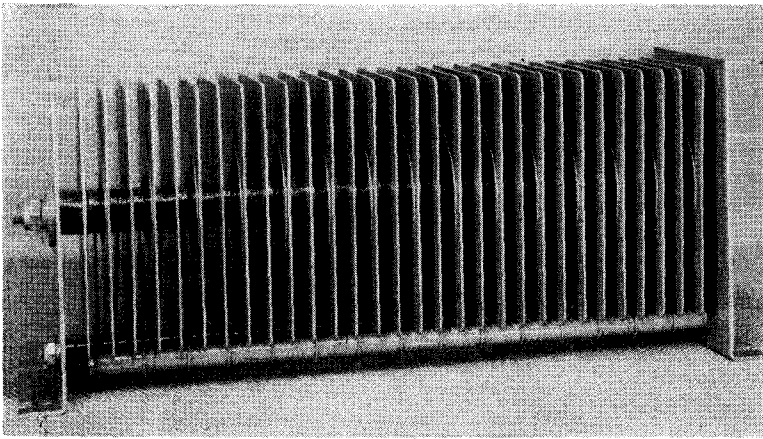


FIG. 7-7 SELENIUM RECTIFIER DISC



SELENIUM RECTIFIER STACK

depend for their operation upon the emission of electrons from a heated cathode. The basic phenomena involved are the same as characterize all types of electron tubes and these are discussed more fully in a later Chapter. Mercury vapor tubes are used quite extensively in telephone plants for charging small storage batteries and similar purposes.

Disc type rectifiers are now widely used in the telephone plant where they serve an increasing variety of purposes. These 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 lower voltage plants (12, 24 and 48-volt) they may be used in place of motor-generator sets in the 100-ampere and 200-ampere sizes.

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.

Similarly the selenium rectifier or cell consists of a back plate or electrode, usually of steel or

aluminum. One side of this plate is coated with a very thin layer of metallic selenium which in turn is in direct contact with a front or counter electrode of conducting metal. Rectification occurs in the so-called "barrier layer" where the selenium is in contact with the front electrode. Current flows readily from the back electrode to the front electrode but only very little current will flow in the reverse direction. A single selenium element is shown in Figure 7-7.

Disc rectifier elements may be assembled in various arrangements to provide the total rectifying capacity required as shown in the accompanying photograph. Selenium rectifier elements are assembled on an insulated center bolt or stud, with spacing washers and contact springs interposed between the cells. The separation allows for the free passage of air for cooling the assembly.

For use in telephone work, selenium rectifier elements are made in sizes ranging from  $9/32$ " to  $4\frac{3}{8}$ " diameter and in rectangular sizes up to  $5 \times 6$  inches. They are rated at maximum a-c input voltages up to 33 volts per element and d-c output current ratings ranging from .010 ampere for the smallest element to 12.0 ampere for the largest element for single phase full wave rectifiers. Contact terminals may be brought out from the elements in various arrangements for series or parallel connections of elements as required.

For power supply purposes in telephone plant, selenium rectifiers have largely superseded the copper-oxide types because of their lower cost and smaller space requirements. As we shall see in later Chapters, however, the copper-oxide type of rectifier has extensive applications in carrier and other circuits where the current values involved are relatively very low.

## CHAPTER 8

### INDUCTANCE AND CAPACITANCE

#### 8.1 Classification of Electric Currents

Thus far we have confined our attention largely to circuits of relatively simple characteristics. We have had a source of direct emf 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 emf generated by a closed loop revolving in a magnetic field; but we have not attempted to analyze the behavior of such an emf 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 electric circuits and their effect on the current set up in them by impressed emf's.

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

- a. The current that results from a constant direct source of emf 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 emf is alternating, having the simplest, most common and most convenient wave form, viz., the sine wave.
- d. Current where the source of emf is alternating, with 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 carry this classification one step further

by noting that *a*, *c*, and *d* are **steady state** currents, while *b* and *e* are 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 important considerations having to do with changes in current values.

In fact, we deal in the telephone plant with all five of the circuit conditions mentioned above. 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 *c* 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.

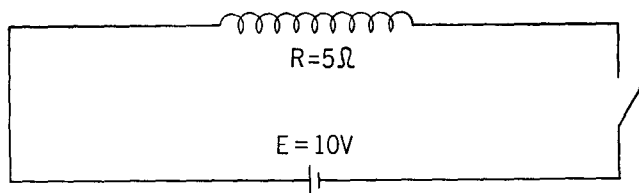


FIG. 8-1 SIMPLE INDUCTIVE CIRCUIT

## 8.2 Changes in Direct-Current Values

Let us 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 8-1, 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 2-2, 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.

Again, if in Figure 8-1 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 electric 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 "electric 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. It may be said that an electric circuit "reacts" to such current changes. But this reaction cannot be explained by our previous understanding of either resistance or emf. 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 "capacitance". **Inductance tends to give the circuit something that is analogous to inertia in a mechanical device, and capacitance something analogous to elasticity.**

## 8.3 Inductance

When an emf 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 emf. 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 electric circuit, for any given emf, 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 electric circuit is that the emf 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-emf generated by a motor opposes the applied voltage, and for an instant a part of the emf 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 emf is applied to overcoming the resistance of the

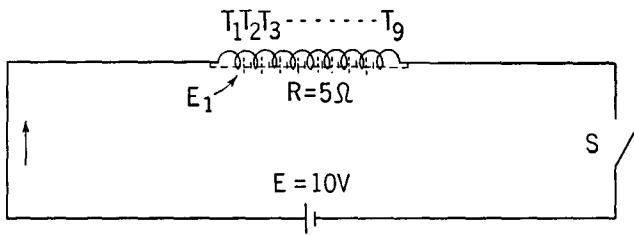


FIGURE 8-2

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 8-2. Immediately after the switch has been closed, the emf  $E$  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 emf tending to establish a current in the opposite direction. 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 emf.

Figure 8-3 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 emf 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 emf of 10 volts, because if it did so, the induced emf would be equal to the applied emf 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 emf which we might represent in Figure 8-2 as a second battery  $E_1$ . The voltage of  $E_1$  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 emf. 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 emf 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 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

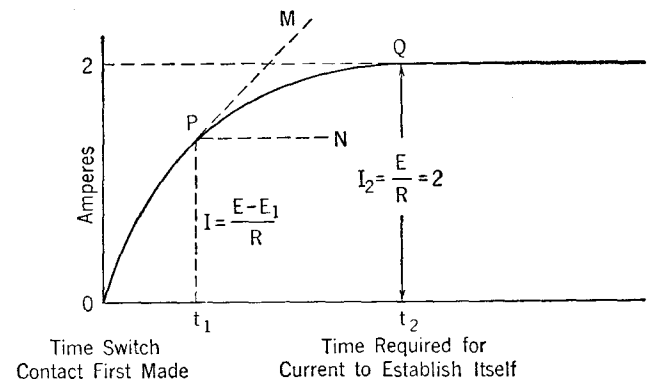


FIG. 8-3 BUILD-UP OF CURRENT IN INDUCTIVE CIRCUIT



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 emf 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 8-3 the "choking" effect of an inductively wound coil to increase in current value.

The case of a decreasing current value, and the emf 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 emf is induced as a result of a decreasing current instead of an increasing one, it aids rather than opposes the existing emf. 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 emf may become much greater than the applied emf besides being additive to it, whereas in the closed circuit it can never be greater than the applied emf. This total emf 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, where as 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 emf opposing that applied to the circuit in the case of an increasing current, and aiding the applied emf in the case of a decreasing current. The value of this induced emf 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 decreasing, eventually becomes zero. The magnitude or influence of the induced emf 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 foregoing property of the circuit as a tool or facility for creating the induced emf.

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

$$E_1 = \frac{LI}{t} \quad (8:1)$$

where  $E_1$  is the symbol for induced emf 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 6 we learned that an emf induced in a conductor cutting through a magnetic field was proportional to the number of lines of magnetic force cut per second. Then we were considering a moving conductor and a stationary field. Now we are considering the stationary conductors of a coil and a moving field in which the magnetic lines threading through the coil will cut each of its turns as they build up or decrease. The voltage induced remains proportional to the number of lines cut per second, or we may write—

$$E_1 = k_1 \frac{\phi N}{t} \quad (8:2)$$

where  $\phi$  is the total flux through the coil,  $N$  is the number of turns, and  $k_1$  is a constant depending on the units employed. Then from equations (8:1) and (8:2) we have

$$\frac{LI}{t} = k_1 \frac{\phi N}{t}$$

or

$$LI = k_1 \phi N \quad (8:3)$$

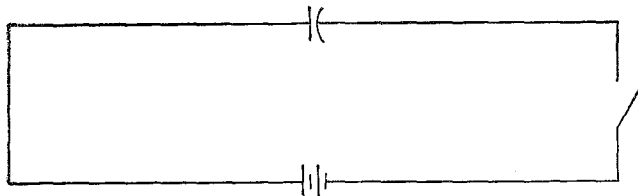


FIG. 8-4 SIMPLE CAPACITIVE CIRCUIT

Now referring to Chapter 4 we find the following two equations—

$$H = k \frac{NI}{l} \quad (4:3)$$

and

$$\frac{\phi}{A} = H\mu \quad (4:5)$$

Substituting  $H$  from equation (4:3) in equation (4:5) we get

$$\phi = k\mu A \times \frac{NI}{l}$$

and substituting this in equation (8:3) we have—

$$LI = k k_1 \frac{\mu AN^2 I}{l}$$

or

$$L = k k_1 \frac{N^2 \mu A}{l} \quad (8:4)$$

Here, if  $L$  is in henrys,  $N$  is the number of turns of the coil,  $\mu$  is the permeability (in the cgs electromagnetic system of units),  $l$  is the length of the coil core in centimeters and  $A$  is the area of the core in square centimeters, the combined value of the constants  $k, k_1$  will be  $1.26 \times 10^{-8}$ , or we may write—

$$L = \frac{1.26 N^2 \mu A}{100,000,000 \times l} \quad (8:5)$$

**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.

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 equations as are parallel resistances. For example, see equation (3:1) 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 emf

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.

## 8.4 Capacitance

There remains that property of the circuit that we have called **capacitance**, which gives it something analogous to elasticity. While a storage battery stores electricity as another form of energy, in a smaller way an electric “condenser” or **capacitor** stores electricity in its natural state.

As a container, a capacitor 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 capacitor is connected to a direct source of emf through a switch as shown in Figure 8-4, and the switch is suddenly closed, there will be a rush of current in the circuit. This will charge the capacitor to a potential equal to that of the battery, but the current will decrease rapidly and become zero when the capacitor is fully charged.

The insulated conductors of every circuit have to a greater or less degree this property of capacitance. 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 capacitance 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 capacitor depends upon the capacitance and the electromo-

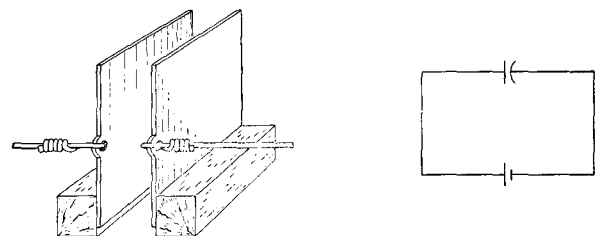


FIG. 8-5 ELEMENTARY CAPACITOR AND CONVENTION

tive force impressed across its terminals. The following equation expresses the exact relation:

$$Q = EC \quad (8:6)$$

where  $Q$  is the quantity of electricity in coulombs,  $E$  is the impressed emf in volts, and  $C$  is the capacitance in **farads**. The farad is a very large unit and is seldom used in practice. The microfarad (from "micro", meaning one-millionth) is the practical unit more commonly used; and with  $C$  expressed in these units, equation (8:6) becomes—

$$Q = \frac{EC}{1,000,000} \quad (8:7)$$

Figure 8-5 illustrates a capacitor in its simplest form together with the convention used for a capacitor connected to a battery. Two wires are connected to two parallel metal plates having a definite separation as shown. This is called an **air capacitor** because air is the **dielectric** medium between the plates. The capacitance of such a capacitor 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 capacitance of the capacitor were increased, the kick would become more noticeable. If now the battery were disconnected and the capacitor 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 capacitor, establishing an instantaneous current in the opposite direction and discharging the capacitor through the winding of the galvanometer.

In addition to the size of its plates and their separation, the capacitance of a capacitor depends upon the insulating medium between the plates. For example, if mica is inserted between the plates of an air capacitor its capacitance is increased about five times. Insulators, in addition to being classified in the order of their insulating properties, may be classified in the order of their "dielectric powers," or "specific inductive capacities," i.e., their ability to increase the ca-

pacitance over that of an air capacitor. Such a classification is given in brief in Table IV.

The equation for the capacitance value of a two-plate capacitor is—

$$C = K \frac{A}{d} \quad (8:8)$$

where  $C$  is capacitance in microfarads,  $K$  is the constant taken from Table IV,  $A$  is inside area of one plate in square centimeters, and  $d$  is separation of the plates in centimeters. There are similar equations for calculating the capacitance per unit length of parallel open wire conductors or cable conductors. These may be found in various 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 IV

DIELECTRIC POWER OF VARIOUS INSULATING MATERIALS

Values are only approximate and are given for value of  $K$  in equation (8:8) rather than compared to air as unity.

SUBSTANCE	$K$ IN EQUATION (8:8)
Air	.09 × 10 <sup>-6</sup>
Alsimag A196	.51 × 10 <sup>-6</sup>
Bakelite (paper-base)	.34 to .50 × 10 <sup>-6</sup>
Bakelite (mica-filled)	.49 × 10 <sup>-6</sup>
Celluloid	.36 to 1.4 × 10 <sup>-6</sup>
Cellulose Acetate	.54 to .72 × 10 <sup>-6</sup>
Fiber	.45 to .64 × 10 <sup>-6</sup>
Formica	.41 to .44 × 10 <sup>-6</sup>
Glass (window)	.68 to .72 × 10 <sup>-6</sup>
Glass (photographic)	.68 × 10 <sup>-6</sup>
Glass (Pyrex)	.38 to .44 × 10 <sup>-6</sup>
Lucite	.23 to .27 × 10 <sup>-6</sup>
Mica	.23 to .72 × 10 <sup>-6</sup>
Mica (Clear India)	.58 to .68 × 10 <sup>-6</sup>
Mycalex	.67 × 10 <sup>-6</sup>
Paper	.18 to .23 × 10 <sup>-6</sup>
Polyethylene	.21 to .22 × 10 <sup>-6</sup>
Polystyrene	.22 to .26 × 10 <sup>-6</sup>
Porcelain	.56 to .68 × 10 <sup>-6</sup>
Rubber (hard)	.18 to .32 × 10 <sup>-6</sup>
Steatite (low-loss)	.40 × 10 <sup>-6</sup>
Wood (dry Oak)	.23 to .61 × 10 <sup>-6</sup>

An inspection of equation (8:8) will show that if two identical capacitors are connected in parallel as shown by Figure 8-6, the effect is that of doubling the plate area of a single capacitor, and

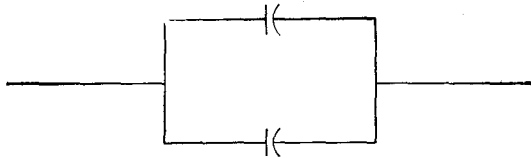


FIGURE 8-6

therefore doubling the capacitance. On the other hand, if two identical capacitors are connected in series as shown by Figure 8-7, the middle or common plates have a neutral potential and the effect is that of doubling the thickness of the dielectric of a single capacitor, which cuts the capacitance in half. It follows that capacitors in parallel and series act inversely to resistors in parallel and series. This may be stated in a single rule covering all conditions—

**Capacitors in parallel should be added to find the total capacitance in the same way that resistors 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 capacitance of capacitors in series in the same way that the reciprocal of the sum of the reciprocals must be taken to find the total resistance of resistors in parallel.**

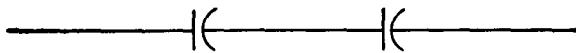


FIGURE 8-7

This rule may be expressed by two simple equations:

For several parallel capacitors—

$$C = C_1 + C_2 + C_3 \text{ etc.} \quad (8:9)$$

For several series capacitors—

$$\frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} \text{ etc.} \quad (8:10)$$

Or, for the simple case of only two series capacitors, equation (8:10) may be expressed as follows:

$$C = \frac{C_1 \times C_2}{C_1 + C_2} \quad (8:11)$$

**Note:** Equation (8:9) may be compared with equation (3:1) and equation (8:11) may be compared with equation (3:4).

### 8.5 Effects of Inductance and Capacitance in Direct-Current Circuits

The circuit reactions resulting from the presence of inductance and capacitance 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 8-8 shows one way to apply the property of capacitance 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 capacitor which prevents excessive arcing when the circuit is opened because the sustained current is charging the capacitor instead of forcing an arc. In practice the capacitor usually has a non-inductive resistance in series, its purpose being to avoid oscillatory effects which are discussed in a later Chapter.

It was noted in Figure 8-3 that when a battery is connected across an inductance, the current builds up slowly to the maximum value determined by the resistance of the circuit. Conversely, when a battery is connected across a capacitor, there is a sudden rush of current into the capacitor, which decreases toward zero as the capacitor charges up to the value of the impressed voltage. This is illustrated in Figure 8-9 by the dashed line indicating the change in current flow with time. In the circuit shown in Figure 8-9, commonly known as an **R-C circuit**, the initial value of current flow is determined by resistance  $R$ . At the instant the circuit is closed, the total voltage drop will be across  $R$  with zero drop across  $C$ . As the current flows into the capacitor, the voltage across  $C$  will increase and, since the current flow is decreasing, the drop across  $R$  will decrease. When the capacitor is fully charged and current has fallen to a negligible value, the total voltage of the circuit will be across the capacitor. At all times, of course, the sum of the voltage drops across  $R$  and  $C$  must be equal and opposite to the impressed emf.

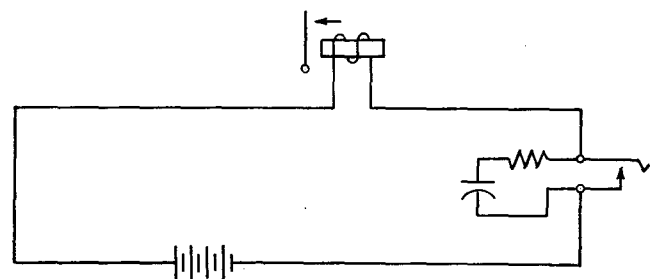


FIG. 8-8 CAPACITOR AS SPARK-KILLER

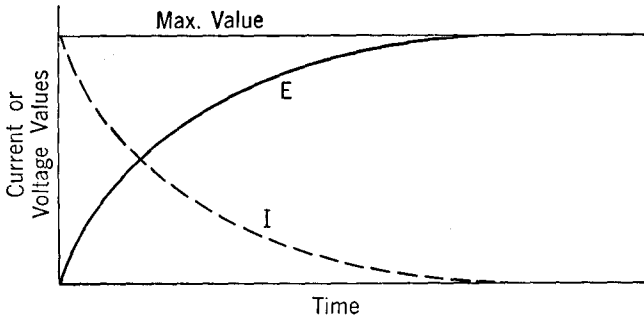
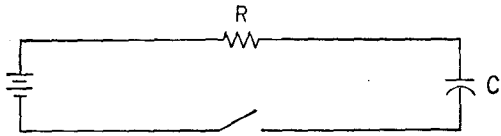


FIGURE 8-9

It is obvious that the larger the resistance  $R$  in series with the capacitor, the slower will be the rate at which the capacitor will build up to full voltage. Also, the larger the capacitance of the capacitor, the longer will be the time involved.

The time required to charge a capacitor to 63 percent, or discharge it to 37 percent, of its final voltage is known as the **time constant** of the circuit. The value of the time constant in seconds is equal to the product of the circuit resistance in ohms and the capacitance in farads, or

$$\text{Time constant} = RC$$

**Note:** The curve of charge or discharge may be exactly expressed mathematically by the exponential equation—

$$E_i = E_{max} \left( 1 - \epsilon^{-\frac{t}{RC}} \right)$$

where  $\epsilon$  is the base of Napierian or natural logarithms, 2.72 . . .

When  $t$  is equal to  $RC$ , the term in the parenthesis reduces to

$$1 - \epsilon^{-1} = 1 - \frac{1}{\epsilon} = 1 - \frac{1}{2.72} = .63$$

For practical working purposes, it is generally considered that the voltage reaches its final value (actually 99+ percent) in a time equal to  $5RC$ .

The time constant is often used in electric circuits to time certain operations. A device that can be “triggered off” at specific voltage values such as certain types of electron tubes, for example, may be connected across a capacitor. The constants of the circuit may then be adjusted so that the voltage across the capacitor reaches a specified value after a required interval of time, thereby setting off the trigger device.

The significance of the time constant of a resistance-inductance, or  $R-L$  circuit is the same as that of an  $R-C$  circuit but is determined by the ratio  $L/R$  or

$$\text{Time constant} = \frac{L}{R}$$

## CHAPTER 9

### TELEPHONE PRINCIPLES AND BASIC APPARATUS

#### 9.1 Sound

The telephone accomplishes the electrical transmission of speech by employing the mechanical energy of the speaker's voice to produce electric energy having similar characteristics, and in turn converting this electric 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**.

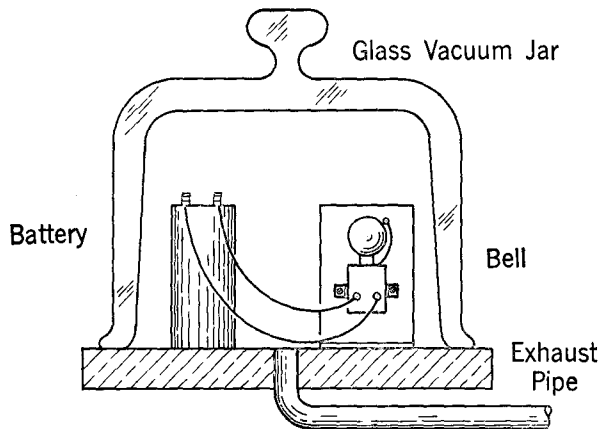


FIGURE 9-1

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.

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 emanate 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 radio 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 electromagnetic wave transmission, sound is an atmospheric disturbance. If, as shown in Figure 9-1, a vibrating bell is placed under an inverted glass bowl resting on a plate that has an outlet 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 a physical medium, usually atmospheric, 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 a tuning fork, and is sustained 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 6.2.) On the other hand, if the

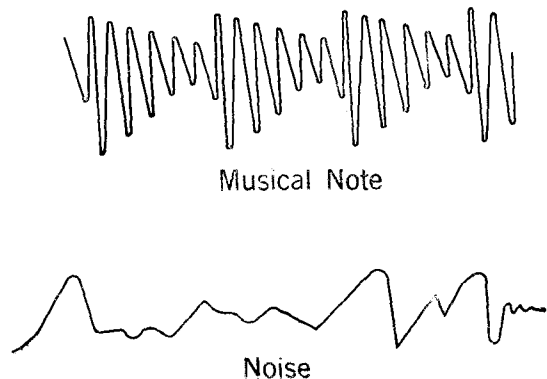
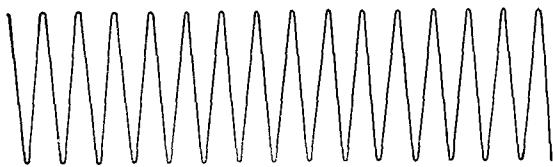
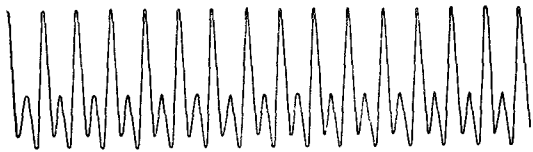


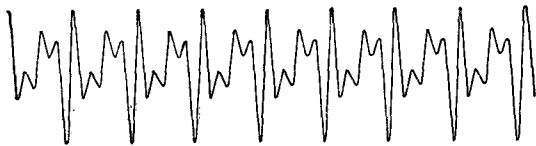
FIGURE 9-2



Simple Sound



oo as in Loose



o as in Low

FIGURE 9-3

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, the latter giving it "quality". A sine wave without harmonics is called a pure 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 normal ear lies somewhere in the octave between 16 and 32 vibrations per second. On the other hand, the normal 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 conveniently defined as a disturbance in the atmosphere whereby a form of wave motion is propagated from some source at a velocity of about 1,075 feet per second, the transmission being accomplished by alternate condensations and rarefactions of the atmosphere in

cycles having a fundamental frequency ranging somewhere between 16 per second and 32,000 per second.

The waves superposed 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 9-2 illustrates wave forms for different kinds of sound and, similarly, Figure 9-3 shows the predominating wave shapes of certain spoken vowels.

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 the most 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 3,500 cycles per second.

## 9.2 The Simple Telephone Circuit

The original telephone, as invented by Bell in 1876, consisted of a ruggedly constructed telephone receiver, which at that time served as both transmitter and receiver. 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 9-4 shows such a circuit.

At the speaker's station, the sound waves of the voice strike the metal diaphragm of the telephone receiver, and the alternate condensations and rarefactions of the air on one 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 corresponding change in the number of lines that thread through



FIG. 9-4 ELEMENTARY TELEPHONE CIRCUIT

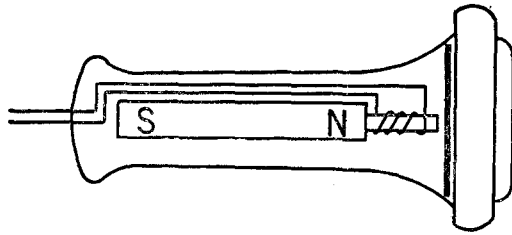


FIG. 9-5 BAR-MAGNET RECEIVER

the receiver winding, resulting in the turns of the winding being cut by these building up and collapsing lines. This establishes a varying electric voltage and current in the winding of the telephone receiver, having wave characteristics similar to the characteristics of the sound wave. This current passes over the connecting wires and through the receiver winding at the distant end. There it 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.

### 9.3 The Telephone Receiver

The earliest forms of telephone receivers were made with a permanent bar magnet as shown in Figure 9-5. The efficiency of the receiver was later greatly increased by the use of a horseshoe magnet as shown in Figure 9-6. This permits the lines of magnetic force to pass in a much shorter path from one magnetic pole to the other through the iron diaphragm. The principle of operation of receivers currently in use in the telephone plant does not differ fundamentally from that of the early types although the receivers themselves are generally quite dissimilar in physical appearance.

Figure 9-7 is a cross-sectional drawing of a bipolar receiving unit which is currently in extensive use in the Bell System. This receiver employs

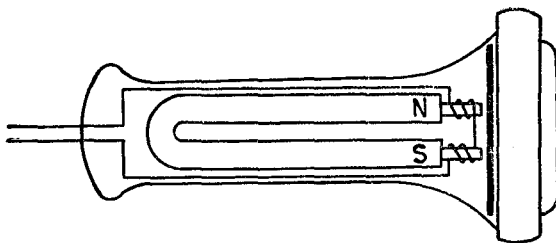


FIG. 9-6 HORSESHOE-MAGNET RECEIVER

in its construction no less than three of the comparatively new magnetic alloys that were mentioned in Article 4.1. 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 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 rayon disc. The latter consists of an air chamber which opens into

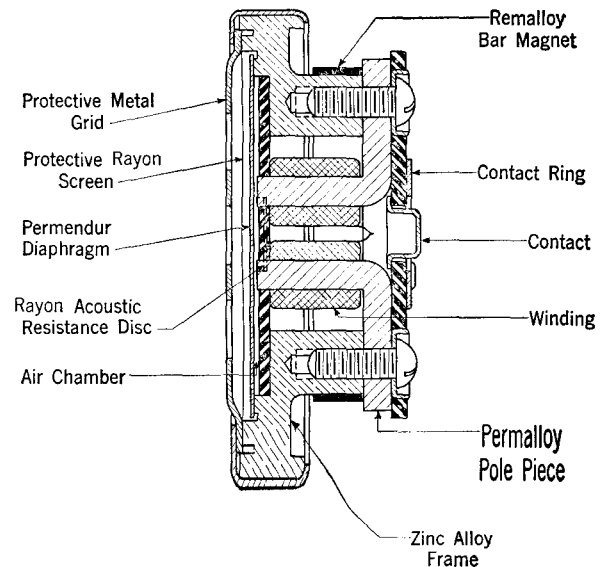
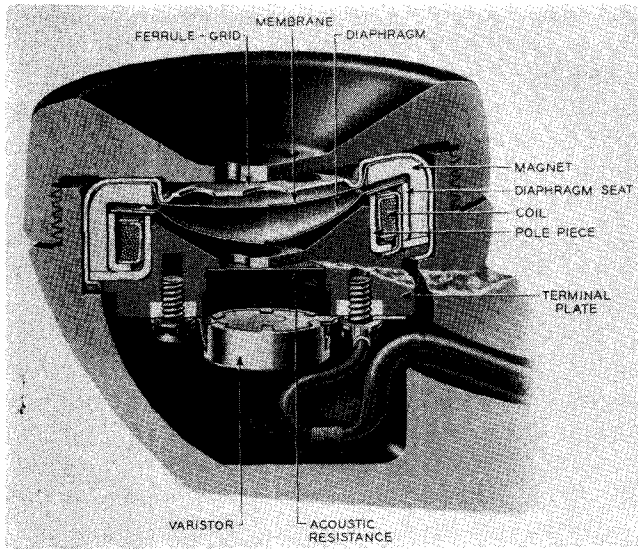


FIG. 9-7 CROSS-SECTION OF BIPOLAR RECEIVER UNIT

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 appreciably. The diaphragm 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 held to close limits. The two permalloy pole pieces are welded to a pair of very strong remalloy bar magnets, and the assembly is fastened to a zinc alloy frame. The whole unit is held together by an aluminum ferrule on the back of which are mounted two silver plated contacts for the electrical connections.

A more recent design of receiver, which is currently standard for new installations, is shown in cross-section in the accompanying photograph. This is known as a **ring-armature** receiver. Differ-





RING-ARMATURE RECEIVER

ing radically in detail design from the bipolar types, it is substantially more efficient chiefly because the dome-shaped diaphragm is much lighter than that of the bipolar receiver. The diaphragm dome is made of lightweight plastic-impregnated cloth cemented at its outer periphery to a flat ring-shaped armature of vanadium permendur. The outer edge of the armature ring rests on a diaphragm seat of non-magnetic material and its inner edge extends into the air-gap between a pole piece of 45% permalloy and the permanent magnet which is made of remalloy. The entire diaphragm structure is thus driven like a piston under the influence of the magnetic fields existing in the air-gap across the inner edge of the armature ring.

Since telephone receivers are equipped with permanent magnets, it is of course important that the magnetism should not be impaired by jarring or other abuse. The permanent magnet is important not only because it increases the amplitude of vibration of the diaphragm when the voice current is flowing through the windings, but also because it prevents the diaphragm vibrating at twice the voice frequency. This principle is illustrated in Figure 9-8. 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.

#### 9.4 The Telephone Transmitter

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 electric current, instead of generating electric energy by means of electromagnetic induction. By this means it was possible to establish an electric 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 and the vibration of the diaphragm acted as a means for regulating or modulating this energy supply, rather than as a generating device.

The principle of the transmitter is illustrated by Figure 9-9. Battery A establishes a direct current in a local circuit consisting of the primary winding of an induction coil  $I$ , and a cup of car-

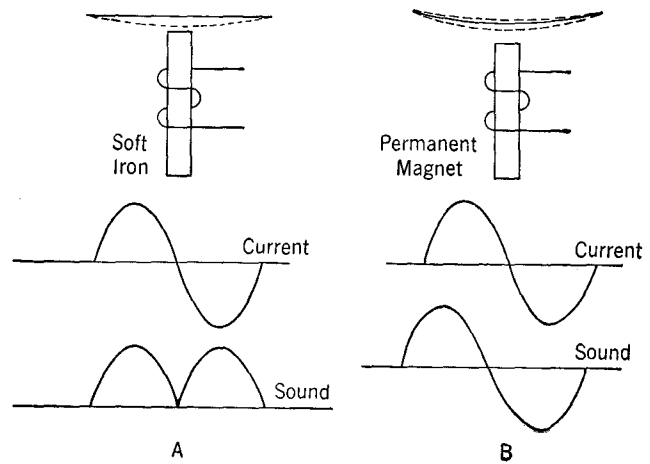


FIGURE 9-8

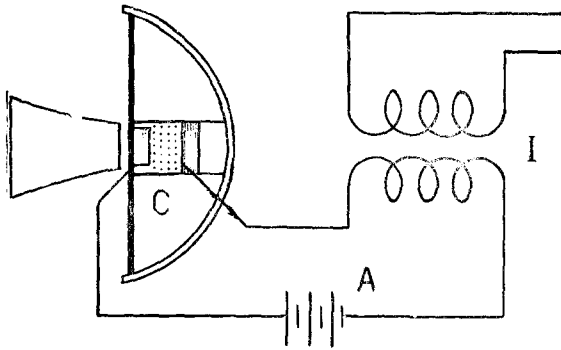


FIG. 9-9 PRINCIPLE OF THE TELEPHONE TRANSMITTER

bon granules *C*. One side of this cup rests against a small carbon disc rigidly connected to the transmitter diaphragm. The vibrating transmitter diaphragm varies the pressure on the carbon granules, which causes the resistance of the electric 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 9-4, establish an alternating emf 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 the same as has been explained in connection with Figure 9-4.

Figure 9-10 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 circuit, giving "side-tone" to the receiver at the home station. It is also transmitted to the distant station, operating the receiver at that point.

Figure 9-11 shows in cross-section a transmit-

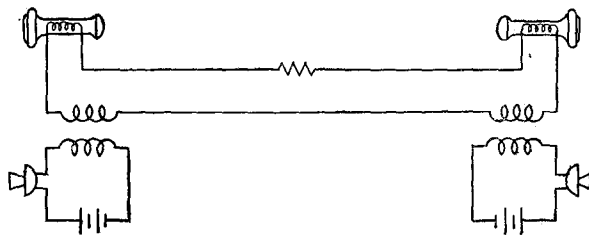


FIG. 9-10 TELEPHONE CIRCUIT WITH LOCAL BATTERY TRANSMITTERS

ter unit that is 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 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 with radial ridges to increase stiffness. Paper books, which consist of a number of thin impregnated paper rings, support the diaphragm at its edges on both sides. The carbon chamber is closed on the front side by a rayon membrane clamped

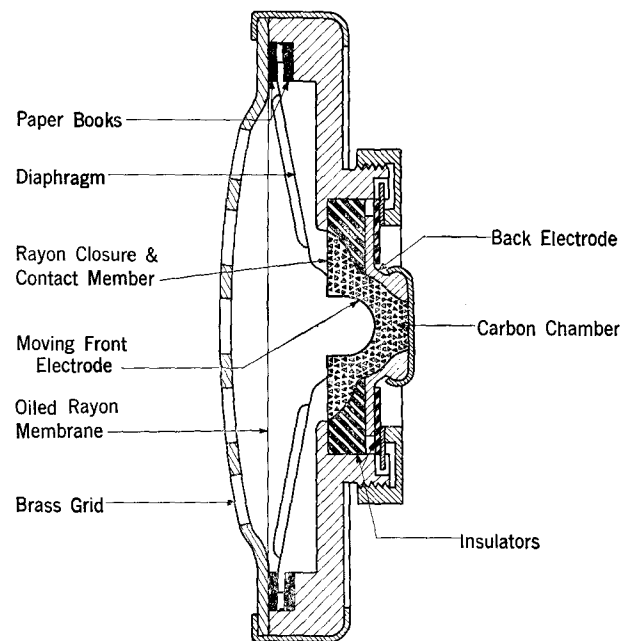


FIG. 9-11 CROSS-SECTION OF TRANSMITTER UNIT

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

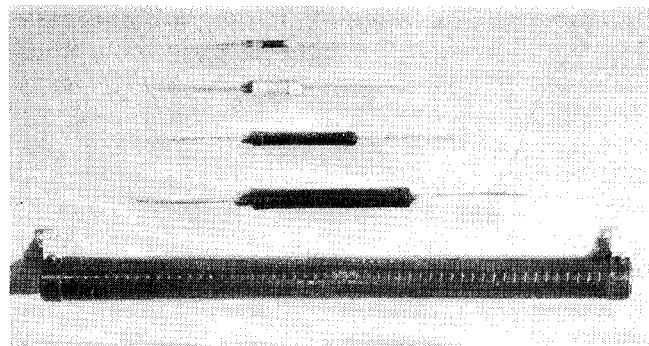
working parts by an oiled rayon moisture-resisting membrane placed between the brass plate and the diaphragm.

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.

## 9.5 Resistors, Inductors and Capacitors

A resistor is a piece of apparatus designed for the specific purpose of inserting resistance into an electric circuit. Resistors are therefore designed in practice to have a definite value of resistance in ohms. They must also be designed in many cases to be capable of dissipating specified amounts of power without heating beyond safe limits. The applications of resistors in telephone and telegraph work are almost innumerable, involving ohmic values ranging from fractions of one ohm to many megohms; and power ratings (i.e., ability to dissipate energy) ranging in general up to about 5 watts.

The great majority of the resistors used can be divided into two principal types—wire-wound and carbon. For relatively low resistances ranging from less than one ohm up to a few thousand ohms, most of the common resistors found in the telephone plant are the so-called flat types coded #18 and #19 and illustrated in Figure 9-12. As indicated, these consist of wire wound on a flat card of insulating material, the flat shape having the advantage of making the resistor relatively non-inductive. Where higher resistances— in the



DEPOSITED CARBON RESISTORS

order of ten thousand ohms and upward—are required, some type of carbon resistor is usually employed. Most common of these is the so-called **composition** resistor in which the resistive element is a combination of finely divided carbon or graphite mixed with a non-conducting filler such as talc, with synthetic resin as a binder. These resistors are well suited for application in relatively high frequency circuits such as radio where great precision in ohmic value ordinarily is not required. They are usually built to tolerances of about 5% and are commercially available in values from 10 ohms to 22 megohms, with power ratings up to 2 watts. Where very high resistances are needed, the **deposited-carbon** resistor may be used. In this type the resistor element is an extremely thin film of carbon deposited on the surface of a ceramic core by thermodecomposition. A helical groove is then cut through the carbon film to leave a ribbon of the carbon film wound around the core between the terminal electrodes at the ends.

The resistance of any ordinary electric conductor, such as the wire used in wire-wound resistors increases as the temperature of the conductor increases. There are, however, many so-called semi-conducting materials in which the opposite effect occurs—that is, the resistance decreases as the temperature increases. Such materials include the oxides of manganese, nickel, cobalt, iron, and zinc, or mixtures of such oxides. Resistors made up of these materials are known as **thermistors**. They have many important applications in telephone work, particularly in controlling amplifier gains to compensate automatically for changes in line loss resulting from temperature changes along the line.

An inductor is a piece of apparatus designed primarily to insert inductance into a circuit. We have already noted the use of certain types of inductors, commonly called retardation or choke coils, in power circuits. Inductors, as was pointed

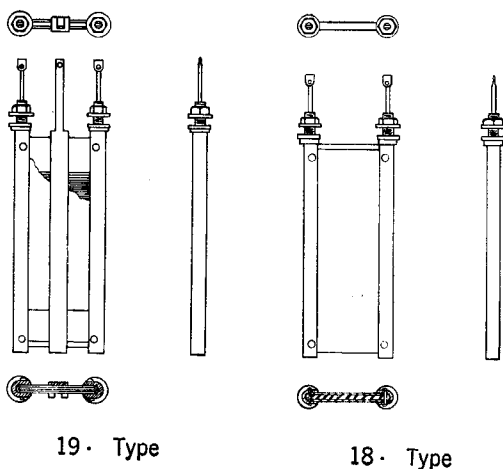


FIG. 9-12 FLAT RESISTORS

out in an earlier Chapter, consist fundamentally of coils of wire, the amount of their inductance depending primarily upon the number of turns in the coil, its size and shape, and the nature of the core material about which the coil is wound. Inductors are usually classified under two main types—those having air cores and those with cores consisting of iron or iron alloys. Both types have numerous uses in communication circuits. As would be expected from our knowledge of magnetism, air core inductors have relatively low inductance values but are comparatively free from the core losses due to magnetic hysteresis and “eddy currents” that are inevitable when metallic cores are used. The principal field of use of air core inductors is in high frequency or radio work. Where relatively large amounts of inductance are needed, as in power circuits and telephone and telegraph circuits at ordinary frequencies, iron core inductors are employed.

Since any inductor consists of many turns of wire, it will always have resistance as well as inductance. In inductor design, however, every effort is made to keep the resistive effect as low as possible in comparison to the inductive effect. The ratio of the inductive effect to the resistive effect is known as the **figure of merit** or “*Q*” of the inductor.

Capacitors (or condensers) were discussed briefly in Chapter 8. It was established then, that the capacitance value depends on the plate areas, and the nature and thickness of the dielectric between the plates. The simplest dielectric is air, and most of the variable capacitors are “air condensers”. Unless such capacitors are extremely large, however, their total capacitance is relatively low and their chief use is, accordingly, in high frequency work. For the many applications in communications work at lower frequencies, capacitors employing impregnated paper or mica as the dielectric material are most commonly used. The paper capacitor is made as a continuous tight roll of aluminum foil strips separated by thin Kraft paper impregnated with oil or wax. They are built in a wide range of capacitance values extending up to 5 microfarads and may be designed to withstand substantial voltages (up to 4,000 volts) without breakdown.

Mica capacitors are used in a smaller range of capacitance values. They will not withstand as high a voltage as paper capacitors but have a higher degree of stability with respect to temperature change and time than do paper capacitors. Their application, accordingly, is to situa-

tions where more precise and constant capacitance values are required.

Within reasonable limits of cost and size, neither the paper nor mica capacitor can be built to have a capacitance value exceeding a very few microfarads. Where larger values of capacitance are needed, an entirely different type known as the electrolytic capacitor is frequently used. Essentially such a capacitor consists of plates of aluminum or tantalum, on which extremely thin oxide films have been electrochemically produced, immersed in a conducting aqueous electrolyte. Here the aluminum or tantalum plate itself is one cathode of the capacitor, the conducting electrolyte is the other cathode, and the oxide film on the plate is the dielectric. Because of the extreme thinness of this film, which is of course itself an insulator, capacitors of this type may have a very high capacitance even though quite limited in total size. For certain applications in power work, such as in voltage stabilizing filters, electrolytic capacitors are built in cells similar in appearance to a storage battery. This is the so-called “wet” type of electrolytic capacitor. There is also a “dry” type, which uses a viscous or paste electrolyte, and is usually built in small cylindrical forms not dissimilar in external appearance to the small paper-type capacitors commonly seen in electronic apparatus. Electrolytic capacitors for general use, however, are neither as stable nor long lasting as paper or mica capacitors.

## 9.6 Relays and Switches

The fundamental instruments of telephony are the telephone transmitters and receivers that are described in some detail in a preceding Article. These instruments, when associated in appropriate circuits with other basic apparatus units such as coils, capacitors, resistors, etc. and connected

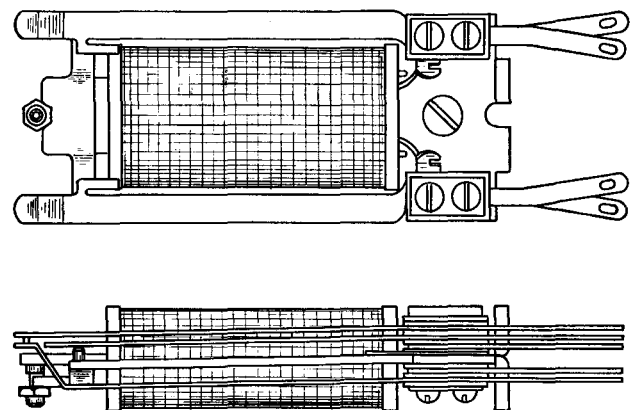
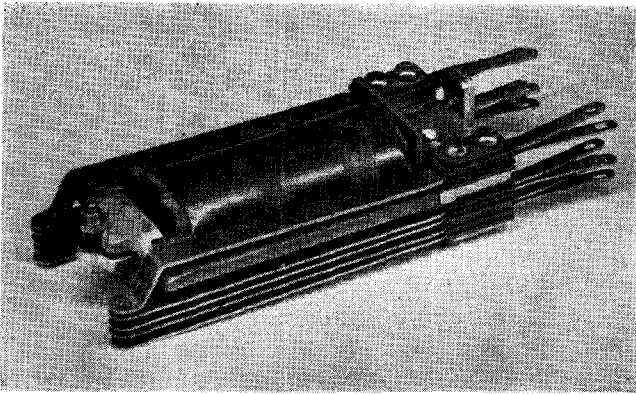


FIG. 9-13 E-TYPE RELAY



E-TYPE RELAY

together by means of appropriate transmission lines, make it possible for persons at different locations to talk with each other.

The problem of telephone service, however, is not just to make it possible for some one person to talk with another person at a distance, but for any subscriber to such a service to be able to talk with any other subscriber. This means that any practical working telephone plant must include vast quantities of **switching apparatus** designed to connect, disconnect and rearrange a great variety of circuits quickly and surely, as may be necessary to meet the ever-changing communication requirements of the subscribers.

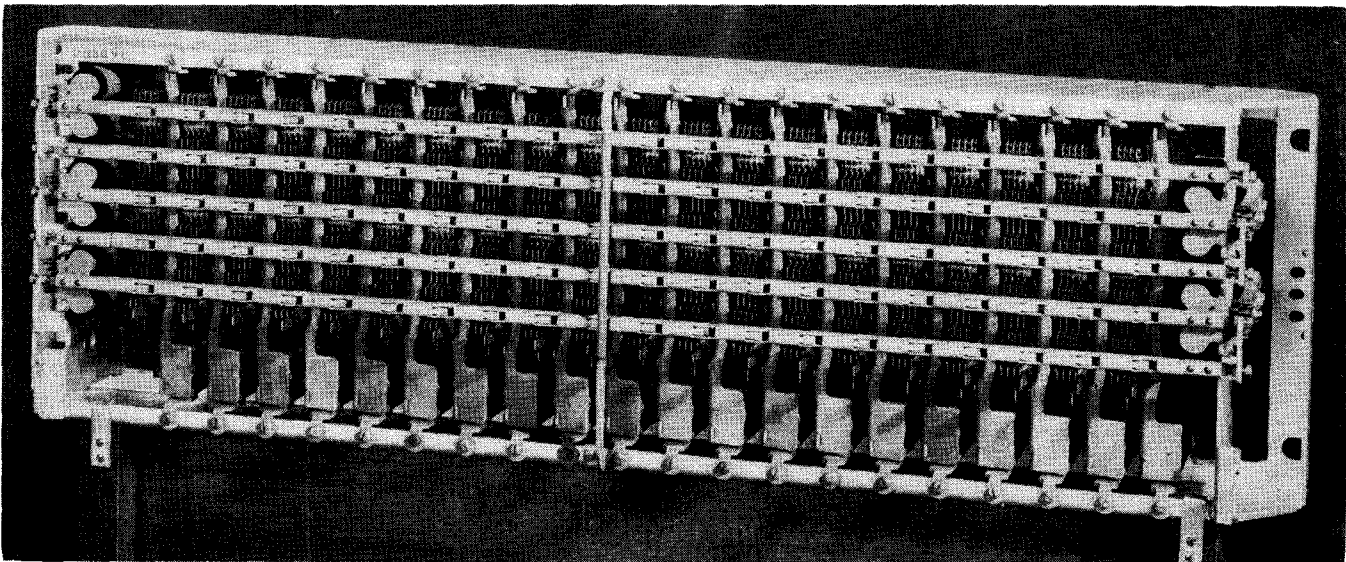
Switching apparatus may be classified first into two fundamental categories—manually operated and electrically operated. Devices in the first category include jacks, plugs, cords, keys, push buttons, etc. Such devices are probably too well known to require any extended discussion here,

although the reader who is familiar with them will also recognize that they can be, and are, built in many different designs to serve particular purposes.

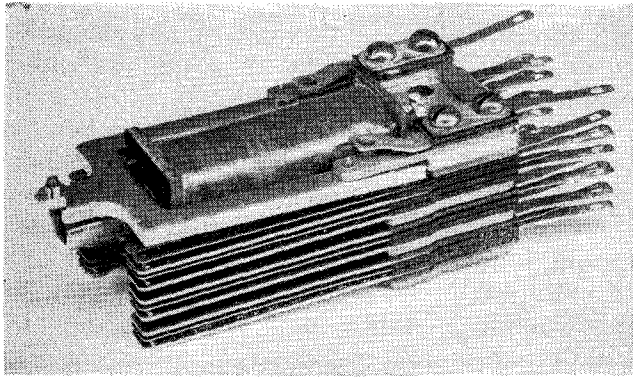
The most commonly used of the electrically operated switching devices is the **relay**. This is a relatively simple piece of apparatus in which an electromagnet driven by one electric circuit may operate to open or close switching contacts in one or more other electric circuits. Although telephone relays have been designed over the years with many different mechanical arrangements, most modern telephone relays are of the general basic design illustrated in Figure 9-13 and the accompanying photograph. This is known in Bell System practice as an “E” type relay. There are a number of other standard types of relays. Some of the most widely used are coded “R”, “U”, “UA”, “UB” and “Y”. Relays of each of these types are built with different winding resistances and different arrangements of the contact springs to meet different requirements. Thus, the total number of relay designs standardized for use runs well up into hundreds.

Common to all relays are the basic requirements that they must operate within specified limits of current in the winding and that they must be unfailing in making or breaking each electrical contact every time they operate; even when, as is quite frequently the case, they are required to operate millions of times per year.

Since a relay is a mechanical device having moving parts with inertia and elasticity, its operation cannot be instantaneous. The operating time

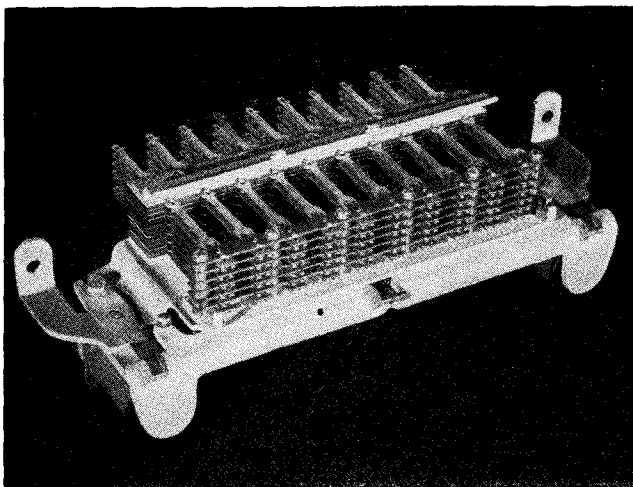


200-POINT CROSSBAR SWITCH



U-TYPE RELAY

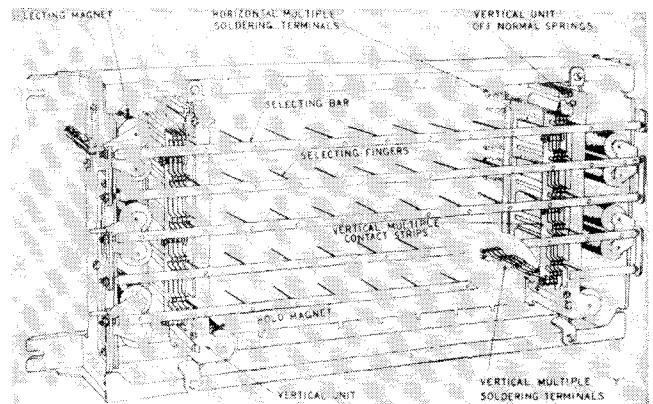
for an ordinary relay is in the order of 50 milliseconds, either on opening or closing. In some telephone circuits, however, it may be desirable to control relay operating time in such a way that it will be longer than this minimum. This can be effected within limits by adding a very low resistance path around the core of the relay in the form of either a short-circuited winding or a solid metal sleeve. When the relay is operated by current flowing through the regular winding, magnetic flux is set up in the core. This flux when it changes, induces a voltage in the surrounding space which will cause current to flow in the added short-circuited winding or metallic sleeve. This current in turn tends to build up flux in the core. When the current in the regular winding is broken, the decaying flux induces a current in the sleeve which in turn tends to set up a flux in the core additional to and in the same direction as the original flux. This will increase the time required for the total flux to decay, thus delaying the time when the relay releases. The lower the resistance of the added sleeve or short-circuited



MULTI-CONTACT RELAY

winding, the greater will be the current and the greater the retarding effect on the relay release time. Thus by varying the resistance of the sleeve, or the short-circuited winding, it is possible to secure a range of delay times varying, as desired, from the minimum of 50 milliseconds up to as much as 500 milliseconds.

Modern relays designed for accurately-timed slow release operation are usually of the "Y" type. These are generally similar in appearance to the "U" type relay shown in the accompanying photograph. Relays of this type are built with as many as 24 springs and their operating reliability is greatly increased by the use of two separate contacts on each spring. This reduces substantially



100-POINT CROSSBAR SWITCH

the possibility of faulty operation due to dust particles lodging between contact points.

Where, as may frequently be the case in the operation of local or toll dialing machinery, it is necessary to operate more than a dozen or so switching contacts simultaneously and under one control, a special type of relay known as a **multi-contact** relay is used. As shown in the photograph, this relay has two coils and two armatures each of which operates half the contacts. The springs are arranged in ten groups, each of which may include as many as six pairs of springs. This provides for a maximum of sixty simultaneous contacts or switching operations when the two armatures are operated together. The complete structure may be used as two independent relays, each having up to thirty contact springs, or as a single relay with double this capacity when the operating coils are connected in parallel. As in the case of the "U" and "Y" type relays, all contacts are doubled to provide greater reliability of operation.

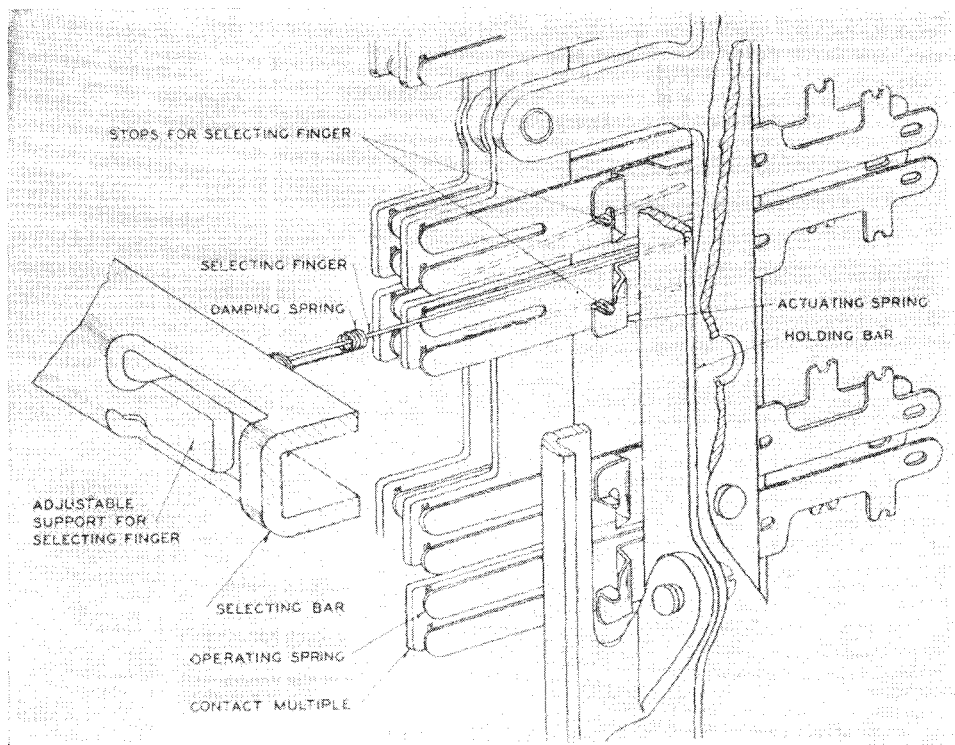


FIG. 9-14 CROSSBAR SWITCH CROSSPOINT

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. Practically all telephone circuits are dependent for their 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.

In many of the more complicated switching operations required in telephone and telegraph work, such as those involved in toll line dialing for example, it is necessary to operate switches which are electrically controlled by two rather than by only one external circuit as in the case of the relays already discussed. A commonly used switch of this type, which has many of the physical characteristics of a relay, is known as the

**crossbar switch.** Such a switch arranged to provide a maximum of 100 different switching connections is shown in the accompanying photograph. (Note, however, that not all of these connections can be made simultaneously, but at each of the 100 switch-points as many as six different contacts may be operated.) The 100 switching points or **crosspoints** of the switch shown are obtained by the use of 10 vertical bars and 10 horizontal crossbars. Many switches are arranged for twice this capacity, or 200 crosspoints, by the employment of 20 vertical bars.

In either case, closure of any one crosspoint is effected by the coordinated operation of one of the horizontal and one of the vertical bars. As shown in the photograph, the horizontal and vertical bars are located in front of the contacts and arranged to be rotated by magnets through a small arc. Each of the vertical bars is provided with a magnet at the bottom of the switch. Only five horizontal bars are used but each bar is equipped with two magnets so that it may be rotated in two directions.

Figure 9-14 shows in more detail how the controlling magnets operate the crosspoint switches. Each horizontal or **selecting bar** carries ten selecting fingers which are normally opposite an open space between two crosspoints. When one of the bars is partially rotated by its selecting magnet so that the fingers are tilted upward or downward, its ten fingers lie across the backs of the contacts as indicated by the dotted lines in the drawing. If one of the ten vertical bars is now rotated so as to press against the vertical row of five selecting fingers, the selecting fingers that are in normal position will be pressed into the spaces between crosspoints, with no effect. The selecting finger that has been tilted upward or downward by the operation of the horizontal bar, however, will be pressed against the adjoining flexible contact spring assembly, thus closing the contacts at

that crosspoint. Once this contact has been made in this way, it will remain closed as long as the vertical bar is held in the operated position even though the horizontal bar has been returned to normal, because the selecting finger will remain

trapped against the contact spring. For this reason the vertical bar is known as a **holding bar** and its associated magnet is known as a **holding magnet**.



## CHAPTER 10

### TELEPHONE CIRCUITS

#### 10.1 The Telephone Subscriber Set

Figure 9-10 of the preceding Chapter showed a simple telephone connection between two telephone sets, each equipped with a transmitter, receiver, induction coil and its own battery for supplying talking power. In most modern telephone station installations, talking battery is supplied to each subset from a common battery at the telephone central office to which each subscriber line is connected. The simplest subscriber station circuit arrangement under these conditions is shown schematically in Figure 10-1. When the receiver is lifted to close the contacts of the hook-switch, and the line is picked up at the central office by an operator or mechanical device, the central office battery is connected in series with the primary winding of the induction coil and the transmitter, and current is sent over the line.

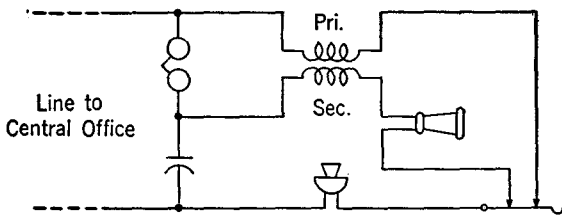


FIG. 10-1 SIMPLE STATION CIRCUIT

Varying currents set up by the transmitter, when it is talked into, add to or modulate the direct current flowing from the central office battery. There is also a path for the varying transmitter currents (which are fundamentally alternating rather than direct) through the capacitor, the secondary of the induction coil and the receiver. Thus, there are varying currents in both windings of the induction coil and each induces a voltage, and a consequent current, in the other winding. The two currents in the primary winding combine to flow out on the line and thence to the distant receiver, and the two currents in the secondary winding combine and flow into the receiver and cause sidetone.

When receiving, the incoming varying (alternating) current, which carries the message,

divides between the two parallel paths containing the transmitter and receiver respectively. In passing through the windings of the induction coil, in these two paths, the current in the primary winding induces a voltage in the secondary winding, which opposes the part of the incoming current flowing into that winding; but the circuit is so designed that the part of the incoming current in the secondary winding is greater than that in the primary so that the resultant current flowing in the receiver is still sufficient to operate it. Since in receiving, a substantial part of the incoming energy is dissipated in the transmitter circuit and in transmitting, energy is likewise dissipated to no useful purpose in the receiver circuit, the subset is fundamentally inefficient. Such inefficiency, however, is inherent in any circuit designed to operate in two directions without using directional switching arrangements.

The transmitting current flowing in the receiver as sidetone obviously serves no useful purpose and may be confusing or annoying to the speaker. The simple subset discussed above is modified generally in modern practice to a circuit arrangement such as that shown in Figure 10-2, which is one of several possible "anti-sidetone" connections. This circuit makes use of a different induction coil having a third winding  $S_2$ , the resistance of which is made high relative to that of winding  $S_1$ . For the receiving condition the current relationships are effectively identical with those of the sidetone circuit shown in Figure 10-1 because the resistance of  $S_2$  is so high. For the transmitting condition there are, as in the sidetone circuit, two parallel paths for the current flow—one directly over the line through the primary of the induction coil and the other through

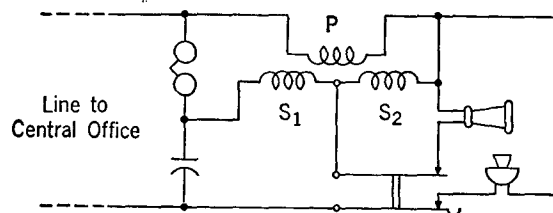


FIG. 10-2 ANTI-SIDETONE STATION CIRCUIT



500-TYPE TELEPHONE SET

the capacitor and  $S_1$  and either through  $S_2$  or the receiver. Because of the relatively high resistance of  $S_2$  most of the current will tend to flow through the receiver, but the induction coil is so designed that current flowing in  $P$  and  $S_1$  will induce a voltage in the winding  $S_2$  which will tend to set up a current approximately equal and opposite in direction to that flowing through the receiver. The net effect is no current in the receiver, and hence no sidetone. This circuit is no more efficient than the sidetone circuit but it seems more so, principally because in receiving the listener is not distracted by extraneous noises coming from his own transmitter.

## 10.2 The Telephone Central Office

In practical telephone systems all subscriber stations are connected to a telephone "central office" by relatively short lines known as subscriber loops. The major purpose of a central office is to provide switching arrangements whereby any subscriber station connected to that office can be connected at will to lines leading to other subscriber stations also connected to that office; or to trunk lines leading to other central offices in the same or distant cities. The central office has numerous incidental functions, including that of providing common talking battery for all of its lines as discussed in the preceding Article. This requires comparatively large installations of storage batteries, power generating machinery and associated control equipment. Installed in the central office too, are numerous frames and racks, testboard and other testing equipment, signaling and control devices, and various other auxiliary equipment. Important, both at the central office and the subscribers' stations, is the installation of

protective devices which will guard both telephone apparatus and people using it from dangerous foreign voltages that may be brought in on the lines as the result of some abnormal condition. This protection is of such basic importance that it may be well to digress here to review it in some detail.

Practically all outside telephone plant, except conductors that are completely underground from terminal to terminal, may be occasionally exposed to excessive voltages from such sources as lightning and other atmospheric disturbances, electric power lines running in close proximity to the telephone lines, high power radio sending apparatus, etc. Accordingly, such exposed conductors leading into a central office or subscriber station are connected first to appropriate 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—open-space cut-outs, 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.

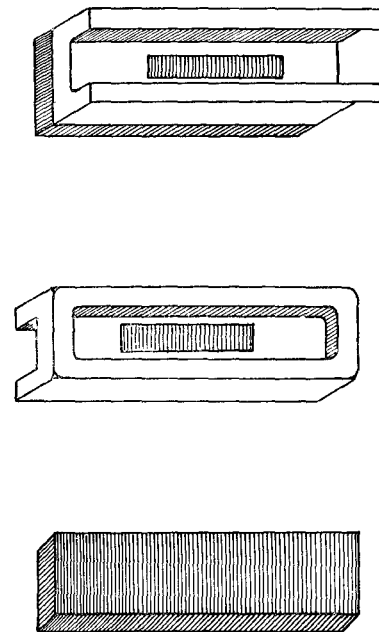


FIG. 10-3 OPEN-SPACE CUTOUT

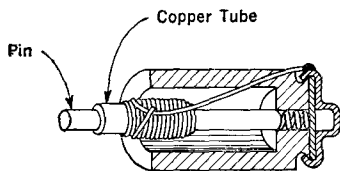


FIG. 10-4 HEAT COIL

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 10-3. 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. In the majority of protector operations, however, the blocks do not become permanently grounded.

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 on 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 10-4, 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

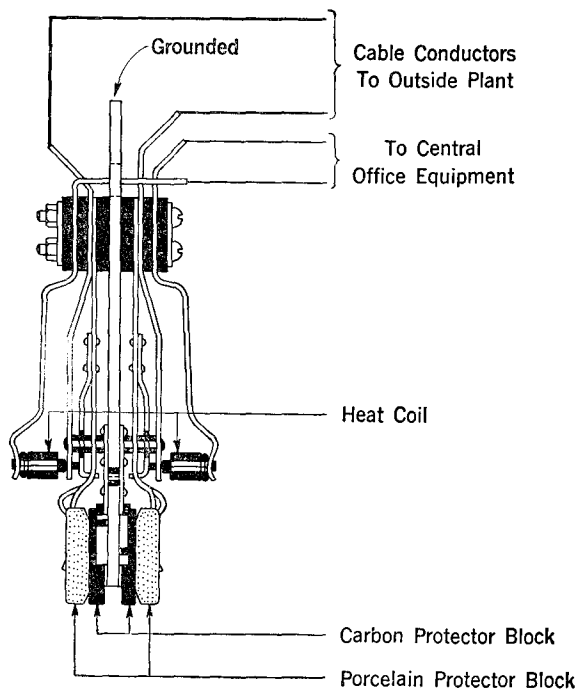


FIG. 10-5 HEAT COILS AND CUTOUTS MOUNTED ON PROTECTOR FRAME IN CENTRAL OFFICE

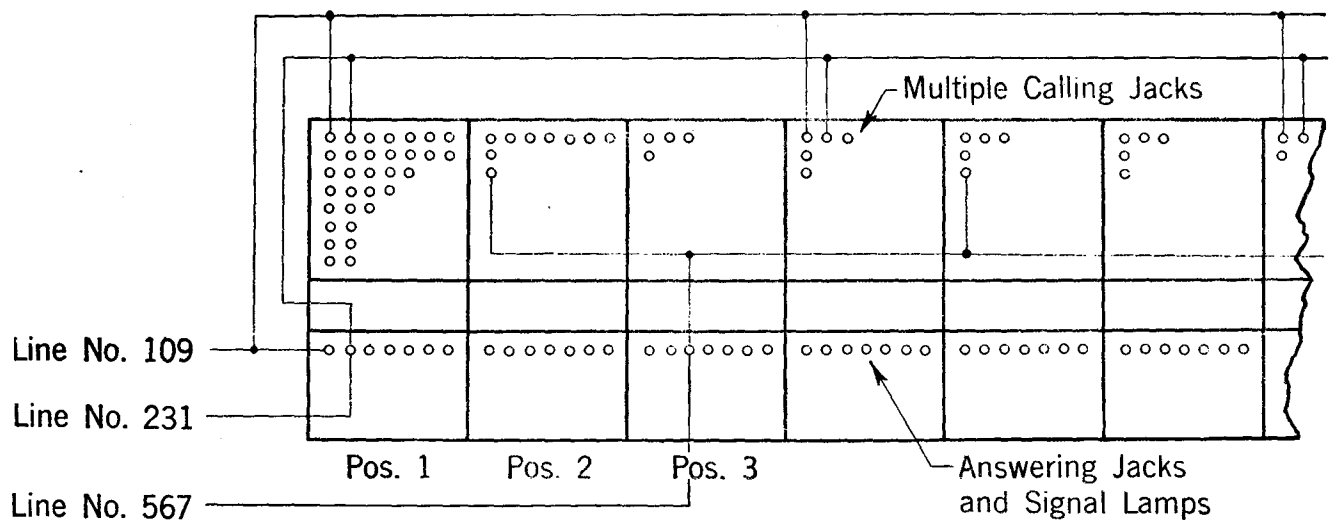


FIG. 10-6 MULTIPLE SWITCHBOARD

the pressure of its mounting spring and thus connect the line to ground. 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 method of mounting heat coils and open-space cutouts on the protector frames in central offices is illustrated in Figure 10-5.

Let us return now to the switching function of the central office, which we have already pointed out as its main purpose. Switching equipment is designated as **manual** or **dial** depending upon whether the switching operations are performed primarily by people or by mechanical devices. Although both types perform the same function, the equipment and methods employed in each case are quite different.

### 10.3 Manual Switching Systems

In the manual central office, switching is performed by operators at switchboards. In a local single office exchange, all subscriber lines terminate in at least two jacks in these switchboards. One jack, mounted in the lower part of the switch-

board panel, has associated with it a signal lamp by means of which a subscriber wishing to make a call can secure the attention of the operator. This is her **answering jack**. The other jack mounted in the upper panel of the board, is used by the operator to complete a connection to this same subscriber's line when it is called by some other station. This is designated the **calling jack**. Except in very small offices, each line appears in parallel at many other jack locations along the board. In other words, the terminations are **multiplied** so that any line is within reach of any operator.

The principle of the multiple switchboard is that the answering jacks and associated 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 multiplied, that is, connected in parallel, with a similarly located jack in the third position to the left and right. 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 10-6. 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 plug into a calling jack already in use somewhere else in the multiple.

Figure 10-7 shows a telephone connection between two 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 which is bridged across the circuit in series with a capacitor. It is a function of the capacitor 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 ground. This energizes the line relay  $R_1$  which connects the central office battery to the answering lamp  $A_1$ , in the face of the switchboard in front of one or more operators. This lamp, lighting, indicates that this particular line is calling. The operator answers the call by inserting plug  $P_1$  into the answering 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 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 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 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 learning the number of the party called, the operator inserts plug  $P_2$  into calling 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 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 finish talking and hang up their receivers, this supervisory relay, as well as the corresponding 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 circuits of the cords are opened at the jacks and the lamps go out.

What we have discussed above is perhaps the simplest example of a telephone switching operation—that is, a connection between two subscribers of the same single exchange established by one operator at an "A" switchboard. Where a

city or other exchange area is of such size as to require more than one central office, a connection between a subscriber whose line terminates in one office and a subscriber whose line terminates in another office cannot be completed by one operator. In such case the "A" operator who picks up the call from a calling subscriber, uses her cord circuit to connect the calling subscriber's line to a trunk leading to a switchboard in the other central office in which the called subscriber's line is terminated. This switchboard is known as a "B" board. It differs somewhat in appearance and in the detail of its circuit arrangements from the "A" board. The incoming trunks are terminated in the "B" operators' key shelf in plugs, and the panels of the switchboard are used almost entirely for calling or multiple jacks.

When the "A" operator at a distant office connects a calling line to a trunk to this "B" board, a lamp associated with the plug and cord in which the trunk terminates in the "B" board key-shelf will light. At the same time the "B" operator's head-set is automatically connected so that the "A" operator at the calling office can pass her the number. The "B" operator then inserts the plug into the called jack. Ringing is usually automatic and the supervisory signal lamp at the "B" board remains lighted until the called subscriber answers. After the conversation is completed and the called subscriber hangs up, the supervisory lamp again lights indicating to the "B" operator that she should take down the connection.

There are other situations in large cities where it is uneconomical to provide direct switching trunks from each central office to every other central office in the exchange area. In these cases a third type of switchboard known as a "tandem" board is provided in some appropriately located intermediate central office. A connection then requires the service of a third operator. The "A" operator at a calling office connects to a trunk leading to the tandem board at the intermediate office; the operator there connects this trunk to another trunk leading to the called office; and the "B" operator there in turn connects the trunk to the called subscriber's line. Naturally, the switching circuits in these three types of switchboards vary in detail, particularly in respect to the supervisory signaling arrangements. Basic methods and principles, however, do not differ fundamentally from these described in connection with Figure 10-7.

The requirements for long distance or toll switching are essentially similar to those for local

switching. Again, there are three principal switching situations—between the calling subscriber and the toll trunk or line, between the toll trunk and the called subscriber, and where necessary, between two toll trunks. In long distance practice, these operations are performed at outward, inward and through switchboards, respectively. Signaling and supervisory facilities are necessarily somewhat different in toll switching than in local switching. In general, the toll switchboard must be designed to permit a larger variety of switching operations and the cord circuits and other equipment facilities associated with it are accordingly somewhat more extensive. Because of the variety of switching situations, there is no one specific design for toll switchboard arrangements that applies alike to all such switchboards. The reader who wishes to study switchboard circuit arrangements in detail may refer to the standard drawings and instructions applying to the particular switchboards in which he is interested.

#### 10.4 Dial Switching Systems

The fundamental operations to be performed by a mechanical switching system are naturally the same as those discussed above. It is hardly necessary to point out, therefore, that any completely mechanical system must be made up of a considerable quantity of intricate electro-mechanical equipment. In the local dial exchange the operation of the mechanical switching system is controlled by a dial installed on the subscriber's telephone set, which transmits to the central office electrical impulses corresponding to each digit dialed by the subscriber.

The oldest type of dial central office machinery in use functions on a "step-by-step" basis. This means that each digit dialed causes the movement of a switch at the central office which will carry the connection a step forward toward its ultimate objective—the called subscriber's line. The basic principle involved can be readily understood by examination of the simplified example shown in Figure 10-8. Here only two steps are shown and there are only five contacts on the switches of each step. If the switches are arranged to move in order through the same number of contacts as the number of electric impulses coming from the dial, any one of twenty-five separate telephones could be reached by dialing only two successive digits between 1 and 5. If, as in the actual telephone dial, there are 10 digits and the system is designed for dialing as many as

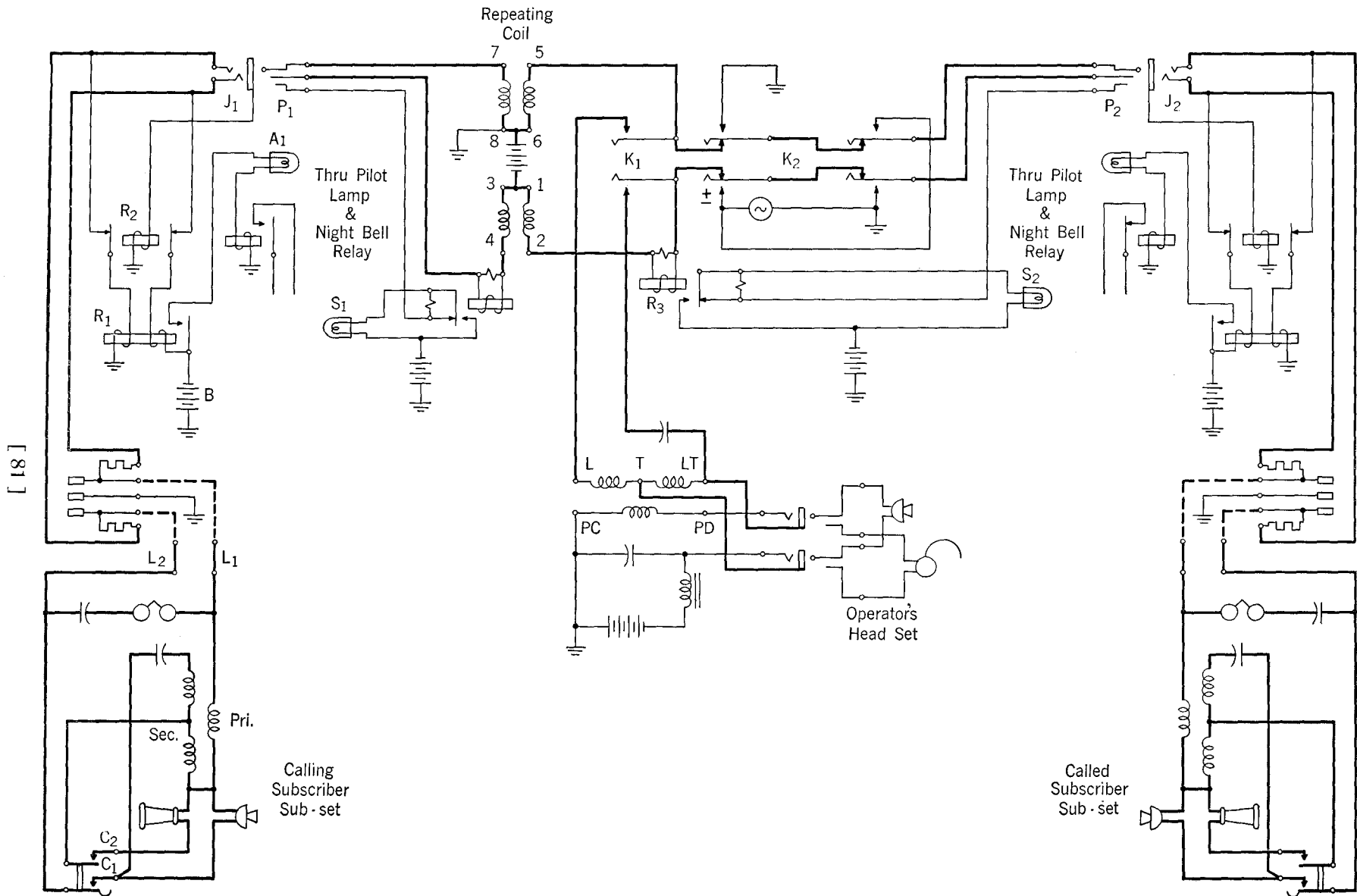


FIG. 10-7 TELEPHONE CONNECTION THROUGH COMMON BATTERY EXCHANGE

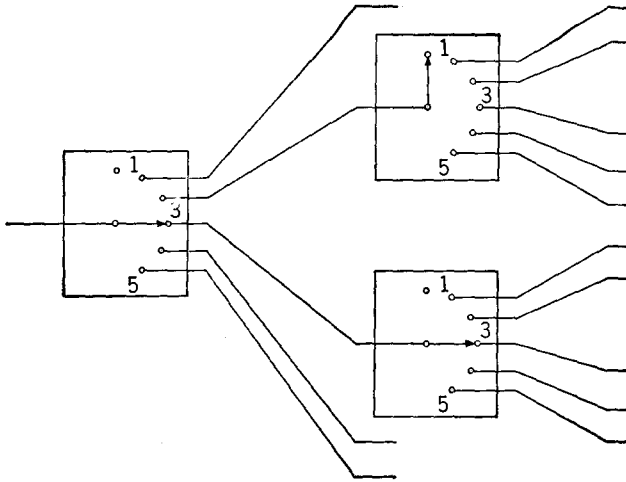


FIG. 10-8 PRINCIPLE OF STEP-BY-STEP SWITCHING

seven digits to reach the desired telephone, the simple diagram of Figure 10-8 could be expanded without any change in principle to show how any one of ten million telephones could be reached in the seven successive dialing steps. The trouble with such a simple arrangement as this is that it would require a tremendous number of switches and connecting wires, and still permit only one telephone to make a call to any of the others.

By suitable arrangements, however, this principle can be employed to permit the origination of the call by any of the telephones and still keep the total amount of switches and other equipment within reasonable bounds. The first step is to eliminate the necessity for having a separate selecting switch associated with each subscriber's line. Instead, switches known as **line-finders** are provided for groups of subscriber lines, each one of which is connected in multiple to the "bank" terminals of the switches. In the **step-by-step** system, the line-finder switch resembles that illustrated in Figure 10-9, which is arranged to move vertically through ten steps and horizontally through ten steps for each vertical step. When a subscriber picks up his receiver, a relay associated with his line causes an idle line-finder switch to go into operation and hunt for the terminal to which his line is connected.

With this arrangement, a simple dial system might take the form illustrated in Figure 10-10. Here the line-finder switch connects the calling subscriber's line to a trunk selecting switch which is installed in the same central office, and in the step-by-step system is of the same design as the line-finder switch illustrated in Figure 10-9. Dial tone is automatically sent back to the calling subscriber and as he begins to dial, the trunk selector

switch in office "A" moves to a terminal at which a trunk corresponding to the initial digit dialed is terminated. This trunk may lead to a distant office or to terminating equipment in the same office. In the simple case illustrated, the first switch at office "B" selects a group of connecting lines under the direction of the second number dialed by the calling subscriber and moves the call on to a final selecting switch, which is operated by the third number dialed to connect to the called subscriber's line.

The above illustration of course is over-simplified as compared to most practical telephone systems. In principle, however, the method indicated could readily be expanded to cover a larger number of telephones (and more digits to be dialed) by including more switches of the same type in the switching chain.

Figure 10-10 illustrates another principle which is of fundamental importance in any practical switching system. It will be noted that it shows two trunks between office "A" and office "B" each of which is multiplied to selector switches in the two offices. This brings in an additional feature

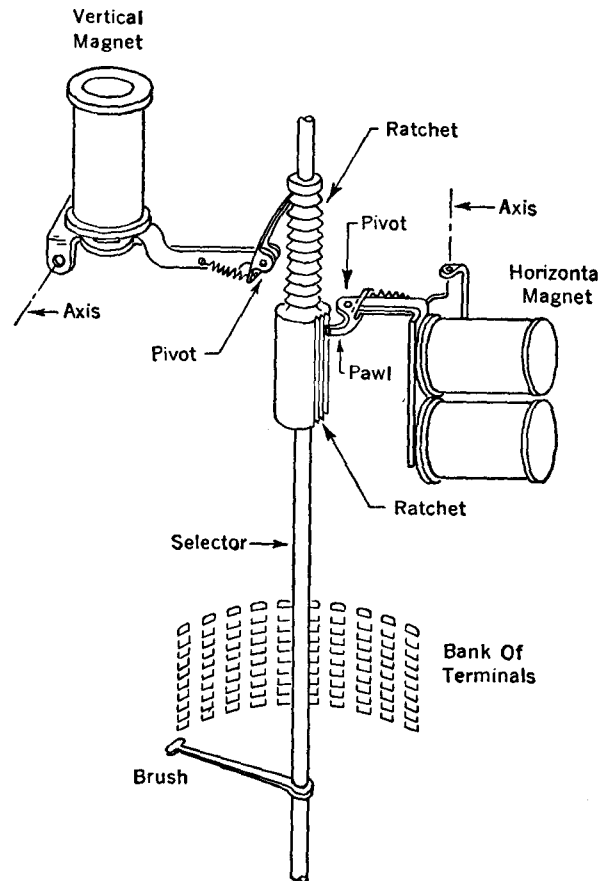


FIG. 10-9 STEP-BY-STEP LINE-FINDER SWITCH



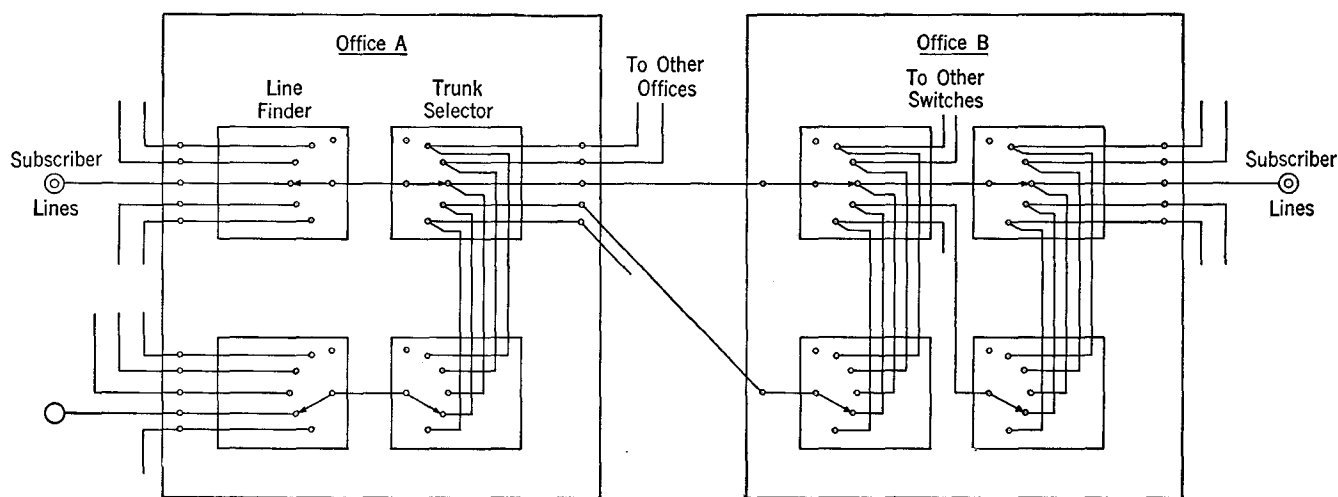


FIG. 10-10 INTER-OFFICE DIAL SWITCHING PRINCIPLE

in the switching apparatus under which a switch directed toward a called trunk, but finding it already busy, will automatically move on and hunt for another trunk leading to the same called office. Just as in manual switching, practical economy requires that the numbers of trunks installed between any two offices (or between any other two points in the switching system) be kept at that minimum which (as determined by applications of the laws of probability) will be sufficient to take care of the busy-hour load under reasonably normal conditions. In the large office or exchange area, this may mean that the total number of trunks between two points in the over-all system may run to a considerable number. To keep the number of switching devices to an economical minimum, it is evident that any switch should be able to hunt automatically over the total number of trunks in a group, no matter what its size. When the number of trunks in a group exceeds ten, a step-by-step dial system becomes inadequate because there may not be time for the selector switch to hunt over a larger number of trunks before the next number is dialed by the calling subscriber. In order, then, to design a mechanical switching system that will operate in an exchange area having more than perhaps four or five central offices (or a total of 50,000 subscriber lines) a fundamentally different approach must be adopted.

Two other types of dial switching systems designed for application in large exchange areas are in use in the Bell System. These are known as the **panel** system and the **crossbar** system. Both depend on the principle of first transmitting the dial pulses from the calling subscriber's telephone

to a "mechanical operator", usually known as a **sender**, which in turn translates them into other electrical impulses that will operate the necessary switches to establish the through connection. The panel system, although functioning today in many locations, no longer is being installed, so that we may confine our attention in what follows to the crossbar system.

If, as mentioned in Article 9.6, the crossbar switch is thought of as a kind of relay, the crossbar system may be considered as an all-relay system. It employs large numbers of these crossbar switches and of ordinary relays of the "U" and "Y" types, as well as the multi-contact relays which also were discussed briefly in Article 9.6. Naturally the detailed arrangements are complicated; it will be practicable here to outline only the broad principles of the system's operation. These are illustrated schematically in Figure 10-11, where the through connection from the calling to the called telephone is indicated by the heavy line. As shown, the connection goes through four different switching or link frames, including the line link frame, which acts as a line finder at the calling office and a final selector at the called office. In each one of these frames the connection routes through crossbar switches, generally known as the primary switch and the secondary switch. In the simplest situation where each switch has 100 crosspoints and the switching frame has 10 primary and 10 secondary switches, the arrangement offers the possibility of connecting any one of 100 incoming lines to any one of 100 outgoing lines.

When a subscriber connected to office "A" lifts his handset to make a call, his line is closed

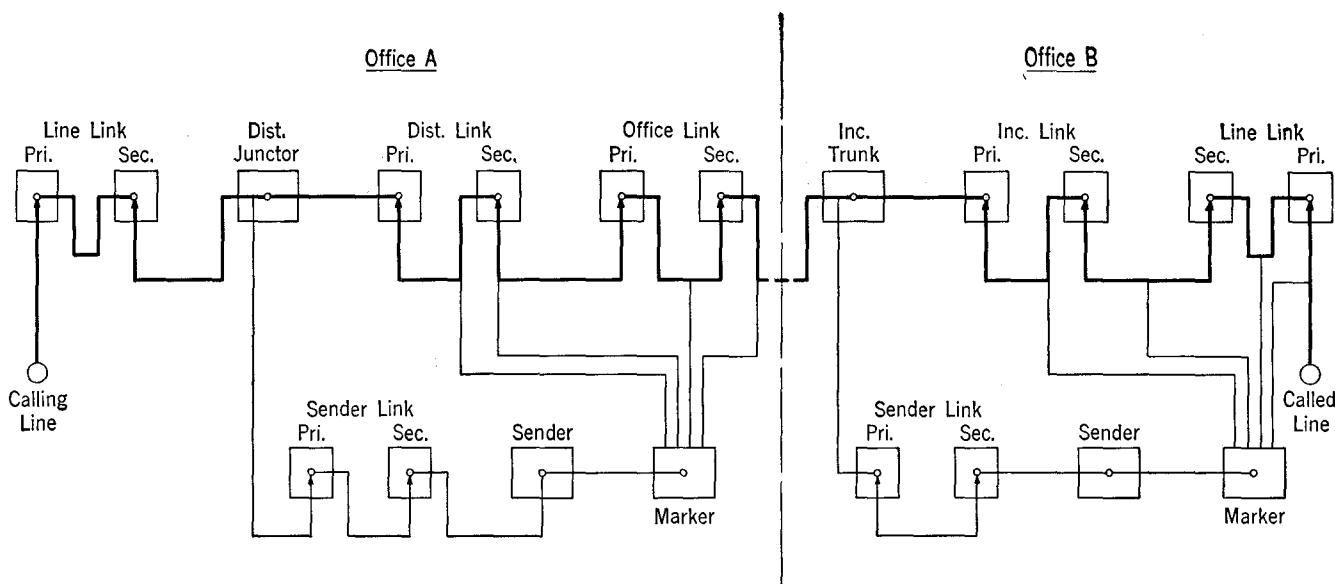


FIG. 10-11 LOCAL CROSSBAR SWITCHING ARRANGEMENT

to a primary vertical in the line link frame at that office. At the same time relays are operated to call into play a **controller** circuit arrangement (not shown in the diagram) consisting of a number of relays of various types, and associated wiring. This device first definitely locates the calling line on the line link frame and operates the proper selecting and holding magnets to connect the line to an idle trunk or **district junctor** leading to the district link frame. The district juncctors extend through a district junctor frame on which is mounted relay equipment that supplies talking battery to the calling line. It also maintains supervision over the call after the connection has been established, times and registers local calls, controls collections or returns of deposited coins at pay stations, etc.

At the same time, the controller hunts for and selects an idle sender and operates selecting and holding magnets in the sender link frame to connect the sender to the calling line. This connection causes dial tone to be sent back to the calling line, indicating that dialing may be started. The dialing pulses are received at the sender and registered there by the operation of relays or crosspoints of a crossbar switch. As soon as the sender has registered the digits indicating the office being called (the first three digits in the case of large multi-exchange areas) the sender operates to locate and connect itself to an idle **marker** through multi-contact relays. It then transmits to the marker information as to the office code dialed, the number of the frame in

which the call is originating, the class of service of the calling line, etc. As soon as this information is registered by relays in the marker, the marker first hunts for and selects and "marks" or makes busy an idle outgoing trunk to the distant office. It also selects an idle junctor between the district link frame and the office link frame and then operates the proper selecting and holding magnets in both frames to complete the connection between the selected inter-office trunk and the district junctor to which the calling line had been previously connected. This total operation of the marker is accomplished in a fraction of a second and as soon as it is completed, the marker is automatically released to work on other calls.

In the meantime, the calling subscriber continues to dial the digits corresponding to the number of the called telephone. These are likewise registered in the sender. As the hundreds digit is dialed, the trunk circuit to the distant office is closed to the incoming trunk frame, which causes a controller at that office to select an idle terminating sender and connect it by the closure of the proper switch-points in the sender link frame. Like the district junctor frame at the calling office, the incoming trunk frame mounts relay equipment which furnishes talking battery to the called subscriber's line, maintains supervision over the terminating end of the connection, furnishes the proper type of ringing current over the called line, and transmits busy tone to the calling line if the called line happens to be busy. After the terminating sender has been connected to the

incoming trunk circuit, the called number registered in the originating office sender is transmitted to the terminating office sender where it is again registered in relays or crosspoints of a crossbar switch. At this point the originating sender having completed its function is disconnected and released for use on another call.

The terminating sender then connects to or "seizes" an idle terminating marker to which it transmits information on the called subscriber number and the number of the incoming trunk frame on which the particular incoming trunk is located. By means of its translator circuit the marker is able to locate the called subscriber's line out of a maximum of about 10,000 such lines terminating at the central office. If this line should be busy, the marker directs the incoming trunk circuit to transmit a busy signal to the calling subscriber. If the called subscriber's line is idle, the marker first makes it busy and then proceeds to test for and select idle links in both the line link and incoming link frames as well as an idle junctor connecting these two frames together. It then operates the proper selecting and holding magnets in the two link frames to operate the necessary crosspoints that will close the circuit through. With this, the incoming trunk circuit transmits ringing current to the called subscriber's line and the terminating sender and marker disconnect and return to normal. As already pointed out, the district junctor at the calling end and the incoming trunk at the called end jointly maintain supervision over the call. When the talking parties disconnect, they function to cause all connections to be restored to normal.

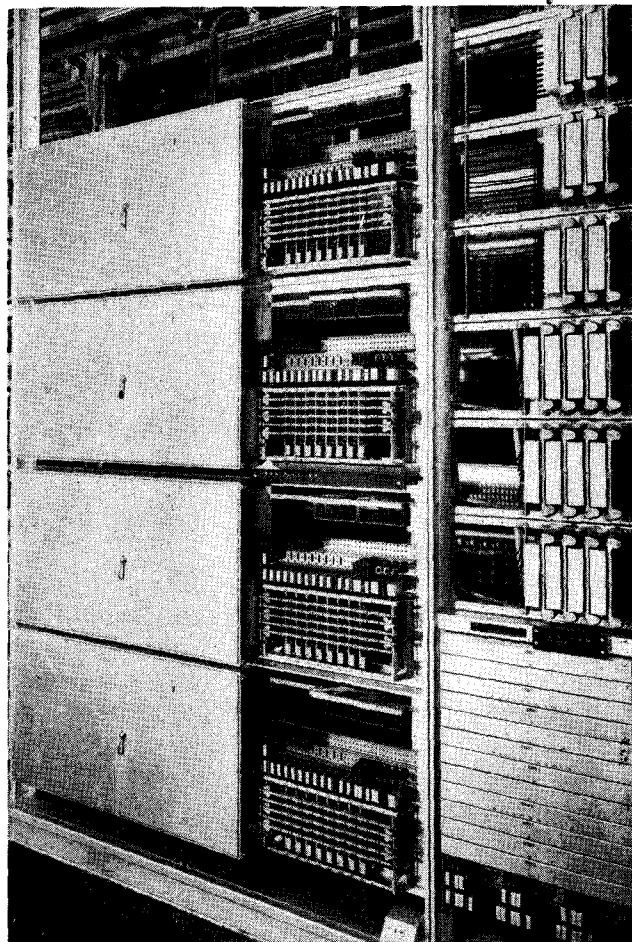
The reader will have noted that in the preceding brief description of a dial call, various apparatus units—particularly the senders and markers—have been spoken of much as if they were persons like manual telephone operators. The functions that they perform are indeed very similar to those performed by operators. Needless to say, therefore, their circuit arrangements are intricate—so much so that their detailed analysis would involve considerably more space than can be allotted to the subject in this book. It should be remembered, however, that they are only machines, having no capacity to exercise either initiative or judgment, but able to do only those things that they were specifically designed to do.

The accompanying photographs give some idea of the physical appearance of a typical sender and a typical marker. The latter may include as many as 1,700 relays. Not all of these operate in estab-

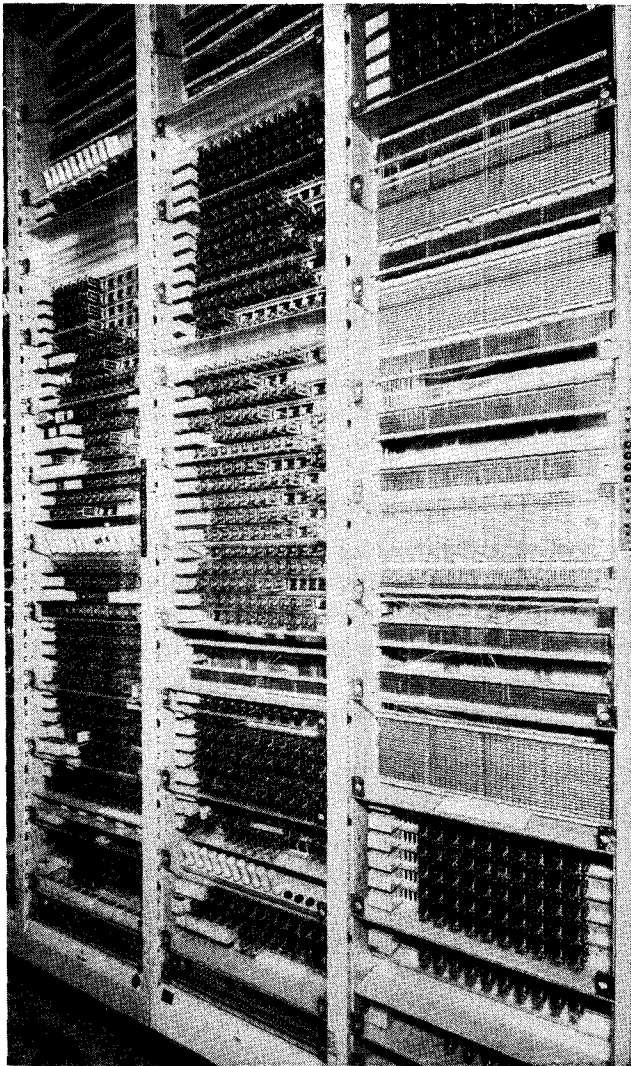
lishing any one connection, although it is interesting to note that a typical local dial connection involves the operation of about 700 relays. Markers and senders are naturally costly pieces of apparatus. As pointed out earlier, however, the marker performs its function in less than a second, which means that it can handle many calls within a short time. Actually it is necessary to provide no more than 5 or 6 markers to handle all the work in one dial system local office. The operating time of the sender is somewhat longer—up to about 15 seconds—so that proportionately a greater number of these apparatus units have to be provided.

### 10.5 Toll Dial Switching

The establishment of toll or long distance connections by mechanical or dialing methods is not essentially different from the setting up of local connections. As was pointed out in the earlier discussion of manual switching arrangements, long distance switching generally involves a



SENDER FRAMES WITH MARKER CONNECTORS AT THE RIGHT



MARKER FRAMES

greater number and variety of different situations than does local switching. Thus the long distance connection will frequently require a greater number of switches for completion from terminal to terminal.

For relatively short toll connections in areas where the total telephone population is not too great, the step-by-step system may be used. This is essentially the same as the local step-by-step connection discussed in the preceding Article, employing a toll trunk instead of an inter-office trunk. On the other hand, the same limitations with respect to dialing time apply. More important in toll connections is the fact that when a call extends beyond the area where toll charges are billed as message units, the details of the call must be recorded for billing purposes. These include the number of the calling line, the name of

the place called, the called number, and the total time of the call. Mechanical equipment which will make such a record is in use in the Bell System. The more usual current practice on long distance calls, however, is to retain the services of an outward toll operator. This operator receives the details of the call from the calling subscriber, records them on a toll ticket, and completes the connection with the aid of dialing machinery direct to the called telephone at the distant point. Thus, if the initial connection from the calling subscriber to the outward toll operator is established mechanically, only one operator is involved in the complete connection.

The outward toll operator establishes the connection to the distant point by inserting the plug of a cord circuit into an appropriate jack of the switchboard and dialing or **key pulsing** a series of digits that will direct the machinery in her own and other offices to make the desired connection. These may include a tandem connection in her own office, one or more through connections at intermediate cities along the route, and an inward connection at the called city, in addition of course to the local office connection to the called subscriber's line at that point.

At each point where such connections are made, the dialing machinery arrangements are quite similar to those at each end of the local crossbar dial connection discussed in the preceding Article. Figure 10-12, for example, might be the arrangement of a toll crossbar tandem at a large calling city. Here the outward operator has plugged into the tandem trunk marked "incoming" in the Figure. This action causes a sender to be connected to the trunk through a sender link frame. As soon as this is done, the sender transmits a signal to the operator indicating she may start key-pulsing. For a typical connection not involving an intermediate switch point, the digits might be 321-WO6-1748. The first three digits select the distant city; the second three select the exchange in that city; and the final four are the called subscriber's number. As soon as the first three digits are received and registered by the sender, it seizes an idle marker and passes along the information that they contain, together with necessary information concerning the location and characteristics of the incoming trunk. The marker then tests for and selects an idle outgoing toll trunk to the called city, following which it tests for and selects idle links "A" and "C" in the incoming and outgoing frames and an idle junctor or "B" link between these two frames. It

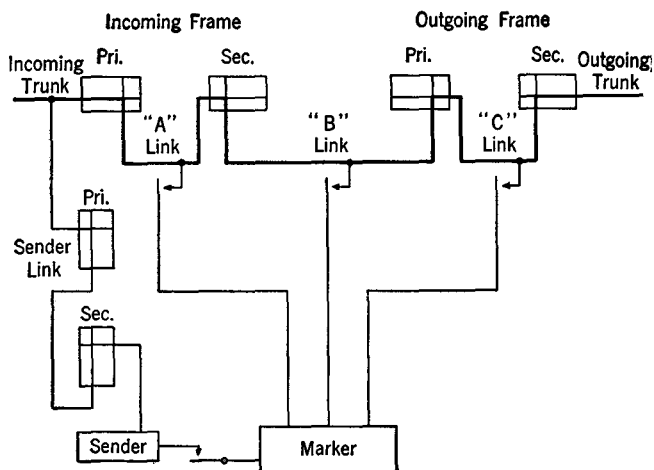


FIG. 10-12 TOLL CROSSBAR SWITCHING ARRANGEMENT

then operates the selecting and holding magnets to close the four crosspoints, and disconnects from the circuit.

In the meantime, the operator will have continued to key into the sender the remaining digits. When the marker closes the circuit through to the outgoing trunk, a signal is automatically relayed to the distant office which causes that end of the trunk to be connected to an idle incoming sender. This connection is made known to the sender at the calling city by an appropriate signal whereupon that sender transmits to the sender at the called city information concerning the digits that it has registered (not including the first three which already have served their purpose). The sender at the calling city then disconnects from the circuit. The incoming sender at the called city proceeds with the aid of a marker to set up a connection through two switching frames between the incoming toll trunk and a switching trunk to the called local exchange. It then passes the called telephone number (four digits in this example) to the dial equipment in the local office, which makes the connection to the called subscriber's line.

In practice there are necessarily many variations in the above procedures, as required to meet the many different switching situations which occur in toll operating work. Thus the call may have to be switched at one or more intermediate points between the calling and called party. This may necessitate the keying of additional digits by the outward operator, but the machine operation at the intermediate switching point is essentially no different from that outlined above. Toll dialing machinery also lends itself quite readily to the automatic selection of alternate routes when the

circuits on the normal route are all busy. Such automatic route selection, however, does not require any different procedure on the part of the outward operator nor does she need to know just what route the call actually follows.

There is an important signaling feature in which toll dial operation differs somewhat fundamentally from manual toll operation. In the latter, ordinary supervision of a call is exercised more or less independently by the several operators concerned by means of the ordinary direct current supervisory signal lights at their switchboards. In a complete dial connection, there is only one operator, who is at the calling end of the connection. She must know, however, when the called party answers the telephone and when he hangs up. This means that some kind of supervisory signal must be transmitted back to her from the distant end of the connection. The procedures used for this purpose are discussed briefly in the Article following.

## 10.6 Signaling

In any telephone switching system, it is naturally necessary to provide some means of signaling subscribers and operators. Thus, the subscriber is signaled by the familiar telephone bell which is ordinarily operated by a 20-cycle alternating current supplied from a local central office when an operator presses a ringing key, or when machinery performs a comparable operation. Such a signaling current is not suitable for use over toll circuits of appreciable length since, among other things, it might interfere with telegraph currents being transmitted over the same circuit. This difficulty may be overcome in relatively short toll circuits by using a signaling current of 135-cycle frequency, which is high enough not to interfere with the lower speeds of telegraph services. The necessity for relaying the 135-cycle signal around repeaters, however, limits its usefulness for application on the longer circuits. More reliable signaling on long circuits is obtained by the use of 1000-cycle signaling current. This is a frequency within the normal voice range which suffers no greater losses in transmission than do the voice currents themselves, and it is amplified at intermediate points by the same telephone repeaters that amplify the voice currents. In order to avoid false operation of the signals by voice currents, the signal receiving apparatus is designed to be operated only by a 1000-cycle current interrupted 20 times per second and sus-

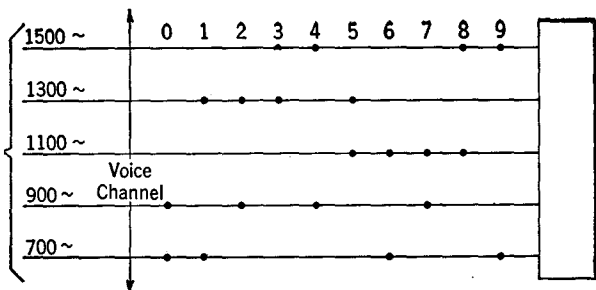


FIG. 10-13 MULTI-FREQUENCY KEY PULSING OF DIGITS

tained for at least several tenths of a second. The signals are usually generated by electron tube oscillators. They are in turn detected at the receiving end by appropriate electron tube circuits and converted into direct current or 20-cycle ringing current as needed.

The signaling methods discussed above are known as **ringdown** systems and are employed on "ringdown trunks". They are not suitable for use on circuits arranged for intertoll dialing operation. Such circuits must be so equipped that both ringing and supervisory signals can be transmitted in both directions. Ringing signals, originally generated by operation of the ordinary dial in the form of dial pulses, may be transmitted over relatively short distances in the form of d-c pulses similar to telegraph signals. In this case, composited or simplexed telegraph channels, separate from the telephone circuits, are used for the transmission. For longer circuits, dial pulses are generally transmitted by interrupting a single frequency tone which is transmitted over the telephone circuit in the same way as the voice signal. This single-frequency signaling method uses a frequency of either 2600 or 2400 cycles on circuits of sufficient bandwidth for such transmission and a frequency of either 1600 or 2000 cycles

on circuits that cannot satisfactorily transmit the higher frequencies.

Where ringing signals are generated by key-pulsing, which is much faster than dial pulsing, multi-frequency signals are employed. This generally applies to the longer toll lines. The frequencies used are 700, 900, 1100, 1300, 1500 and 1700 cycles. Each digit is represented by a combination of two of these frequencies. Figure 10-13 shows the combinations for each of the digits from 0 to 9, inclusive. The 1700-cycle tone, not shown in the Figure, is used in combination with 1100 to provide a priming or "get ready" signal; and in combination with 1500 for a "start" signal.

Supervisory signals are transmitted either in the form of direct current over separate telegraph channels, or by means of a single voice frequency transmitted over the telephone circuit. In the latter case, the supervisory signaling tone should of course not be present when the circuit is being used for conversation, since it is within the voice range. The supervisory signal tone is applied continuously to the line as long as the called telephone is in the "on-hook" condition. When the receiver is taken off the hook, the signaling tone is removed from the line. Its presence or absence causes the operation of relays at the calling end which control the signal lamps at the outward operator's switchboard and thus enables her to know when the distant telephone answers or hangs up. These supervisory signal tones must of course be applied to a circuit from both ends in order that a circuit may be used for calls originating at either end. For circuits that are actually or effectively 4-wire, the single frequency of 2600 (or 1600) cycles can be used in both directions. For 2-wire circuits, a 2600 (or 1600) cycle tone is used in one direction and a 2400 (or 2000) cycle tone in the opposite direction.

## CHAPTER 11

### TELEGRAPH CIRCUITS

#### 11.1 Means of Obtaining Telegraph Circuits

As a means of communication, telegraphy depends upon the transmission of electric signals which are arranged according to some definite code that can be readily translated into language form by an operator or a machine.

The great majority of long-haul telegraph circuits today are obtained through the use of "voice-frequency carrier systems" by means of which up to 18 two-way telegraph circuits are derived from a transmission facility that would otherwise be capable of use as a single telephone circuit. Description of these carrier systems must be left to later Chapters of this book that follow the general study of alternating current circuits. At this point, since thus far we have been concerned essentially only with direct current phenomena, it will be necessary to confine our study of telegraphy to its direct current aspects. This will permit an analysis of the more fundamental principles of telegraph circuits and telegraph transmission, although the reader should recognize that modern telegraph practice involves many alternating current applications, including the use of electron tubes and other electronic devices.

Before the advent of carrier systems, practically all long telegraph circuits in the Bell System were derived from open wire or cable wire conductors that were used simultaneously for telephone transmission. This is still standard practice although, as noted above, only a minority of the total of telegraph circuits is now so obtained.

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 11-1. 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.

Referring to Figure 11-1, 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 capacitances 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 voltage in the "drop winding" of the coil, which will not be neutralized by the voltage induced by the lesser part.

Differing radically in principle from the simplex set, the **composite set**, illustrated by Figure 11-2, 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.

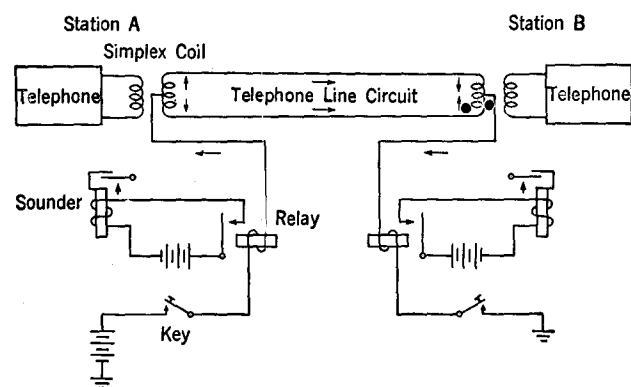


FIG. 11-1 TELEGRAPH CIRCUIT ON SIMPLEXED TELEPHONE CIRCUIT

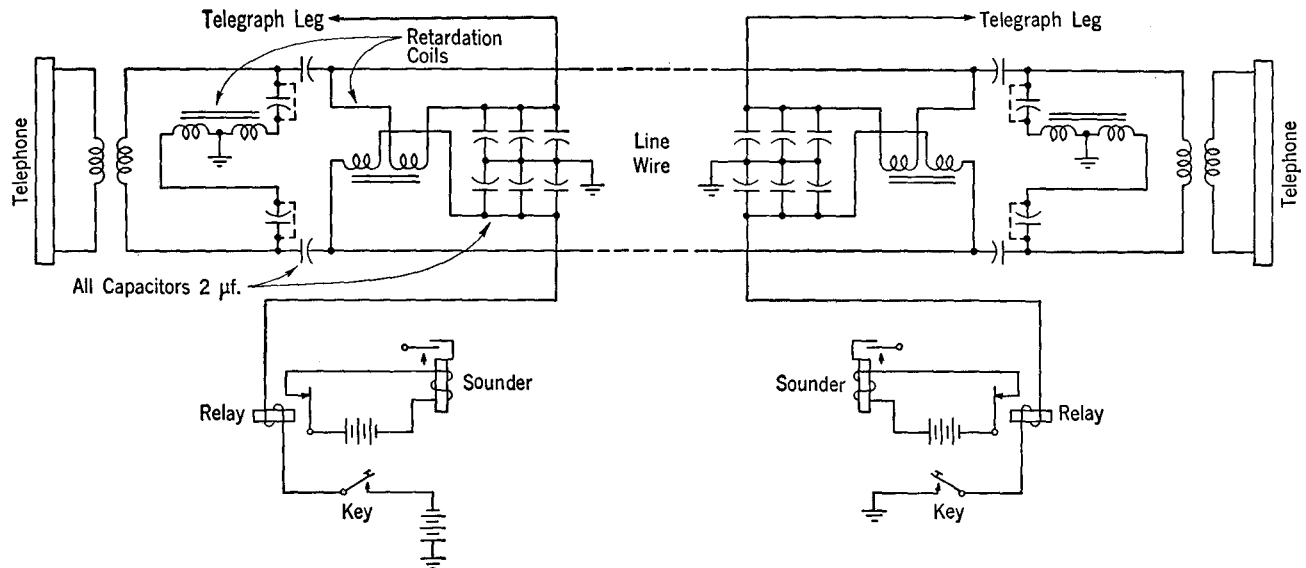


FIG. 11-2 TELEGRAPH CIRCUIT ON COMPOSITED TELEPHONE CIRCUIT

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  $2\mu\text{f}$  capacitors 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 capacitors 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  $2\mu\text{f}$  capacitor in series with the telephone drop, which prevents the direct telegraph currents from reaching the telephone equipment. In some cases, there is a bridge across the telephone circuit on the drop side of the series capacitors to provide a path for any unbalance currents to "leak" to ground. This tends to prevent "crossfire", a condition where telegraph sig-

nals 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. Where, as is normally the case, the two wires shown in the Figure make up one-half of a phantom circuit and a repeating coil is used on the phantom itself, the two  $2\mu\text{f}$  capacitors in the crossfire branch are unnecessary and are strapped out as shown by the dotted lines.

## 11.2 The Elementary Telegraph Circuit

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. The circuit's 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 the 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 11-3 illustrates such a simple 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,



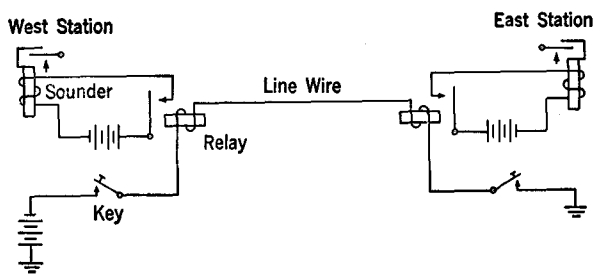


FIG. 11-3 ELEMENTARY TELEGRAPH CIRCUIT

current flows through 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 longer, 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.

A telegraph circuit of this kind is limited to comparatively short distances. The amount of current that could be sent over the usual types of long circuits with this arrangement might not be sufficient to operate the receiving relays; or the signal distortion caused by such a long line might introduce errors. It is necessary, therefore, to break long circuits into links of shorter circuits, with each link relaying the signals into the next adjacent one. Figure 11-4 represents a telegraph circuit similar to that shown in Figure 11-3 but with an intermediate relay at a central point in the circuit. This will permit the West Station to

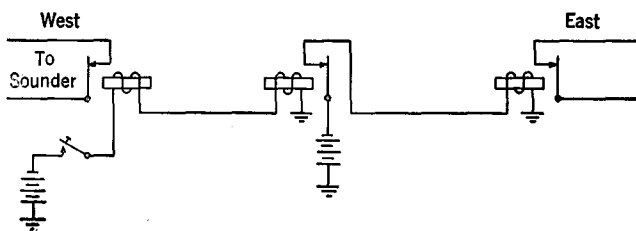


FIGURE 11-4

send to a 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 11-5. 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 tele-

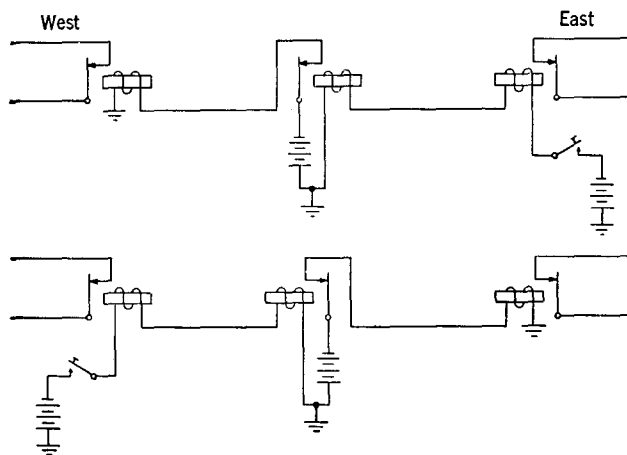


FIGURE 11-5

graph subscriber's standpoint, the class of service which it would provide would be entirely different from that given by the circuit of Figure 11-3 and might or might not be preferable to it. With the two-circuit arrangement, it would 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 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 mes-

\* 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... .

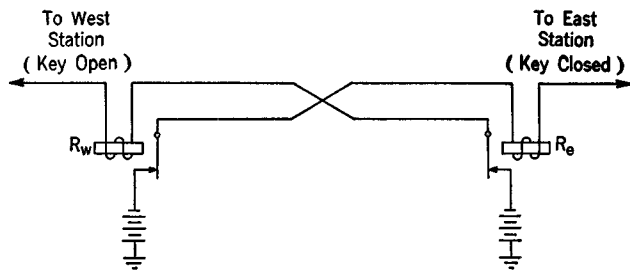


FIGURE 11-6

sages in the manner of a conversation rather than for simultaneous transmission in both directions.

### 11.3 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 11-5 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 into a single circuit as shown in Figure 11-6, 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,

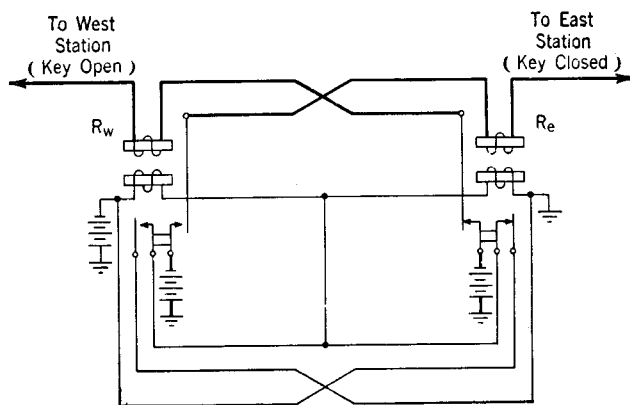


FIG. 11-7 SIMPLIFIED SINGLE LINE REPEATER CIRCUIT

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 to overcome this condition **an additional coil is added to each relay, so wired that its own armature will be held closed while the armature of the other relay is released, regardless of whether the circuit through its own main winding is open.**

These coils are called "holding coils" and Figure 11-7 represents the same connections as shown in Figure 11-6 but with the additional holding coil

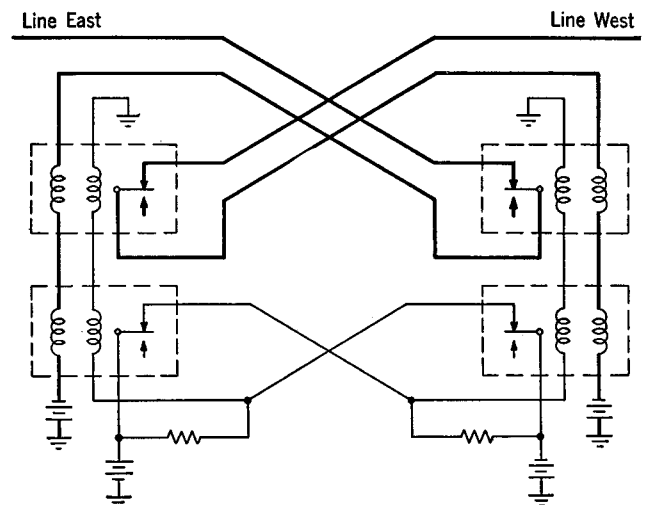


FIG. 11-8 SINGLE LINE REPEATER WITH POLAR RELAYS

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 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

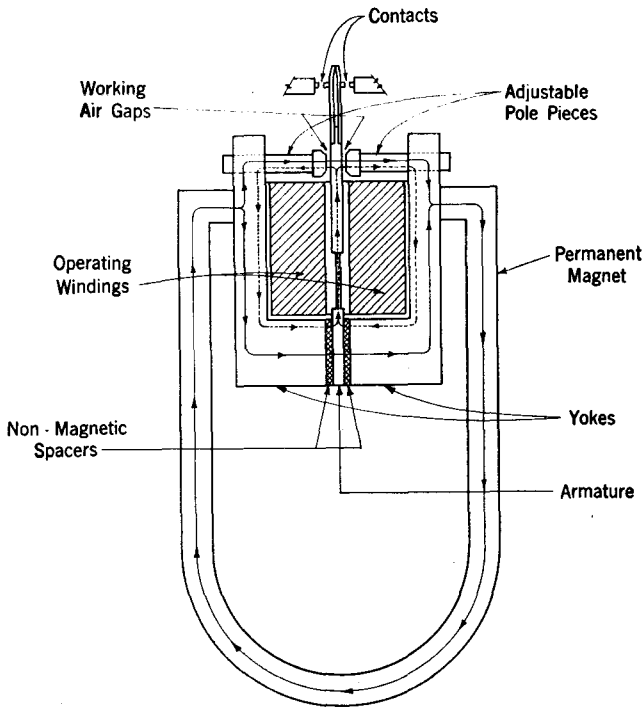


FIG. 11-9 MAGNETIC CIRCUIT OF THE DIFFERENTIAL POLAR RELAY

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 further development of the single-line telegraph repeater is shown in Figure 11-8. The general principle of operation is not different from the repeater already discussed except that instead of "neutral" relays it employs "polar" relays. It will be better understood after studying the discussion of how relays of this latter type are used in differential duplex systems.

#### 11.4 Differential Duplex Systems

The principle of the differential relay depends upon winding a magnetic core with two equal but opposite windings; when 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, i.e., made to operate in one direction or the other depending upon the direction of the current flow in its winding. With such relays, telegraph systems may be operated on the basis of marking and spacing signals being represented by current flow in opposite directions. This is in counter-distinction to the system previously discussed where current flow always represents a marking signal and the absence of current flow represents a spacing signal. As is pointed out in a later Chapter, the use of current flow for both types of signal has important transmission advantages in long telegraph circuits.

The magnetic circuit of a typical differential polar relay is illustrated in Figure 11-9. The manner in which such relays are employed to repeat signals in differential duplex telegraph operation may be best understood by referring to Figure 11-10, which is a simplified schematic drawing of a terminal differential duplex set arranged for full-duplex service. The differential polar receiving relay used in this repeater has its two windings on one side connected respectively to the real line and to a balancing network sometimes called an artificial line. The other ends of the windings are connected together to the armature of the sending relay. When the artificial 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

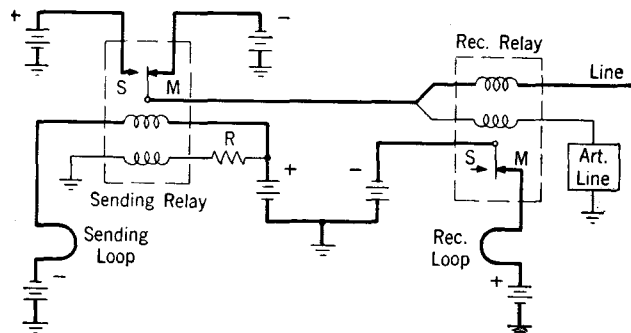


FIG. 11-10 TERMINAL DIFFERENTIAL DUPLEX SET ARRANGED FOR FULL DUPLEX SERVICE

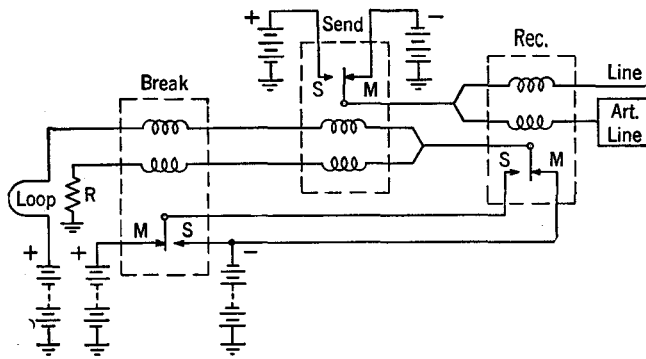


FIG. 11-11 TERMINAL DIFFERENTIAL DUPLEX SET ARRANGED FOR HALF DUPLEX SERVICE

other of its contacts, depending upon the polarity of the incoming current. Thus, 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 but in practice a differential polar relay is used for this purpose also.

Referring again to Figure 11-10, 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 smaller applied voltage, together with the resistance  $R$ , however, limits the current in the biasing winding to a substantially smaller value than that in the operating winding and the armature is 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.

Figure 11-11 shows the arrangement for differential “half-duplex” operation, which provides for the transmission of signals alternately in two directions over the same circuit (as contrasted with simultaneous transmission in full-duplex). This method of operation naturally only requires one loop to the subscriber. But the circuits must be so arranged that the operator at either end, when receiving, may interrupt or break the operator sending at the other end. In order to both send and receive on the same loop, the circuit is designed with the armature of the receiving relay connected to the apex of the sending relay windings and an artificial line or network  $R$  is added to balance the electrical characteristics of the loop.

An additional relay, known as the break relay, is also added.

The sending operation with this circuit arrangement is no different from that in the full-duplex operation of Figure 11-10. When receiving, the operation of the receiving relay armature impresses alternate positive and negative signals on the subscribers’ loop, provided that the break relay armature remains in the marking position. An analysis of the current flow in the winding of the sending and break relays will show that the armatures of both these relays will remain on the marking contacts as long as the loop is closed. Thus when the receiving armature is in the marking position, the positive battery of the loop is reinforced by the negative battery connected to the marking contact of the receiving relay, and a relatively large current flows in the operating windings of the break and sending relays. At the same time, a current of about half this value is flowing in the same direction through the biasing windings of the two relays. Because they are differentially connected, this will oppose the effect of the current in the operating windings but, since the latter is much larger, it will be controlling and the relay armatures will be held in the marking position. When the receiving armature moves to the spacing contact, positive battery at the marking contact of the break relay counteracts the marking battery at the loop and practically no current flows in the operating windings of the break and sending relays. The current through the biasing windings of these relays is now in the opposite direction to that which flowed when the receiving relay was on the marking contact. The biasing windings will therefore still hold the break and sending relay armatures on their marking contacts.

The purpose of the break relay is to assure a rapid and positive break when the loop is opened. If it were not included (and the positive and negative batteries were connected directly to the marking and spacing contacts of the receiving relay instead of as shown in the drawing) the operation of the receiving relay when the loop was opened would cause the sending relay to transmit inverted signals back over the line. This would not result in the most effective breaking action. With the break relay in the circuit, opening the loop will cause its armature to move to the spacing contact when the receiving relay is marking because there will then be no current in the operating winding. This will result in marking battery being connected to both contacts of the re-

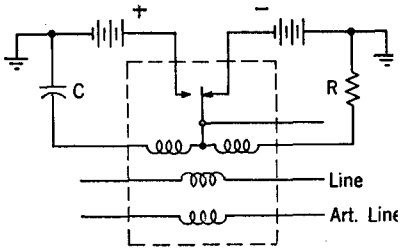


FIG. 11-12 VIBRATING CIRCUIT OF DIFFERENTIAL RELAY

ceiving relay, which compels the latter to transmit a marking signal to the apex of the loop circuit. The resultant current in the biasing winding of the sending relay causes its armature to move to the spacing contact which sends a continuous spacing signal over the line to stop the sender at the distant end.

### 11.5 Principle of the Vibrating Circuit

Polar relays are sometimes equipped with a special third winding which is designed to form a part of a **vibrating circuit** that adds to their sensitivity and rapidity of response. The principle of the vibrating circuit may be understood by referring to Figure 11-12, which is a schematic drawing of one of several possible types, all of which have the same purpose. In this arrangement, the mid-point of the third winding is connected to the armature while the two ends of the windings are connected to ground through a capacitor,  $C$ , and a resistor,  $R$ , respectively.

For the purpose of analyzing the circuit, it may be assumed that there is no current in either of the two main relay windings and that at the instant shown in the diagram the capacitor,  $C$ , is charged to its full capacity. Current is then flowing through the resistor,  $R$ , 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 capacitor 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 capacitor in the opposite direction. This current flows in the same direction in this 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 capacitor 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 resistor. The armature then pulls away from the left contact; the current flowing from the capacitor, 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 capacitor, thus holding the armature solidly against the contact. When the capacitor is charged, the cycle is completed and it continues to repeat itself indefinitely.

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

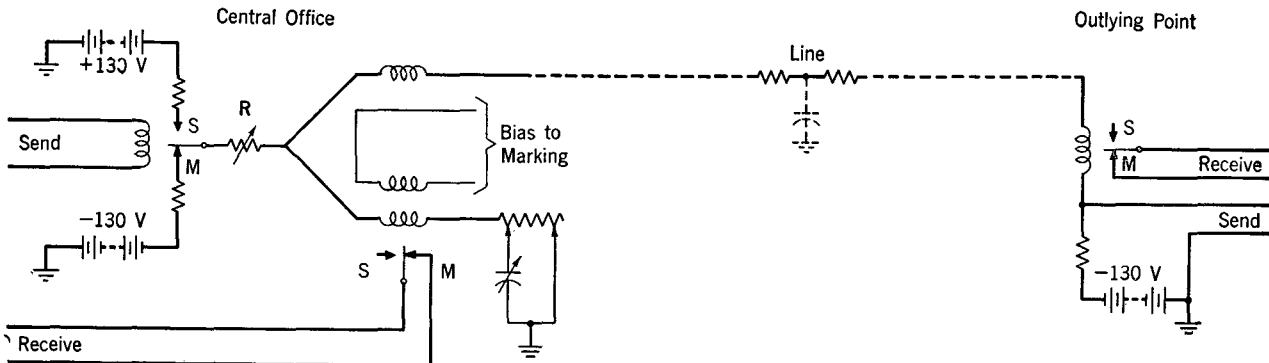


FIG. 11-13 TYPE A POLARENTIAL TELEGRAPH CIRCUIT

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 present.

### 11.6 Polarential and One-Way Polar Systems

For furnishing service to subscribers at outlying points, two special types of grounded telegraph systems are frequently used. These are known respectively as "One-way Polar" and "Polarential, types A and B".

The polarential system permits true polar operation from the central office out and differential operation from the outlying point into the central office. Thus the advantages of polar transmission are secured and at the same time only one battery is required in the subscriber's office. A schematic diagram of one arrangement of a Type "A" polarential system is shown in Figure 11-13, with only the essential elements included for the sake of simplicity.

By inspection of this diagram it may be seen that when the outlying sending loop is closed, 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 wind-

ing. 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 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 battery at the outlying point has more effect at the apex of the repeater circuit than does the battery 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.

Because transmission from the outlying point is not polar, changes in line resistance will cause some distortion of the telegraph signals unless the repeater is lined up in a particular manner. By adjusting the values of the resistance of the artificial line and apex branches of the home repeater so that the line currents are made equal and opposite for marking and spacing, transmission from the outlying point will be made free from distortion due to line resistance variations. This is due to the design characteristics of the repeater circuit which cause the marking and spacing effects on the home receiving relay to be reduced equally when the line resistance is increased and to be increased equally when the line resistance is reduced. The Type "A" polarential system is accordingly most useful in circuits where line resistance variations are a controlling factor.

Figure 11-14 shows a diagram for Type "B"

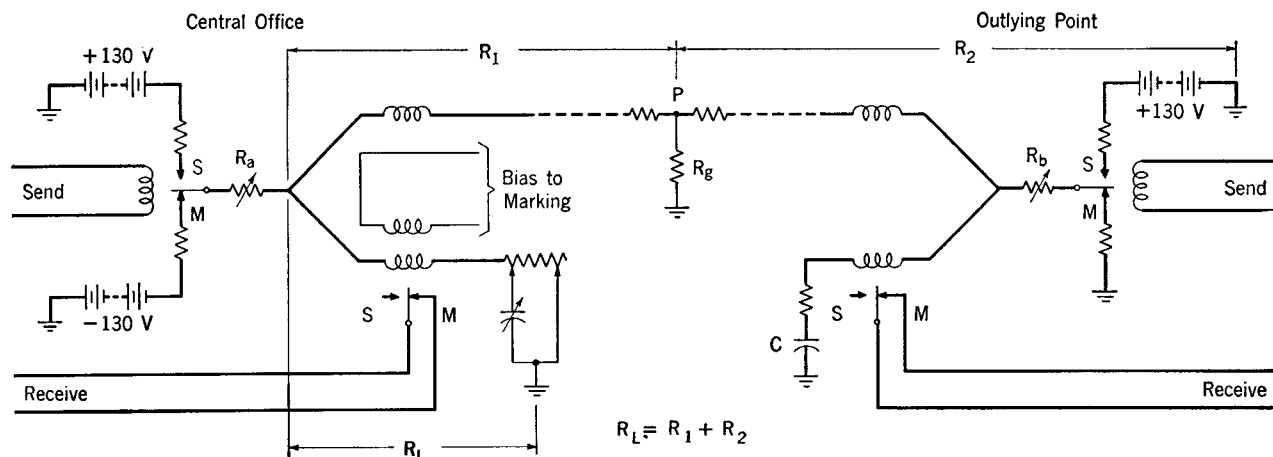


FIG. 11-14 TYPE B POLARENTIAL TELEGRAPH CIRCUIT

polarential operation. The important advantage of this method of operation is that the circuit is self-compensating to a considerable extent for line leakage. It is therefore frequently used on open-wire circuits where weather conditions cause considerable variation in the amount of leakage.

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 Type "A" 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,  $E/R_L$ , 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  $E/2R_L$ . 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:

$$R_a = \frac{R_L (2R_2 - R_1)}{2R_1 + R_2} \quad (11:1)$$

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,

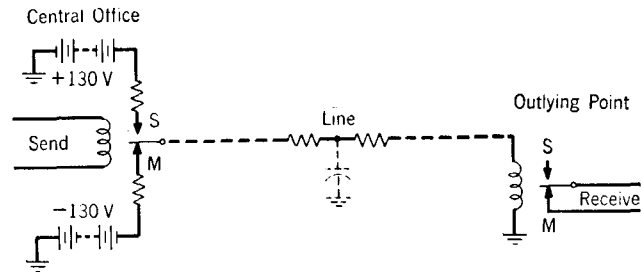


FIG. 11-15 ONE-WAY POLAR TELEGRAPH CIRCUIT

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 point to "kick off" and mutilate the home copy or produce false breaks. To neutralize the effect of this "kick", the bridge arm containing the capacitor  $C$  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.

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. As indicated in Figure 11-15, this circuit arrangement also gives the advantage of polar operation.

### 11.7 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

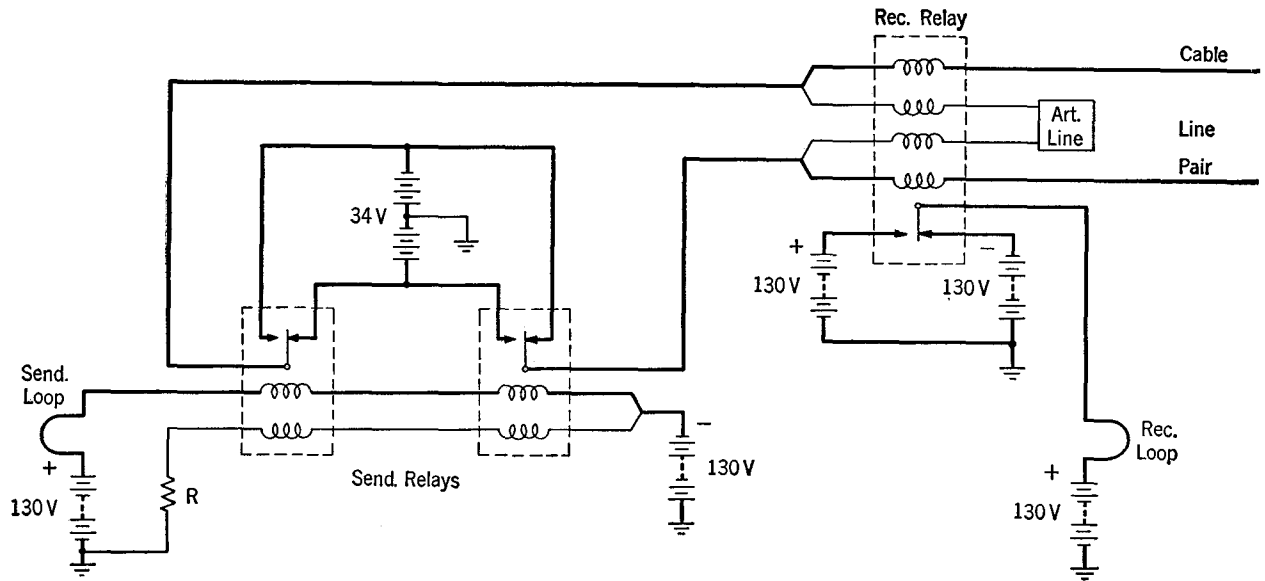


FIG. 11-16 TERMINAL METALLIC DUPLEX SET ARRANGED FOR FULL DUPLEX SERVICE

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. Figure 11-16 is a schematic drawing of a two-wire 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

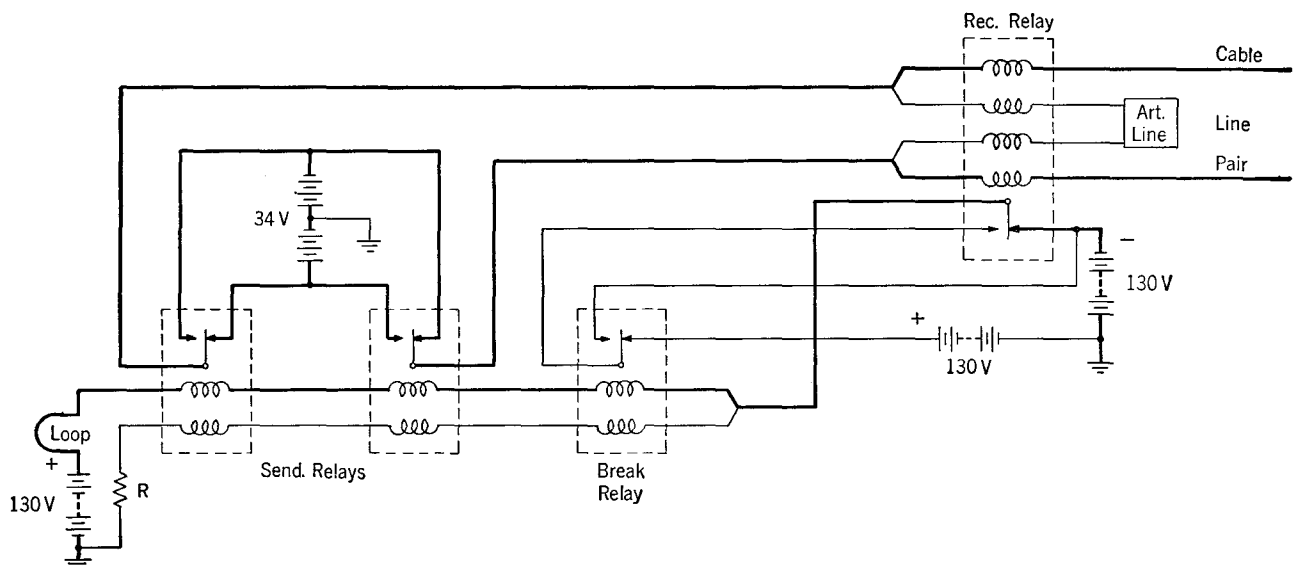


FIG. 11-17 TERMINAL METALLIC DUPLEX SET ARRANGED FOR HALF DUPLEX SERVICE



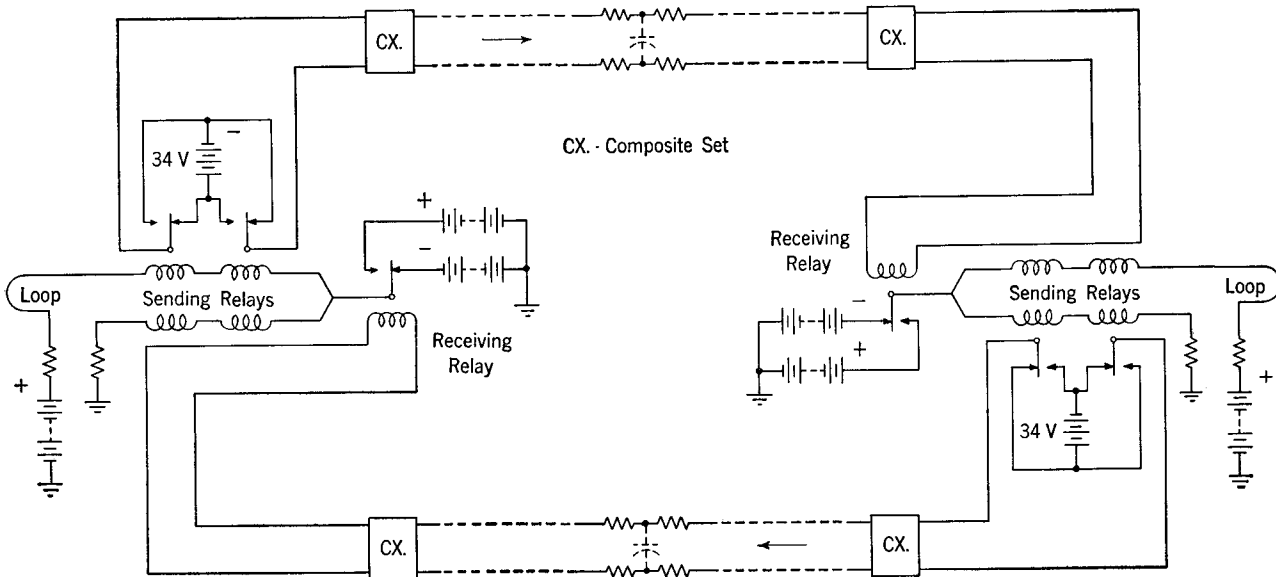


FIG. 11-18 FOUR-WIRE METALLIC TELEGRAPH CIRCUIT

19-gage conductors, is in the order of 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 biasing windings are energized and the armatures are accordingly operated to spacing.

In half-duplex operation, the addition of a break relay, as shown in Figure 11-17, facilitates 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. 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.

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 11-18, which is a schematic of a 4-wire metallic telegraph circuit between two terminal type repeaters, the sending and receiving paths of transmission over the line are separated from each other. 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 transmission 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 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 compositing throughout their entire length because of low-frequency "delay distortion" introduced by the composite sets.

## CHAPTER 12

### TELEGRAPH CIRCUITS—(Continued)

#### 12.1 Principles of the Teletypewriter

In discussing telegraph circuits in the preceding Chapter, we have tacitly assumed that the signals to be transmitted were produced manually by hand operation of an ordinary telegraph key. As a matter of fact, the great majority of telegraph circuits now in use 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 mechanism designed to type 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 some 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 12-1, 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 corresponding to the code for that letter are sent out on the line; and when the 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 12-2. 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

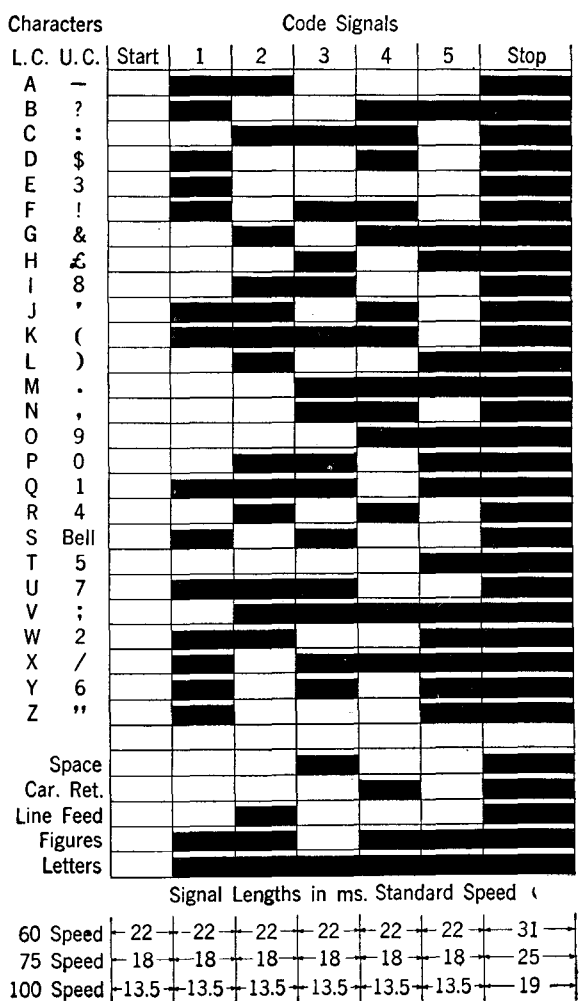


FIG. 12-1 TELETYPEWRITER CODE

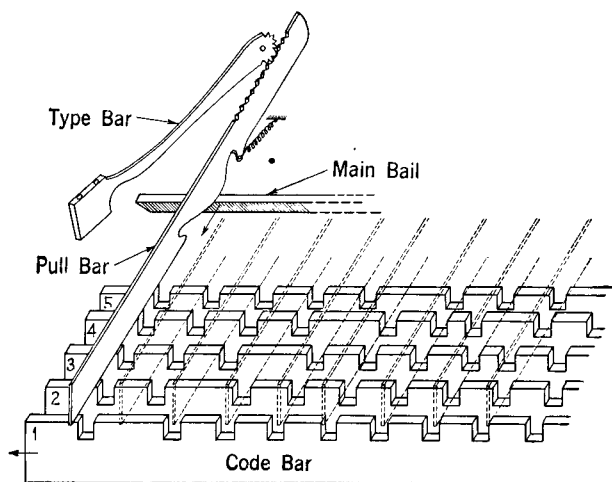


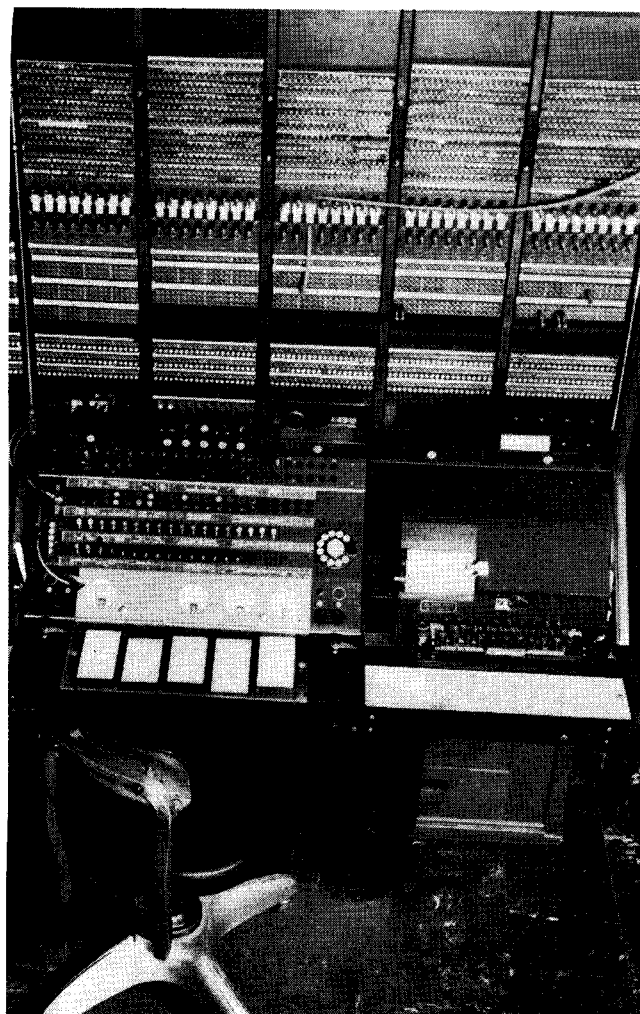
FIG. 12-2 SELECTING MECHANISM OF RECEIVING MACHINE

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.

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 selecting pulses comprising the signal for one character. This insures that the two machines will be in exact synchronism at the beginning of transmission of each

character; in other words, it corrects so frequently any speed differences that may exist, 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 permanently connected to the line and to the electrical elements of the sending and receiving machines. Temporary connections between segments are then made periodically and 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



TELEGRAPH SERVICE BOARD POSITION

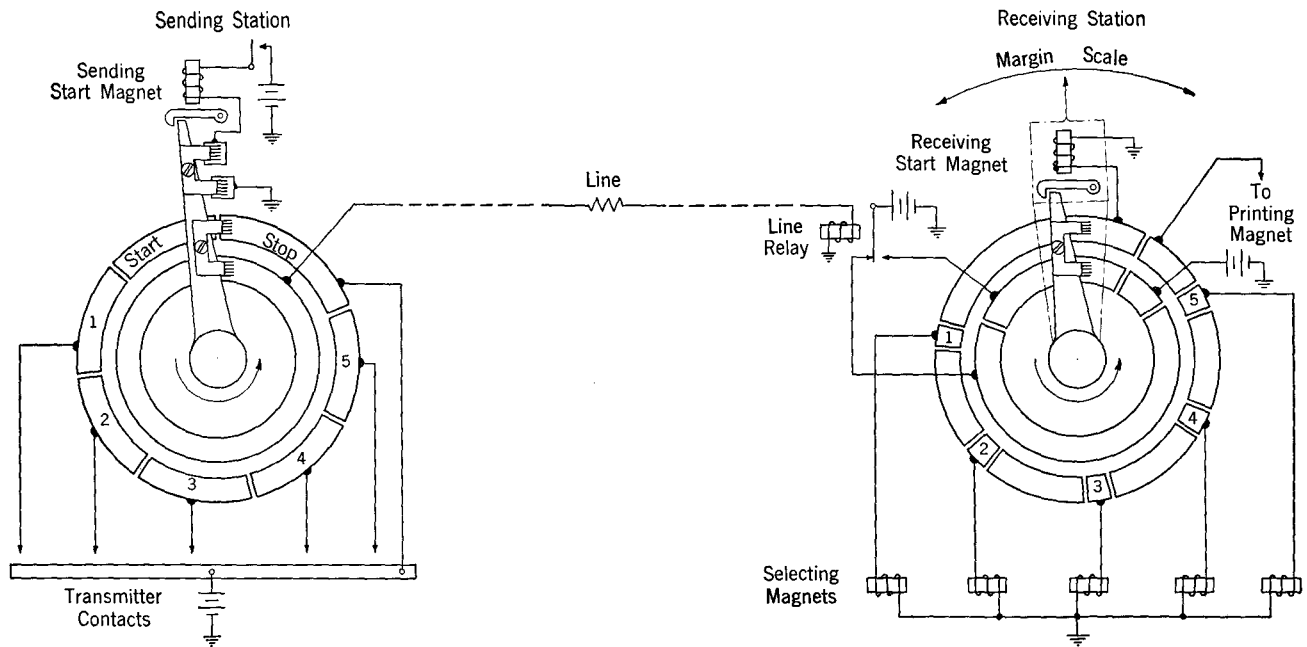


FIG. 12-3 THEORY OF START-STOP TELETYPEWRITER SYSTEM

pair of these commutator devices, known as **distributors**, is given in Figure 12-3.

To follow the operation, let us assume that the letter, D, is to be transmitted. By referring to Figure 12-1, 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 stop pulse 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 winding, 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 accor-

dance 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 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

stop 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.

Machines very similar to the teletypewriter, known as teletypesetters, are employed for automatically operating linotype or other typesetting machines at a distant point. Because the number of characters and "stunts" required for such a service is greater than is used for ordinary teletypewriter service, it is necessary to employ a six-unit signal code in the teletypesetter. This makes available a maximum of 64 signal combinations as compared with a maximum of 32 in the five-unit code of the teletypewriter. The principle of operation of these machines, however, is the same as that of the teletypewriter just described.

## 12.2 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. It was replaced by a device which was quite different mechanically but employed 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, this machine employs spring contacts which 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 cause 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.

The several selecting magnets employed in the receiving equipment of the teletypewriter mechanism discussed immediately above were replaced in later 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 12-4. 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 start 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 with the first code bar and rotates it slightly in a

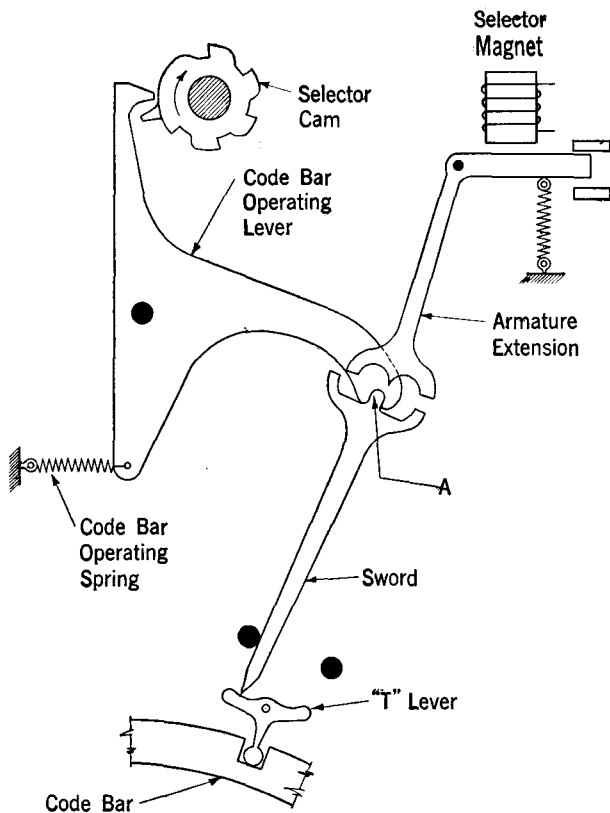


FIG. 12-4 RECEIVING MECHANISM OF SINGLE MAGNET TELETYPEWRITER

counter-clockwise direction. The effect of this movement depends upon whether or not the selector 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 selector 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 two standard operating speeds, namely, 60 and 75 words per minute. Service at a speed of 100 words per minute also may be furnished with certain limitations. 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 designed with the maximum possible operating margin. By referring to Figure 12-3, 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 12-5 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.

### 12.3 The Regenerative Repeater

The high speed operation commonly used with teletypewriters is possible only if the total distortion of the received signals is held within the limits outlined above. Since telegraph signals are invariably distorted to a greater or less extent in the process of transmission and since 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 a **regenerative repeater**. The primary function of such a repeater is to retime and retransmit received signal impulses. It is

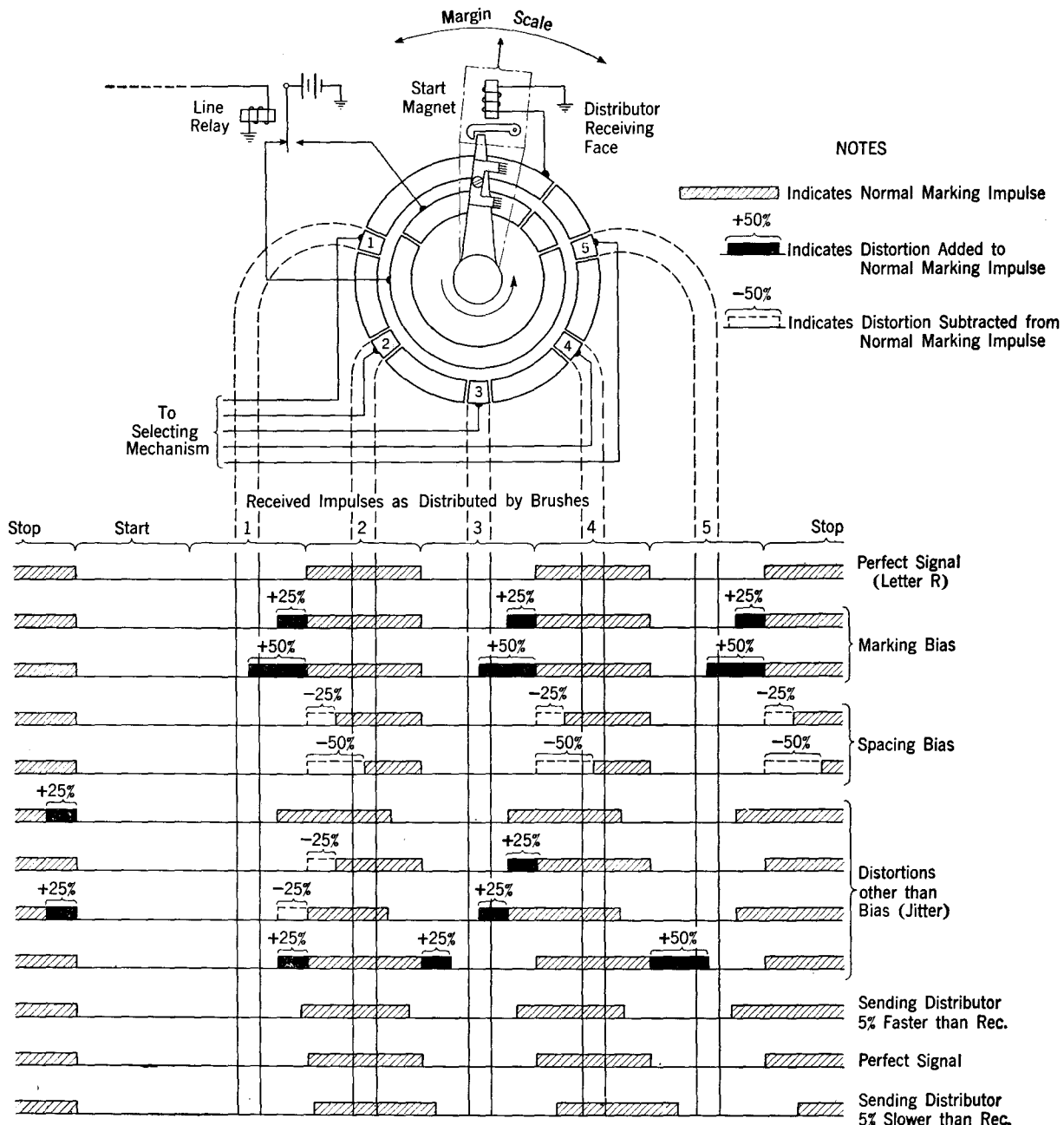
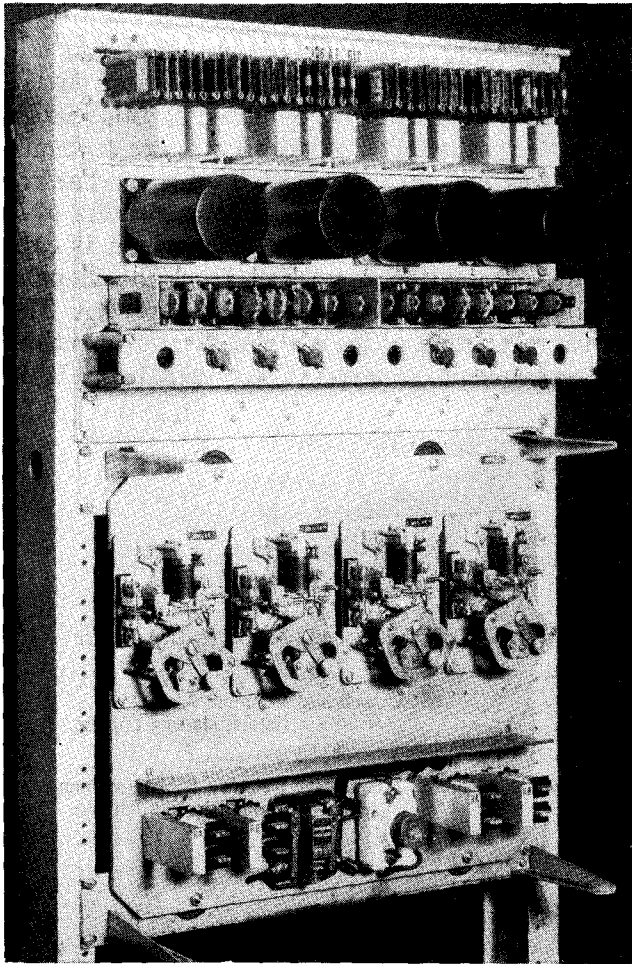


FIG. 12-5 TIME RELATIONSHIPS OF RECEIVED SIGNALS



CAM-TYPE REGENERATIVE TELEGRAPH REPEATERS

capable of receiving without error any set of signals that would be satisfactorily received by an ordinary teletypewriter, and of sending these same signals out in exactly the same form as the signals produced by the sending teletypewriter.

Figure 12-6 shows schematically the simplified arrangement of an early 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 12-1, 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 released (through a mechanism not shown in

the drawing) and the 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 capacitor 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 sending brush moves on to No. 1 segment of the sending face and the positively charged right storing capacitor 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 capacitor negatively in accordance with the incoming marking signal. This capacitor is in turn discharged through No. 2 segment of the sending face while the right capacitor 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.

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

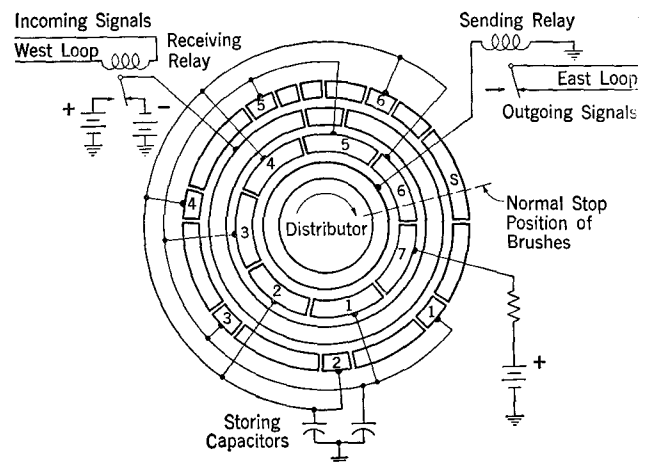


FIG. 12-6 PRINCIPLE OF THE REGENERATIVE REPEATER



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.

A later type of regenerative repeater is illustrated in the adjacent photograph. It operates on the same principle as the distributor arrangement first discussed but like the modern teletypewriter employs rotating cams instead of rotating brush arms to simplify its mechanical construction. The basic mechanism is shown schematically in Figure 12-7. Here the essentials are a receiving magnet, the armature of which is associated with a reed which operates sending contacts in the outgoing telegraph circuit. The position of this armature at any given time is under the joint control of the receiving magnet and two rotating cams which are driven through a clutch at a constant speed by a small motor.

When a start pulse is received the cams start to rotate through the action of a relay not shown in the drawing. The armature cam then moves the armature gradually toward the pole face of the receiving magnet, which is of a type known as a holding magnet. Its magnetic force when current is flowing in its windings is sufficient to prevent release of the armature once it is presented to the pole face, but is not powerful enough

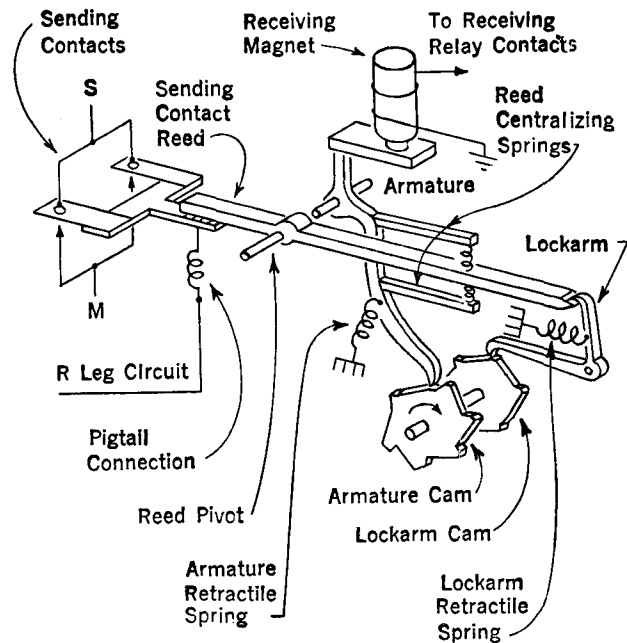
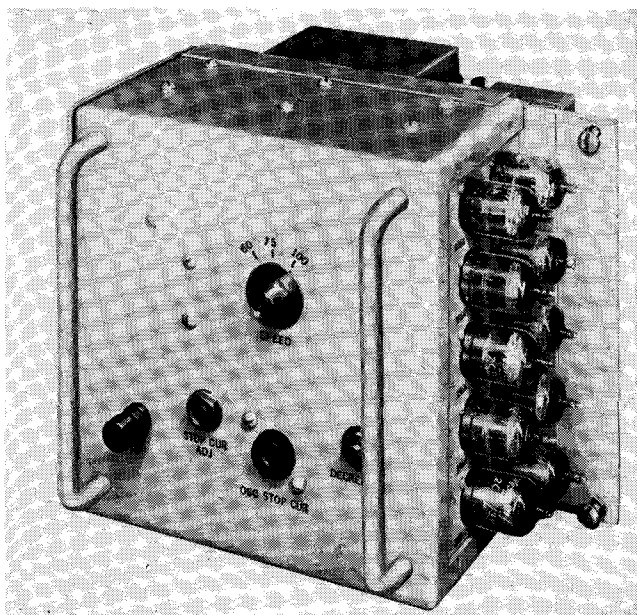


FIGURE 12-7

to attract it from a released position. The armature cam is so designed that the armature is first presented to the pole face of the magnet slightly before the middle of the first signal pulse reaches its windings. At this time the lockarm cam reaches a position in its rotation where it will release the lockarm for a very brief period during which time the armature cam also moves forward so that the position of the armature is controlled only by the receiving magnet and the armature retractile spring. If the signal pulse in the receiving magnet is marking, the armature will remain operated and will move the sending contacts to the marking position. If the received pulse is spacing the armature releases under the influence of the retractile spring and causes the reed to move the sending contacts to the spacing position. In either event, the lockarm will be restored very quickly to the locked position thus holding the reed in whichever position it has taken. Thus the length of the "retransmitted signal" is timed entirely by the lockarm cam, which operates for a very short time at regular intervals during the reception of the middle of each signal pulse. The length of the retransmitted signal pulses are therefore independent of any distortion that may be present in the received pulses.

Other designs of regenerative telegraph repeaters now in general use obtain the same results as discussed above through the employment of



ELECTRONIC REGENERATIVE TELEGRAPH REPEATER

various electronic circuits and devices, some of which are discussed in later Chapters.

#### 12.4 Teletypewriter Switching Systems

Telegraph and teletypewriter circuits in practical use may be set up simply as a single connection between two fixed points or as a network of circuit connections that may reach simultaneously into many points. Service requirements may also make it desirable to change the particular circuit connections at frequent intervals. The Bell System provides a Teletypewriter Exchange Service (TWX) similar to ordinary telephone message service. Switching in this case is handled by operators at manual switchboards in the telephone central offices. The switchboards and other apparatus used are essentially similar to those used in manual long distance telephone switching except that exchange of information between subscribers and operators is effected by teletype rather than by voice.

In the case of private line telegraph or teletype networks, switching operations may be performed by either manual or automatic methods. For use with some of the more elaborate teletypewriter networks involving several lines and a multiplicity of stations, a completely automatic switching system may be used. Such systems employ crossbar switches and associated relay apparatus somewhat comparable to that used in telephone toll line dialing. All lines terminate at one or both ends in a switching office, of which there may be several in a system. These offices consist not only of

the switching equipment to guide the messages to their destinations but include machines for handling messages to and from the office, for storing messages briefly while the switching operations are taking place, and for general supervision and control of traffic flow on the network. Such switching offices can serve a maximum of 200 individually coded stations.

The entire system works in such a manner that messages may be sent from any sending station to any receiving station or group of stations without requiring manual handling other than the original perforation of tape at the sending station and the removal of the typed message at the receiving station or stations. The switching operations are under the control of directing characters or codes punched into the original transmitting tape at the head of each message.

Messages from the various sending stations on each multipoint line are under the control of the switching office, where they are recorded in perforated tape, then transmitted through crossbar switching mechanisms to machines associated with outgoing line or trunk circuits, which record them again in perforated tape for retransmission on the required circuit when available. Each receiving station is assigned an individual two-letter code which is used as a directing code preceding single-address messages intended for a particular station. Messages sent to a group of receiving stations are preceded by a common multi-address code and may then include as many single station codes as desired, or a group code, or any combination of these.

## CHAPTER 13

### TELEGRAPH TRANSMISSION PRINCIPLES

#### 13.1 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 properties of these circuits and apparatus are such that certain characteristics of the transmitted signals may be altered somewhat in the process of transmission over the line.

As previously indicated two general methods of transmitting telegraph signals are commonly used in land-line practice—(1) neutral or open and close transmission in which current is sent over the line to operate the relays to the marking position and the current is stopped or cut off 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 opposite directions of current flow rather than to the open and close condition. 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.

#### 13.2 Wave Shapes in Neutral Telegraph Systems

In any neutral telegraph circuit, if we could ignore the time 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 13-1(A). As a practical matter, however, every telegraph circuit has some series inductance. 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 13-1(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 capacitors of the composite set, we have a further sloping of the pulse as shown by Figure 13-1(C). Here the shaded portion represents the

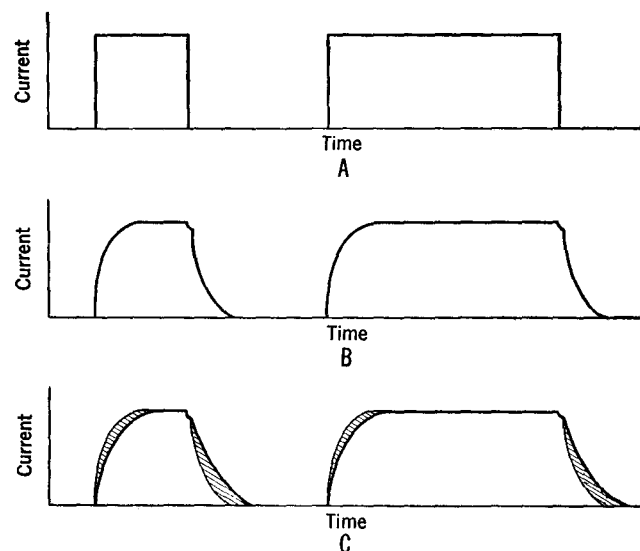


FIGURE 13-1

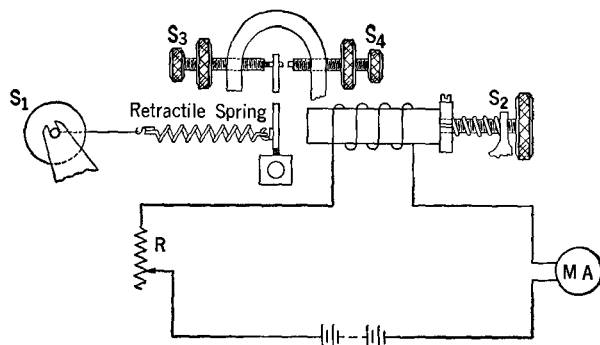


FIGURE 13-2

effect of the capacitors 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 capacitors 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 capacitors 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 13-2 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 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 armature releases that did not exist before the relay was operated.

We may represent by the points  $O$  and  $R$  in Figure 13-3(A) the operating and release current values respectively for a relay like that illustrated in Figure 13-2. 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 retractile spring with the screw  $S_1$ , lessening the air gap between the pole piece 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 13-3(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 13-1(A). Here it is evident that we could neither increase nor decrease the length of the signal by relay adjustments.

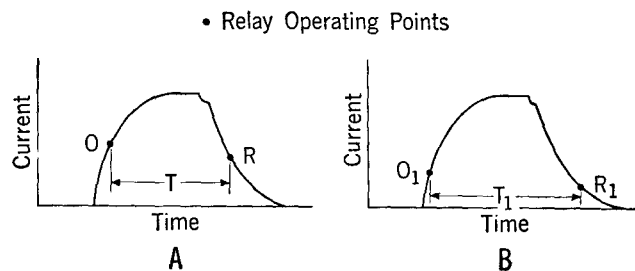


FIG. 13-3 EFFECT OF RELAY ADJUSTMENT ON TELEGRAPH SIGNAL LENGTHS

To a degree this explains the frequent adjustments that are necessary on telegraph apparatus of this general type 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.

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

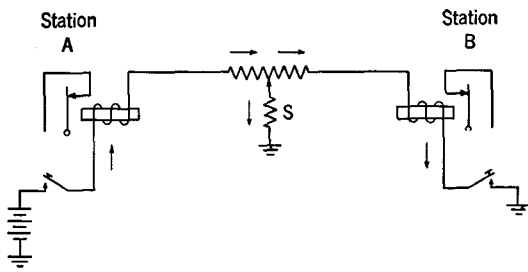


FIGURE 13-4

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 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 13-4 where *S* represents a leak to ground along the line, either distributed or

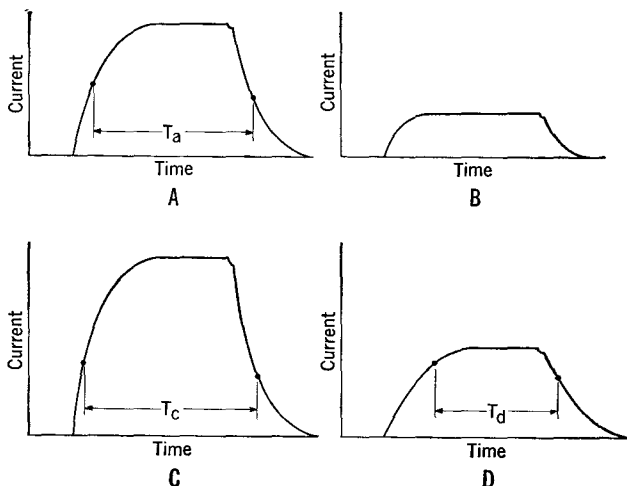


FIGURE 13-5

otherwise. Let us suppose that Figure 13-5(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 13-5(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 13-5(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,

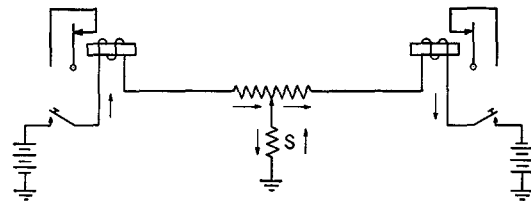


FIGURE 13-6

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 13-6 tend to neutralize each other because they are flowing in opposite directions.

The shunting effect to ground of a leak, such as is shown by Figure 13-4, is a special case. On every telegraph wire, regardless of insulation conditions, we have in effect a leak to ground through the capacitance between the wire and ground, or a condition that might be illustrated by substituting a capacitor for the resistance *S* in Figure 13-4. Since the telegraph current wave shapes are somewhat similar to alternating-current cycles, the capacitor may properly be considered as shunting the current to some extent. Furthermore, this capacitance not only decreases the current value that reaches the distant station but tends further to distort the wave shape, thereby limiting the distance over which satisfactory signals can be sent without additional repeaters.

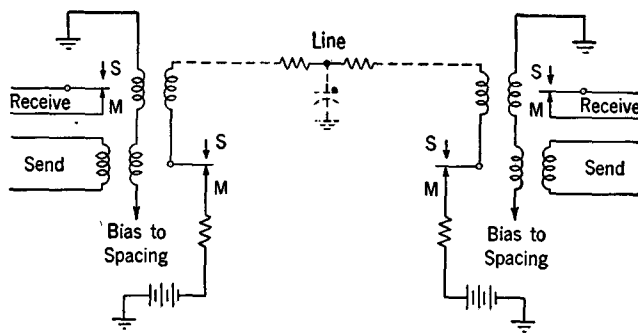


FIG. 13-7 OPEN AND CLOSE TELEGRAPH CIRCUIT EMPLOYING BIASED RELAYS

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 circuit schematically illustrated in Figure 13-7, using electrically biased receiving relays as discussed in Article 11.4. It will be seen from an inspection of this circuit that when the sending key is closed, the capacitance 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 13-8 (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 design and adjustments, we will consider for simplicity that it operates on an effective  $\pm 3$  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.

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.

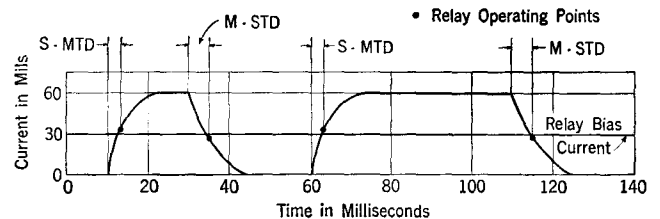


FIG. 13-8 SIGNAL WAVESHAPES IN TELEGRAPH CIRCUIT OF FIG. 13-7

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.

### 13.3 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 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 equation for millisecond bias is:

$$\text{M-STD} - \text{S-MTD} = \text{ms. bias} \quad (13:1)$$

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

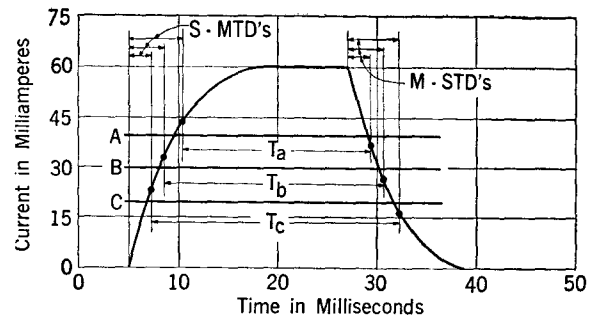


FIG. 13-9 EFFECT OF RELAY BIASING CURRENT ON SIGNAL LENGTH

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 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.

The wave shape of a typical mark signal in a circuit operating with a line current of 60 mils and having capacitance to ground is shown in Figure 13-9. 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 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

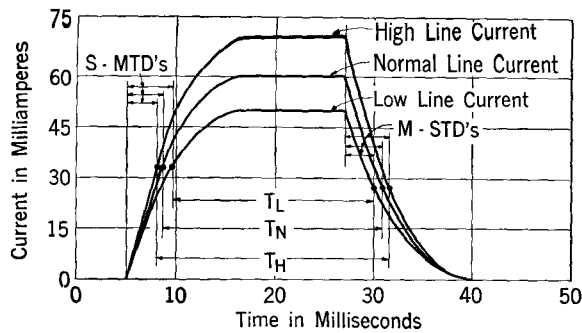


FIG. 13-10 EFFECT OF LINE CURRENT MAGNITUDE ON SIGNAL LENGTH

impulse is shortened. This is illustrated by the wave shape for the "Low Line Current" in Figure 13-10. 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 13-10. In other words, increasing the line current tends to produce marking bias and decreasing it tends to produce spacing bias.

### 13.4 Wave Shapes in Polar Telegraph Systems

A one-way polar circuit using a ground return is shown schematically in Figure 11-15. 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 capacitance between the line and ground. When the line current is marking, the voltage on the capacitor, representing the capacitance between the line and ground, is negative. On the other hand, when the line current is spacing, the voltage on this capacitor is positive. The change of the line current from marking to spacing then involves a change of the voltage on the line capacitance from a negative value down to zero and then up to a positive value. The part of the discharge current

from the capacitor 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 capacitor 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 13-11. 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.

The fact that the M-S and S-M wave shapes are identical in form is a valuable feature of polar

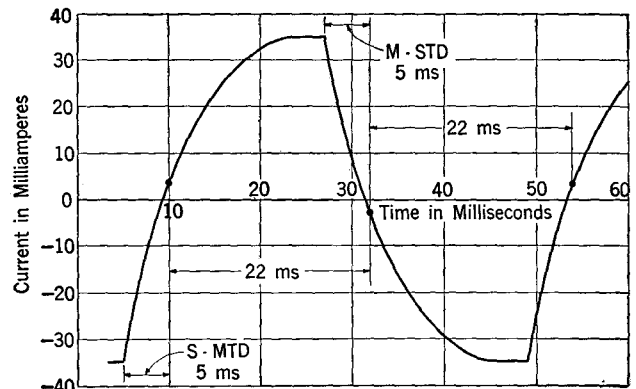


FIG. 13-11 SIGNAL WAVESHAPES IN POLAR TELEGRAPH CIRCUIT

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



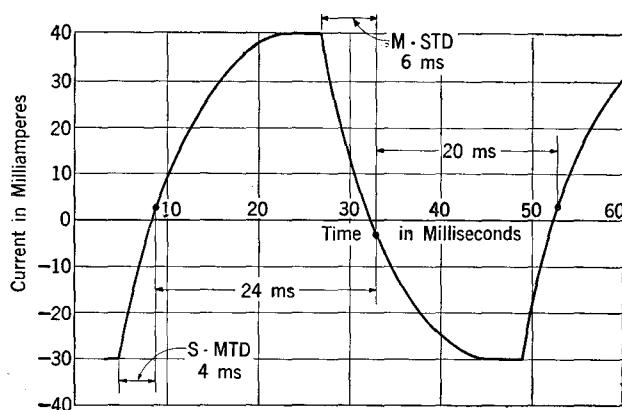


FIG. 13-12 EFFECT ON SIGNAL LENGTHS OF UNEQUAL POLAR LINE CURRENTS

wave shape in order that equal transition delays will be secured.

Figure 13-12 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.

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 14

### TELEGRAPH TRANSMISSION PRINCIPLES—(Continued)

#### 14.1 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 13-11 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 14-1 (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 change 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.

The net effect on the marking impulse being

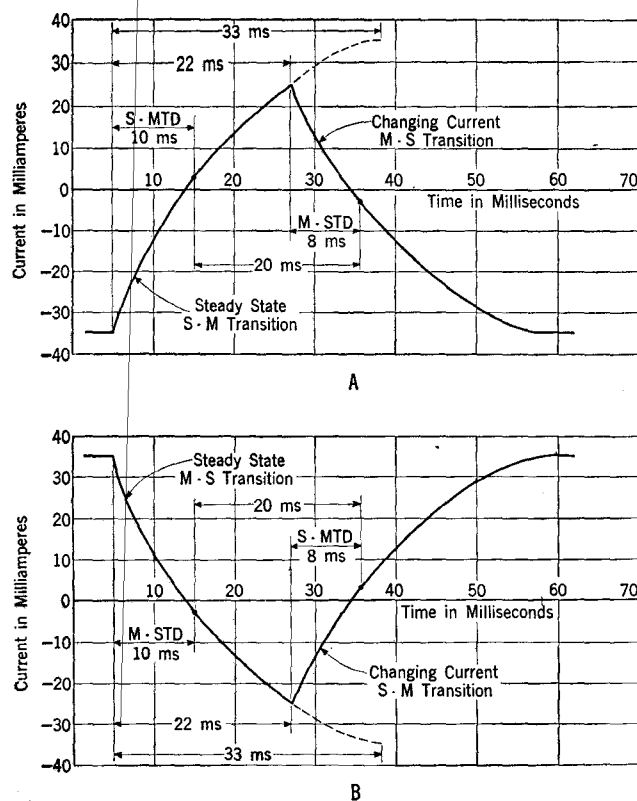


FIGURE 14-1

transmitted in Figure 14-1(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  $+35$  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 also 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 14-1(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 14-1(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 14-2 which shows wave shapes

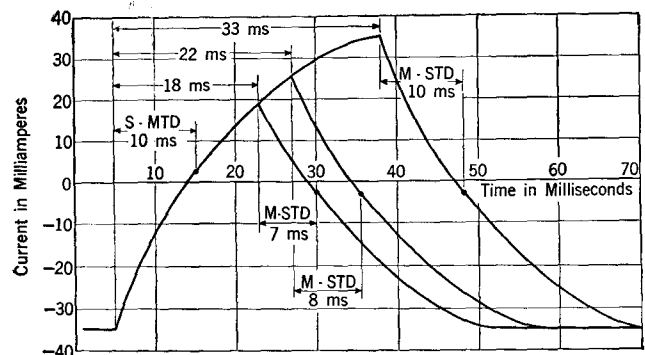


FIG. 14-2 CHARACTERISTIC DISTORTION EFFECTS ON SIGNAL LENGTHS AT DIFFERENT OPERATIONAL SPEEDS

of marking impulses for three different 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, 22 ms. long and 33 ms. long. 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.

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 de-

termines 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 factors of characteristic distortion, **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**

**impulses 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 capacitance, 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 capacitance, 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.

#### 14.2 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".

#### 14.3 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. Where such circuits are operated for teletypewriter service, the teletypewriter itself may be used as a measuring instrument for this purpose within certain limits.

As illustrated by Figure 12-1, 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 60 and 75 words per minute. 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 a speed of 60 words per minute, therefore, there are 60 times 6 or 360 revolutions per minute of the brush arm.

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 14-3(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.

The received signals shown in Figure 14-3(B) have the same time lengths as those produced by the sending distributor of Figure 14-3(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 12-3).

The selecting mechanism, when "oriented" cor-

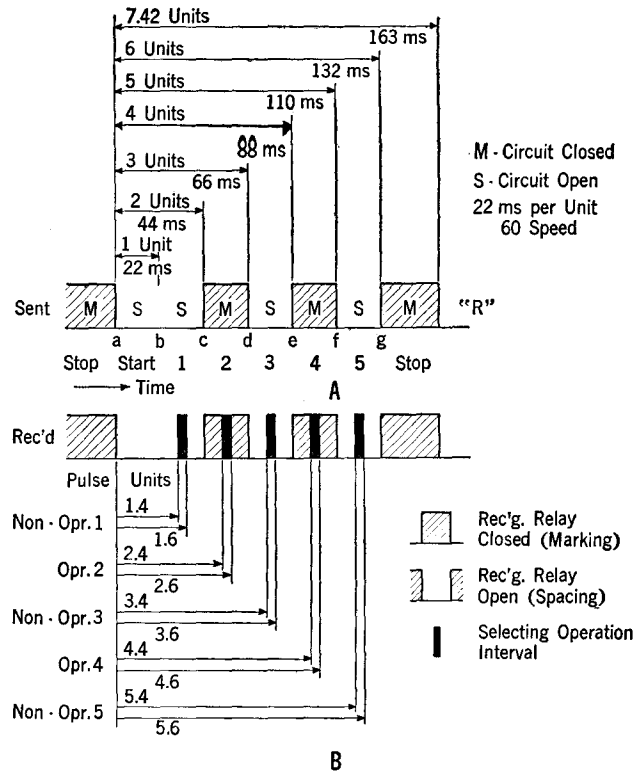


FIGURE 14-3

rectly, 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 14-3(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 14-3(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

As illustrated by Figure 12-1, 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.

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The selecting mechanism, when "oriented" cor-

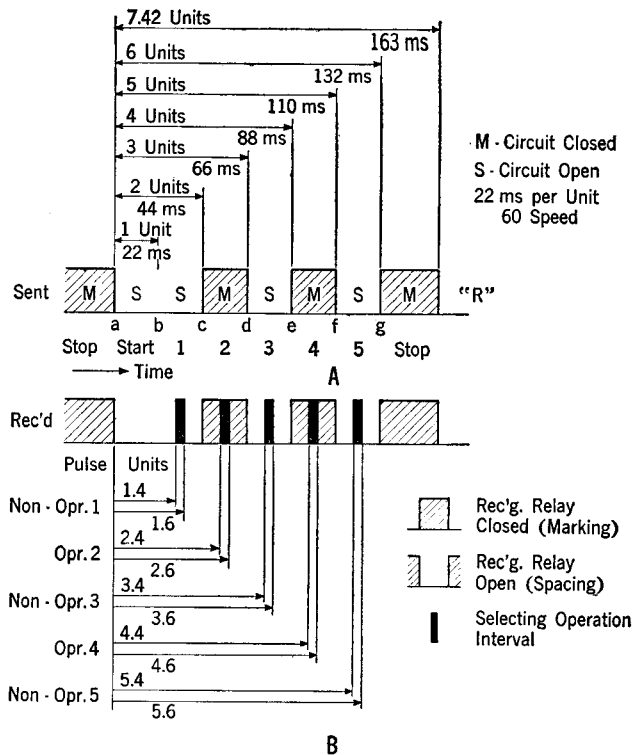


FIGURE 14-3

rectly, 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.

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Inasmuch as the selecting functions take place only during the intervals shown by the solid blocks of Figure 14-3(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.

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 indi-

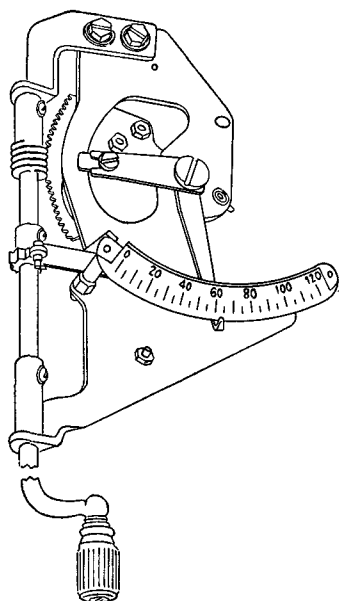


FIG. 14-4 TELETYPEWRITER RANGE FINDER MECHANISM

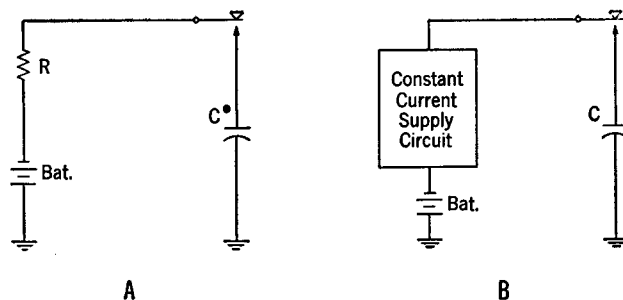


FIGURE 14-5

cated in Figure 14-4. 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.

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.

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. 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.

#### 14.4 Telegraph Transmission Measuring Sets

A standard telegraph transmission measuring set used in the Bell System employs the capacitor charge principle for the measurement of maximum or total distortion, as well as bias, in terms of per cent length of a unit signal interval. This measuring set may be compared to the teletypewriter used as a measuring device. The essential difference is that the teletypewriter measurement depends upon shifting the range finder

mechanism while noting the presence or non-presence of errors in the copy, and the resultant range of shifting is used as a measure of distortion. The measuring set provides direct meter readings of per cent distortion.

In order to consider its underlying operating principle, let us refer to the simple  $RC$  circuit of Figure 14-5 (A) which includes a battery and key. When the key is closed there will be a flow of electricity into the capacitor until its voltage rises to be equal and opposite to that of the battery. The rate at which the capacitor voltage rises depends upon the value of the resistance  $R$ . Thus, a measurement of the value of the capacitor voltage would also give a measure of the time elapsed since the key was closed in accordance with the time constant of the circuit.

In the measuring set, the time-charge relationship is simplified by using an electron tube circuit which has a constant output current to charge the capacitor, so that the voltage increase is the same for each millisecond of time that the capacitor is allowed to charge. This is known as a **constant current supply circuit** and is connected in the circuit as schematically indicated in Figure 14-5 (B).

Figure 14-6 illustrates schematically the principle of a telegraph transmission measuring set. For simplicity of explanation this Figure includes a mechanical distributor as actually used in the earliest designs of this type of device, although modern measuring sets employ electron tube cir-

cuits to accomplish the same result. As shown, the incoming teletypewriter signals to be measured are repeated by the master relay into a simple form of one-way polar circuit, in which are located three other polar relays. The start relay closes the start magnet circuit when the M-S transition beginning a start pulse is received. This releases the brush arm of the distributor which is arranged to rotate at a speed to correspond with the incoming signal. The armatures of the transfer relays are connected to measuring capacitors  $C_1$  and  $C_2$ . On each M-S transition,  $C_1$  is disconnected from its charging current supply and connected to a voltage indicating circuit. Similarly, on each S-M transition  $C_2$  is disconnected from its charging supply and connected to a voltage indicating circuit.

The transfer relays thus serve to interrupt the capacitor charging at the desired time and to immediately transfer the charged capacitors to a circuit which will measure the voltage existing across their terminals. In addition, some arrangement must be used to discharge the capacitors when they are transferred back to the charging circuit, so that they may start charging again at the right time. This is accomplished by the distributor which is divided into fourteen alternate long and short segments, the long segments being three times the length of the short ones. As already noted the speed of the distributor or brush arm must be adjusted to that of the incoming signals. Thus for 60 speed teletype signals it would

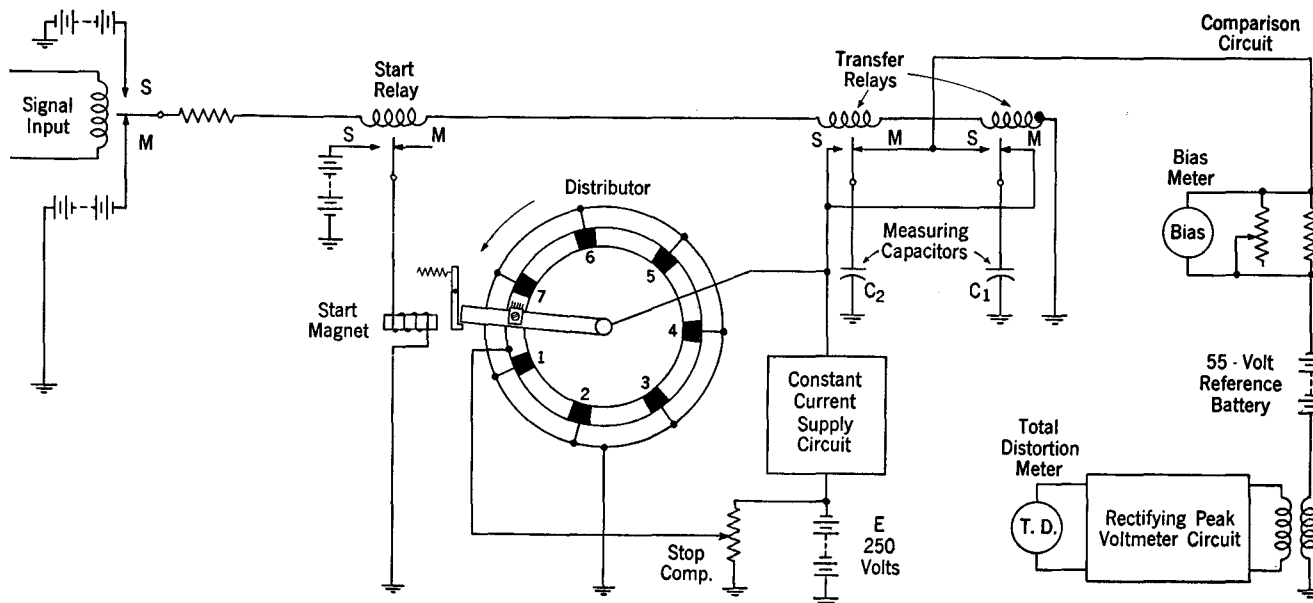


FIG. 14-6 PRINCIPLE OF TELEGRAPH TRANSMISSION MEASURING SET



take 22 milliseconds for the brush arm to pass over a long segment and the adjacent short one. (For different signal speeds capacitors  $C_1$  and  $C_2$  must also be adjusted accordingly.)

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 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 Constant Current Supply so that when the brush arm is in contact with a short grounded segment, the capacitor then 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 will be just half way between two successive short grounded segments when any transition occurs. The charging current is adjusted to such a value that when the measuring capacitor 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 middle of the long segment. Thus whenever a perfect transition occurs, a relay interrupts the charging of the measuring capacitor and transfers it to the circuit containing the bias meter and reference battery just as the voltage across its terminals has reached 55 volts.

If there is no transition when the brush is passing over a long segment the capacitor 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 capacitor voltage, there will be no flow of current if the voltage of the measuring capacitor is also 55 volts when transferred to this circuit by a relay. If the capacitor voltage is greater than the battery voltage, there will be a discharge of electricity from the capacitor through the opposing battery; but for a capacitor voltage less than 55, the battery will cause a current flow into the capacitor. The greater the voltage difference, the greater the surge of current. Since the capacitor voltage is being compared to the battery, the 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 at the beginning of the start pulse and the brush cannot rest on a grounded segment in its stopped position, it is necessary to control the voltage across the capacitor  $C_1$  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, capacitor  $C_1$  will be charged to the reference battery voltage when the brush is on the "Stop" segment. Accordingly, there will be no indication in the comparison circuit for the first or start pulse 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 capacitor 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 capacitor 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 capacitor 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 capacitor, but in the opposite direction.

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 electron tube circuit known as a "rectifying peak voltmeter circuit" is used for this purpose. The indicating meter included in this 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 capacitor  $C_2$  to rise to a voltage higher than 55 volts; therefore, there will be a discharge current out of the capacitor 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 capacitor  $C_2$  from rising to 55 volts. Therefore, there will be a charging current in the opposite direction from the reference battery into the capacitor 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; and on six transition characters, it will read one and one-half times what it should.

For miscellaneous 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.

As pointed out earlier, modern transmission measuring sets employ an electronic distributor to perform the functions of the mechanical distributor shown in Figure 14-6. The electronic distributor consists of three parts; a character-timer, pulse oscillator and capacitor discharger. The character-timing portion of the circuit starts to operate at the beginning of the start pulse. This operation starts the pulse oscillator and at the end of the character, the character-timer furnishes the correct timing for the stopping of the pulse oscillator. The pulse oscillator supplies correctly timed pulses to the capacitor discharger for the purpose of discharging the measuring capacitors at the correct intervals. Electronic circuits arranged for performing functions of this type are discussed in later Chapters.

## CHAPTER 15

### ALTERNATING CURRENTS

#### 15.1 Source of Alternating Emf

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 electric energy, and thence to our ultimate aim, which is the application of these in electrical communications systems. 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 emf.

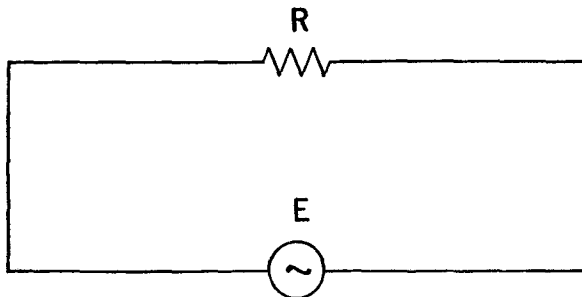


FIGURE 15-1

Figure 15-1 represents an alternating-current circuit in its simplest form. In this Figure we have a new convention for source of emf, which represents a single alternating cycle. Unlike the battery or other simple form of direct emf, we cannot completely describe such a source of emf by simply giving its voltage, for example  $E = 10$  volts. Here we have a voltage which gradually increases to a maximum value, then decreases to zero, to increase again to a maximum value in the opposite direction, and then 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 15-2 illustrates cycles of alternating emf's all very different in this respect.

Furthermore, we should not know the rapidity with which the alternations are taking place. For

example, Figure 15-3 represents two cycles of identical emf 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 emf we must know the following:

- The wave shape of the alternating cycle.
- The value of the emf at some specified point in the cycle.
- The length of time to complete the cycle, or the frequency of its repetition in a given time.

In classifying electric currents in Chapter 8, 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 steady-state alternating-current circuits deals with sine waves. Complex waves may be analyzed into combinations of sine waves of different frequencies, just as complex tones are analyzed into fundamentals and harmonics.

#### 15.2 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 emf 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 practically all vibratory motion. It greatly simplifies alternating-current circuits because—a **sine wave emf impressed upon a circuit having a network of any number and arrangement of resistances, inductances, and capacitances 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 emf.

The above rule holds in all its applications since the sine wave possesses the following properties:

- Sine waves of the same frequency can be added (or subtracted) either in or out of

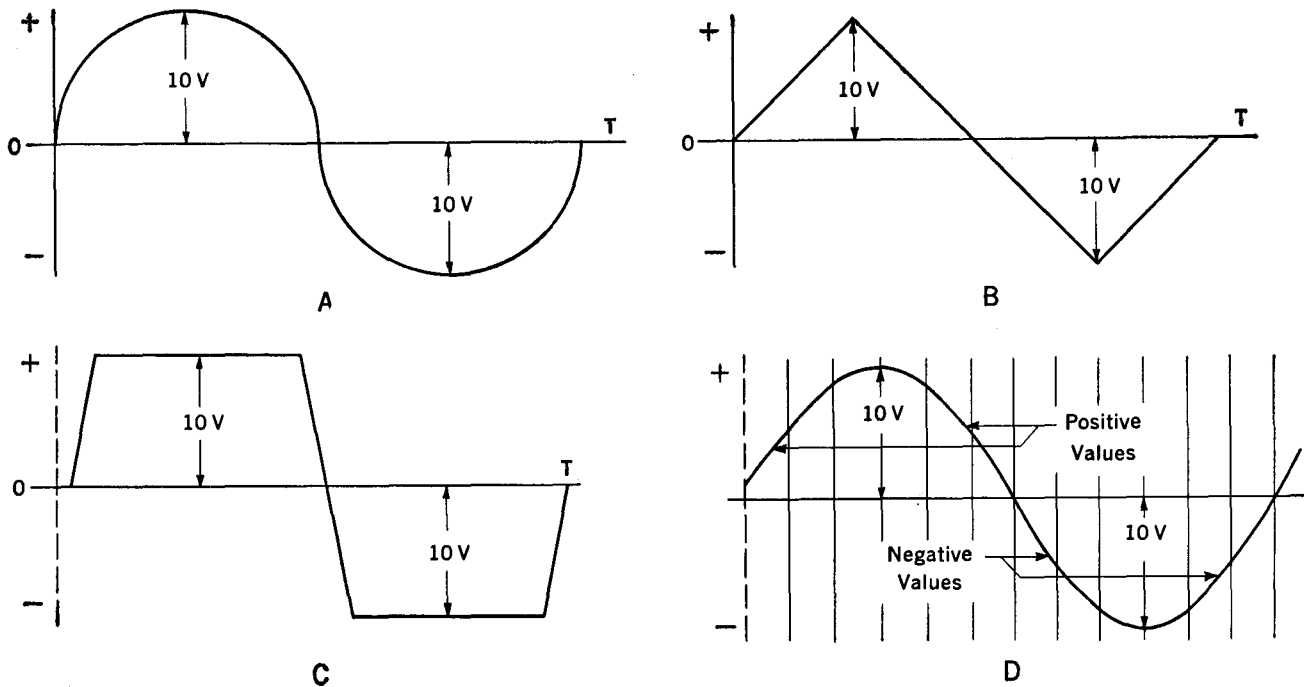


FIG. 15-2 THEORETICAL ALTERNATING CURRENT WAVESHAPES COMPARED WITH SINE WAVE

“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 emf across a resistance, inductance or capacitance gives a sine wave current through the resistance, inductance or capacitance (though not necessarily in phase).
- c. Whenever an emf is induced on account of the ever-changing value of a sine wave current, this induced emf is a sine wave (though not in phase).

### 15.3 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 emf's with vectors. Figure 15-4 shows the graphical construction of a sine wave as described in

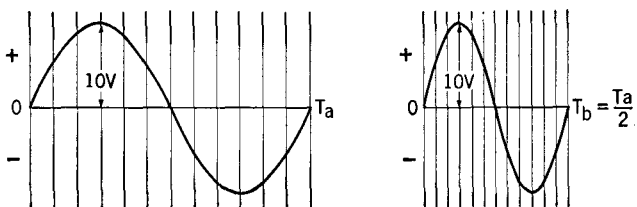


FIG. 15-3 SINE WAVES OF DIFFERENT FREQUENCY

Chapter 6. In this figure the horizontal scale (abscissa) represents time and the vertical scale (ordinate) 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 15-4, 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 maxi-

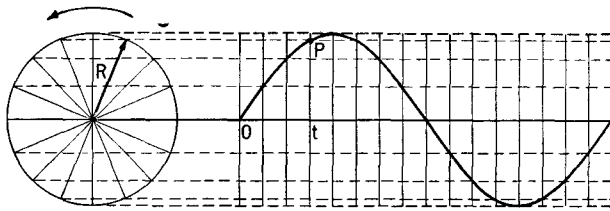
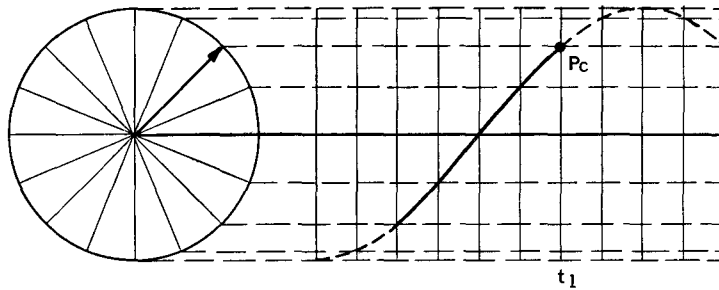


FIG. 15-4 VECTOR REPRESENTATION OF INSTANTANEOUS CURRENT VALUE

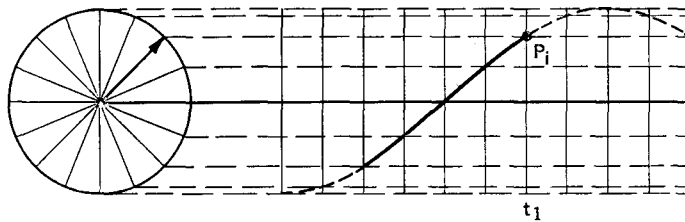
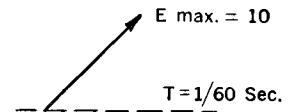
mum current or emf value of the sine wave in question. The angle this vector makes with the horizontal gives the position of point *P*. If we assume a direction of rotation for the vector, the instantaneous value of current at that time is

measured by the vertical distance *Pt* which is equal to the length of the vector times the sine of the angle which it makes with the horizontal. Furthermore, we can determine by the position of the vector whether the value of the current or emf is increasing or decreasing, and its direction. The accepted convention for direction of rotation is counter-clockwise and will be understood hereafter, without the arrow being used to indicate it.

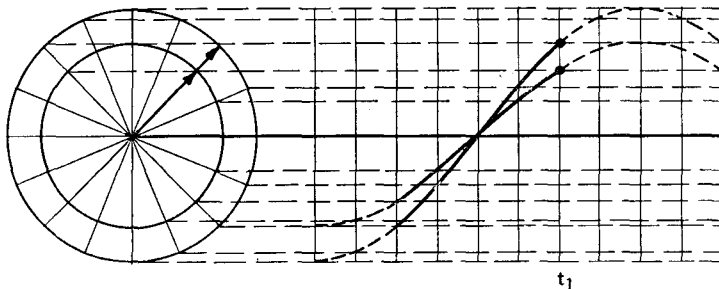
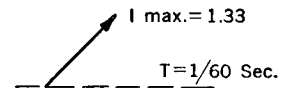
In Figure 15-1, let us assume that the maximum value of *E* is 10 volts, the frequency is 60 cycles per second, and the value of *R* is 7.5 ohms. Also let us assume the circuit to have negligible capacitance and inductance. By arbitrarily adopting a scale, we can represent the emf at a given



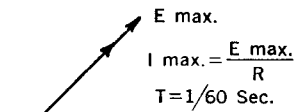
A - Voltage



B - Current



C - Current and Voltage



VECTOR REPRESENTATION

FIG. 15-5 CURRENT AND VOLTAGE IN PHASE

instant by Figure 15-5(A). Since the inductance and capacitance 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 emf 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 emf becomes a maximum of 10 volts, and becoming zero at exactly the same time that the emf 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".

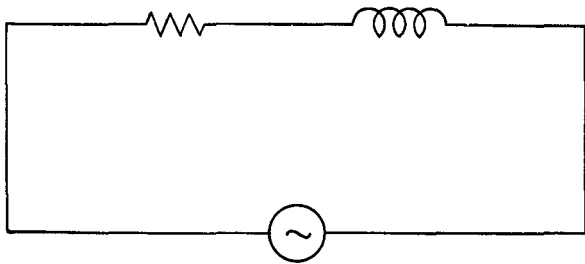


FIGURE 15-6

But if, instead of a circuit such as that shown by Figure 15-1, we have the circuit shown by Figure 15-6, it will be necessary to consider the effect of the inductance. 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 affect materially 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 emf has reached its maximum value. Figure 15-7 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

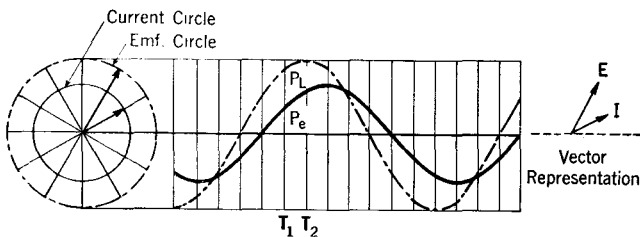


FIG. 15-7 CURRENT LAGGING IMPRESSED EMF

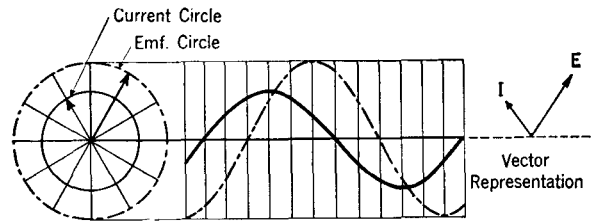


FIG. 15-8 CURRENT LEADING IMPRESSED EMF

of phase. This is accomplished by having the voltage vector ahead of the current vector in its rotation (in the conventional counter-clockwise direction) 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 positions of the radii of the two circles.

In the case of a circuit having a series capacitance instead of an inductance, the circuit reactions are the reverse. The current vector then is ahead of, or "leads", the emf vector as shown by Figure 15-8. Electrical conditions in circuits containing inductance or capacitance, therefore, can be represented by current and voltage vectors, which will, in general, be out of phase.

In direct-current networks, we used equations based on Kirchoff's Laws which called for adding or subtracting current or emf 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 15-9, 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 currents in the parallel branches.

#### 15.4 Effective Emf 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

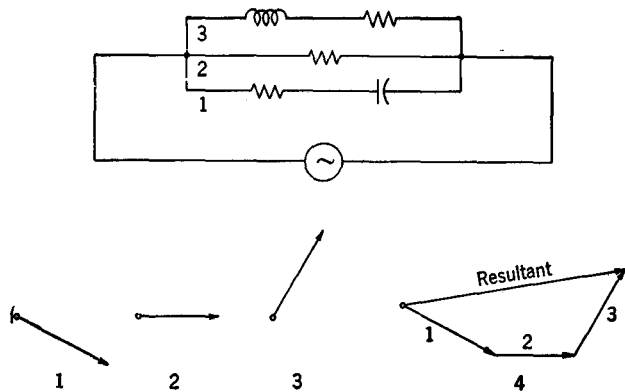


FIG. 15-9 GRAPHICAL ADDITION OF CURRENT VECTORS

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 counter-clockwise sense, indicated the particular instant being considered.

For practical purposes, however, it would be inconvenient to always state 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 resistor will be exactly the same as the heating effect of the same value of direct current in the same resistor. 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 the square root of the average of the squares of the instantaneous values over a complete cycle. This evaluates to .707 times the maximum value, or—

$$I = .707 I_{max}. \quad (15:1)$$

and

$$E = .707 E_{max}. \quad (15:2)$$

where  $E$  and  $I$  without subscripts indicate effective values. Unless specifically stated otherwise, values of alternating currents and voltages are al-

ways given in terms of their effective values. Likewise, in most practical applications, 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 phase relationship of the current and voltage with reference to each other, or to some other current or voltage in the same circuit.

### 15.5 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—

$$p = ei \quad (15:3)$$

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 15-10.

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 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—

$$P = EI \quad (15:4)$$

and, as always,—

$$P = I^2R \quad (15:5)$$

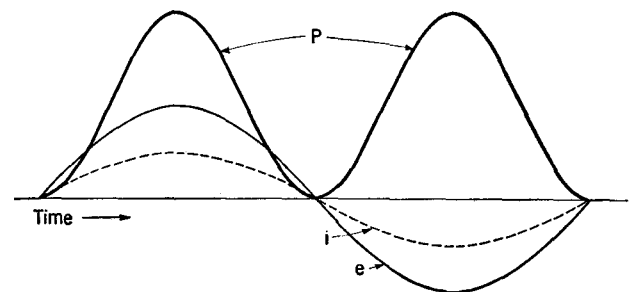


FIG. 15-10 POWER IN A-C CIRCUIT WHEN CURRENT AND VOLTAGE ARE IN PHASE

The condition where the circuit contains either inductance or capacitance 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 15-11. Here the product  $ei$  gives both positive and negative values and we have the 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 the voltage. The power now is—

$$P = EI \cos \theta \quad (15:6)$$

where  $\theta$  is the phase angle between the current and voltage. The term,  $\cos \theta$ , is known as the **power factor** and has a value of 1 when  $\theta$  is zero, or the current and voltage are in phase. This is the situation obtained in Figure 15-10. When  $\theta$  is 90 degrees, or the voltage and current are one-quarter cycle out of phase,  $\cos \theta = 0$  and power consumption is zero. For intermediate phase angles  $\cos \theta$  will be between zero and one.

It may be noted that the expression,  $P = I^2R$ , remains true in this case and confirms with equation (15:6) because as we shall learn,  $R = Z \cos \theta$  and  $I = E/Z$ , from whence—

$$\begin{aligned} P &= I^2R = I \times I \times R \\ &= I \times E/Z \times Z \cos \theta = IE \cos \theta. \end{aligned}$$

### 15.6 Ohm's Law and Alternating-Current Calculations

In Chapter 2 we learned that the relation between the voltage and 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., capacitance 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.

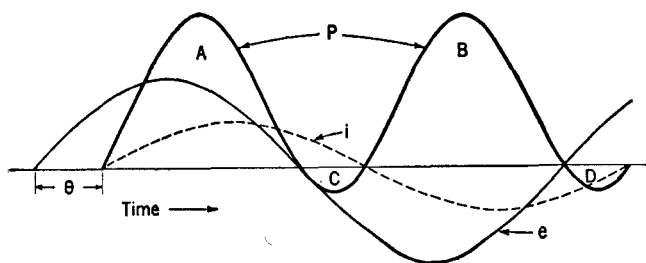


FIG. 15-11 POWER IN A-C CIRCUIT WHEN CURRENT AND VOLTAGE ARE NOT IN PHASE

Again, the effects of inductance and capacitance depend entirely upon the particular frequency which we wish to consider. We must therefore

TABLE V

#### Conventional Symbols Commonly Used in Alternating-Current Work

Symbol	Stands for
$P$	Average power for a cycle of emf and current.
$E$	Effective emf.
$I$	Effective current.
$E_{ave}$	Average emf.
$I_{ave}$	Average current.
$e$	Emf at some specific instant, or instantaneous voltage.
$i$	Current at some specific instant, or instantaneous current.
$E_1$	Induced emf.
$T$	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.
$X_C$	Capacitive reactance in ohms.
$Y$	Admittance in mhos.
$\theta$	Angle between current and impressed emf, or between impedance and resistance, etc.

introduce some new quantity that will express in ohms not only the resistance to current but the **combined effects** of resistance, capacitance and inductance at a definite stated frequency. This quantity is called **impedance**, and Ohm's Law is adjusted to read—

$$Z \text{ (ohms)} = \frac{E \text{ (volts)}}{I \text{ (amperes)}} \quad (15:7)$$

where  $Z$  is the symbol for **impedance** or the com-



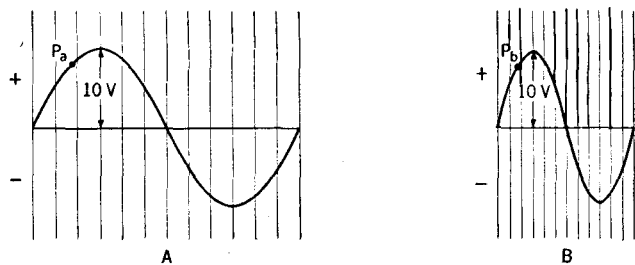


FIGURE 15-12

combined effect of the circuit's resistance, inductance and capacitance 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 capacitance 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 capacitance 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 capacitance values in a given circuit branch.

### 15.7 Inductive Reactance

Referring to Chapter 8, it will be recalled that we consider 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,

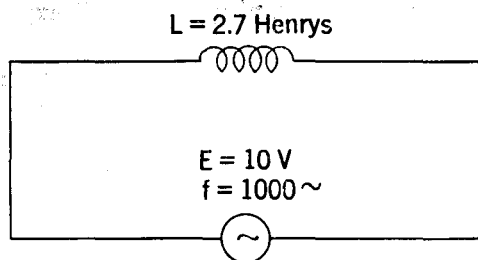


FIGURE 15-13

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 15-12 this becomes apparent. Here are two current cycles of the same effective value but 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 the measure of current change and we would expect, therefore, that the induced emf 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  $2\pi$  or—**

$$X_L = 2\pi fL \quad (15:8)$$

where  $X_L$  is the inductive reactance in ohms,  $\pi$  is 3.1416,  $f$  is the frequency expressed in cycles per second, and  $L$  is the inductance in henrys.

For practical use this becomes—

$$X_L = 6.2832fL \quad (15:9)$$

**Example:** In Figure 15-13 assume that the source of alternating emf 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.

**Solution:**

$$\begin{aligned} X_L &= 6.2832fL = 6.2832 \times 1000 \times 2.7 \\ &= 16964 \text{ ohms} \\ I &= \frac{E}{16964} \\ &= \frac{10}{16964} \\ &= .00059 \text{ ampere, ans.} \end{aligned}$$

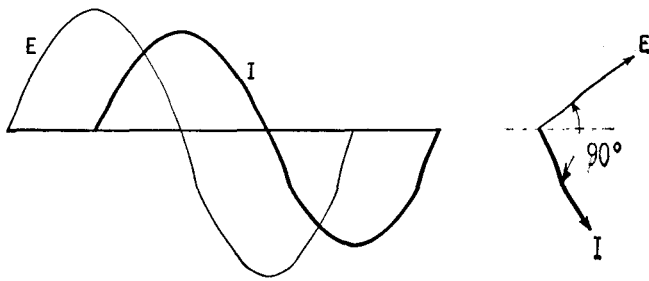


FIG. 15-14 EFFECT OF INDUCTIVE REACTANCE

In this example, the current will be 90° behind the impressed voltage, as shown in Figure 15-14, because the induced emf due to the current must be equal and opposite to the impressed emf and the induced emf is the rate of change of current times the inductance and must, therefore, be 90° behind the current.

### 15.8 Capacitive Reactance

The effect of capacitive reactance in the circuit is exactly opposite to that of inductive reactance—in fact, the two tend to neutralize each other. Capacitive reactance decreases with increasing frequency and capacitance values. It also tends to make the current lead instead of lag the voltage (see Figures 15-14 and 15-16). Accordingly, if inductive reactance is assumed as positive, capacitive reactance must be taken as negative.

This time relation of the voltage and current in a circuit containing capacitance may be seen by referring to Figures 15-15 and 15-16. Here when the impressed voltage  $E$  is at its maximum positive value, the capacitor 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 capacitor 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, it charges the capacitor in the opposite (positive) direction. During this time, the negative current decreases to zero as the capacitor becomes fully charged. Then as the negative impressed voltage decreases from its maximum, the capacitor 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 capacitive reactance is as follows:

$$X_c = -\frac{1}{2\pi fC} \quad (15:10)$$

where  $C$  is capacitance in farads. Converting  $C$  to the customary capacitance unit, microfarad, we have—

$$X_c = -\frac{1,000,000}{2\pi fC} \quad (15:11)$$

or with 3.1416 substituted for  $\pi$ —

$$X_c = -\frac{1,000,000}{6.2832fC} \quad (15:12)$$

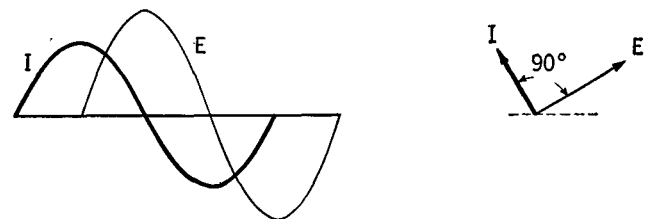


FIG. 15-16 EFFECT OF CAPACITIVE REACTANCE

**Example:** In Figure 15-15,  $E$  is 10 volts,  $f$  is 1000 and  $C$  is  $2\mu\text{f}$ . What is the current in amperes?

**Solution:**

$$\begin{aligned} I &= \frac{E}{X_c} \\ X_c &= -\frac{1,000,000}{6.2832 \times 1000 \times 2} \\ &= -\frac{1,000}{6.2832 \times 2} \\ &= -79.5 \text{ ohms} \\ I &= -\frac{10}{79.5} \\ &= -.126 \text{ ampere, ans.} \end{aligned}$$

(minus sign here means leading current)

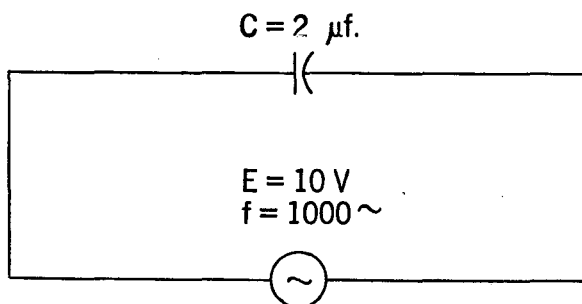


FIGURE 15-15

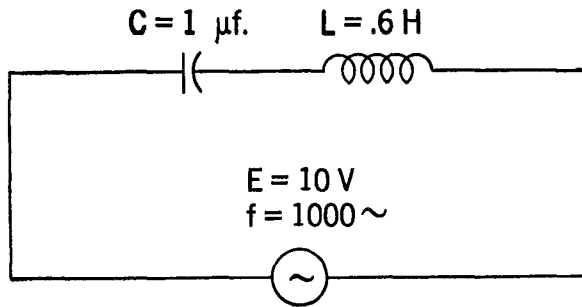


FIGURE 15-17

### 15.9 Combination of Inductive and Capacitive Reactance

If we wish to get the combined or total reactance of an inductance in series with a capacitance, such as that shown in Figure 15-17, we may combine the reactances as follows:

$$X = X_L + X_C$$

or, from equations (15:9) and (15:12)—

$$X = 6.2832fL - \frac{1,000,000}{6.2832fC} \quad (15:13)$$

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 capacitive reactance predominates.

**Example:** Calculate the current in the circuit shown by Figure 15-17.

**Solution:**

With no resistance in the circuit—

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

and

$$X = X_L + X_C$$

$$= 6.2832fL - \frac{1,000,000}{6.2832fC}$$

$$X = 6.2832 \times 1000 \times .6 - \frac{1,000,000}{6.2832 \times 1000 \times 1}$$

$$= 3770 - 159$$

$$= 3611 \text{ ohms}$$

$$I = \frac{10}{3611} = .0028 \text{ ampere, ans.}$$

### 15.10 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 15-18(A), and for the second condition, as shown in Figure 15-18(B). In a circuit containing resistance only, the current and voltage are in phase. In a circuit containing only reactance, the current either “leads” or “lags” the voltage by  $90^\circ$ . Thus, the effect of reactance on the phase of the current is at an angle of  $90^\circ$  to the effect of resistance. Therefore, in a series combination of the two, the resultant effect cannot be found by simple addition or subtraction.

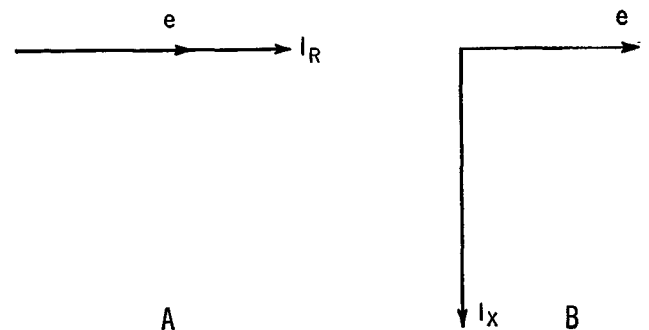


FIGURE 15-18

For the purpose of this discussion, let us assume a series combination of resistance and inductive reactance as in Figure 15-19. The vector diagram in Figure 15-20 shows a horizontal line  $R$  to represent resistance, and since the effect of reactance is at right angles to the resistance, line  $X_L$  is drawn at right angles from the end of  $R$  to represent reactance. The line  $Z$ , then represents the resultant of  $R$  and  $X_L$ , or the impedance.

Since the lines  $R$ ,  $X_L$  and  $Z$  always form a right triangle, of which  $Z$  is the hypotenuse, an equation can be formed to find the value of  $Z$  directly. The equation in this case is—

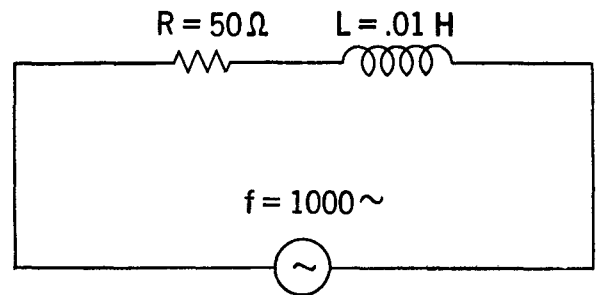


FIGURE 15-19

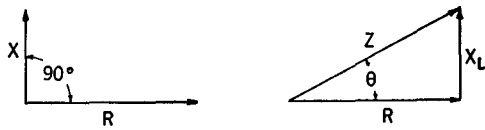


FIGURE 15-20

$$Z^2 = R^2 + X_L^2$$

or

$$Z = \sqrt{R^2 + X_L^2} \quad (15:14)$$

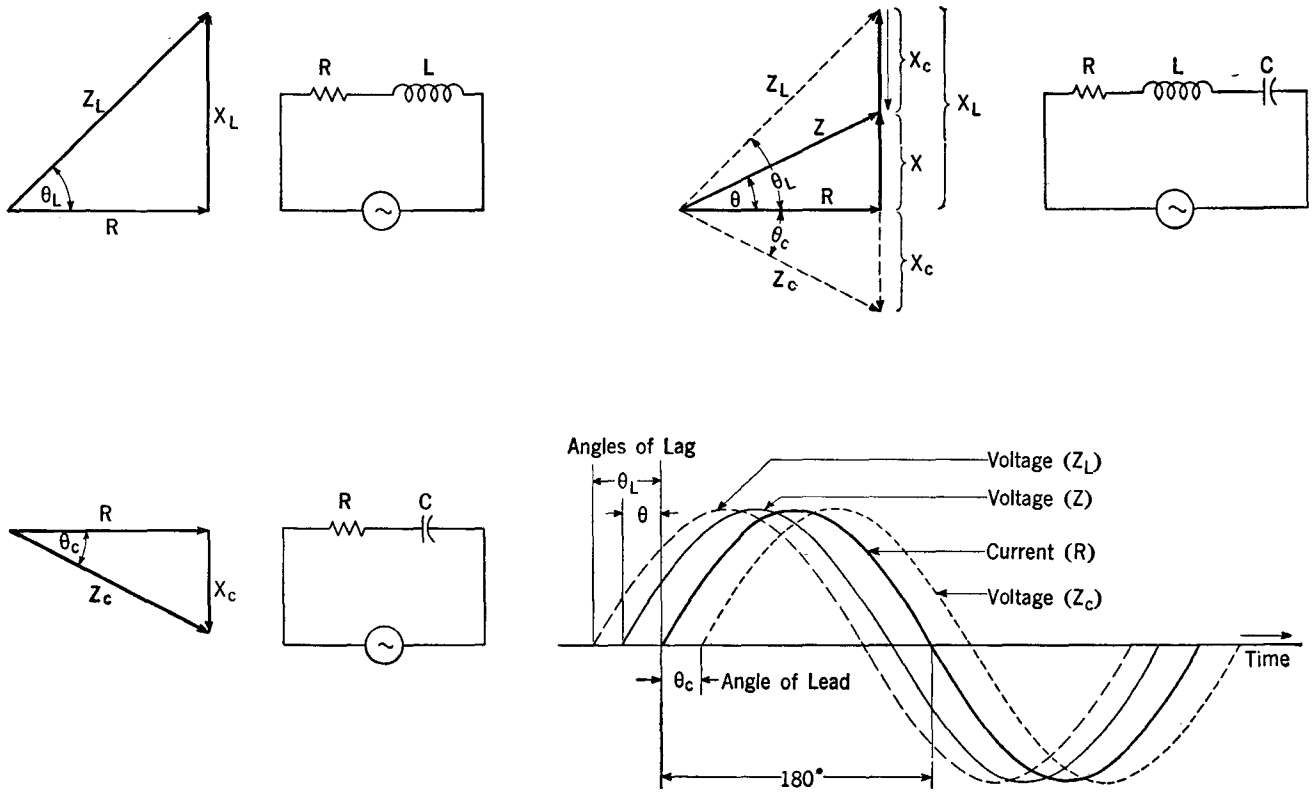
If the reactance is a capacitive reactance, the combined effect of resistance and capacitance, as shown in Figure 15-21, may be represented by the vector diagram of Figure 15-22 in which the reactance is shown  $90^\circ$  behind the resistance since it acts in the opposite direction to inductive reactance. The mathematical result and the equation are the same as for inductive reactance—

$$Z = \sqrt{R^2 + X_c^2} \quad (15:15)$$

In these diagrams if  $R$  is represented by the

TABLE VI

CHART OF VECTOR RELATIONS



PROPERTY	REACTANCE	IMPEDANCE	PHASE ANGLE
Inductance ( $L$ ) . . . . .	$X_L = 2\pi fL$	$Z_L = \sqrt{R^2 + X_L^2}$	$\theta_L = \text{Tan}^{-1} \frac{X_L}{R}$
Capacitance ( $C$ ) . . . . .	$X_C = -\frac{1,000,000}{2\pi fC}$	$Z_C = \sqrt{R^2 + X_C^2}$	$\theta_C = \text{Tan}^{-1} \frac{X_C}{R}$
Net Effect . . . . .	$X = X_L + X_C$	$Z = \sqrt{R^2 + X^2}$	$\theta = \text{Tan}^{-1} \frac{X}{R}$

- NOTES: 1. If lines  $Z_C$ ,  $Z_L$  or  $Z$  represent phase of voltage, line  $R$  will indicate lead or lag of current and  $\theta_C$ ,  $\theta_L$  and  $\theta$  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  $Z/\theta$ ; for example,  $Z/\theta = 15/30^\circ$ , etc.

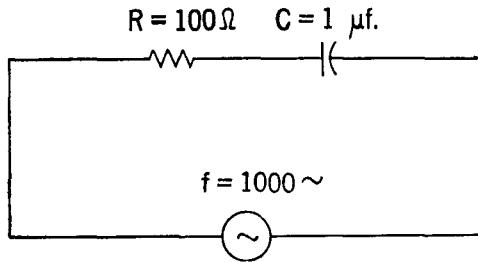


FIGURE 15-21

same line as the current,  $Z$  will be represented by the same line as the impressed emf; consequently the angle  $\theta$  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 emf for positive angle as shown in Figure 15-20, and current leading impressed emf for negative angle as shown in Figure 15-22.

We can now consider a simple series circuit with all three properties, resistance, inductance,

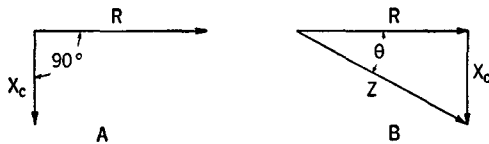


FIGURE 15-22

and capacitance, as shown in Figure 15-23. Here we have two reactances acting in opposite phase as shown in Figure 15-24(A). In constructing the impedance triangle,  $X_c$  must be considered as negative and subtracting from  $X_L$  as shown in Figure 15-24(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 15-25,  $X$  will be negative.

Having the relation of impedance to its component parts fixed in mind by the foregoing graphical construction, we can calculate its value

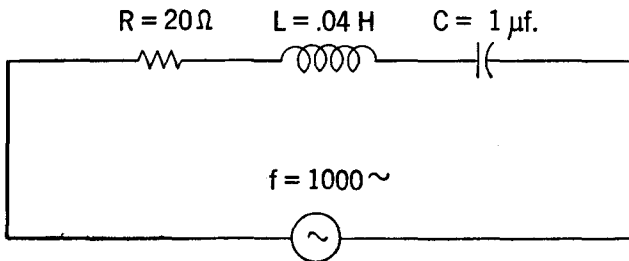


FIGURE 15-23

in the same manner as we calculate the length of the hypotenuse of any right triangle, as has been explained. Thus, we may combine resistance, inductive reactance and capacitive reactance in one general equation for impedance where

$$X = X_L + X_c = 2\pi fL - \frac{1,000,000}{2\pi fC}$$

and

$$Z = \sqrt{R^2 + \left(2\pi fL - \frac{1,000,000}{2\pi fC}\right)^2} \quad (15:16)$$

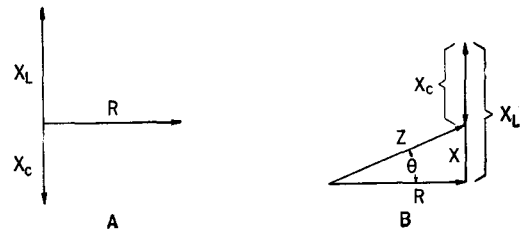


FIGURE 15-24

**Example:** In Figure 15-23,  $R$  is 20 ohms,  $f$  is 1000 cycles per second,  $L$  is .04 henry and  $C$  is 1  $\mu$ f. What is the numerical value of the impedance in ohms?

**Solution:**

$$\begin{aligned} Z &= \sqrt{R^2 + \left(2\pi fL - \frac{1,000,000}{2\pi fC}\right)^2} \\ &= \sqrt{(20)^2 + \left(6.28 \times 1000 \times .04 - \frac{1,000,000}{6.28 \times 1000 \times 1}\right)^2} \\ &= \sqrt{(20)^2 + (251 - 159)^2} \\ &= \sqrt{400 + 8464} \\ &= \sqrt{8864} \\ &= 94 \text{ ohms, ans.} \end{aligned}$$

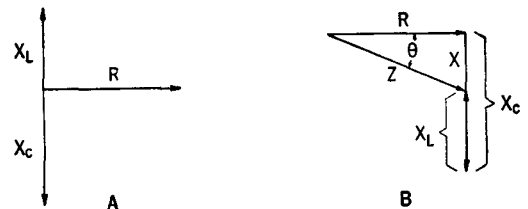


FIGURE 15-25

In these calculations we have only determined the numerical value of the impedance. This does not completely describe it, however, since there

could be any number of resistance, capacitance 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  $\theta$  in Figures 15-24 and 15-25.

Impedance is customarily expressed, accordingly, in the form  $Z/\theta$  ( $Z$  at an angle  $\theta$ ) where  $Z$  is the magnitude of the impedance and  $\theta$  is the angle of

lag or lead between any emf impressed across the impedance and the resultant current. As may be seen from Figure 15-24,

$\theta$  is equal to  $\tan^{-1} \frac{X}{R}$  (the angle whose tangent is  $\frac{X}{R}$ )

Also, by simple trigonometry we know that  $R = Z \cos \theta$  and  $X = Z \sin \theta$ . Thus, with the impedance expressed in the form  $Z/\theta$  it is completely defined and we may readily determine the magnitude of its resistance and reactance components.

## CHAPTER 16

### ALTERNATING CURRENTS—(Continued)

#### 16.1 Series Networks

In Chapters 2 and 3, means of solving direct-current networks for the current values in the various branches were described. **The same methods and equations 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.

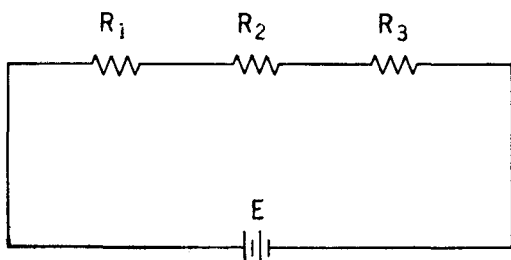


FIGURE 16-1

We learned in Chapter 3 that the total resistance in a d-c series circuit such as is shown in Figure 16-1 is equal to the **arithmetic** sum of the individual resistances, or—

$$R = R_1 + R_2 + R_3, \text{ etc.} \quad (3:1)$$

Similarly in an a-c series circuit, as shown in Figure 16-2, the total impedance is equal to the **vector** sum of the individual impedances or—

$$\bar{Z} = \bar{Z}_1 + \bar{Z}_2 + \bar{Z}_3 \quad (16:1)$$

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  $\theta_1 = 30^\circ$ ,  $Z_2 = 15$

ohms with  $\theta_2 = 45^\circ$  and  $Z_3 = 20$  ohms with  $\theta_3 = 60^\circ$ ; we then have the three vectors represented by Figure 16-3(A) which, when added, give the value of  $Z$  shown in Figure 16-3(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 16-4. By comparing Figure 16-4(C) with Figure 16-4(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—

$$Z = \sqrt{(R_1 + R_2 + R_3)^2 + (X_1 + X_2 + X_3)^2} \quad (16:2)$$

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.

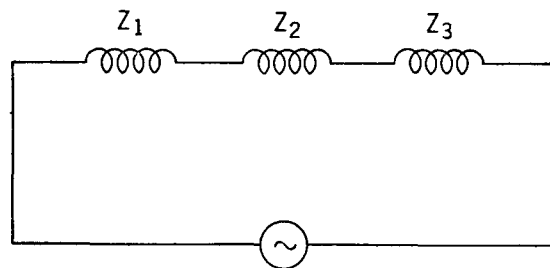


FIGURE 16-2

**Example:** Find the total impedance of the circuit of Figure 16-2 using the values for  $Z_1$ ,  $Z_2$  and  $Z_3$  given above.

**Solution:**

$$X = Z \sin \theta$$

$$R = Z \cos \theta$$

This gives

$$X_1 = 10 \times .500 = 5 \text{ ohms}$$

$$R_1 = 10 \times .866 = 8.7 \text{ ohms}$$

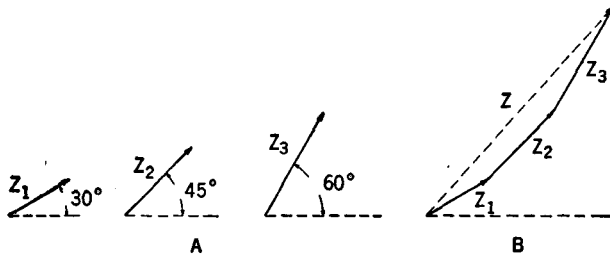


FIG. 16-3. GRAPHICAL ADDITION OF IMPEDANCE VECTORS

The other values can be determined in the same way, and we find that—

$$\begin{aligned} X_2 &= 10.6 \text{ ohms} \\ R_2 &= 10.6 \text{ ohms} \\ X_3 &= 17.3 \text{ ohms} \\ R_3 &= 10.0 \text{ ohms} \end{aligned}$$

Applying equation (16:2)—

$$\begin{aligned} Z &= \sqrt{(R_1 + R_2 + R_3)^2 + (X_1 + X_2 + X_3)^2} \\ &= \sqrt{(8.7 + 10.6 + 10)^2 + (5 + 10.6 + 17.3)^2} \\ &= \sqrt{(29.3)^2 + (32.9)^2} \\ &= 44.0 \text{ ohms} \end{aligned}$$

$$\begin{aligned} \theta &= \tan^{-1} \frac{32.9}{29.3} \\ &= \tan^{-1} 1.12 \\ &= 48^\circ. \end{aligned}$$

Therefore

$$Z = 44/48^\circ \text{ ohms, ans.}$$

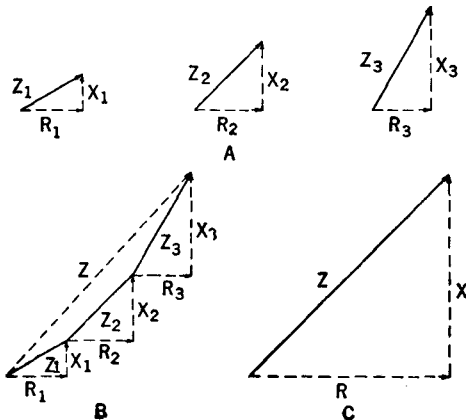


FIG. 16-4. ANALYSIS OF IMPEDANCE VECTOR ADDITION

The foregoing calculation covers a general case. In practice, however, we usually have given the inductance, capacitance and resistance values rather than the individual impedances with their respective angles.

**Example:** Find the impedance of the series circuit shown by Figure 16-5.

**Solution:**

$$Z = \sqrt{(R_1 + R_2 + R_3)^2 + (X_c + X_L)^2}$$

where

$$\begin{aligned} X_c &= -\frac{1,000,000}{2\pi fC} \\ &= -\frac{1,000,000}{6.28 \times 1000 \times .2} \\ &= -796 \text{ ohms} \end{aligned}$$

and

$$\begin{aligned} X_L &= 2\pi fL \\ &= 6.28 \times 1000 \times .02 \\ &= 125.6 \text{ ohms} \end{aligned}$$

Then

$$\begin{aligned} Z &= \sqrt{(70 + 60 + 100)^2 + (-796 + 125.6)^2} \\ &= \sqrt{(230)^2 + (-670.4)^2} \\ &= 709 \text{ ohms} \end{aligned}$$

and

$$\begin{aligned} \theta &= \tan^{-1} \frac{-670.4}{230} \\ &= \tan^{-1} (-2.9) = -71^\circ, \end{aligned}$$

whence

$$Z = 709/-71^\circ, \text{ ans.}$$

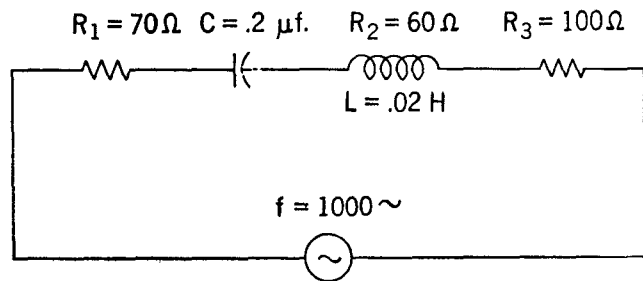


FIGURE 16-5

## 16.2 Parallel and Series-parallel Networks

In Chapter 3 we learned that the combined resistance of two parallel resistances was equal to—

$$R = \frac{R_1 R_2}{R_1 + R_2} \quad (3:4)$$

or that if more than two resistances are in parallel, the combined resistance may be found by add-



ing together the reciprocals of each resistance (called conductance) and taking the reciprocal of this value. That is—

$$G = G_1 + G_2 + G_3 \quad (3:6)$$

or

$$\frac{1}{\bar{R}} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}$$

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—

$$\bar{Z} = \frac{\bar{Z}_1 \bar{Z}_2}{\bar{Z}_1 + \bar{Z}_2} \quad (16:3)$$

or, for more than two in parallel,

$$\frac{1}{\bar{Z}} = \frac{1}{\bar{Z}_1} + \frac{1}{\bar{Z}_2} + \frac{1}{\bar{Z}_3}, \text{ etc.} \quad (16:4)$$

which latter may also be written—

$$\bar{Y} = \bar{Y}_1 + \bar{Y}_2 + \bar{Y}_3, \text{ etc.} \quad (16:5)$$

where  $Y$  represents the reciprocal of impedance and is called **admittance**.

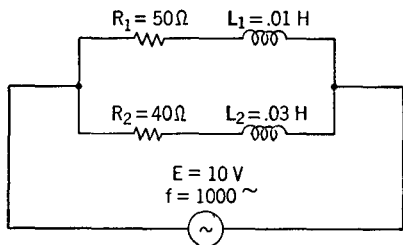


FIGURE 16-6

The mathematical solution of such equations as (16:3) 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  $\bar{Z} = Z/\theta = R + jX$  where “ $j$ ” indicates a rotation of  $90^\circ$  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 equations 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 16-6 and the phase angle of this current with the generator emf.

By Ohm’s Law we know that the total current delivered by the generator is—

$$I = \frac{E}{Z/\theta}$$

where  $Z/\theta$  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  $Z_1/\theta_1$  and  $Z_2/\theta_2$ . Then from the usual equation for parallel circuits—

$$Z/\theta = \frac{Z_1/\theta_1 \times Z_2/\theta_2}{Z_1/\theta_1 + Z_2/\theta_2} \quad (16:3)$$

The first step is to find the values of  $Z_1/\theta_1$  and  $Z_2/\theta_2$ . We know that—

$$\theta_1 = \tan^{-1} \frac{X_1}{R_1}$$

where

$$R_1 = 50 \text{ ohms}$$

and

$$X_1 = 2\pi fL_1 = 6.28 \times 1000 \times .01 = 62.8 \text{ ohms}$$

Then

$$\theta_1 = \tan^{-1} \frac{62.8}{50} = \tan^{-1} 1.255 = 51^\circ 27'$$

Then, since  $R_1 = Z_1 \cos \theta_1$ —

$$Z_1 = \frac{R_1}{\cos \theta_1} = \frac{50}{\cos 51^\circ 27'} = \frac{50}{.6232} = 80.3 \text{ ohms}$$

Impedance of branch 1

$$= Z_1/\theta_1 = 50 + j 62.8 = 80.3/51^\circ 27'$$

Likewise  $R_2 = 40$  ohms

$$X_2 = 2\pi fL_2 = 6.28 \times 1000 \times .03 = 188.4 \text{ ohms}$$

$$\theta_2 = \tan^{-1} \frac{188.4}{40} = \tan^{-1} 4.71 = 78^\circ 1'$$

$$Z_2 = \frac{40}{\cos 78^\circ 1'} = \frac{40}{.2076} = 193.0 \text{ ohms}$$

Impedance of branch 2

$$= Z_2/\theta_2 = 40 + j 188.4 = 193.0/78^\circ 1'$$

Then using equation (16:3) and expressing the vectors of the numerator in the  $Z/\theta$  form, since multiplication is involved, and the vectors of the denominator in the  $R + jX$  form, since addition is involved, we have—

$$\begin{aligned}
Z/\theta &= \frac{Z_1/\theta_1 \times Z_2/\theta_2}{(R_1 + jX_1) + (R_2 + jX_2)} \\
&= \frac{Z_1 Z_2 / \theta_1 + \theta_2}{(R_1 + R_2) + j(X_1 + X_2)} \\
&= \frac{80.3 \times 193.0 / 51^\circ 27' + 78^\circ 1'}{(50 + 40) + j(62.8 + 188.4)} \\
&= \frac{15500 / 129^\circ 28'}{90 + j251.2} \\
&= \frac{15500 / 129^\circ 28'}{90} \bigg/ \cos \left[ \tan^{-1} \frac{251.2}{90} \right] \bigg/ \tan^{-1} \frac{251.2}{90} \\
&= \frac{15500 / 129^\circ 28'}{90} \bigg/ \cos (\tan^{-1} 2.79) \bigg/ \tan^{-1} 2.79 \\
&= \frac{15500 / 129^\circ 28'}{90} \bigg/ \cos 70^\circ 17' \bigg/ 70^\circ 17' \\
&= \frac{15500 / 129^\circ 28'}{.3374} \bigg/ 70^\circ 17' \\
&= \frac{15500 / 129^\circ 28'}{267 / 70^\circ 17'} \\
&= \frac{15500}{267} \bigg/ 129^\circ 28' - 70^\circ 17' \\
&= 58.0 / 59^\circ 11'
\end{aligned}$$

And

$$\begin{aligned}
I &= \frac{E}{Z/\theta} = \frac{10}{58 / 59^\circ 11'} \\
&= \frac{10}{58} / 0^\circ - 59^\circ 11' = .173 / -59^\circ 11'
\end{aligned}$$

Thus we find that the generator will deliver a current of .173 ampere and this current will lag the generator voltage by  $59^\circ 11'$ .

With a little practice it will be found that sev-

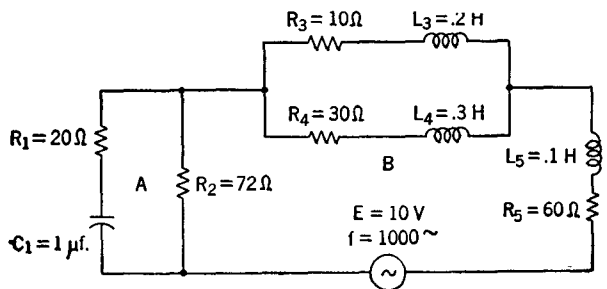


FIGURE 16-7

eral of the detailed steps given above can be performed in a single operation. This may be illustrated by solving the circuit of Figure 16-7 to find the current delivered by the generator and its phase relationship with the emf.

**Solution:**

$$\begin{aligned}
Z_1 &= R_1 + jX_1 \\
X_1 &= -\frac{10^6}{2\pi f C_1} = -\frac{1,000,000}{6.28 \times 1000 \times 1} = -159.3 \\
Z_1 &= 20 - j159.3 = 160.7 / -82^\circ 51' \\
Z_2 &= R_2 + jX_2 = 72 + j0 = 72 / 0^\circ \\
Z_A &= \frac{Z_1 Z_2}{Z_1 + Z_2} = \frac{160.7 / -82^\circ 51' \times 72 / 0^\circ}{20 - j159.3 + 72 + j0} \\
&= \frac{11,570 / -82^\circ 51'}{92 - j159.3} = \frac{11,570 / -82^\circ 51'}{184 / -60^\circ 1'} \\
&= 62.8 / -22^\circ 50' = 58.0 - j24.4 \\
Z_3 &= R_3 + jX_3 \\
X_3 &= 2\pi f L_3 = 6.28 \times 1000 \times .2 = 1256 \\
Z_3 &= 10 + j1256 = 1256 / 89^\circ 33' \\
Z_4 &= R_4 + jX_4 \\
X_4 &= 2\pi f L_4 = 6.28 \times 1000 \times .3 = 1884 \\
Z_4 &= 30 + j1884 = 1884 / 89^\circ 5' \\
Z_B &= \frac{Z_3 Z_4}{Z_3 + Z_4} = \frac{1256 / 89^\circ 33' \times 1884 / 89^\circ 5'}{10 + j1256 + 30 + j1884} \\
&= \frac{2,365,000 / 178^\circ 38'}{40 + j3140} \\
&= \frac{2,365,000 / 178^\circ 38'}{3140 / 89^\circ 16'} \\
&= 753 / 89^\circ 22' = 8 + j753
\end{aligned}$$

$$Z_C = R_5 + jX_5$$

$$X_5 = 2\pi f L_5 = 6.28 \times 1000 \times .1 = 628$$

$$Z_C = 60 + j628$$

The total impedance  $Z = Z_A + Z_B + Z_C$

$$\begin{aligned}
Z &= 58.0 - j24.4 + 8 + j753 + 60 + j628 \\
&= 126 + j1356 = 1360 / 84^\circ 40'
\end{aligned}$$

$$I = \frac{E}{Z} = \frac{10}{1360 / 84^\circ 40'} = .00735 / -84^\circ 40'$$

The current delivered by the generator has a value of .00735 ampere and lags the impressed voltage by  $84^\circ 40'$ .

### 16.3 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. 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 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”, in the case of a coil winding for example, represents not only the actual resistance of the conductor used to wind the coil but includes also a factor due to power losses within the iron core. These losses are 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. 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. Certain comparable losses may occur in the dielectric materials of capacitors that may also have the effect of increasing the apparent resistance of the circuit.

The effective resistance of a long wire such as an ordinary transmission line also increases as the frequency of the current through it increases, due to a cause known as **skin effect**. It is a result of the fact that any current flowing through a wire is accompanied by a magnetic field set up around the wire. This field exists to some degree at least within the wire as well as in the space surrounding it. As the frequency of the current in the wire increases, however, the reaction in the interior of the wire between the current flowing and the counter-voltages induced by the magnetic field tends to prevent the field from penetrating deeply into the wire. The net effect is to force the current in the wire to flow nearer the surface instead of being evenly distributed through the cross-sectional area of the conductor. Since the current thus flows through an effectively smaller conductor area the resistance increases. At low frequen-

cies this effect is small but at the high frequencies used in radio work the a-c resistance of a conductor due to skin effect may be many times the d-c value.

### 16.4 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 capacitance value, on the other hand, the negative reactance,  $X_C$ , is reduced when the frequency is increased. This is illustrated by Figure 16-8 where the inductive reactance  $X_L$  and the capacitive reactance  $X_C$  are plotted against frequency. If a series circuit contains both inductance and capacitance, as shown in Figure 16-8(A), there is therefore some frequency at which the negative reactance,  $X_C$ , becomes equal but opposite in value to  $X_L$ . Where the dotted line crosses the zero axis in Figure 16-8 (B), the combined reactance is equal to zero. **This is called the frequency to which the circuit is resonant, or where—**

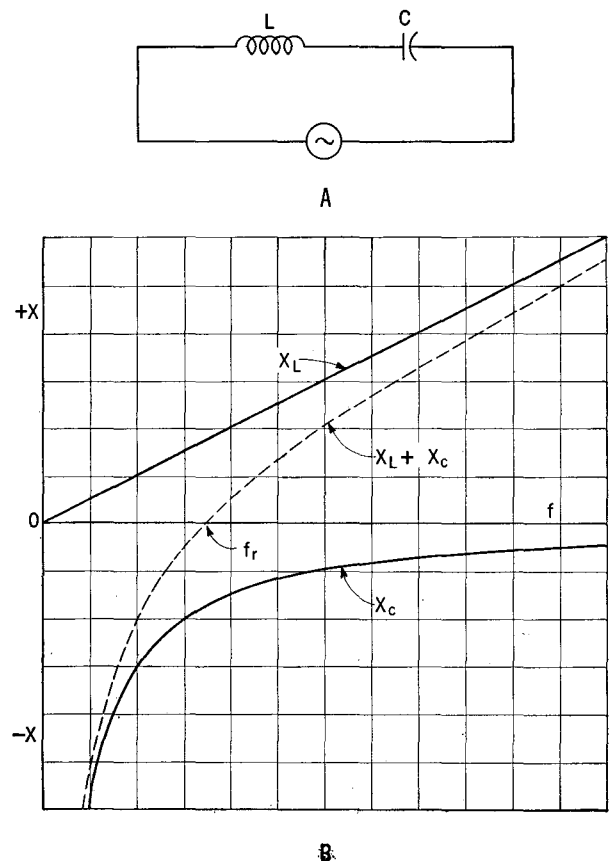


FIG. 16-8 SERIES RESONANT CIRCUIT

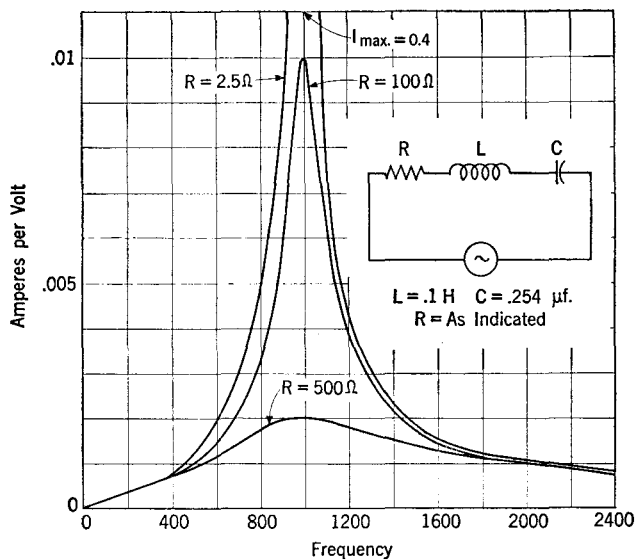


FIG. 16-9 CURVES OF CURRENT VALUES IN SERIES RESONANT CIRCUIT

$$0 = 2\pi fL - \frac{1,000,000}{2\pi fC}$$

The value of the resonant frequency,  $f_r$ , can be determined in terms of the inductance and capacitance by solving this equation for  $f$  as follows:

$$f_r = \frac{1,000}{2\pi\sqrt{LC}} \quad (16:5)$$

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 16-9 illustrates the behavior of a series resonant circuit similar to that shown in Figure 16-8(A), but including some resistance, when the applied emf is varied through a band of frequencies. The curves were plotted by assuming an impressed emf of 1 volt for each frequency of the band, and three different values of resistance. As will be noted, the peak current 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. 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. Below the resonant frequency, the capacitance in the circuit will have the major effect in limiting the current, and the circuit will tend to look like capacitance with

the current leading the voltage. Above resonance the inductance will limit the current and the circuit will look like inductance with the current lagging the voltage. At resonance the current and voltage are in phase although the current may be relatively very large. The actual voltages across the inductor and capacitor may therefore be many times as great as the voltage applied to the circuit.

**Example:** To what frequency is the circuit shown by Figure 16-9 resonant if  $C$  is .254  $\mu\text{f}$ ,  $L$  is .10 H; what current will flow at resonance when  $R$  is 4 ohms and  $E$  is 1.0 volt and what is the voltage drop across the inductance?

**Solution:**

$$f_r = \frac{1,000}{6.28\sqrt{.10 \times .254}}$$

$$= \frac{1,000}{6.28\sqrt{.0254}}$$

$$= \frac{1,000}{6.28 \times .159}$$

$$= 1,000 \text{ cycles per sec. Ans.}$$

$$I = \frac{E}{R} = \frac{1.0}{4} = .25 \text{ amp. Ans.}$$

$$E_L = IX_L$$

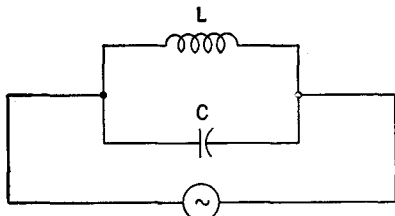
$$X_L = 2\pi fL = 6.28 \times 1,000 \times .10 = 628$$

$$E_L = .25 \times 628 = 157 \text{ volts. Ans.}$$

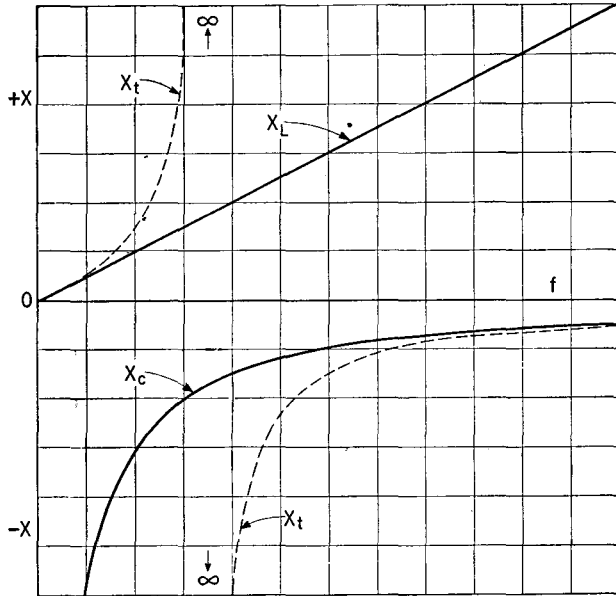
The resonance principle has numerous and interesting uses in connection with communication circuits. One application is the use of a capacitor of proper value in series with a telephone receiver winding, repeating coil winding, or other winding having inductance, where it is desired to increase the current.

A much more common use of the resonance principle is the so-called "tuned" circuit which is so extensively employed in radio and other high frequency applications. 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 small current for any other frequency. Figure 16-9 illustrates this principle.

In many tuned circuits, capacitance and inductance are connected to form a **parallel-resonant** circuit as shown in Figure 16-10(A). For this



A



B

FIG. 16-10 PARALLEL-RESONANT CIRCUIT

connection, when the positive reactance is equal and opposite to the negative reactance, and the resistance of the inductor is low, 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 16-10(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 capacitor 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 capacitance 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 inductor and capacitor, with the emf 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 the generator current is actually zero.

Figure 16-11 illustrates the selectivity of a parallel-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 parallel-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.

In radio and other high frequency work the parallel-resonant circuit is often called a "tank circuit". This is because it acts as a storage reservoir for electric energy. Here it may be more helpful to think in terms of energy transferring back and forth between the electric field of the capacitor and the magnetic field of the inductor rather than merely of current oscillating back and forth in the parallel circuit.

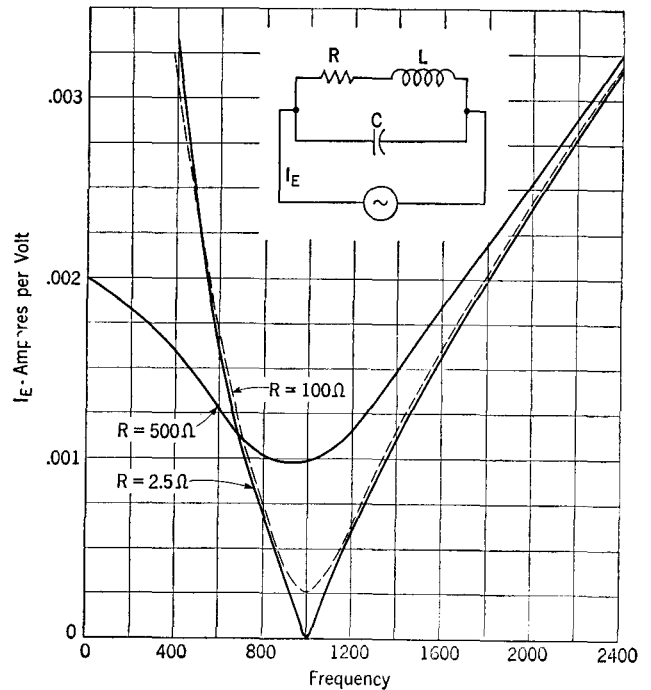


FIG. 16-11 CURVES OF CURRENT VALUES IN PARALLEL-RESONANT CIRCUIT

## 16.5 Q

Where resonant circuits are used for purposes of tuning or frequency selection, it is important that the resistance of the circuit be held to a minimum. This is illustrated by Figures 16-9 and 16-11 where it is shown that the sharpness of tuning is greater the lower the resistance. Since the resistance in the circuit is largely contained in the inductor, the objective is to have the ratio of the reactance of the inductor to its a-c resistance as high as possible. This ratio is known as the **Quality** or  $Q$  of an inductor and is usually expressed by the equation—

$$Q = \frac{X_L}{R} = \frac{2\pi fL}{R} \quad (16:6)$$

The  $Q$  of resonant circuits used in practice varies from values in the order of a hundred in the case of iron-core coils to values as high as 20 thousand or more in the case of certain cavity resonators.

In a series resonant circuit at the resonant fre-

quency, the voltage developed across the capacitor is  $Q$  times the net voltage across the coil and capacitor, or  $E_c = QE$ . In a parallel-resonant circuit when  $Q$  is large, the total impedance of the combination at the resonant frequency is  $Q$  times the coil or capacitor reactance, or  $Z = X_L Q$  (at resonance  $X_c = X_L$ ).

**Example:** In a parallel-resonant circuit having an inductance of 50 microhenries and a resistance of 10 ohms, what is the  $Q$  of the circuit and what is the impedance at a resonant frequency of 1000 kc?

**Solution:**

50 microhenries =  $5 \times 10^{-5}$  henries; 1000 kc =  $10^6$  cycles.

$$X_L = 2\pi fL = 6.28 \times 10^6 \times 5 \times 10^{-5} = 314 \text{ ohms}$$

$$Q = \frac{X_L}{R} = \frac{314}{10} = 31.4$$

$$Z = X_L Q = 314 \times 31.4 = 9860 \text{ ohms. Ans.}$$

## CHAPTER 17

### REPEATING COILS AND TRANSFORMERS

#### 17.1 Mutual Induction

The inductive effects discussed in Chapter 8 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** emf was superposed upon the current resulting from the **impressed** emf.

In practice, we may experience inductive effects in circuits other than the one in which the current due to the impressed emf 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 electric circuit that is responsible for the effect is known as its mutual inductance.**

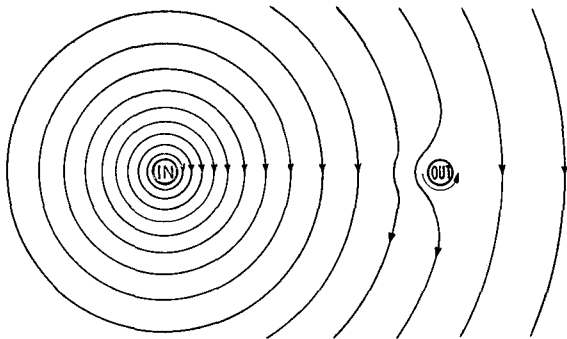


FIGURE 17-1

#### 17.2 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 17-1 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. This induces an emf 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.

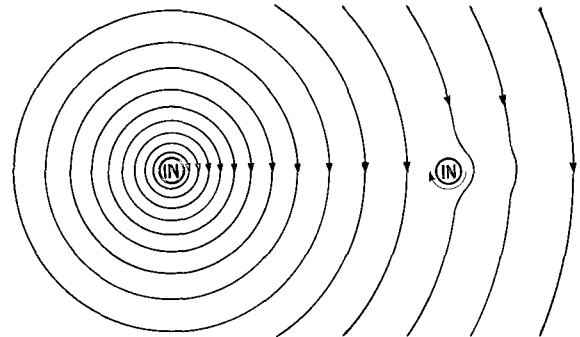


FIGURE 17-2

If the current in the first conductor is decreased, we have the reverse condition, or that shown in Figure 17-2. Here the lines, instead of expanding and moving outward, are contracting and moving inward, again cutting the second conductor, 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 emf may be expressed as follows: **For any two parallel conductors, a current in one increasing in value induces an emf in the other, tending to establish a current in the opposite direction, and a current decreasing in one will induce an emf in the other, tending to produce a current in the same direction.**

Instead of the two single conductors shown in Figures 17-1 and 17-2, let us consider two separate coils, one inside the other, as in Figure 17-3.

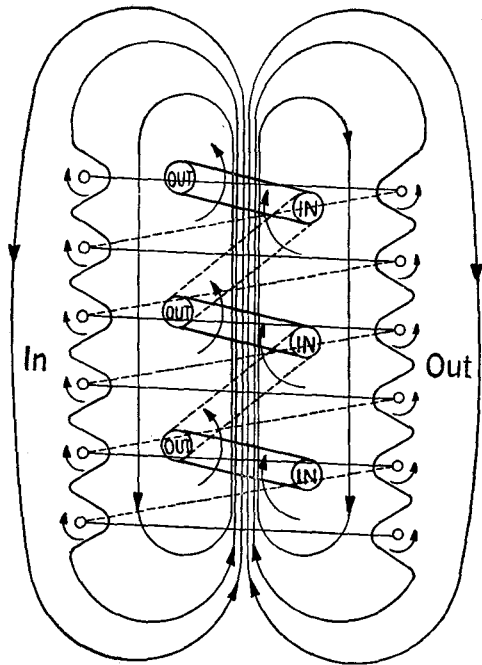


FIG. 17-3 PRINCIPLE OF INDUCTION COIL

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 relatively 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.

If now the two separate coils of Figure 17-3 are wound on the same iron core in the manner indicated by Figure 17-4, 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 emf of the secondary

winding is equal to the impressed emf 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:

$$\frac{E_P}{E_S} = \frac{N_P}{N_S} \quad (17:1)$$

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 emf 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 will 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—

$$E_S = N_S \times v \quad (17:2)$$

where *N<sub>s</sub>* is the number of turns on the secondary.

In the primary the induced emf must be exactly

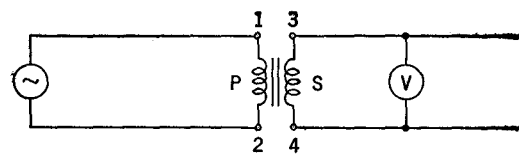


FIG. 17-4 ELEMENTARY TRANSFORMER CIRCUIT



equal and opposite to the impressed emf since the emf due to  $IR$  drop is practically negligible. This could be expressed by an equation similar to equation (17:2), thus—

$$E_P = N_P \times v \quad (17:3)$$

Since  $v$  is the same in both equations (17:2) and (17:3), we may derive equation (17:1) by dividing (17:3) by (17:2).

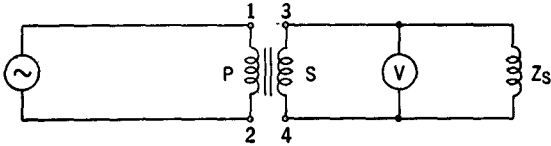


FIGURE 17-5

If a load in the form of a shunting impedance is connected to the secondary of the transformer, as shown in Figure 17-5, the induced emf in the winding  $S$  causes a current to flow through the impedance  $Z_s$ , which can be expressed as follows—

$$I_s = \frac{E_s}{Z_s}$$

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 emf 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 emf 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.

The 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—

$$N_P \times I_P = N_S \times I_S \text{ or } \frac{I_S}{I_P} = \frac{N_P}{N_S} \quad (17:4)$$

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—

$$E_S I_S = E_P I_P$$

from which—

$$\frac{I_S}{I_P} = \frac{E_P}{E_S} \text{ or } \frac{I_S}{I_P} = \frac{N_P}{N_S} \quad (17:4)$$

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 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, that—

$$Z_s = \frac{E_s}{I_s}$$

But the current and voltage of the generator with the transformer inserted between it and the load  $Z_s$  are  $I_P$  and  $E_P$ , 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—

$$Z_P = \frac{E_P}{I_P}$$

We find, then, that—

$$\frac{Z_P}{Z_s} = \frac{E_P}{I_P} \div \frac{E_s}{I_s} = \frac{E_P}{I_P} \times \frac{I_s}{E_s} = \frac{E_P}{E_s} \times \frac{I_s}{I_P} = \frac{N_P}{N_s} \times \frac{N_P}{N_s}$$

or—

$$\frac{Z_P}{Z_s} = \left[ \frac{N_P}{N_s} \right]^2 \quad (17:5)$$

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 17-5 has or may have accomplished.

- a. The characteristics of the electric 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 electric 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 more often concerned with *b* and *c*.

First, let us illustrate the power case by referring to Figure 17-6 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—

$$R = \frac{E}{I} = \frac{10}{100} = \frac{1}{10} \text{ ohm}$$

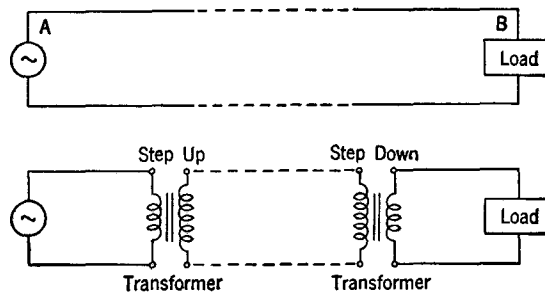


FIG. 17-6 USE OF TRANSFORMERS IN POWER CIRCUIT

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.

### 17.3 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 17-6, 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 transformer 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

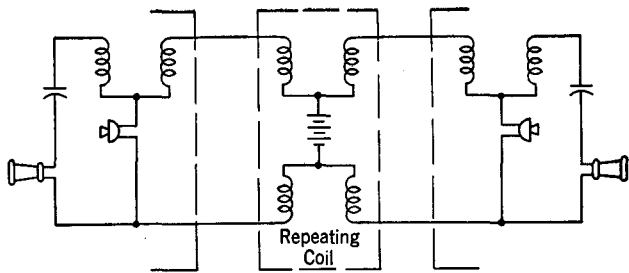


FIG. 17-7 TRANSFORMERS IN COMMON BATTERY TELEPHONE CONNECTION

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 match impedances to permit maximum energy transfer, as explained in later Chapters.

One of the most common applications of the repeating coil in telephone work is in connection with the common battery cord circuit, as illustrated by Figure 17-7. 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. 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 electric 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 as impedance matching devices.

#### 17.4 The Phantom Circuit

Figure 17-8 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 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

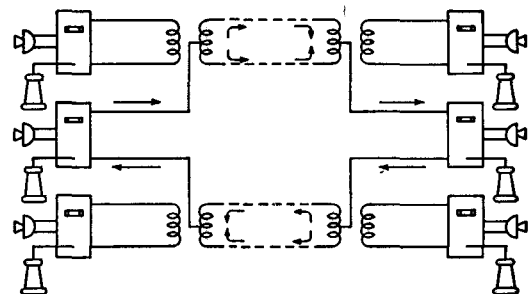


FIG. 17-8 PRINCIPLE OF THE PHANTOM CIRCUIT

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 circuits 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.

### 17.5 Standard Repeating Coils

A number of general types of repeating coils are currently standard in the Bell System. One principal type, illustrated by the 62 and 93 series, has four windings, the terminals of which are designated by numbers as shown by Figure 17-9 (A). The other type, illustrated by the 173 series, has six windings which may be connected as shown in Figure 17-9(B) with four windings on the line

TABLE VII  
STANDARD REPEATING COILS

IMPEDANCE RATIO LINE TO DROP 4-3 AND 8-7 TO 2-1 AND 6-5	93-TYPE	62-TYPE
1:1	93-A	62-A
1:1.62	93-B	62-B
1.62:1	93-F	62-C
2.66:1	93-G	62-E
1.24:1	93-H	
2.28:1	93-J	
1:1.28		62-F
1:2.34		62-G

side, or with the 9-10, 11-12 windings not used, depending on the impedance ratio required. In all types, the windings which are used to form the line side are precision manufactured so as to be as nearly identical electrically as possible. This balance is required, as we have already seen, to avoid crosstalk where the coils are used in phantom operation. The drop windings (that is, 1-2 and 5-6) do not need to be so well balanced in normal use.

The 62- and 93-type coils have toroidal cores made of many turns of fine-gage silicon-steel wire sawed through at one point to introduce a gap in the magnetic circuit. In the 93-type coil this gap is filled with compressed powdered iron which, while increasing slightly the reluctance of the core gives it a high degree of magnetic stability, preventing permanent magnetization under abnormal service conditions. In the 62-type coil the gap in the magnetic circuit is unfilled which tends to make the coil even more stable. This coil is especially well adapted for use on circuits composited for d-c telegraph operation. The same feature, however, tends to make the 62 series inefficient at low frequencies and they cannot be used on circuits employing 20-cycle signaling, whereas the 93 series may be used for such purposes. Standard 173-type coils are built with permalloy cores.

The types of repeating coils discussed above are manufactured with a number of different turn ratios to provide various impedance matching combinations. Table VII gives the standard impedance ratios for 93- and 62-type coils. The 173-type coils are likewise available in a wide range of impedance ratios. The impedance ratio obtained in their use depends on whether all four of the line windings are used and on how those

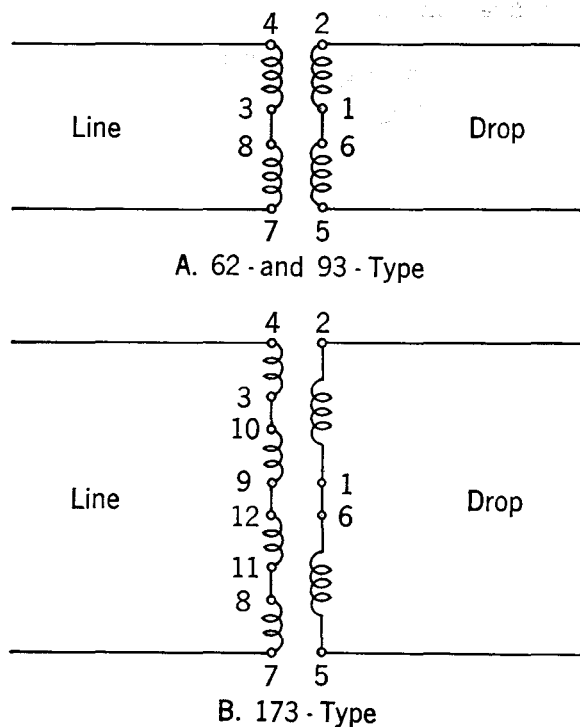


FIG. 17-9 STANDARD REPEATING COILS

used are connected. The impedance ratios that can be obtained accordingly do not lend themselves readily to tabular presentation, but various ratios line-to-drop ranging from as low as 0.6:1 to as high as 2.52:1 may be obtained.

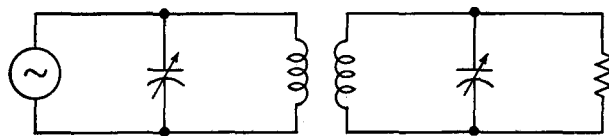


FIGURE 17-10

### 17.6 Air-Core Transformers

At high frequencies such as used in radio, it is impractical to employ transformers having iron cores because the losses set up by circulating or eddy currents in the core would be excessive. Radio frequency transformers are accordingly built without cores and are usually referred to as **air core** transformers. It is not possible, however, to build such a transformer so that all of the flux lines set up by one coil will cut all of the turns of the other coil because the reluctance of air to the flow of magnetic flux is so high. The simple voltage, current and impedance relationships that apply to iron-core transformers as outlined in a preceding Article therefore no longer hold true. In practice, air-core transformers are designed largely on an empirical or "cut and try" basis and large step-ups of voltage cannot ordinarily be obtained.

When air-core transformers are used for coupling between radio frequency or other high-frequency circuits, the primary, secondary, or both coils, are often tuned to resonate at the operating

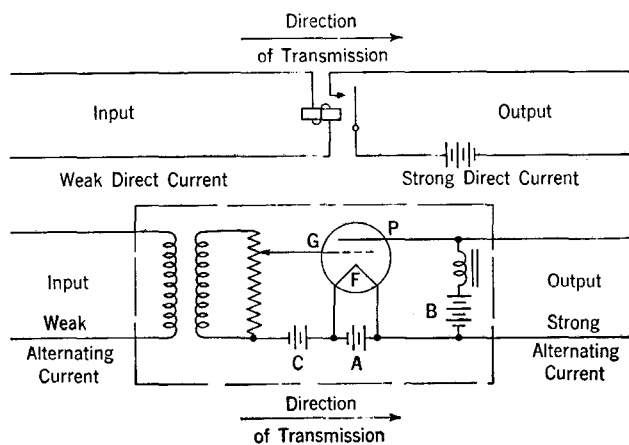


FIGURE 17-11

frequency. Such a circuit is shown in Figure 17-10 where both primary and secondary are tuned by means of variable capacitors. The tuning is affected by adjusting the values of capacitance so as to satisfy the resonance equation—

$$f = \frac{1}{2\pi\sqrt{LC}}$$

In this way parallel-resonant circuits of high impedance are obtained. In addition, a circuit of this sort will act as a filter and pass a band of frequencies the width of which is determined by the  $Q$  of the circuit.

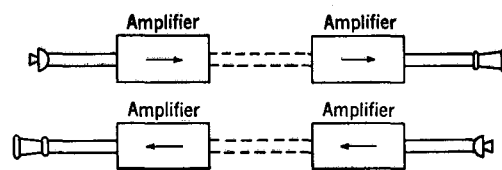


FIGURE 17-12

### 17.7 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 necessary because the device used for amplifying voice-frequency currents operates in one direction 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 17-11).

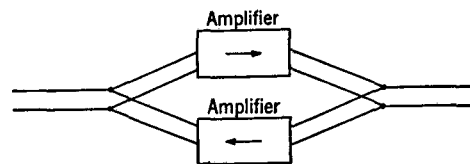


FIGURE 17-13

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 17-12. This would require not only twice the circuit facilities for each long distance connection, but also special telephones at each terminal. It would not be possible for two such amplifiers to be connected at the same point in a telephone circuit as shown in Figure 17-13,

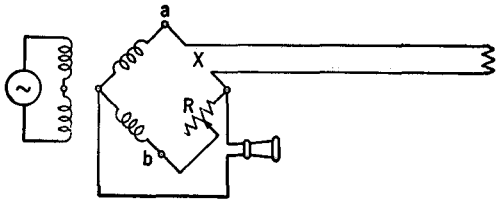


FIGURE 17-14

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.

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, 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 balance principle of the Wheatstone bridge, 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 17-14 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 emf is then impressed across *a* and *b* by mutual induction instead of by direct connection. In place of a 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

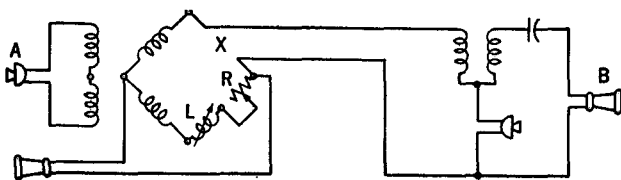


FIGURE 17-15

the *X* terminals. We can 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 17-14, but arranged to measure impedance as well as 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 17-15. 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 variable arm of the bridge adjusted to give perfect balance, any voice current in the transmitter circuit at

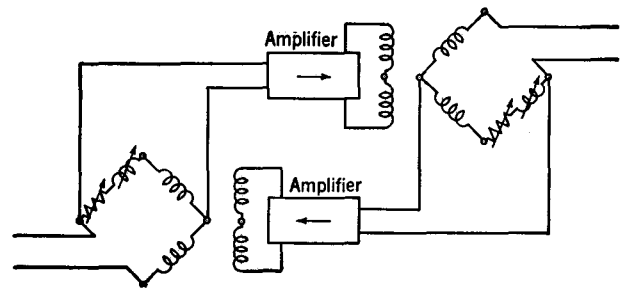


FIGURE 17-16

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.

The coil that takes the place of the bridge mechanism in Figures 17-15 and 17-16, is known as a **hybrid** coil or three winding transformer. 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. The conventional schematic for the hybrid coil is shown in Figure 17-17, which, it will be observed, shows the same circuit connections as Figure 17-15 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 17-18, in order that symmetry in the wiring of the talking circuit may be maintained.

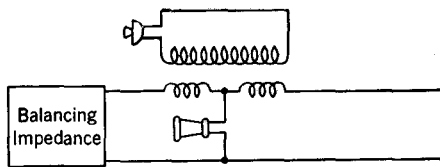


FIGURE 17-17

Both sets of windings, of course, are inductively coupled to the third winding. Figure 17-19 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 17-20, where for convenience the coil is shown as in Figure 17-17 instead of as in Figure 17-18. If  $Z_1$  is the impedance of the telephone line and  $Z_2$  the impedance of the balancing network,

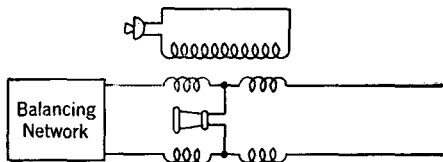


FIG. 17-18 THE HYBRID COIL

$Z_1$  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.

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

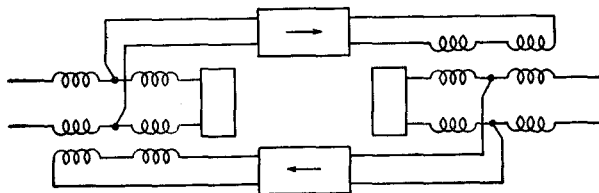


FIG. 17-19 TWO-WIRE TELEPHONE REPEATER CIRCUIT

the impedance of the network circuit, accomplishes no useful purpose and is lost. For this reason alone, the amplifier must 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.

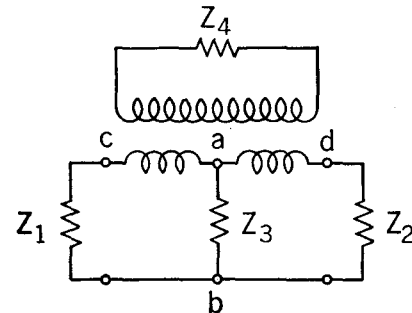


FIGURE 17-20

Hybrid coils are used 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 17-21.

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.

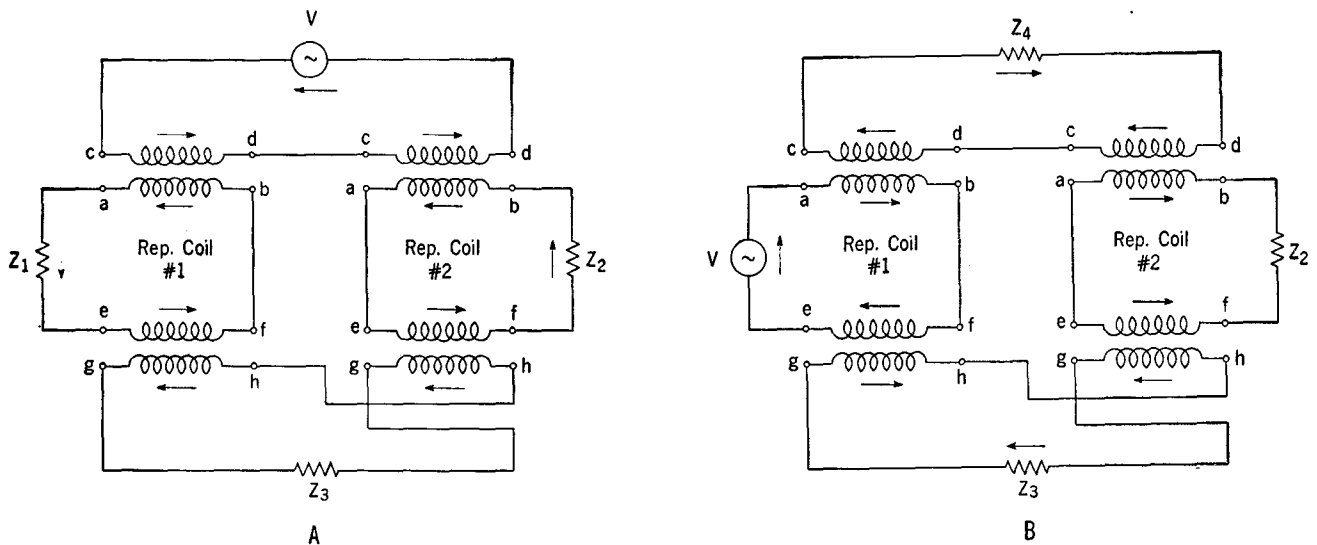


FIG. 17-21 FOUR-WIRE TERMINATING SET

This is illustrated by Figure 17-21 (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 17-21 (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  $a-b$ .

Where repeating coils having six windings are used in a hybrid arrangement, the connections are naturally somewhat different. Thus Figure

17-22 shows the arrangement at one side of a repeater for a two-wire circuit. Here it is assumed that the line and balancing network impedances are equal and that the impedances connected at HYB OUT and HYB IN are equal.

Current received from the line flows through the line windings of both coils and induces equal voltages in the network windings (9-10 and 11-12) of both coils. Because of the reversed poling of the network windings of coil (B) with respect to those of coil (A) the resultant voltage across the network is zero. Equal voltages are also induced in the 2-1 and 6-5 windings of both coils and the power received from the line divides equally between the impedances connected at HYB OUT and HYB IN.

Current received from the output of one amplifier at HYB IN flows through windings 2-1 and 6-5 of the (B) coil and induces equal voltages in the line and network windings. The currents in these windings flow through the corresponding windings of the (A) coil to line and balancing network respectively. The currents in the line and network windings of coil (A) induce equal voltages in the 2-1 and 6-5 windings but these voltages are opposite in phase because of the reversed poling of the network windings and the resultant voltage is zero. The power received from the amplifier divides equally between the line and the balancing network.

This arrangement requires that the windings of a given coil be mutually balanced to a high degree of precision in order to secure good trans-hybrid balance but does not require a high degree of balance between separate coils.

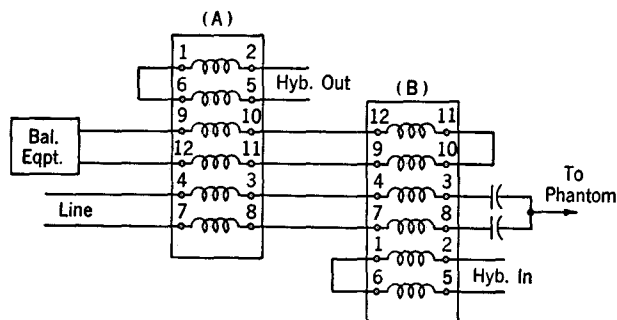


FIG. 17-22 REPEATING COIL HYBRID ARRANGEMENTS FOR 2-WIRE CIRCUITS



## CHAPTER 18

### TELEPHONE TRANSMISSION THEORY

#### 18.1 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 electric energy into some form more useful. In a power transmission line, an electric 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 electric 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 small electric currents into audible vibrations of air called sound waves.

In this and the following two Chapters we shall be concerned with the transmission line or other medium over which energy is carried from the transmitting to the receiving device. Our study will begin with an analysis of the characteristics of ordinary wire conductors at what we may call relatively low frequencies—i.e., voice frequencies and carrier frequencies ranging up to approximately 150 kc. Later we shall consider transmission at higher frequencies, including radio frequencies extending up to several thousand megacycles. This will require us to make some study of certain types of transmission mediums other than the usual wire line. Before beginning our analysis of the properties of the transmission line, however, we may devote some attention to certain basic theorems regarding networks and power transfer that will be useful in the subsequent analysis.

#### 18.2 Transfer of Power

If a transmission system is to accomplish its purpose, it must be so designed that the receiving device is successfully operated as a result of the

application of energy at the source. 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, to keep at a minimum the power lost in transmitting the energy from the source. 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 operating the receiving device**, regardless of what percentage of power may be lost.

The **power efficiency** of a transmission system may be determined by dividing the received power by the power delivered to the system at the generator end. This efficiency will increase as the resistance of the receiving device increases, and the most efficient operation is accordingly obtained when the receiving circuit has high resistance. In most telephone applications, on the other hand, we are interested in receiving all the power possible regardless of the efficiency. In a system containing only resistance **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 circuit the maximum amount of power, but in doing so will operate at an efficiency of only 50 per cent.

Figure 18-1 shows a curve which represents the

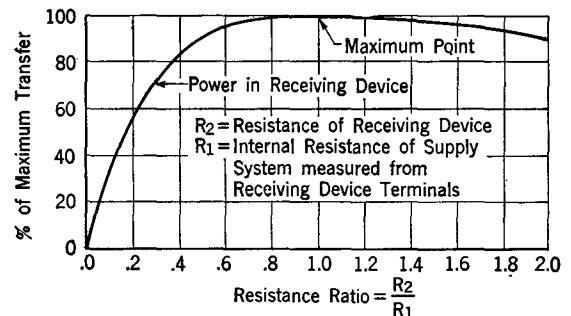


FIG. 18-1 POWER TRANSFER AS A FUNCTION OF RESISTANCE RATIO

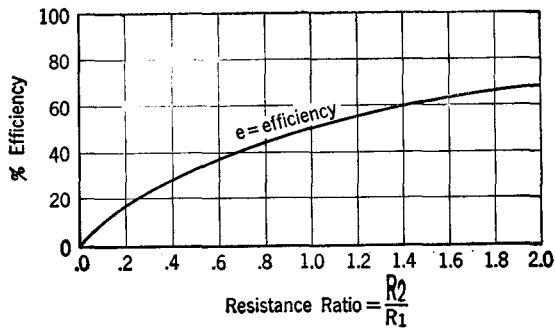


FIG. 18-2 EFFICIENCY AS A FUNCTION OF RESISTANCE RATIO

power in a receiving circuit for various values of its resistance expressed as a ratio of the fixed resistance of the supplying circuit. Figure 18-2 shows the efficiency for the same conditions.

Maximum power transfer in a-c circuits, where reactances are involved, is obtained in the general case when the impedance of the receiving device is conjugate to the impedance looking back into the network—that is, when the resistances are equal and the reactances are equal and opposite in sign.

### 18.3 Thevenin's Theorem

For the purpose of simplifying electrical calculations, we can consider any electric system as one network supplying energy to another. **The first of these networks may then be replaced by an equivalent simplified circuit, which consists of an emf and an impedance in series.**

This means that regardless of how complicated an electric 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 an impedance. This principle is called **Thevenin's Theorem** and Figure 18-3 illustrates its use. Here  $E$  is a source of electromotive force connected to a complicated network. If it is desired to connect some receiving device to terminals  $A$  and  $B$ , the effect of this electric system on the receiving device will be the same as that of the electric system shown by Figure 18-4 where  $E'$  is the voltage between ter-

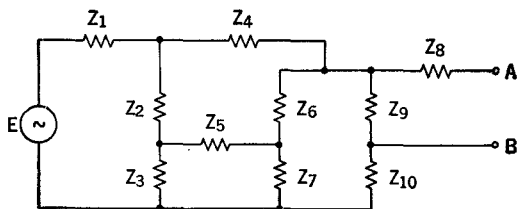


FIGURE 18-3

minals  $A$  and  $B$  of Figure 18-3, and  $R'$  is the impedance measured or calculated from the same terminals with the electromotive force  $E$  considered as having zero value. Thevenin's Theorem may be briefly stated as follows:

**The current supplied to a receiving device connected to two terminals of any electrical system is equal to the voltage between these terminals before the device is connected, divided by the impedance of the receiving device plus the impedance measured or calculated across the terminals with any sources of emf considered as having zero value.**

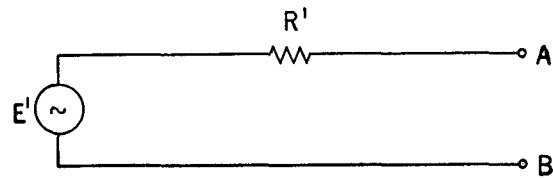


FIG. 18-4 APPLICATION OF THEVENIN'S THEOREM TO THE NETWORK OF FIG. 18-3

### 18.4 Equivalent Networks

Thevenin's Theorem gives us a method of substituting a simple equivalent circuit for any complicated electric 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 electric system, which will supply to some receiving device the same current as the electric system and will take from the same source of electromotive force the same current as the electric system.** A network consisting of three

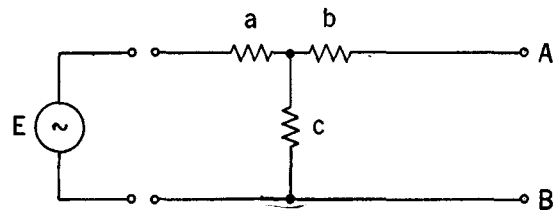


FIGURE 18-5

impedances of proper value arranged in the form of a T as shown by Figure 18-5 may be substituted for such a complex network as that shown by Figure 18-3, 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  $\pi$ .

In determining values for the three impedances in an equivalent T-network such as is shown by Figure 18-5, the following equations may be used:

$$\text{Impedance of } a = Z_1 - c \quad (18:1)$$

$$b = Z_3 - c \quad (18:2)$$

$$c = \sqrt{(Z_1 - Z_2)Z_3} \quad (18:3)$$

where  $Z_1$  is the impedance of the complicated network at the terminals connected to the source of emf with the receiving device terminals open;  $Z_2$  is the same with the receiving device terminals short-circuited; and  $Z_3$  is the impedance of the complicated network at the receiving device terminals with the source terminals open.

At a single frequency, a long transmission line can be exactly represented electrically by a simple three-element equivalent network, although the determination of the values of the three arms involves in this case the use of certain higher branches of mathematics. For many purposes, it is convenient to deal with a long transmission line by considering it as being made up of a number of separate T-sections, each identical and each representing approximately a short section of the transmission line. Obviously the degree of approximation will be higher the larger the number of sections used to represent a given line. Figure 18-6 illustrates such a multisection uniform network where each T-section consists of two series arms of value  $Z_1/2$  and a shunt arm  $Z_2$ .

### 18.5 Characteristic Impedance

One thing that we will ordinarily wish to know about a transmission line, or a multisection uniform network representing such a line, is the input impedance of the line. If a line is very long, this input impedance will be found to depend entirely upon the characteristics of the line itself and will not be affected by what, if anything, is connected to the far end of the line.

If we should measure or calculate the impedance

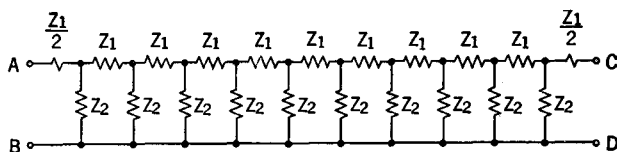


FIG. 18-6 MULTISECTION UNIFORM NETWORK

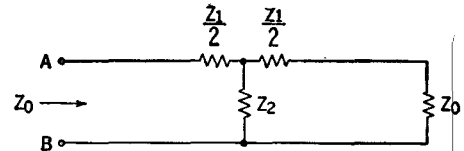


FIGURE 18-7

of the network of Figure 18-6 across the input terminals  $A$  and  $B$ , we would obtain a certain impedance value which we may call  $Z'_0$ . Now let us connect this impedance  $Z'_0$  across the receiving terminals  $C, D$ . As seen from the sending end, this will make the network 20 sections long instead of 10 sections long. If we then repeat this process, we will arrive at a new impedance which connected across the terminals  $C, D$  will make the network 30 sections long. Very soon we will arrive at an impedance value which will not change; in other words, the network will become infinitely long for all practical purposes. This impedance is the **characteristic impedance**  $Z_0$  of the network. That  $Z_0$  depends only on the impedance values of the unit sections of the network is obvious from the fact that those were the only impedance values used in determining it. This being so, it follows that if we take one section of the network and terminate it with its characteristic impedance  $Z_0$  as shown in Figure 18-7, the input impedance as shown across  $A$  and  $B$  will also be  $Z_0$ .  $Z_0$  is then readily calculated from the Figure as follows:

$$Z_0 = \frac{Z_1}{2} + \frac{Z_2 \left( \frac{Z_1}{2} + Z_0 \right)}{Z_1 + Z_2 + Z_0}$$

$$\frac{Z_0 Z_1}{2} + Z_0 Z_2 + Z_0^2 = \frac{Z_1^2}{4} + \frac{Z_1 Z_2}{2} + \frac{Z_0 Z_1}{2} + \frac{Z_1 Z_2}{2} + Z_0 Z_2$$

$$Z_0^2 = \frac{Z_1^2}{4} + Z_1 Z_2$$

$$Z_0 = \sqrt{\frac{Z_1^2}{4} + Z_1 Z_2} \quad (18:4)$$

As has already been pointed out, the more identical T-sections that are used in the multisection network to represent a given length of transmission line, the more closely does the network approximate the line. If we take a transmission line having a series impedance  $Z_1$  per mile and a shunt impedance  $Z_2$  per mile and represent each mile by  $n$  T-network sections, the impedance of each series T-arm will be  $Z_1/2n$  and each shunt arm will be  $nZ_2$ . The characteristic impedance of the network may then be written from equation (18:4) as—

$$Z_0 = \sqrt{\frac{Z_1^2}{4n^2} + \frac{Z_1}{n} n Z_2}$$

From this it may be seen that as the number of sections  $n$  is increased to a very large number, the first term under the radical approaches zero and the equation reduces to—

$$Z_0 = \sqrt{Z_1 Z_2} \quad (18:5)$$

Actually the use of higher mathematics, as previously mentioned, will show that this is the exact expression for the characteristic impedance of a uniform transmission line.

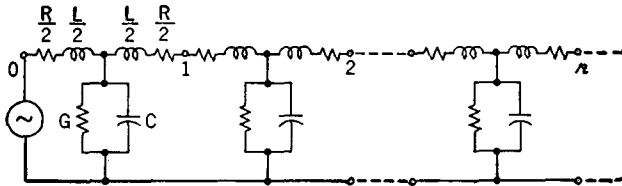


FIGURE 18-8

The fundamental transmission characteristics of the wire transmission line are determined by four basic properties known as the “primary constants” of the line. These are the series resistance of the conductors,  $R$ ; the inductance of the conductors,  $L$ ; the capacitance between the two conductors,  $C$ ; and the leakage resistance or conductance between the two conductors,  $G$ .  $R$  and  $L$  are series constants and  $C$  and  $G$  are shunt constants. In terms of these constants, a multisection uniform network simulating the transmission line may be represented as shown by Figure 18-8. The series impedance of a single section of this network may be written as—

$$Z_1 = R + j\omega L \quad (18:6)$$

where the symbol  $\omega$  is used for convenience to represent  $2\pi f$ . The impedance of the parallel shunt branch similarly is—

$$Z_2 = \frac{1}{G + j\omega C} \quad (18:7)$$

From equation (18:5) the characteristic impedance may then be written as—

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (18:8)$$

Here  $R$  is resistance in ohms per unit length of line;  $L$  is inductance in henries;  $G$  is conductance in mhos;  $C$  is capacitance in farads.

As the above equation shows, the value of characteristic impedance is dependent solely on the

primary constants of the line and the frequency. At any given frequency, therefore, this impedance has a fixed value for any given type of line regardless of the length of the line or what may be connected to the line terminal. One of its most useful applications in practical telephone work lies in the fact that receiving devices to be connected to a line may be designed with impedances equal to the characteristic impedance of the line, thus permitting a maximum transfer of power out of the line.

## 18.6 Propagation Constant

If a transmission line having characteristic impedance  $Z_0$  is infinitely long, or is terminated at the receiving end with an impedance equal to its characteristic impedance, the current entering the line when a voltage  $E$  is applied at the sending end will obviously be  $I_0 = E/Z_0$ . Our next main interest in the line will be to know what part of this current or applied energy will reach the receiving device. We may also be interested in knowing how long it will take it to get there. To answer these questions we may again simulate the line by a multisection network and analyze what happens in each individual section. Clearly, since there is a shunt path across the line in each section, the current leaving a section will be less than that entering it in the direction of transmission. But since the sections are all identical, the ratio of leaving to entering current will be the same for all sections; that is—

$$\frac{I_1}{I_0} = \frac{I_2}{I_1} = \frac{I_3}{I_2} \text{ etc.}$$

Accordingly, we may write generally for the current at the end of the  $n$ th section—

$$\frac{I_n}{I_0} = \frac{I_1}{I_0} \times \frac{I_2}{I_1} \times \frac{I_3}{I_2} \times \dots \times \frac{I_n}{I_{n-1}}$$

or

$$\frac{I_n}{I_0} = \left[ \frac{I_n}{I_{n-1}} \right]^n \quad (18:9)$$

An equation such as (18:9) is most conveniently dealt with by means of logarithms. Expressed in this way it may be written—

$$\log_e \frac{I_n}{I_0} = -n\gamma \quad \text{or} \quad 2.303 \log \frac{I_n}{I_0} = -n\gamma \quad (18:10)$$

where  $\gamma$  is defined as the **propagation constant** of the transmission line. Its value is dependent on the primary constants of the line and the fre-

quency. It may be determined to any desired degree of approximation by an analysis of the current relationships in a single network section terminated in  $Z_0$  as illustrated by Figure 18-7. Its exact value will thus be found to be—

$$\begin{aligned} \gamma &= \sqrt{\frac{Z_1}{Z_2}} = \sqrt{(R + j\omega L) \div \frac{1}{G + j\omega C}} \\ &= \sqrt{(R + j\omega L)(G + j\omega C)} \quad (18:11) \end{aligned}$$

The propagation constant  $\gamma$  is a vector quantity because it is equal to  $\sqrt{Z_1/Z_2}$  and both  $Z_1$  and  $Z_2$  are vector quantities. It may accordingly be separated into a “real” and “imaginary” component thus—

$$\gamma = \alpha + j\beta \quad (18:12)$$

where  $\alpha$ , the real part of  $\gamma$ , is a measure of the amount by which the current or voltage is reduced in each unit length of line and is called the **attenuation constant**. Similarly  $\beta$ , the imaginary part of  $\gamma$ , provides a measure of the phase shift, and is related to the time required for the energy to traverse the unit length of line. It is called the **phase** or **wavelength constant**. It should be noted that while  $\beta$  represents an angle of phase shift, it should not be confused with any phase angle which may exist between the current and voltage at a particular point on a line. In practice it is usually easier to evaluate  $\alpha$  and  $\beta$  by making use of equations (18:11) and (18:12). They may be expressed, however, directly in terms of the primary constants— $R$ ,  $L$ ,  $G$  and  $C$ , as follows:

$$\alpha = \sqrt{\frac{1}{2}\sqrt{(R^2 + \omega^2 L^2)(G^2 + \omega^2 C^2)} + \frac{1}{2}(GR - \omega^2 LC)} \quad (18:13)$$

$$\beta = \sqrt{\frac{1}{2}\sqrt{(R^2 + \omega^2 L^2)(G^2 + \omega^2 C^2)} - \frac{1}{2}(GR - \omega^2 LC)} \quad (18:14)$$

The value of  $\beta$  gives directly in radians the phase shift in current or voltage over a unit length of line, such as one mile, having primary constants  $R$ ,  $L$ ,  $G$  and  $C$ . (A radian is  $360/2\pi$  or 57.3 degrees.) For a length of line comprising  $n$  unit lengths the total phase shift will obviously be  $n\beta$  radians. The value of  $\alpha$  may be used to determine the current or voltage attenuation by applying the real part of equation (18:10). Thus, for a line  $n$  units in length—

$$-n\alpha = 2.303 \log \frac{I_n}{I_0} = 2.303 \log \frac{E_n}{E_0} \quad (18:15)$$

where the current and voltage ratios express mag-

nitudes only without regard to any phase differences that may exist. The power entering the line is—

$$P_0 = E_0 I_0 \cos \theta$$

and the power at the end of the  $n$ th unit length is—

$$P_n = E_n I_n \cos \theta$$

$\theta$  in both cases being the angle of the characteristic impedance of the line. The magnitude of the power loss or attenuation may then be determined from—

$$2.303 \log \frac{P_n}{P_0} = -2n\alpha \quad (18:16)$$

The power, therefore, is seen to die out or attenuate in a ratio which is the square of the current and voltage ratio.

In the foregoing we find for the most part a mathematical significance of  $\alpha$  and  $\beta$ . 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 emf impressed on the sending end of a multisection network, and consider separately the effects of inductance and capacitance on the propagation.

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 18-9 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 emf  $E_1$ , which is the sum of the  $IR$  drop

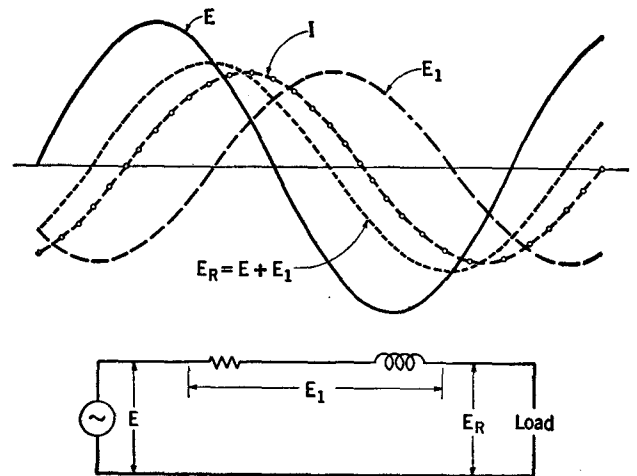


FIG. 18-9 VOLTAGES AND CURRENT IN AN INDUCTIVE CIRCUIT

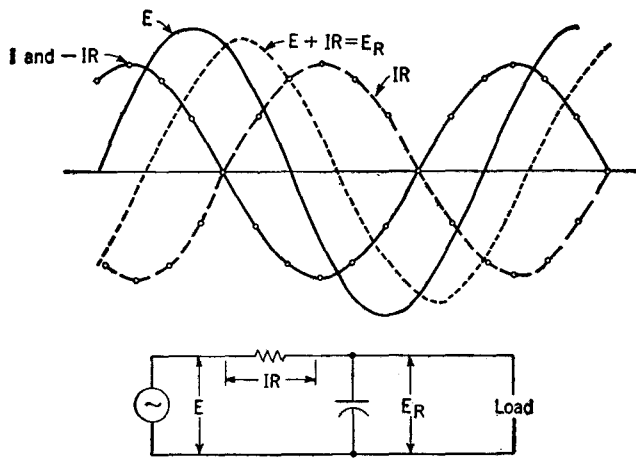


FIG. 18-10 VOLTAGES AND CURRENT IN A CAPACITIVE CIRCUIT

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 behind  $E$ , the original voltage. A circuit containing resistance and capacitance, on the other hand, produces a leading current as shown by Figure 18-10, and this current produces an  $IR$  drop which is opposite in phase to the current. Now if we combine the  $IR$  drop and the voltage, we obtain the resultant voltage  $E_R$ , which exists across the capaci-

tor and the load. This voltage likewise lags behind  $E$ , the original voltage.

In both cases we have obtained a resultant voltage which lags behind the impressed voltage. Bridged capacitance assists series inductance in the phase retarding effect. Due to the presence of reactance, therefore, the voltage has been "delayed in time", so that the maximum voltage reaches any given point along the network later than it would if the reactance were not present. Each section of the network, due to resistance and leakage, absorbs energy and thereby 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 **time** and **geographical distance** traveled along the line.

To bring this out clearly, let us assume that we take 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

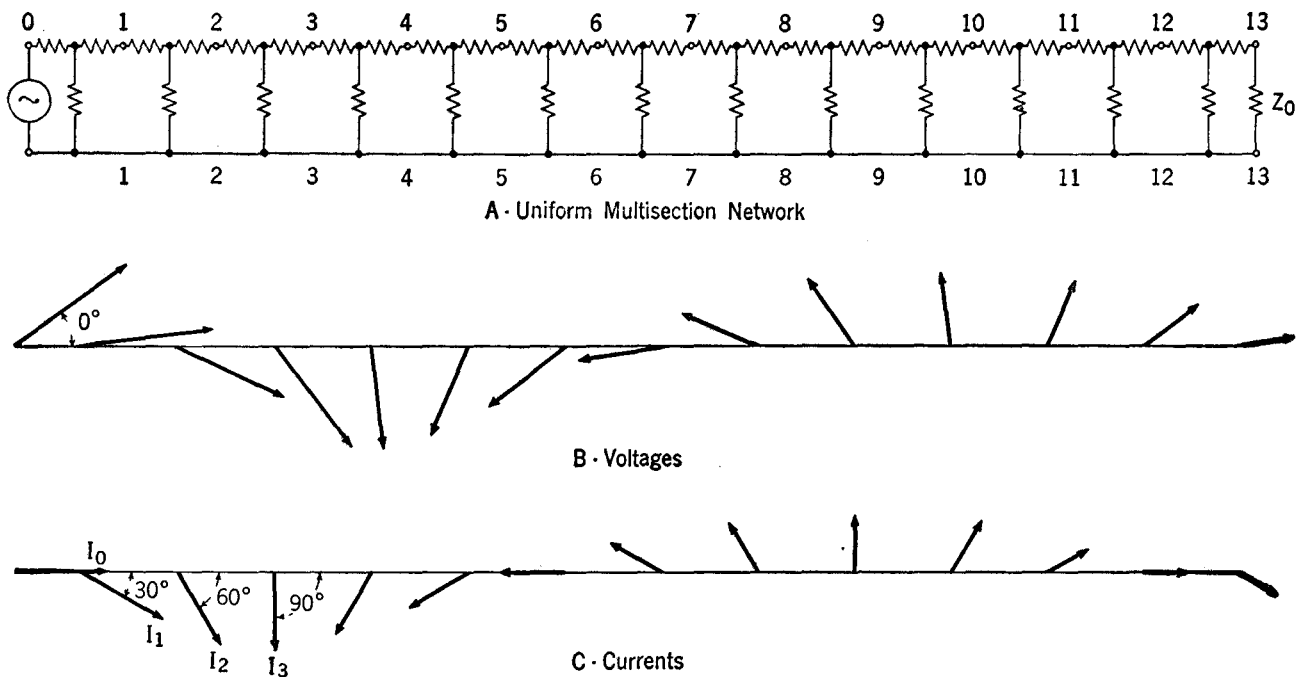


FIG. 18-11 TRANSMITTED CURRENTS AND VOLTAGES AT JUNCTIONS OF A MULTISECTION UNIFORM NETWORK

such as to give a ratio of 0.895 between the end and beginning of each 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  $30^\circ$  behind  $E_0$ .  $E_2$ , at the end of the second section, will be  $0.895 \times 8.95$  or 8.01 volts, lagging  $30^\circ$  behind  $E_1$  or  $60^\circ$  behind  $E_0$ . If we represent the voltages at various points by vectors, we will obtain a system of vectors as shown in Figure 18-11(B), where the multisection network is shown as Figure 18-11(A) and the voltage acting at each junction is directly below.

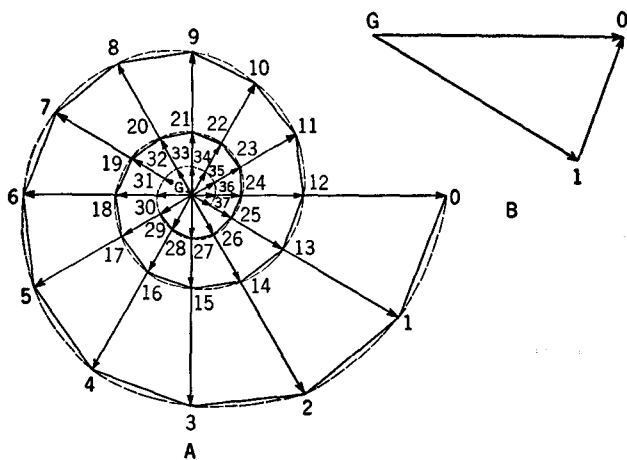


FIG. 18-12 POLAR DIAGRAM OF THE VECTORS OF FIG. 18-11(C)

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  $\theta$  from the corresponding voltage vector, where  $\theta$  is the angle of the characteristic impedance  $Z_0$ . Thus we may treat a similar Figure such as 18-11(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 18-12(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 18-12(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  $90^\circ$  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  $30^\circ$  for every section traversed and diminishing in value about 10% in each section.

Figures 18-11 and 18-12 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 18-13 which shows clearly how the current reverses in direction as it passes through the various sections. It also shows how the current decreases in value or is attenuated as it passes along the line. It will be noted that at the instant shown the input current is at a maximum value flowing in a positive direction; at a point three sections along the line no current is flowing; at a point six sections along the line maximum current is flowing in a direction opposite to that of the current at the input end; at nine sections there is again no current while at 12 sections a maximum positive current is flowing.

The distance between adjacent points where maximum positive (or negative) current is flowing is known as the wavelength and is represented by the symbol  $\lambda$ . The time required to set up one wavelength along the line is naturally equal to the time required for the impressed voltage to complete one cycle. The velocity of propagation of the energy along the line is therefore equal to the wavelength divided by the time required to establish the wave, which is the reciprocal of the applied frequency; thus—

$$W = \frac{\lambda}{\frac{1}{f}} = f \lambda \quad (18:17)$$

This velocity may range in value in telephone circuits from as little as 10,000 miles per second

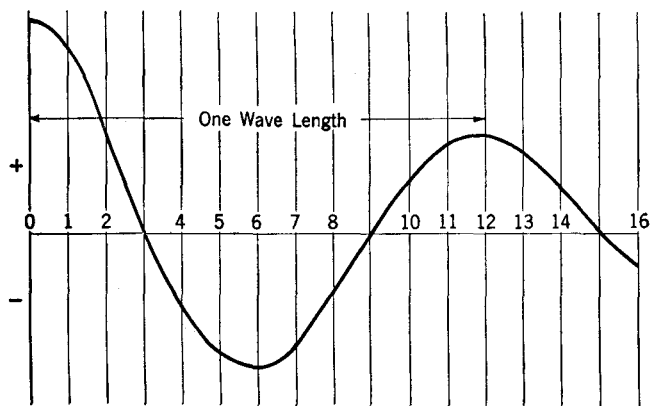


FIG. 18-13 INSTANTANEOUS CURRENT ALONG UNIFORM TRANSMISSION LINE

to a maximum approaching the speed of light in a vacuum, some 186,000 miles per second.

The value of the wavelength  $\lambda$  is readily determined from the value of  $\beta$  which, as we have seen, depends on the primary constants of the line and the frequency.  $\beta$  gives in radians the phase shift or delay per unit section. The total phase shift for one wavelength is obviously  $360^\circ$  and since  $360^\circ$  equals  $2\pi$  radians we may write the equation—

$$\lambda = \frac{2\pi}{\beta} \quad (18:18)$$

The equation for velocity may then be rewritten

$$W = \frac{2\pi f}{\beta} = \frac{\omega}{\beta} \quad (18:19)$$

**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 wavelength, and (6) the velocity of propagation. The equivalent distributed constants per loop mile of this particular circuit at 1000 cycles are as follows:

$$\begin{aligned} R &= 89.7 \text{ ohms} & C &= .062 \text{ } \mu\text{f} \\ L &= .040 \text{ henry} & G &= 1.5 \text{ } \mu\text{mhos} \end{aligned}$$

**Solution:**

$$\omega = 2\pi f = 2 \times 3.1416 \times 1000 = 6283$$

From equation (18:8)

$$\begin{aligned} Z_0 &= \sqrt{\frac{R + j\omega L}{G + j\omega C}} \\ &= \sqrt{\frac{89.7 + j6283 \times .040}{1.5 \times 10^{-6} + j6283 \times .062 \times 10^{-6}}} \\ &= \sqrt{\frac{266.8 / 70^\circ 21'}{389.5 \times 10^{-6} / 89^\circ 47'}} \\ &= \sqrt{684,980 / -19^\circ 26'} \\ &= 827.5 / -9^\circ 43' \text{ Ans. (1)} \end{aligned}$$

The input power,  $P_0$ , is

$$P_0 = E_0 I_0 \cos \theta$$

Substituting

$$I_0 = \frac{E_0}{Z_0} \text{ in the above,}$$

$$P_0 = \frac{E_0^2 \cos \theta}{Z_0}$$

or

$$E_0^2 = \frac{P_0 Z_0}{\cos \theta}$$

(Note: When  $Z_0$  is a pure resistance,  $\theta$  is zero and its cosine is one. Therefore, when  $\theta$  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

$$\begin{aligned} I_0 &= \frac{E_0}{Z_0} \\ &= \frac{2.90}{827.5} = .0035 \text{ ampere or } 3.5 \text{ milliamperes} \end{aligned}$$

From equation (18:11)

$$\begin{aligned} \gamma &= \sqrt{(R + j\omega L)(G + j\omega C)} \\ &= \sqrt{266.8 / 70^\circ 21' \times 389.5 \times 10^{-6} / 89^\circ 47'} \\ &= \sqrt{103,920 \times 10^{-6} / 160^\circ 8'} \\ &= .3222 / 80^\circ 4' \end{aligned}$$

From equation (18:12)

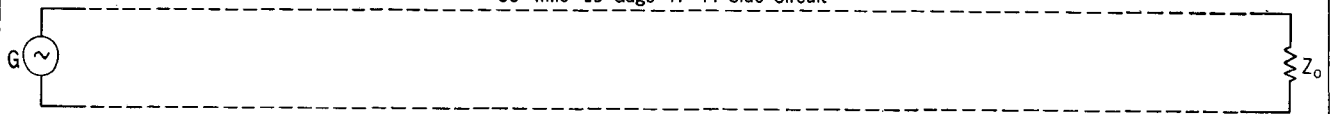
$$\begin{aligned} \gamma &= \alpha + j\beta \\ &= .3222 \cos 80^\circ 4' + j.3222 \sin 80^\circ 4' \\ &= .0556 + j.3174 \end{aligned}$$

From equation (18:15)

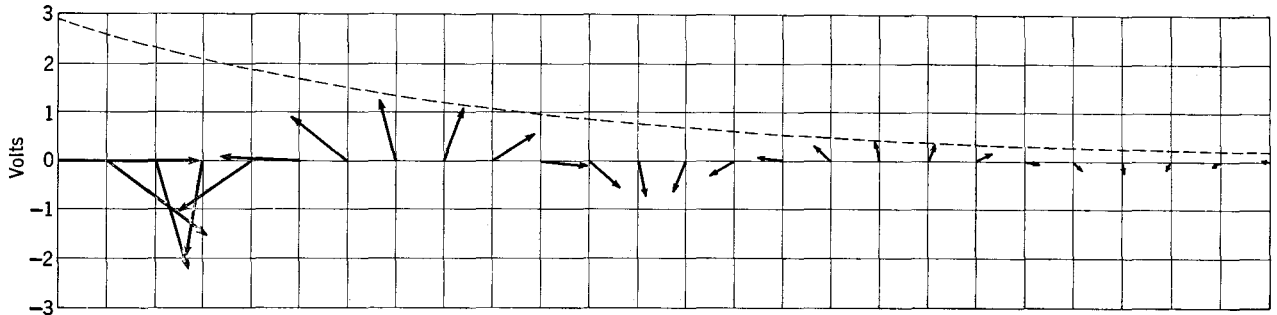
$$2.303 \log \frac{I_n}{I_0} = -n \alpha$$



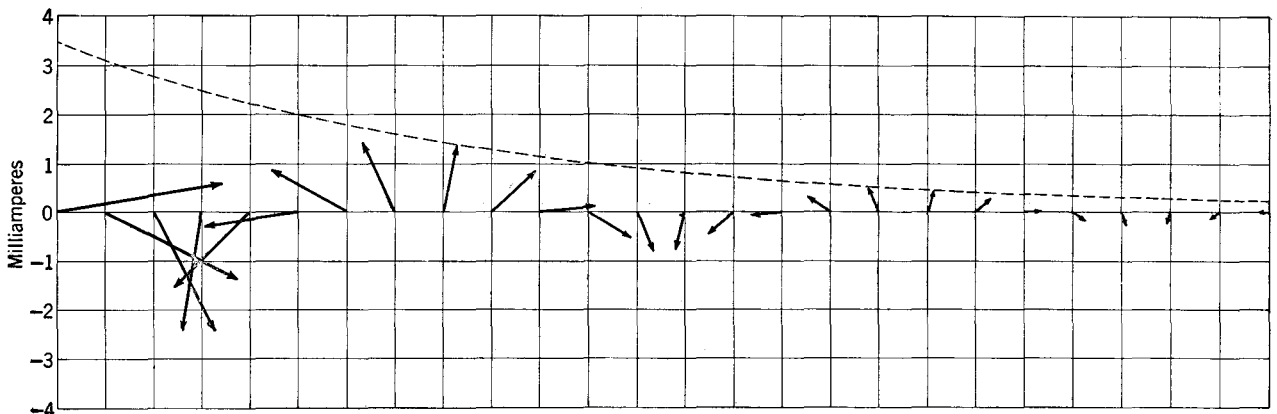
50 - Mile 19 Gage H - 44 Side Circuit



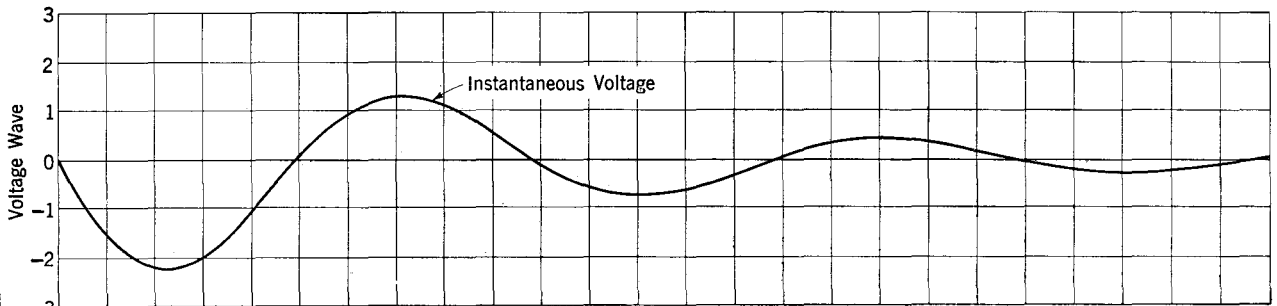
A



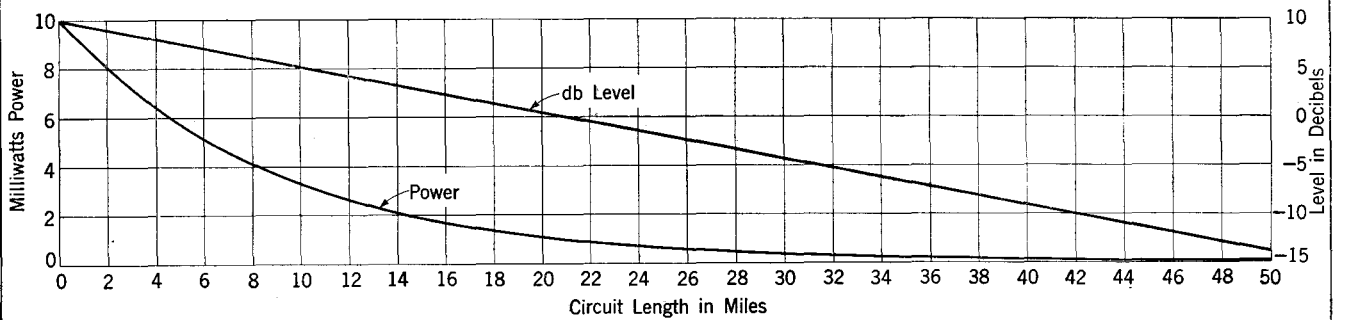
B



C



D



E

FIGURE 18-14  
[ 162 ]

or

$$2.303 \log \frac{I_0}{I_n} = n a$$

$$\log \frac{3.50}{I_n} = \frac{50 \times .0556}{2.303} = 1.207$$

$$\frac{3.50}{I_n} = 16.11$$

$$I_n = \frac{3.50}{16.11} = .22 \text{ milliamperes} \quad \text{Ans. (2)}$$

Phase shift per mile

$$\beta = .3174 \text{ radian or } 18.2^\circ$$

Total phase shift for 50-mile circuit is

$$50 \times 18.2 = 910^\circ \quad \text{Ans. (3)}$$

Then  $I_n = .22 / -910^\circ$  milliamperes

From equation (18:15)

$$2.303 \log \frac{E_0}{E_n} = n a$$

$$\log \frac{2.90}{E_n} = \frac{50 \times .0556}{2.303} = 1.207$$

$$\frac{2.90}{E_n} = 16.11$$

$$E_n = \frac{2.90}{16.11} = .18 \text{ volt}$$

Then

$$P_n = E_n I_n \cos \theta$$

$$= .18 \times .00022 \times .9856$$

$$= .000039 \text{ watt or } .039 \text{ milliwatt. Ans. (4)}$$

From equation (18:18)

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

$$= \frac{6.283}{.3174} = 19.79 \text{ miles} \quad \text{Ans. (5)}$$

From equation (18:17)

$$W = f\lambda$$

$$= 1000 \times 19.79$$

$$= 19,790 \text{ miles per second} \quad \text{Ans. (6)}$$

The conditions along the line are graphically illustrated by Figure 18-14. 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 instantaneous voltage on the line is shown in *D*. 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 percentagewise than either the effective voltage or current values illustrated by the dashed curves in *B* and *C*.

## CHAPTER 19

### TELEPHONE TRANSMISSION THEORY—(Continued)

#### 19.1 Loading

From the equations for attenuation and velocity developed in the preceding Article it will be seen that in the general case both quantities are functions of frequency. For telephonic transmission over long lines variation of these quantities with frequency is obviously undesirable because it results in distortion of the transmitted signal. Thus frequencies at the upper end of the voice range might suffer more attenuation than frequencies at the lower end of the range; also they might differ from the lower frequencies in the time taken in reaching the receiving point. The seriousness of such distortions in practice of course would depend on the extent of the attenuation and phase shift variations with frequency. Optimum transmission conditions would be obtained with a line where  $\alpha$  was as low in value as possible and entirely independent of frequency; and  $\beta$  was so related to frequency that  $W$  would be independent of that factor. Such a condition can be realized in theory with a line whose primary constants are of such values that  $LG = RC$ . When this is true, the value of  $\alpha$  as determined from equation (18:13) reduces to—

$$\alpha = \sqrt{RG} \quad (19:1)$$

and  $\beta$  from equation (18:14) becomes—

$$\beta = \omega\sqrt{LC} \quad (19:2)$$

Here  $\alpha$  is independent of frequency as desired, and  $\beta$  is proportional to frequency in such a way that—

$$W = \frac{\omega}{\beta} = \frac{\omega}{\omega\sqrt{LC}} = \frac{1}{\sqrt{LC}} \quad (19:3)$$

which is also independent of frequency.

Under the same optimum if somewhat hypothetical condition, the value of the characteristic impedance works out to be approximately—

$$Z_0 = \sqrt{\frac{L}{C}} \quad (19:4)$$

Again this is independent of frequency, which is desirable because the characteristic impedance of the line can then be matched by pure resistance.

Unfortunately, a transmission line having such optimum characteristics is not readily attainable in practice. The value of  $G$  is actually very small in normal transmission lines and it would not be desirable to increase it artificially because that would tend to increase the attenuation correspondingly. The value of  $C$  likewise cannot be changed much because of practical considerations of wire spacing. To attain the optimum condition where  $LG = RC$ , therefore, it would be necessary either to increase the value of  $L$  substantially or to decrease the value of  $R$ . The latter is not practical beyond rather narrow limits because of economic considerations. It is possible, however, to increase the value of  $L$  to a limited degree and thus approach the conditions sought. This practice is known as **loading**.

The artificial addition of inductance or loading to a circuit is practicable only to an extent that will approximate the optimum situation discussed above. Reasonably close approaches may be made to the values given for  $\beta$ ,  $W$  and  $Z_0$  in equations (19:2), (19:3) and (19:4), but the value for  $\alpha$  in practice will be more nearly

$$\alpha = \frac{R}{2} \times \sqrt{\frac{C}{L}} \quad (19:5)$$

which may also be written—

$$\alpha = \sqrt{\frac{R}{2L} \times \frac{RC}{2}} \quad (19:5)$$

In a cable circuit, which is the only type of circuit to which loading is now applied, the conductance  $G$  is negligible and the inductance of the conductors may also be neglected because of their close spacing. If we introduce these assumptions into equation (18:13), then the attenuation constant for the non-loaded cable circuit will approximate—

$$\alpha = \sqrt{\omega \frac{RC}{2}} \quad (19:6)$$

If this is compared with equation (19:5) it will be seen that the addition of inductance reduces the attenuation to the extent that the value of the ratio  $R/2L$  is less than  $\omega$ . Since it is generally

not feasible in practice to increase inductance without also adding resistance, loading also increases the resistance of the circuit, and this partially offsets the beneficial effect of the added inductance. Nevertheless, material net reduction in the value of  $a$  can be obtained. Practical loading also decreases phase distortion as pointed out above; and increases the value of the characteristic impedance, as may be seen by inspection of equation (19:4). This latter gives the effect of transmission at a higher voltage and lower current, which also may be helpful to an understanding of why the attenuation is decreased.

Despite the advantages discussed above, there are important practical limitations to the usefulness of loading in modern telephone practice. One disadvantage is its effect in decreasing the velocity of propagation. Where the loading is "heavy"—i.e., relatively large amounts of inductance are added—the time delay of propagation over very long circuits may become great enough to introduce disturbing effects. Much more important is the fact that practical loading imposes a sharp limitation on the total range of frequencies that can be transmitted. This would not be true if it were feasible to add the required inductance to the circuit on a continuous uniform basis by wrapping each conductor with a spiral of magnetic material such as permalloy. This can be and is done in cases such as long submarine telegraph cable where only a single conductor is involved. But it would obviously be impracticable to apply a similar technique to the many conductors of ordinary telephone cables or open-wire lines. Instead, "lumped" loading is applied by inserting inductance coils at regularly spaced intervals along the lines. This effectively breaks the loaded circuit up into network sections, the major electrical constants of each of which are the series lumped inductance and the shunt capacitance.

Such a network has the essential characteristics of a "low-pass filter" which means that it will tend to block the transmission of frequencies above some critical value. This filtering action is due to resonance effects which are explained in a later Chapter, but it may be noted at this point that the equations given above for the characteristics of a loaded circuit have a definite resemblance to the resonance equation given in Article 16.4. The critical frequency where the attenuation of the loaded circuit begins to increase rapidly is known as the cutoff frequency and may be determined from the following equation—

$$f_c = \frac{1}{\pi\sqrt{LC}} \quad (19:7)$$

where  $L$  is the inductance of the loading coil in henries and  $C$  is the total capacitance of the length of circuit between the adjacent coils in farads.

In the design of loading systems the value of this cutoff frequency may be varied considerably by varying the spacing of the loading points and the amount of inductance inserted. It is not feasible in practice, however, to design a loading system in which the cutoff frequency is much higher than about 30,000 cycles. Loading, therefore, cannot be applied to circuits on which broadband carrier systems are to be superimposed. Its present application in long distance telephone practice is limited to toll cable circuits on which carrier systems are not superimposed, and to the relatively short toll entrance cables connecting to open-wire facilities.

## 19.2 Total Losses in the Transmission System

It was pointed out previously that a complete transmission system includes at least three parts—a generating or sending circuit, a transmission line, and a load or receiving circuit. We have seen that there will be attenuation losses on the line in any practical case. Unless the impedance of the sending circuit is exactly equal to the sending end impedance of the line, and the impedance of the receiving circuit is exactly equal to the impedance of the receiving end of the line, there will be other losses known as **reflection losses** at these junction points. Figure 19-1 represents a simple transmission system in which a sending circuit having an impedance  $Z_s$  is connected at points 1 and 2 to a line of length  $l$ , of characteristic impedance  $Z_0$ , and propagation constant  $\gamma$ ; which in turn is connected at points 3 and 4 to a receiving circuit of impedance  $Z_R$ . Assuming first that the line is so long that both its sending end and receiving end impedance is equal to its characteristic impedance  $Z_0$ , there will be reflection losses at both junction points if neither  $Z_s$  nor  $Z_R$  is equal to  $Z_0$ . Considering the junction at 1,2 the ratio of the

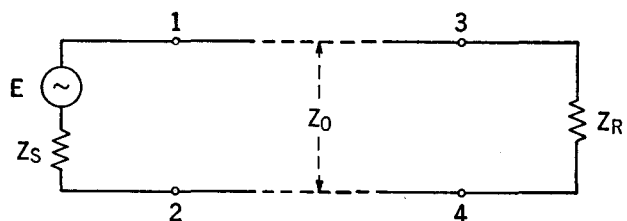


FIGURE 19-1

current  $I_0$  entering the line to the current  $I_0'$  that would enter the line if  $Z_s$  were equal to  $Z_0$  is—

$$\frac{I_0}{I_0'} = \frac{2\sqrt{Z_s Z_0}}{Z_s + Z_0} \quad (19:8)$$

Similarly the reflection loss at point 3,4 may be determined from—

$$\frac{I_R}{I_R'} = \frac{2\sqrt{Z_0 Z_R}}{Z_0 + Z_R}$$

It may be noted that the criterion for no reflection loss is that the impedances at each side of the junction points should be identical and that this is not the same as the criterion for maximum power transfer, which requires that the impedances be conjugate. The total loss in the transmission system, remembering that an effectively infinite line is assumed, is now equal to the sum of the two reflection losses, determined as above, plus the loss in the line itself, which is  $\epsilon^{-\gamma l}$

If, as often may be the case, the line is not electrically “infinite” in length, the sending and receiving end impedances will be somewhat different from  $Z_0$  because of the different impedances connected to the distant ends. Any such differences will of course reduce the accuracy of the reflection losses as calculated from the equations given above. An exact general expression for the total loss of the system, however, may be written as follows:

$$\frac{I_R}{I_R'} = \frac{Z_s + Z_R}{2\sqrt{Z_s Z_R}} \times \frac{2\sqrt{Z_0 Z_s}}{Z_0 + Z_s} \times \frac{2\sqrt{Z_0 Z_R}}{Z_0 + Z_R} \times \epsilon^{-\gamma l} \\ \times \frac{(Z_0 + Z_R)(Z_0 + Z_s)}{(Z_0 + Z_R)(Z_0 + Z_s) - (Z_0 - Z_R)(Z_0 - Z_s) \epsilon^{-2\gamma l}}$$

Here the first three terms represent the two reflection losses as above, and an inverse term cor-

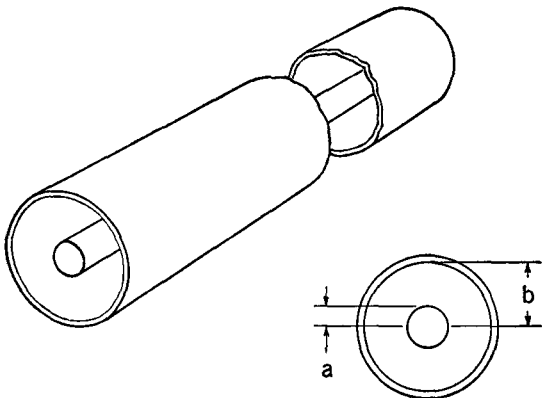


FIGURE 19-2

recting for the fact that  $Z_s$  is not equal to  $Z_R$ ; the fourth term  $\epsilon^{-\gamma l}$  represents the line attenuation and phase shift; and the final term represents an “inter-action factor” which takes care of the fact that the line may be so short that the impedance looking into it from either end is not  $Z_0$ . Inspection will show that this inter-action factor will approach unity as the length of the line is increased because  $\epsilon^{-2\gamma l}$  becomes very small. Similarly, if  $Z_s$  equals  $Z_R$  the first term of the equation becomes unity.

The desirability of matching the impedances at each junction point in a transmission system whenever possible is evident.

### 19.3 Coaxial Conductors

Our consideration of transmission lines thus far has been confined to lines made up of two parallel wire conductors. An entirely different configuration of conductors may be used to advantage where high, and very high, frequencies are involved. This configuration is known as **coaxial** and the conducting pair consists of a cylindrical tube in which is centered a wire as shown in Figure 19-2. In practice the central wire is held in place quite accurately by insulating material which may take the form of a solid core, discs or beads strung along the axis of the wire or a spirally wrapped string. In such a conducting pair equal and opposite currents will flow in the insulated central wire and the outer tube just as equal and opposite currents flow in the more ordinary parallel wires.

At high frequencies, a unit length of coaxial in which the dielectric loss in the insulation is negligible (effectively gaseous) will have an inductance which is about one-half the inductance of two parallel wires separated by a distance equal to the radius of the coaxial tube. The capacitance of the same coaxial is approximately twice that of two parallel wires separated by the same distance and having the same diameter as that of the central coaxial conductor. If the outside radius of the central conductor is designated  $a$  and the internal radius of the tube is  $b$ , the characteristic impedance at high frequencies neglecting leakage may be shown to be approximately—

$$Z_0 = \sqrt{\frac{L}{C}} = 138 \log \frac{b}{a} \quad (19:9)$$

The attenuation constant per mile, where both conductors are of the same material, varies as the square root of frequency and is approximately—

$$\alpha = \frac{R}{2Z_0} = .24 \times 10^{-4} \frac{\sqrt{f} \left( \frac{1}{a} + \frac{1}{b} \right)}{\log \frac{b}{a}} \quad (19:10)$$

where  $a$  and  $b$  are in centimeters. From equation (19:10) it may be determined that minimum attenuation is obtained when the coaxial is so designed that  $b/a = 3.6$ . With this configuration  $Z_0$  is about 77 ohms. The present standard coaxial used for transmission in the Bell System employs a copper tube .375 inches in inside diameter and a copper center wire .1004 inches in diameter. This, it will be noted, approximates the optimum ratio specified above for minimum attenuation. The nominal impedance is about 75 ohms. Velocity of propagation in the coaxial approaches closely the speed of light. A study of the basic characteristics of the coaxial shows that at the high frequencies assumed, the attenuation is substantially less than that of a parallel wire line of comparable dimensions. More important is the fact that the shielding effect of the outer cylindrical conductor prevents interference from external sources of electric energy, and also prevents radiation losses of the energy being transmitted over the coaxial.

#### 19.4 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 mile and a capacitance of .054  $\mu$ f per mile. In this cable the series inductance and the shunt leakage were negligible, while the bridged capacitance 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  $\alpha$  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 a convenient 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**

TABLE VIII  
RELATION BETWEEN DECIBELS AND POWER RATIOS FOR GAINS AND LOSSES

DECIBELS	APPROXIMATE POWER RATIO		
	FOR LOSSES		FOR GAINS
	FRACTIONAL	DECIMAL	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

**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—

$$\frac{P_1}{P_0} = 10^{-1} \quad (19:11)$$

where  $P_1$  is input power and  $P_0$  is output power. This corresponds to a current ratio of  $10^{-0.5}$  and to an attenuation constant value of  $a = .115$ . Table VIII showing the power ratios for several values of decibels will aid in forming a clear conception of the magnitude of the unit.

For any given power ratio the number of decibels corresponding can be determined by the following simple equation—

$$\text{No. of db} = N = 10 \log \frac{P_1}{P_0} \quad (19:12)$$

or, if the current ratio rather than the power ratio is known—

$$N = 20 \log \frac{I_1}{I_0} \quad (19:13)$$

or from equation (18:15)—

$$N = 20 \log \frac{I_1}{I_0} = 20 \times \frac{1}{2.303} \times a = 8.68 a \quad (19:14)$$

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, for example, if the circuit in which it was used was effectively shortened, or had its net attenuation reduced, to that extent.

## CHAPTER 20

### TELEPHONE TRANSMISSION THEORY—(Continued)

#### 20.1 Wave Propagation

In considering the transfer of electric energy along circuits at the lower frequencies, we ordinarily think of a continuous flow of current through the conductors under the pressure of an applied electromotive force. Furthermore, in simple networks or relatively short transmission lines, we ordinarily think of this energy transfer as being practically instantaneous. In the study of long transmission lines and higher frequencies these concepts tend to become inadequate, if not incorrect, as may have already been noted. Thus, even at a frequency as low as 1000 cycles, we would find that the current in a non-loaded 19-gage cable circuit is flowing in opposite directions at points about 88 miles apart, and that an appreciable time is required for the transmission of energy over that distance along the circuit.

Now if we increase the frequency tremendously, we may reach a point where the current is flowing in the line in opposite directions at points only a few feet or even a few inches apart. In other words, the flow of electrons in the conductors is now back and forth within a very limited distance range. Furthermore, because of skin effect this oscillating movement of the electrons is occurring only at, or very nearly at, the surfaces of the conductors. Under these conditions it is somewhat difficult to see any real relationship between the movement of the electrons or current flow and the transfer of energy along the line. It is easier now to think of a **wave of energy** traveling along the line in the space between and surrounding the conductors, and setting up "ripples" of electron motion in the wires as it goes along.

This traveling wave of energy is usually thought of as an electromagnetic field which, in the present case, is associated with or bounded by the wire conductors but which can also exist in free space where no conductors are present. Under this concept the electromagnetic field is considered as being made up of (1) the electric field whose lines of force radiate at right angles from the charges on the surfaces of the conductors and (2) the magnetic field whose lines of force encircle the conductors and are always at right

angles to the force lines of the electric field. The total energy of the wave is always equally divided between the electric and magnetic fields and passes back and forth between them at a rate determined by the frequency of the applied emf. The entire electromagnetic wave travels along the line at the speed of light if not retarded by inductance or capacitance effects in the conductors.

#### 20.2 Reflections on Transmission Lines

If a transmission line were actually of infinite length and without loss, an electromagnetic wave of energy would obviously travel along the line forever; and this would be true even though the sending-end source of energy was disconnected some time after the wave was started. Practically, of course, we are concerned with delivering power to receiving devices over lines of finite length. It becomes necessary, therefore, to consider what will happen when a traveling wave reaches the end of a uniform line.

For purposes of such analysis, it is convenient to assume a line of such low resistance as to be practically without loss, a situation which is actually approximated in a relatively short line at very high frequencies. Let us investigate first the situation where such a lossless line is open at the far end. In that case, there can be no current flow at the open end and the magnetic field therefore must disappear. But the energy which the magnetic field contained must be conserved by being transferred to the electric field. This energy, added to the energy already contained in the electric field approaching the end of the line, doubles the total electric field at that point. In other words, the voltage at the open end of the lossless line becomes twice as great as the voltage applied at the sending end.

We may then consider that the increased voltage starts a wave of energy back along the line towards the sending end. Thus the traveling electromagnetic wave reaching this total discontinuity in the line is **reflected** in somewhat the same way as a light wave is reflected from a mirror. If the initial or incident wave is a sine wave the reflected wave must also be a sine wave. Back over the



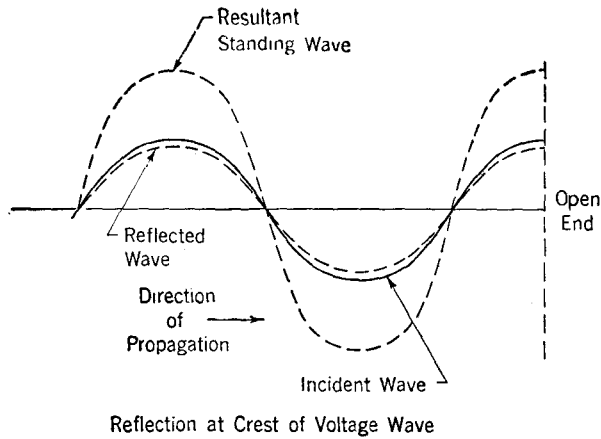


FIGURE 20-1

line, the reflected wave will add to the incident wave at all points and the sum must also be a sine wave. This is graphically illustrated in Figure 20-1 where an incident voltage wave is shown reflected from the open end of the line at its positive crest. The reflected wave then adds in phase with the incident wave to give a wave having crest values approximately twice those of the incident wave. The resultant wave is known as a **standing wave** because it does not travel longitudinally along the line. In other words, if a voltmeter is used to make measurements along the line it would read zero at the nodal points and maximum values of twice the effective value of the applied voltage at the points where the positive and negative voltage crests are shown. Instantaneously at these points the voltage would be varying from maximum positive value through zero to maximum negative value in accordance with the frequency of the applied voltage. This is perhaps better illustrated by Figure 20-2.

A reflection from an open-end line that results in the development of a standing voltage wave of approximately double magnitude will occur only when the distance from the sending end to the open end is equal to an odd number of quarter-wave lengths. The illustrations given, for example, show a line one and a quarter wave lengths long. It may also be noted that if, as we have assumed, the line is lossless the impedance at the sending end in this case is theoretically zero. Energy is nevertheless surging back and forth in the line. The line may accordingly be thought of as behaving like a series resonant circuit. For this reason, lines having discontinuities which cause reflections and consequent standing waves are known as **resonant lines**. A uniform line of infinite length or one terminated in its characteristic

impedance, on the other hand, is a non-resonant line. Of course the complete no-loss condition we have assumed for discussion purposes never actually exists, but at very high frequencies the inductive and capacitive reactances may be so large as compared with the resistance as to cause the actual situation to approach closely the no-loss condition.

A similar analysis of a line consisting of an odd number of quarter wavelengths and shorted at its distant end will show that the standing wave of voltage will have its maximum value equal to the maximum value of the applied voltage at the sending end and will have a nodal point or zero voltage at the shorted end. The sending end impedance of such a line approaches infinity. Lossless lines whose length is an even multiple of quarter wavelengths, on the contrary, have extremely high impedance when open at the distant end and impedance near zero when shorted at the distant end.

The extreme situations that we have discussed above would not occur in a transmission line designed to carry energy from one point to another. We have considered only lines that were open or shorted, and only at points that were some integral multiple of a quarter-wavelength at some specific frequency. As will be brought out in a later Chapter, short resonant lines of these types have numerous useful applications in very high frequency work, but they are not useful as transmission lines in the ordinary sense. Nevertheless, the phenomena of wave reflection must be taken into consideration in any transmission line where the impedances are not matched at all junction points. Where there is any such impedance mismatch on any type of transmission line, and regardless of the frequencies being transmitted, there will be some reflections. In general, the impedance ir-

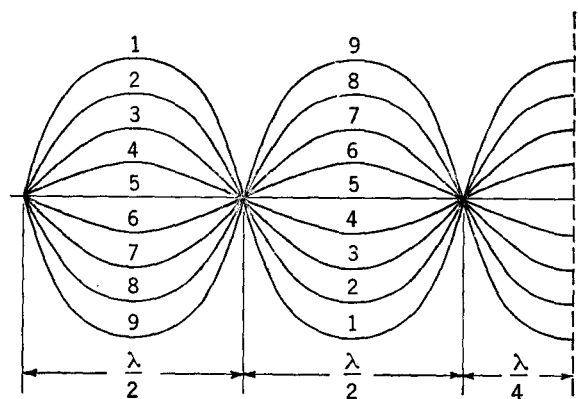


FIGURE 20-2

regularity will not only prevent a maximum power transfer at the point of mismatch but will also result in actual loss of energy. This is due to the fact that any actual transmission line must have some resistance which will cause  $I^2R$  losses due to the reflected currents in addition to the  $I^2R$  losses of the useful current transmitted to the load at the receiving end of the line. The extent of a reflection loss of course depends on the extent of the mismatch and can be calculated from the equation given in Article 19.2.

### 20.3 Electromagnetic Radiation

In analyzing the total energy losses that may occur in the transmission of electric power from one point to another, it is necessary now to consider another phenomenon which has, up to this point, been ignored. This is the loss due to **radiation**.

In 1864, James Clerk Maxwell undertook to set up a series of mathematical equations that would provide a general statement of the relationships between electric and magnetic fields under any and all conditions. In rounding out this series of equations to achieve mathematical symmetry, he was led to some very interesting conclusions. The equations seemed to indicate that the hitherto existing assumption that all of the energy contained in the electric and magnetic fields accompanying the flow of current in a conductor returned into the conductor when the source of emf was cut off was not wholly true. Some part of the field, it appeared, would detach itself entirely and escape into space in the form of electromagnetic radiation, carrying with it a comparable part of the total energy.

This led to the further conclusion that a moving electric field can exist in the absence of any electric charges, despite the fact that an electric field is usually thought of as being made up of lines of electric force always terminating on electric charges. Furthermore, if a moving electric field can exist independently in space, it must be thought of as being equivalent in certain ways to a flow of electric current. In other words, the moving electric field in free space must set up, or be accompanied by, a moving magnetic field just as it would be in the case of current flow along conductors.

The Maxwell equations cannot be written in any form that does not involve branches of mathematics which are beyond the scope of this book. With respect to radiated energy, however, the

equations indicate, and experience confirms, that for a given current the amount of energy radiated depends upon the square of the frequency. Naturally the amount of radiation also depends upon the intensity of the current. What is of major significance from a practical viewpoint is the fact that the amount of radiation goes up very rapidly as the frequency increases. There is always radiation whenever there are changing current values but at voice frequencies, and at frequencies well up into the ordinary telephone "carrier range", the amount of radiation is negligible for most practical purposes. At frequencies that are measured in megacycles, on the other hand, radiation may cause losses that are much greater than any  $I^2R$  losses in the conductor.

### 20.4 Radio Transmission

It was not until some twenty years after Maxwell developed his famous equations that Heinrich Hertz demonstrated experimentally the truth of the electromagnetic radiation hypothesis; and it was some years later before experimenters began to develop methods for taking advantage of this radiation phenomenon to transmit electric energy through space for useful purposes. In this case the objective was not to avoid energy losses by radiation but to do everything possible to facilitate maximum radiation. Since, as we have seen, the amount of radiation increases with frequency at a geometric rate, purposeful radio transmission naturally involves the use of high frequencies and transmission lines or antennas designed to radiate maximum energy.

If the transmission of the radiated energy were through unobstructed space in the form of electromagnetic waves like light waves, there would be no loss of energy "along the line" because there would be nothing to absorb the energy. If energy could be radiated from a given point and confined in a narrow beam extending directly to the receiving point, this means of transmission could be far superior to any wire transmission because of this lossless quality. However, the natural tendency of any radiator, isolated in space, is to send out energy in practically all directions although only such energy as actually reaches any receiving point is useful.

"Line losses" in radio transmission should largely be thought of not as energy losses in the transmission path itself, such as occur in wire lines, but as energy escaping entirely from the effective transmission path. The basic transmis-

sion problem, accordingly, is to devise methods that will direct the path of the radiated energy. Such methods are concerned primarily with the radiating antennas and it is customary in radio work to measure the effectiveness of antennas in controlling the directivity of radio propagation in terms of **antenna gain**. This is merely a measure in decibels, or other appropriate units, of the amount of energy received at a given point from a given transmitting antenna compared with what would have been received if the transmitting antenna radiated with uniform strength in all directions. Thus, high antenna gain in radio transmission corresponds to low line loss in wire transmission.

As is discussed briefly in a later Chapter, it is possible to design antennas with quite high gains—particularly in the superhigh-frequency range (thousands of megacycles). There is no practical possibility, however, of designing antennas with such directivity that all, or even a major part, of the transmitted energy will reach a receiver located at any great distance from the transmitter. In other words, there must always be a very substantial effective loss of energy. What is perhaps worse, this lost energy may be absorbed elsewhere where it may interfere with other communication circuits, or be received at unauthorized points in such a way as to militate against the privacy of the transmission. There remain, therefore, obvious advantages in the employment of physical facilities that guide the energy directly to the desired receiving point. An ordinary wire line is one type of such a guide but is satisfactory at only relatively low frequencies because its radiation and other losses become too great at high frequencies. A coaxial is a better guide because the outer tube acts as a shield to prevent any of the electromagnetic energy transmitted within the tube from radiating into space. At frequencies in the thousands of megacycles, however, the losses of any practical design of coaxial also become very high.

The usefulness of radio transmission in practice is found in two principal situations. The first is where it is impossible or economically impractical to construct physical facilities that will guide the energy. Here the application is in transmission over large bodies of water or to moving points such as ships at sea, motor vehicles, trains, and aircraft. The second situation is point-to-point service over land where radio transmission is considered more economical on an overall basis than

other methods. A major application here is to microwave radio relay systems. But even in this case, it is necessary to use some form of physical transmission medium to guide the energy through the relatively short distances from the transmitter itself to the transmitting antenna, and from the receiving antenna to the receiver. Within limits, shielded wire lines or coaxials can be used for this purpose. More effective, however, are the simple hollow metal tubes commonly known as **waveguides**.

## 20.5 Waveguides

It is probably easiest to think of a waveguide as a device which merely isolates a particular path in space. Propagation of energy through the guide is then essentially no different from ordinary radio propagation, except that it is confined to this particular isolated path. However, the transmitting path, i.e., the space of the interior of the guide, must be large enough to permit the traveling electric and magnetic fields to assume configurations comparable to those that they would naturally assume when traveling in free space. It can be shown that for effective propagation a waveguide must have a maximum cross-sectional dimension at least equal to one-half the wavelength. This automatically limits the practical use of waveguides to the transmission of very high frequencies, where wavelengths are of the order of a few inches.

Waveguides commonly used in telephone practice are rectangular in shape as illustrated in Figure 20-3, with the dimension  $b$  greater than one-half wavelength but not greater than one wavelength. The dimension  $a$  is not critical but is usually about half as large as  $b$ . In such a guide the electric field tends to arrange itself as shown in Figure 20-4(A) with the lines of electric force extending vertically between the top and bottom guide walls and having maximum intensity at the center, and tapering off to zero at the sides of the

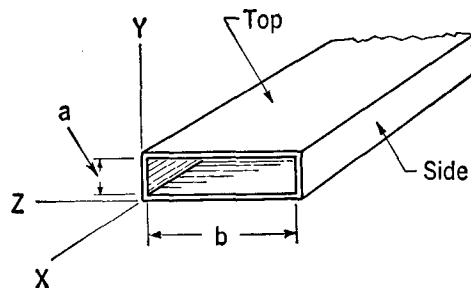


FIG. 20-3 RECTANGULAR WAVEGUIDE

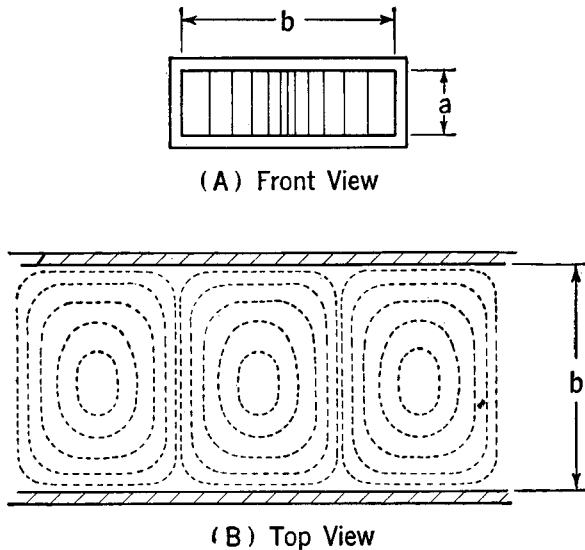


FIGURE 20-4

guide. The magnetic field is at right angles to the electric field as shown in Figure 20-4(B). The lines of magnetic force, as indicated, are closed loops and the magnetic field has its greatest intensity along the sides of the guide with minimum intensity in the center where the electric field is the greatest.

If the interior walls of the guide are perfectly conducting, the lines of force of the electric field must always be perpendicular to the top and bottom walls; and the field at the side walls must be zero because the field there would be short-circuited. Similarly, the magnetic field must always be parallel to the side walls of the guide and can have no perpendicular component which will cut through the conducting surfaces because any such component would set up a current that would in turn set up a magnetic field exactly opposite to the exciting field. All this is to say that if the inner surface of the guide is a perfect conductor, the traveling wave in the guide would at all times be a plane wave without any curvature whatever. Under these conditions energy would travel in the guide with practically no loss since no appreciable energy can be absorbed by the air dielectric. Actually, the inner surfaces of the guide are not perfect conductors, although copper and sometimes silver-plating is used in their construction. As a result, the actual configuration of the fields in the guide tend to deviate slightly from the ideal plane form. This results in some movement of the electrons in the surfaces of the guide and some consequent  $I^2R$  energy losses.

At 4000 megacycles the loss in a  $1\frac{1}{4} \times 2\frac{1}{2}$  inch bronze guide is about 1.5 db per 100 feet. In

terms of the losses that we encounter at low frequencies in ordinary wire transmission lines, this is extremely high. It is nevertheless substantially lower in the microwave region than would be caused by the usual types of wire line or coaxial cable. The velocity of propagation of energy in a waveguide approaches but is always somewhat less than the speed of light.

Like any other transmission line, the waveguide should be uniform in structure along its total length. Any discontinuity such as changes in its size or shape, holes in the guide walls, or foreign conducting materials in its interior, will cause reflections and consequent energy losses. For the same reason care has to be used in designing bends or twists in the guides. In connecting different types of guides together, or in connecting them to antennas or energy sources, methods of impedance matching similar in general principle to those applying in wire or coaxial lines must be employed to avoid reflection losses.

If a waveguide is "shorted" (i.e., closed by a conducting end plate) at its far end, total reflection will occur just as when a wire transmission line is shorted. Similar to the wire line case discussed in Article 20.2, if the waveguide is shorted at a point which is an odd multiple of a quarter wavelength distant from the energy source, the reflected wave will add in phase to the incident wave to set up a standing wave in the guide. In other words, the guide is now a **resonant line** and relatively large amounts of energy will surge back and forth between its contained electric and magnetic fields when energized or excited by the appropriate resonant frequency. This is the phenomenon that is taken advantage of to produce the **cavity resonators** that we shall encounter in later Chapters. Thus, if a section of waveguide

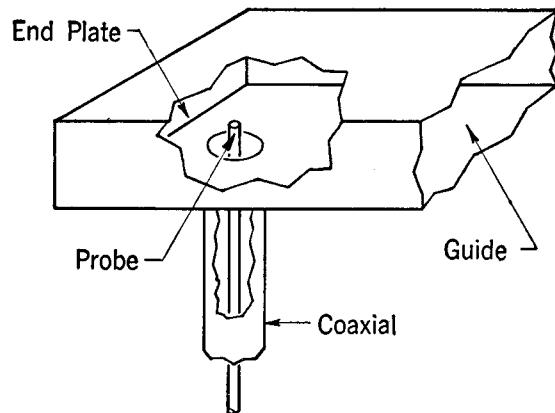


FIG. 20-5 PROBE TYPE COAXIAL TO WAVEGUIDE TRANSDUCER

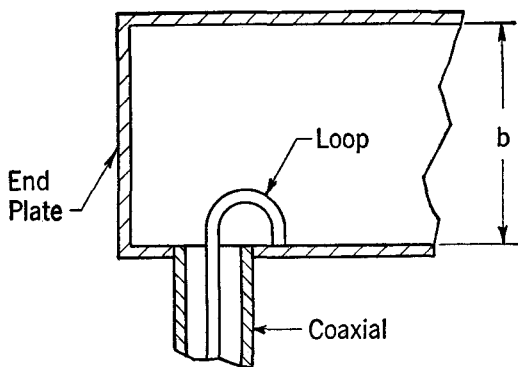


FIG. 20-6 LOOP TYPE COAXIAL TO WAVEGUIDE TRANSDUCER

one half wavelength long is closed at both ends to form a small box and energized at the center point, it may be considered as two quarter wave lines, each of which is resonant. The cavity then acts in the same manner as an ordinary parallel resonant circuit made up of an inductor and a capacitor where likewise at the resonant frequency relatively large amounts of energy may surge back and forth between the magnetic field of the inductor and the electric field of the capacitor.

The most common method for energizing or exciting a waveguide is indicated in Figure 20-5.

Here the central conductor of a coaxial is inserted vertically into the center of the guide, where the electric field has its maximum value. This probe acts like an antenna in delivering energy to the guide, or in removing energy from it. The outer tube of the coaxial is connected electrically and mechanically to the wall of the guide and the impedances are matched by locating the probe at a point slightly less than one-quarter wavelength from the closed end of the guide so that reflections from the end plate will be in phase with, and add to, the traveling wave in the tube. Guides can also be energized by inserting a loop as shown in Figure 20-6 at a point in the wall of the guide where the magnetic field is most intense.

In the above discussion we have deliberately confined our attention to a single configuration of the electric and magnetic fields in the waveguide. The configuration discussed is designated technically as the  $TE_{0,1}$  "mode". This mode will permit the transmission of the lowest frequency for a given size guide. If a guide is made larger so that its maximum dimension is greater than one wavelength, other configurations or modes of the fields become possible. The **dominant mode** discussed, however, is the one of major applicability in telephone practice.

## CHAPTER 21

### CHARACTERISTICS OF CIRCUIT FACILITIES

#### 21.1 Classification of Wire Facilities

Transmission facilities or media employed in telephone and telegraph work, to be cataloged completely, would have to include both waveguides and free space, which is the medium of radio transmission. In this Chapter, however, we shall confine our attention to various types of wire conductors, including coaxials. The kind of wire line facility to be used in a particular case depends upon economic considerations and the transmission requirements to be met. Ordinary wire facilities may be classified in several ways according to their uses, or on the basis of their physical or electrical characteristics.

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 greater 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 economically desirable 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. 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 some cases it may be necessary to use telephone repeaters as well.

The principal types of toll or long distance wire facilities are considered separately in the following Articles.

#### 21.2 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. At any given time, accordingly, the working plant may include facilities ranging from earlier types to newly developed types which are barely out of the experimental stage. Before the advent of the telephone repeater, the 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 hard drawn copper wire and each group of four wires was usually arranged to carry a phantom circuit.

The wires were carried on crossarms in the manner indicated in Figure 21-1. 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

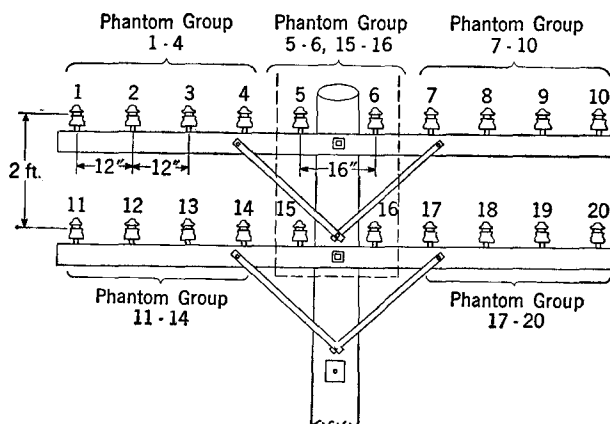


FIG. 21-1 WIRE CONFIGURATION FOR OPEN WIRE LINE CARRYING VOICE-FREQUENCY SIDE AND PHANTOM CIRCUITS

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 electrical characteristics than 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 Figure 21-1, 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—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.

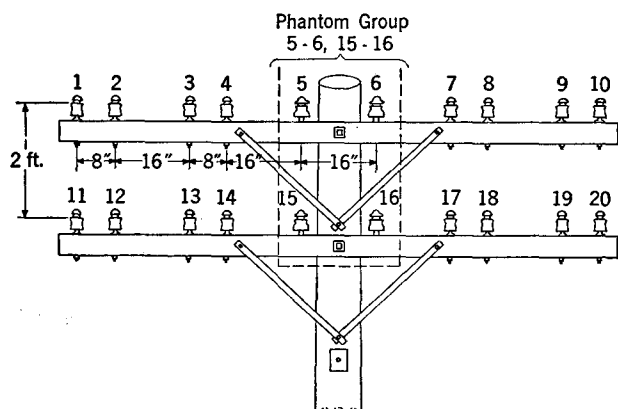


FIG. 21-2 WIRE CONFIGURATION FOR OPEN WIRE LINE ON WHICH TYPE-C CARRIER SYSTEMS ARE SUPERIMPOSED

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 to open wire lines has led to other changes in open wire facility arrangements. On account of the higher frequencies employed in carrier systems, the probability of crosstalk is increased. Since the greatest crosstalk hazard is between the side and phantom cir-

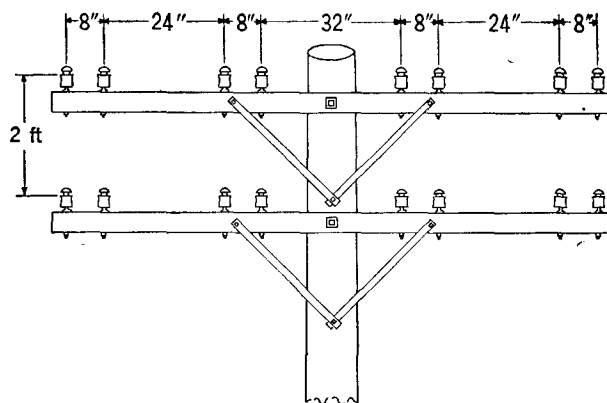


FIG. 21-3 WIRE CONFIGURATION OF 8-INCH SPACED OPEN WIRE LINE FOR TYPE-J CARRIER OPERATION

cuits of a phantom group, it is 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 21-2 shows a wire configuration used to a considerable extent on lines carrying Type-C telephone carrier systems (frequencies up to 30 kc) 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 which is designated 8-16-8 includes a pole-pair phantom group which ordinarily would be used only for voice frequencies. The change in spacing from 12 inches to 8 inches reduces the linear inductance of the pair and increases its linear capacitance 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.

Where open wire line facilities are designed to carry broad-band carrier systems (Type-J) em-

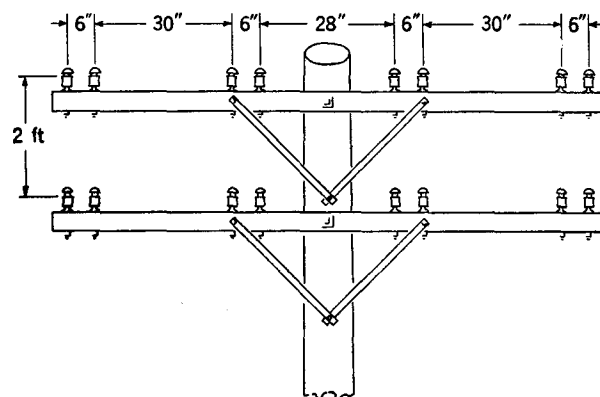


FIG. 21-4 WIRE CONFIGURATION OF 6-INCH SPACED OPEN WIRE LINE FOR TYPE-J CARRIER OPERATION

ploying frequencies up to 140 kc, 8 or 6 inch spacing between wires of a pair is employed, and the pole-pair groups are usually dispensed with. Each crossarm then carries 8 wires, with spacings and configurations as indicated in Figures 21-3 and 21-4, and no phantom circuits are provided for. These configurations are designated 8-24-8 and 6-30-6 respectively.

Open wire facilities are subject to the effects of leakage which increase attenuation losses, particularly at carrier frequencies, and which must be adequately controlled to obtain satisfactory transmission. This is done by insulating the wires from their supporting structure with glass insulators. The effectiveness of such insulators under given conditions of weather varies with their size, shape, and the kind of pin employed.

When new open wire facilities are placed on existing lines and are likely to be used for carrier operation at frequencies above 10 kc, it is necessary to take into account the manner in which the other pairs on the line are insulated. This arises from the fact that the wet weather attenuation of similar gage facilities equipped with different types of insulators is unequal and that as a result energy level differences may occur, which cause crosstalk. When this is the case, it may be desirable that all of the open wire facilities of the same gage which are to be used for carrier operation at frequencies above 10 kc be equipped with the same type insulators.

Table IX gives the more important physical and electrical constants of the commonly used types of open wire circuits. The values given are cal-

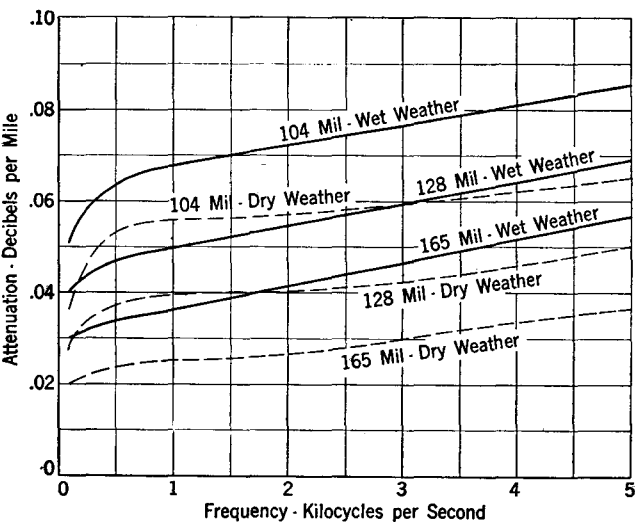


FIG. 21-5 ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE SIDE CIRCUITS OVER THE VOICE RANGE

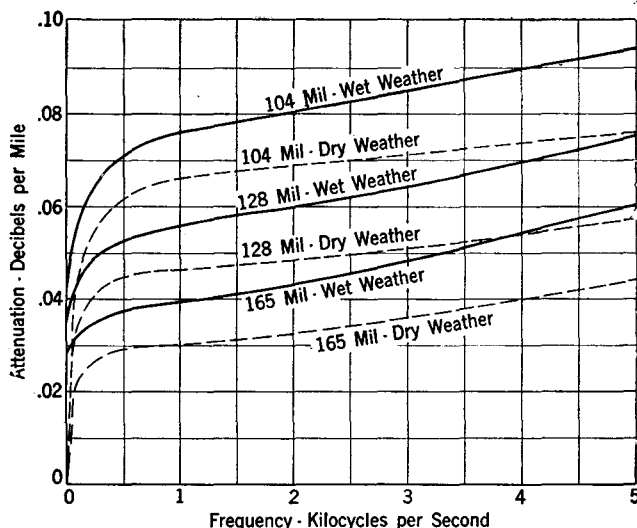


FIG. 21-6 ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHANTOM CIRCUITS OVER THE VOICE RANGE

culated 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, wavelength, 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 21-5 and 21-6 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 21-1, 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



TABLE IX  
CHARACTERISTICS OF STANDARD TYPES OF OPEN WIRE TELEPHONE CIRCUITS AT 1000 CYCLES PER SECOND

TYPE OF CIRCUIT	GAGE OF WIRES (MILS)	SPACING OF WIRES (IN.)	CONSTANTS PER LOOP MILE				PROPAGATION CONSTANT				LINE IMPEDANCE				WAVE-LENGTH MILES	VELOCITY MILES PER SECOND W	TRANS-MISSION EQUIVALENT DB PER MILE
			R Ohms	L Henrys	C Mt.	G M.Mho.	Polar		Rectangular		Polar		Rectangular				
							Magni-tude	Angle Degrees +	$\alpha$	$\beta$	Magni-tude	Angle Degrees -	R Ohms	X Ohms -			
Non-Pole Pair Side	165	12	4.11	.00337	.00915	.29	.0352	84.36	.00346	.0350	612	5.35	610	57	179.5	179,500	.030
Pole Pair Side	165	18	4.11	.00364	.00863	.29	.0355	84.75	.00325	.0353	653	5.00	651	57	178.0	178,000	.028
Non-Pole Pair Phan.	165	12	2.06	.00208	.01514	.58	.0355	85.34	.00288	.0354	373	4.30	372	28	177.5	177,500	.025
Pole Pair Phan.	165	18	2.06	.00207	.01563	.58	.0359	85.33	.00293	.0358	366	4.33	365	28	175.5	177,500	.025
Non-Pole Pair Phys.	165	8	4.11	.00311	.00996	.14	.0353	83.99	.00370	.0351	565	5.88	562	58	179.0	179,000	.032
Non-Pole Pair Phys.	165	6	4.11	.00292	.01070	.14	.0356	83.63	.00394	.0353	529	6.25	526	58	177.8	177,800	.034
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 Phys.	128	6	6.74	.00308	.01011	.14	.0361	80.33	.00606	.0356	568	9.53	560	94	176.6	176,600	.053
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
Non-Pole Pair Phys.	104	6	10.15	.00322	.00967	.14	.0371	76.60	.00859	.0361	610	13.26	594	140	174.2	174,200	.075

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.).

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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 21-7 gives curves for 8 inch spaced, physical circuits, transposed for Type-C carrier, 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 21-8, the 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 21-9 gives a representative example of the measured effect of melting glaze of an estimated diameter of  $\frac{1}{2}$  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.

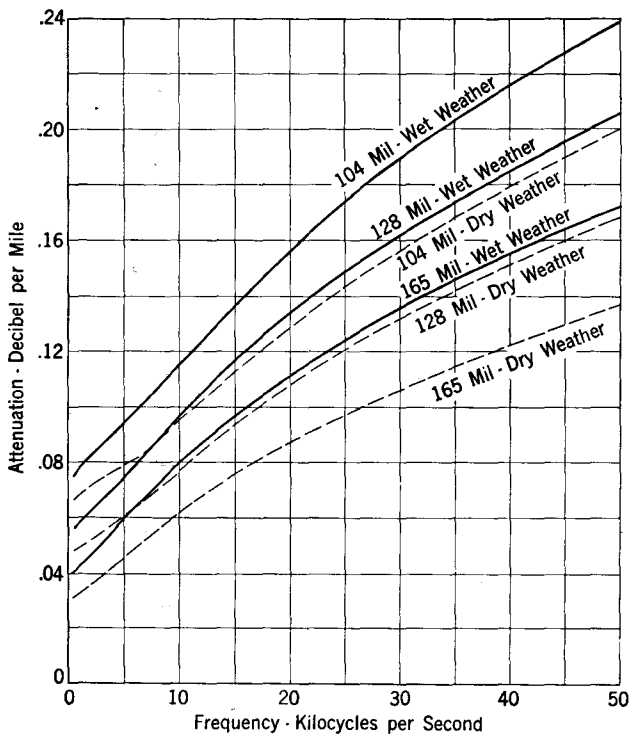


FIG. 21-7 ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHYSICAL CIRCUITS OVER THE TYPE-C CARRIER RANGE

### 21.3 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

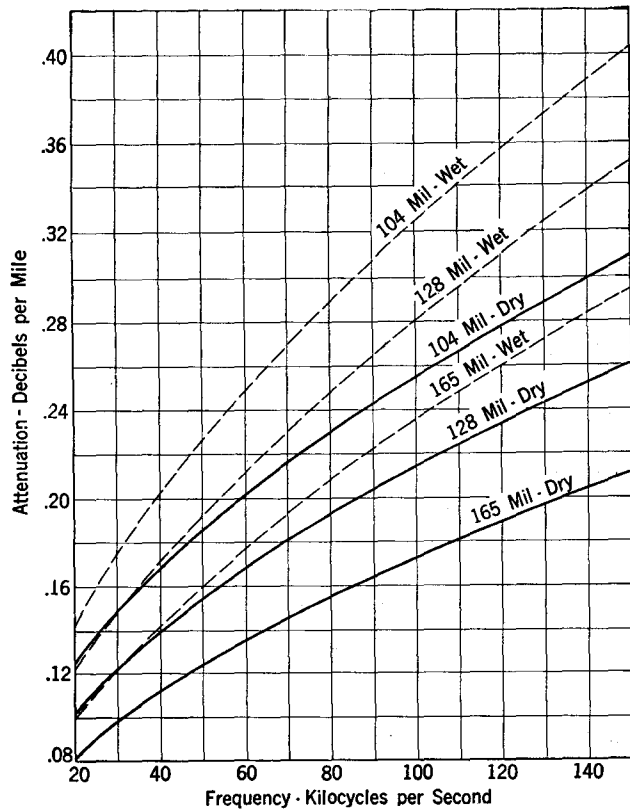


FIG. 21-8 ATTENUATION-FREQUENCY CHARACTERISTICS OF OPEN WIRE PHYSICAL CIRCUITS OVER THE TYPE-J CARRIER RANGE

obvious economic reasons, wire conductors in cables are of considerably finer gage than open wire conductors, which of course increases their attenuation per unit length. The much higher capacitance, 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 are loaded.

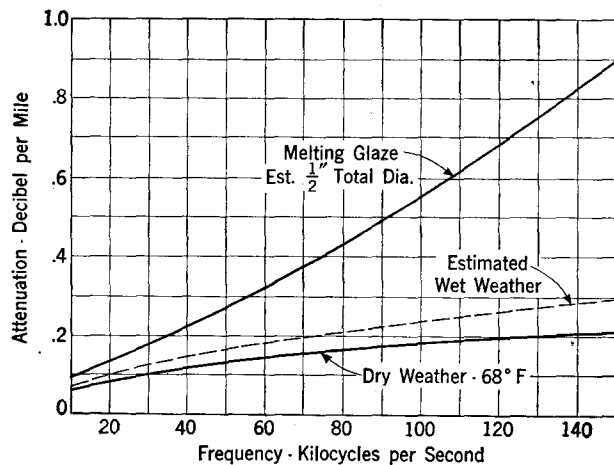


FIG. 21-9 CURVE SHOWING THE EFFECT OF SLEET DEPOSIT ON ATTENUATION OF OPEN WIRE CIRCUIT

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 19, however, such heavy loading, while effective in reducing the attenuation, has some undesirable effects. In the first place, it reduces the velocity of propagation to relatively low values which may seriously interfere with effective transmission over the longer circuits. Also, such a loaded circuit acts as a low-pass filter with a relatively low cutoff frequency—in the neighborhood of 2500 cycles. Requirements for high quality transmission demand that cutoff points be higher than this.

It is desirable that the ordinary cable telephone circuit transmit frequencies up to at least 3000 cycles without substantial attenuation 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 such cables are now of either 16- or 19-gage.

The computed constants at 1000 cycles of the types of circuits extensively used in toll cables are given in Table X. This Table is similar in make-up to Table IX and the general comments made in the preceding Article regarding the former Table also apply to this. As Table X 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.

The loading designations given in the third column of Table X make use of a standardized code. The first letter in the code indicates the spacing between coils as shown in Table XI. 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 millihenrys 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 millihenry 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 millihenry inductance and the side circuits are loaded with coils of 88 millihenry 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 millihenrys, 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 cutoff point is relatively small. This may be seen by referring to the curves of Figure 21-10. These curves also show how the cutoff frequency takes higher and higher values as lighter loading is employed. The lightest toll cable circuit loading is used for program circuits. Figure 21-10 shows the attenuation-frequency characteristic of one such type of loading (B-22) which permits the transmission of frequencies up to about 8000 cycles without serious distortion. For program transmission up to 15 kc still lighter loading is used. One system employs 11 millihenry coils spaced at 1500-foot intervals; another uses 7.5 millihenry coils spaced at 1000-foot intervals. Either system may be applied to either 16 or 19 gage conductors. 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 21-11, but this is offset by the use of closely spaced high-gain amplifiers.

Modern toll cables include coaxials as well as the all-wire types that we have been considering. As was noted in Chapter 19, the present standard

CHARACTERISTICS OF STANDARD TYPES OF PAPER CABL

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				Pol Magnitude
					R Ohms	L Henrys	R Ohms	L Henrys	C Mf.	G M. Mho.	
Side	19	N.L.S.	—	—	—	—	85.8	.001	.062	1.5	.1830
"	19	H-31-S	M-4	1.135	2.7	.031	88.2	.028	.062	1.5	.2769
"	19	H-44-S	M-2 & M-3	1.135	4.1	.043	89.4	.039	.062	1.5	.3188
"	19	H-88-S	M-11	1.135	7.3	.088	92.2	.078	.062	1.5	.4408
"	19	H-172-S	M-1	1.135	13.0	.170	97.3	.151	.062	1.5	.6095
"	19	H-174-S	D-99318	1.135	16.1	.171	100.0	.152	.062	1.5	.6116
"	19	B-88-S	M-9	0.568	7.3	.088	98.7	.156	.062	1.5	.6195
"	16	N.L.S.	—	—	—	—	42.1	.001	.062	1.5	.1288
"	16	H-31-S	M-4	1.135	2.7	.031	44.5	.028	.062	1.5	.2659
"	16	H-44-S	M-2 & M-3	1.135	4.1	.043	45.7	.039	.062	1.5	.3148
"	16	H-88-S	M-11	1.135	7.3	.088	48.5	.078	.062	1.5	.4380
"	16	H-172-S	M-1	1.135	13.0	.170	53.6	.151	.062	1.5	.6084
"	16	H-174-S	D-99318	1.135	16.1	.171	56.3	.152	.062	1.5	.6105
"	16	B-88-S	M-9	0.568	7.3	.088	54.9	.156	.062	1.5	.6184
Phantom	19	N.L.P.	—	—	—	—	42.9	.0007	.100	2.4	.1646
"	19	H-18-P	M-4	1.135	1.4	.018	44.1	.017	.100	2.4	.2695
"	19	H-25-P	M-2 & M-3	1.135	2.1	.025	44.7	.023	.100	2.4	.3082
"	19	H-50-P	M-11	1.135	3.7	.050	46.2	.045	.100	2.4	.4243
"	19	H-63-P	M-1	1.135	6.1	.063	48.3	.056	.100	2.4	.4724
"	19	H-106-P	D-99318	1.135	8.2	.107	50.1	.095	.100	2.4	.6135
"	19	B-50-P	M-9	0.568	3.7	.050	49.4	.089	.100	2.4	.5939
"	16	N.L.P.	—	—	—	—	21.0	.0007	.100	2.4	.1161
"	16	H-18-P	M-4	1.135	1.4	.018	22.2	.017	.100	2.4	.2618
"	16	H-25-P	M-2 & M-3	1.135	2.1	.025	22.8	.023	.100	2.4	.3032
"	16	H-50-P	M-11	1.135	3.7	.050	24.3	.045	.100	2.4	.4223
"	16	H-63-P	M-1	1.135	6.1	.063	26.4	.056	.100	2.4	.4709
"	16	H-106-P	D-99318	1.135	8.2	.107	28.2	.095	.100	2.4	.6128
"	16	B-50-P	M-9	0.568	3.7	.050	27.5	.089	.100	2.4	.5931
Physical	19	B-22	620	0.568	1.28	.022	88.1	.040	.062	1.5	.3221
"	16	B-22	616	0.568	1.25	.022	44.3	.040	.062	1.5	.3153

NOTE. The values for cut-off frequency and transmission equivalent per mile, as given in the last These values accordingly may not be identical with the measured values given in standard formal in

TABLE X  
CABLE TELEPHONE CIRCUITS AT 1000 CYCLES PER SECOND

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.)	TRANSMIS- SION EQUIV- ALENT DB PER MILE (CALCU- LATED)
Polar		Rectangular		Polar		Rectangular							
Angle Degrees +	$\alpha$	$\beta$	Magnitude	Angle Degrees -	R Ohms	X Ohms							
0	46.98	.1249	.1338	470.1	42.80	344.9	319.37	46.93	—	—	46930	—	1.08
1	76.58	.0643	.2693	710.0	13.20	691.2	162.17	23.33	20.55	20555	23331	6700	.56
2	79.87	.0561	.3138	818.0	9.91	805.8	140.80	20.02	17.64	17638	20022	5705	.49
3	84.56	.0418	.4388	1131.0	5.22	1126.3	102.83	14.32	12.61	12615	14319	3997	.36
4	86.96	.0323	.6085	1564.7	2.82	1562.8	76.90	10.33	9.10	9098	10326	2878	.28
5	86.90	.0331	.6107	1570.0	2.84	1568.0	78.83	10.29	9.59	9586	10288	2870	.29
6	87.01	.0322	.6186	1590.2	2.76	1588.3	76.70	10.16	17.88	17882	10157	5655	.28
7	49.13	.0842	.0974	330.7	40.65	250.9	215.39	64.51	—	—	64506	—	.73
8	82.79	.0334	.2638	682.5	6.99	677.4	83.02	23.82	20.99	20985	23818	6700	.29
9	84.61	.0296	.3134	808.0	5.17	804.7	72.83	20.05	17.66	17663	20048	5705	.26
10	87.64	.0224	.4374	1124.0	2.71	1122.8	53.09	14.36	12.66	12656	14365	3997	.19
11	88.27	.0183	.6082	1562.0	1.51	1561.5	41.06	10.33	9.10	9102	10331	2878	.16
12	88.20	.0191	.6102	1567.0	1.58	1566.4	43.11	10.30	9.06	9062	10297	2870	.17
13	88.29	.0185	.6181	1587.4	1.49	1586.9	41.35	10.17	17.90	17897	10165	5655	.16
14	47.78	.1106	.1219	262.1	41.97	194.8	175.23	51.53	—	—	51525	—	.96
15	78.67	.0529	.2642	428.8	11.11	420.8	82.61	23.78	20.95	20952	23781	6959	.46
16	81.30	.0466	.3047	490.7	8.48	485.3	72.39	20.62	18.16	18158	20621	5916	.40
17	85.25	.0351	.4228	675.2	4.53	673.1	53.33	14.86	13.09	13093	14861	4193	.30
18	85.98	.0331	.4712	751.8	3.80	750.1	49.81	13.33	11.75	11748	13334	3738	.29
19	87.49	.0269	.6129	976.4	2.29	975.6	38.98	10.25	9.03	9033	10252	2871	.23
20	87.37	.0273	.5933	945.2	2.41	944.4	39.83	10.59	18.64	18645	10590	5936	.24
21	50.02	.0746	.0890	184.8	38.98	143.7	116.29	70.60	—	—	70604	—	.65
22	84.02	.0273	.2604	416.7	5.76	414.6	41.83	24.13	21.26	21259	24129	6959	.24
23	85.41	.0243	.3022	482.5	4.37	481.1	36.80	20.79	18.32	18319	20792	5916	.21
24	87.43	.0189	.4218	672.1	2.35	671.5	27.52	14.90	13.12	13124	14896	4193	.16
25	87.74	.0185	.4705	749.4	2.04	748.9	26.63	13.35	11.77	11766	13354	3738	.16
26	88.54	.0156	.6126	975.2	1.24	975.0	21.13	10.26	9.04	9037	10257	2871	.14
27	88.48	.0157	.5929	943.9	1.30	943.7	21.39	10.60	18.66	18657	10597	5936	.14
28	80.23	.0546	.3174	826.8	9.55	815.3	137.20	19.79	34.85	34850	19790	11276	.47
29	84.89	.0281	.3140	809.4	4.89	806.4	69.01	20.01	35.26	35260	20010	11276	.24

Last two columns, are calculated from the primary constants which are assumed as uniformly distributed.  
Instructions. The values given in Bell System Practices should therefore be used for engineering work.

TABLE XI

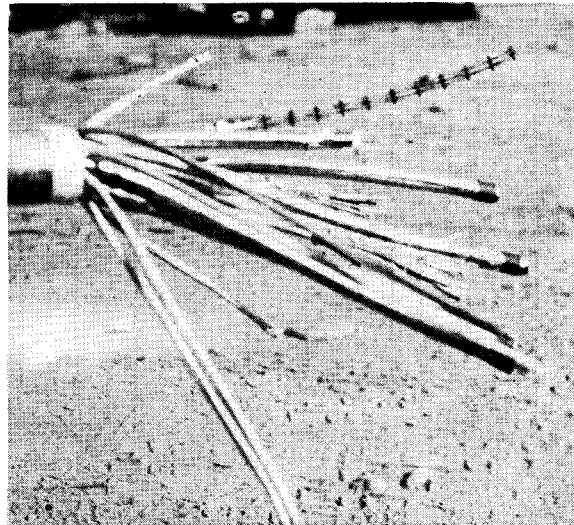
LOADING COIL SPACING CODE

CODE DESIGNATION	LOAD COIL SPACING IN FEET
A	700
B	3000
C	929
D	4500
E	5575
F	2787
J*	640
H	6000
X	680
Y	2130

\* Spiral-four cable.

coaxial conductor used in the Bell System has a tube diameter of .375 with a central wire .1004 inches in diameter. Early and more or less experimental types of coaxial having an outer tube diameter of .27 inches and a 13-gage central wire are also in use. Coaxial cables in present practice may be manufactured to include 2, 4, 6, or 8 tubes; and in all cases they also include a number of ordinary wire pairs or quads. In the manufacturing process some "twisting" of the tubes with respect to each other and the cable sheath occurs so that the tubes are somewhat longer than the cable. In the case of an eight tube cable this extra length is about 1.54%.

Figure 21-12 shows the attenuation loss at a



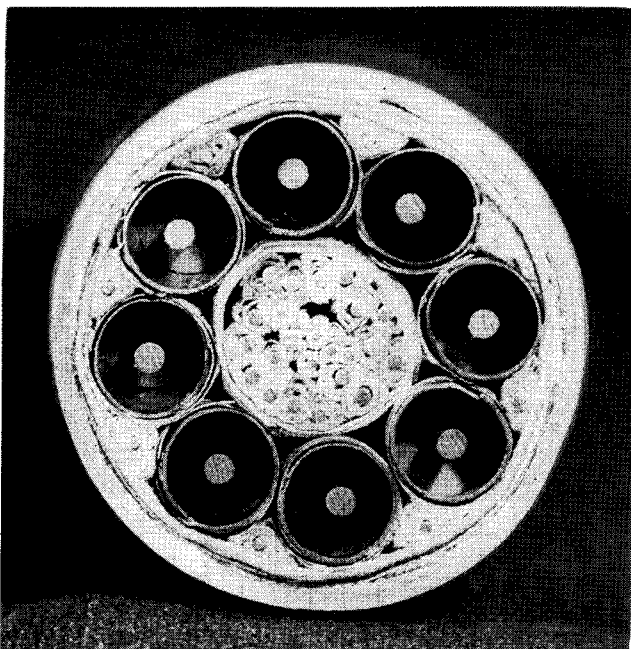
SPIRAL-FOUR, DISC-INSULATED TOLL ENTRANCE CABLE

temperature of 55° F of a coaxial in an 8-tube cable through frequencies ranging up to 3200 kc.

#### 21.4 Toll Entrance and Intermediate Cable Facilities

It is seldom practicable to extend open wire line facilities into the central sections of the larger towns and cities. Instead, the open wires are terminated at a pole somewhere on the outskirts where they 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. There are also numerous situations, such as river crossings, where it is necessary to insert relatively short lengths of cable in the long open wire line. Such cable lengths are known as intermediate cables.

In order to meet the overall transmission requirements of the long distance circuits, it is of course desirable to keep the attenuation of toll entrance and intermediate 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. The cables usually contain three gages of wire—namely, 13, 16, and 19, and the larger gages are connected to



COAXIAL CABLE CROSS SECTION

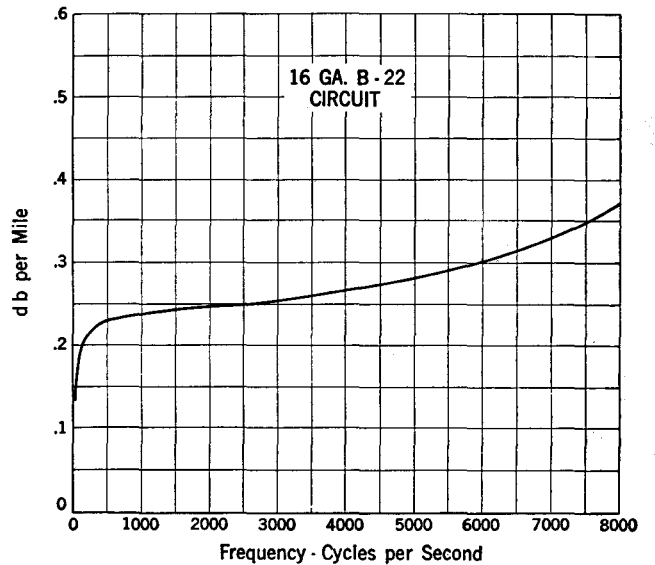
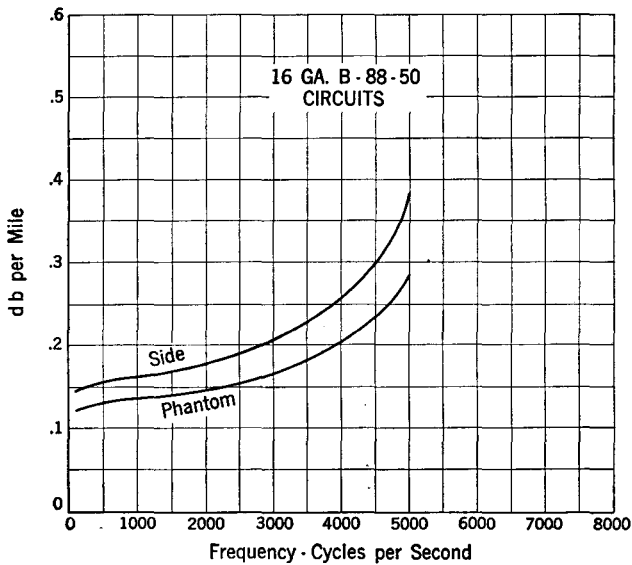
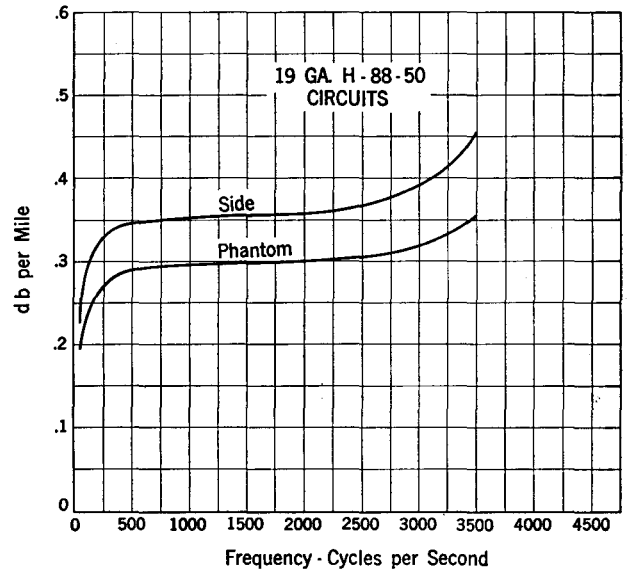
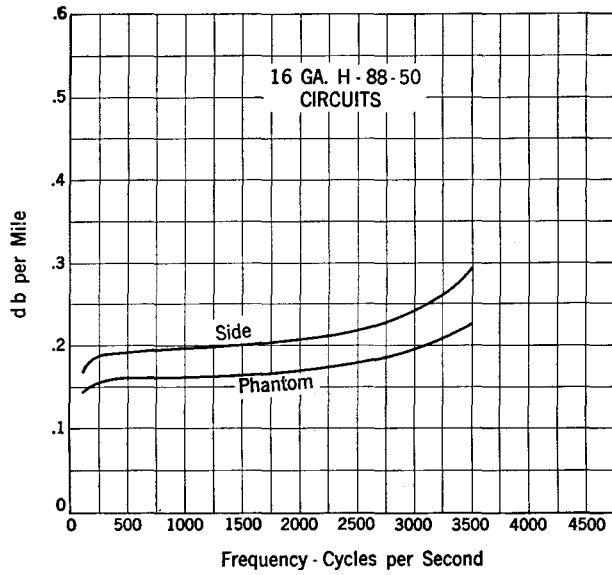
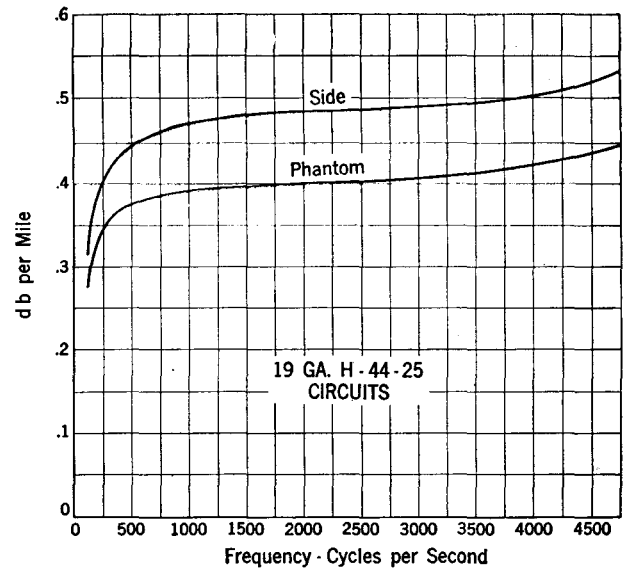
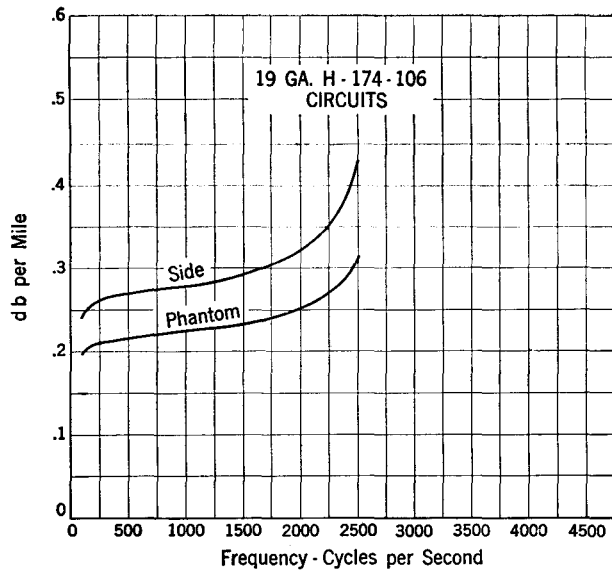


FIG. 21-10 ATTENUATION-FREQUENCY CHARACTERISTICS OF VARIOUS TYPES OF LOADED CABLE CIRCUITS [ 182 ]

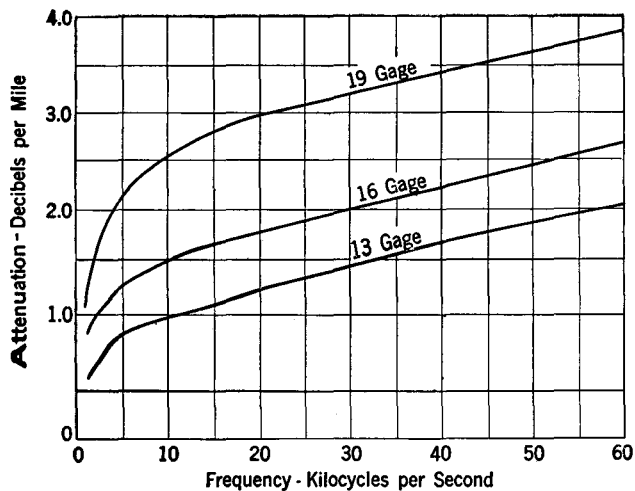


FIG. 21-11 ATTENUATION-FREQUENCY CHARACTERISTICS OF NON-LOADED CABLE CIRCUITS OVER THE TYPE-K CARRIER RANGE

the larger gages of open wire. That is, 165 open wire will be connected to a 13-gage pair in the cable while 104 will be connected to a 19-gage pair. For voice-frequency open wire circuits, the standard loading for toll entrance and intermediate cables is H-31-18. Loading may be applied to cable lengths as short as 2000 feet, but short lengths may require artificial building-out in order that the loading section shall be of the proper electrical length to give the desired characteristic impedance.

Where open wire lines carry Type-C carrier systems, the 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.

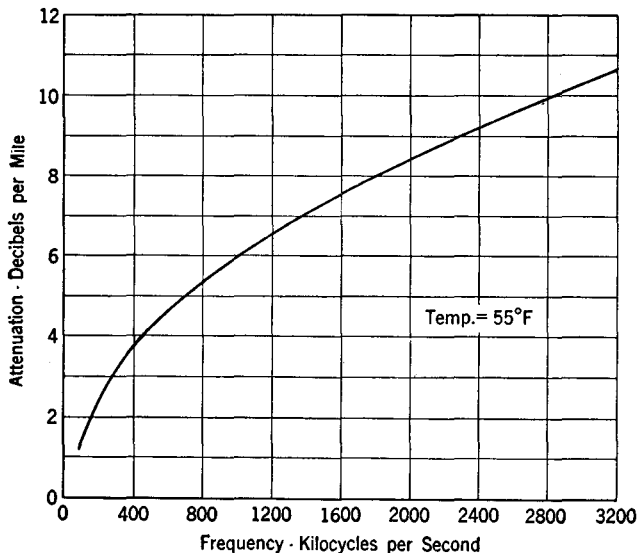


FIG. 21-12 ATTENUATION-FREQUENCY CHARACTERISTIC OF TYPICAL .375 COAXIAL

This loading may be modified by means of capacitance and resistance building-out to be satisfactory with 8-inch spaced open wire. For very short lengths of toll entrance cable and for long lengths of office cable used with Type-C carrier systems, X-2.7 loading is sometimes used.

In the case of open wire lines carrying Type-J carrier systems, the top frequencies are so high that it is impracticable to load entrance or intermediate cable conductors of the usual type properly for these systems. Non-loaded conductors are therefore used in most cases for toll entrance and the resulting higher attenuation is compensated by additional repeater gain. In certain cases, particularly in intermediate cables, however, a special

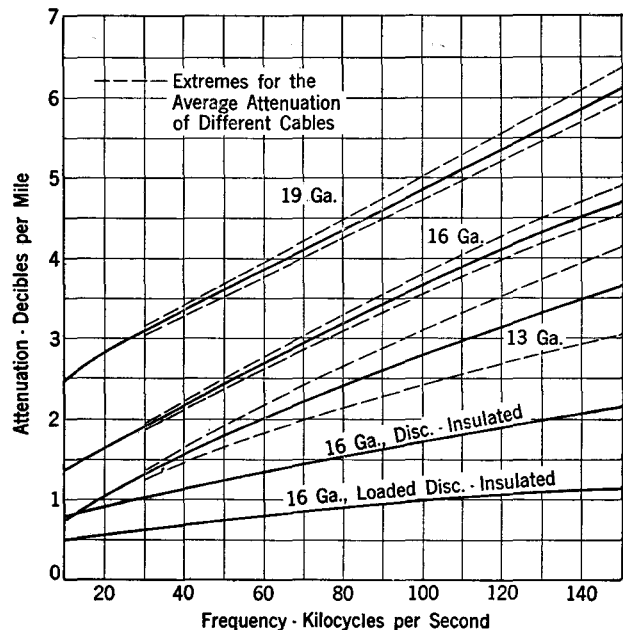


FIG. 21-13 ATTENUATION-FREQUENCY CHARACTERISTICS OF DISC-INSULATED TOLL ENTRANCE CABLE CIRCUITS COMPARED WITH NON-LOADED PAPER INSULATED CIRCUITS

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 capacitance 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 21-13. These disc-insulated quads may also be loaded to improve still further their attenuation and impedance characteristics.



## CHAPTER 22

### ATTENUATING, IMPEDANCE MATCHING AND EQUALIZING NETWORKS

#### 22.1 Pads or Attenuators

In the operation of various telephone circuits, it is frequently necessary to reduce artificially 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 obvi-

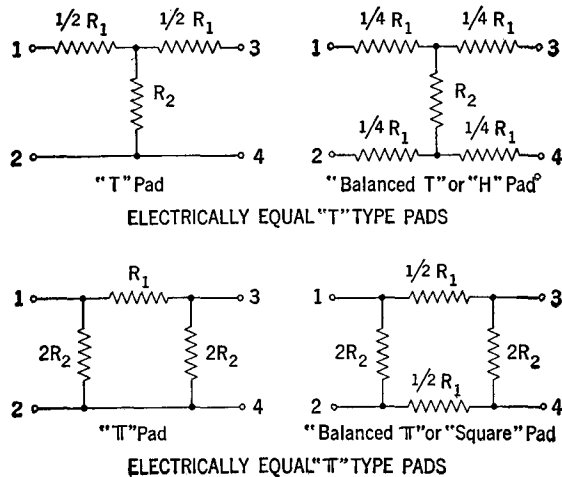


FIGURE 22-1

ously must be made up of 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 22-1.

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 22-1, 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 22-1. 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.

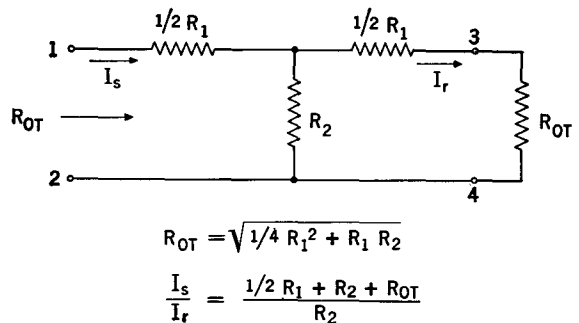


FIGURE 22-2

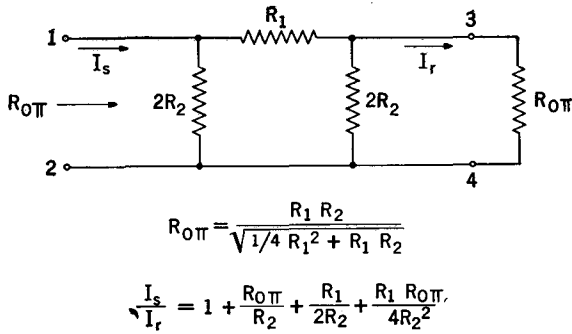


FIGURE 22-3

These relations for both the characteristic resistance and the current ratio are given for the T- and  $\pi$ -pads in Figure 22-2 and 22-3 respectively.

If the same values of series ( $R_1$ ) and shunt ( $R_2$ ) resistances are used in T- and  $\pi$ -pads, their characteristic resistance and db losses will be different. However, by using proper resistances of different values, two such pads can be made electrically identical. In Article 18.4, equations (18:1), (18:2) and (18:3), 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  $\pi$ -pad in Figure 22-4, the equivalent T-pad can be calculated. The

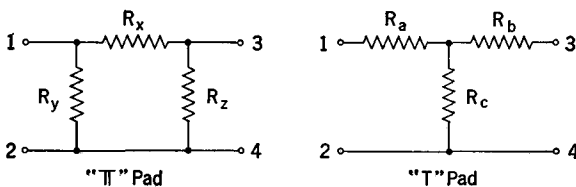


FIGURE 22-4

resistances  $R_1$ ,  $R_2$ , and  $R_3$  in Chapter 18 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 22-4, these resistances now become:

$$R_1 = \frac{(R_x + R_z) R_y}{R_x + R_y + R_z} = \frac{R_x R_y + R_y R_z}{R_x + R_y + R_z} \quad (22:1)$$

$$R_2 = \frac{R_x R_y}{R_x + R_y} \quad (22:2)$$

$$R_3 = \frac{(R_x + R_y) R_z}{R_x + R_y + R_z} = \frac{R_x R_z + R_y R_z}{R_x + R_y + R_z} \quad (22:3)$$

By substituting these values in equations (18:1), (18:2) and (18:3), and solving for the resistances of the T-network, we get—

$$R_a = \frac{R_x R_y}{R_x + R_y + R_z} \quad (22:4)$$

$$R_b = \frac{R_x R_z}{R_x + R_y + R_z} \quad (22:5)$$

$$R_c = \frac{R_y R_z}{R_x + R_y + R_z} \quad (22:6)$$

Where the  $\pi$ -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  $\pi$  type, we can make use of three similar equations which can be developed from equations (22:4), (22:5) and (22:6) 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 (22:4) and (22:5), (22:4) and (22:6), and (22:5) and (22:6) which gives—

$$R_a R_b + R_a R_c + R_b R_c = \frac{R_x^2 R_y R_z + R_x R_y^2 R_z + R_x R_y R_z^2}{(R_x + R_y + R_z)^2}$$

This becomes—

$$R_a R_b + R_a R_c + R_b R_c = \frac{R_x R_y R_z}{R_x + R_y + R_z}$$

Then by dividing this equation by each of equations (22:6), (22:5), and (22:4), we get—

$$R_x = \frac{R_a R_b + R_a R_c + R_b R_c}{R_c} \quad (22:7)$$

$$R_y = \frac{R_a R_b + R_a R_c + R_b R_c}{R_b} \quad (22:8)$$

$$R_z = \frac{R_a R_b + R_a R_c + R_b R_c}{R_a} \quad (22:9)$$

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  $\pi$  type and from these values, each of the types illustrated in Figure 22-1 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 22-2,

$$600 = \sqrt{\frac{R_1^2}{4} + R_1 R_2}$$

or

$$\frac{R_1^2}{4} + R_1 R_2 = 360,000$$

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

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

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

Also from Figure 22-2

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

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

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

Substituting  $R_2$  in the first equation above—

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

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

From this

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

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

and

$$R_2 = .231 R_1 + 277.8$$

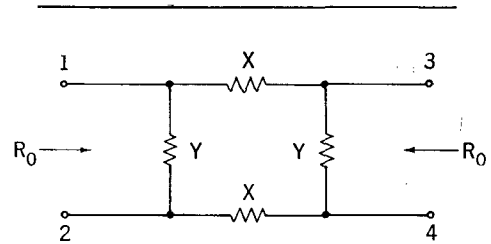
$$= 144 + 277.8 = 421.8 \text{ ohms.}$$

The H-pad will have a shunt resistance ( $R_2$ ) of 421.8 ohms, and each of the four series arm resistances ( $R_1/4$ ) will be 155.8 ohms.

A common use of pads is as an auxiliary method of controlling the net gain of telephone repeaters. Such pads are usually of the square type. Table XII gives the series and shunt resistance values for symmetrical square pads of 300-ohm impedance 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 ad-

TABLE XII  
SERIES (X) AND SHUNT (Y) RESISTANCE VALUES  
OF BALANCED  $\pi$  OR SQUARE PADS  
 $R_0 = 300$  ohms



Resistance Values for Square Pads

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

justed 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.

## 22.2 Impedance Matching

Because it can be designed to have any desired impedance values looking in either direction, a simple resistance pad can be used to match any two resistive impedances. The attenuation loss of such a pad, however, would generally be greater than the reduction in reflection loss obtained by impedance matching, although its use might still be desirable for other reasons.

Within certain limits, impedances may be matched by the insertion of a simple T- or  $\pi$ -pad made up entirely of inductors and capacitors. Such a pad would introduce relatively little attenuation loss but its effectiveness as an impedance matching device would of course be limited to the frequency for which its reactance values were determined. Repeating coils or transformers

are therefore generally used for impedance matching purposes in voice-frequency and other relatively low frequency circuits.

As we learned in Chapter 17, however, iron-core transformers cannot be used at the higher carrier and radio frequencies because of excessive core losses. Air-core transformers also have definite shortcomings at very high frequencies. Other devices accordingly must be devised for matching unlike radio frequency transmission lines or for matching such lines to their loads (as, for example, the connection between an antenna and the transmission line from a radio transmitter). Among the most effective of such devices is the quarter-wavelength resonant line.

If an impedance  $Z_R$  is connected to one end of a quarter-wave line whose characteristic impedance is  $Z_0$ , it can be shown that at the resonant frequency (i.e., the frequency at which the line is a quarter-wavelength long) the impedance  $Z_S$  looking into the other end of the line will be—

$$Z_S = \frac{Z_0^2}{Z_R}$$

from which, we may write—

$$Z_0 = \sqrt{Z_S Z_R} \quad (22:10)$$

A quarter-wave line designed to have the characteristic impedance  $Z_0$  as determined from this equation will therefore serve to match two unlike impedances,  $Z_S$  and  $Z_R$ , so that there will be no reflection losses at the junction points.

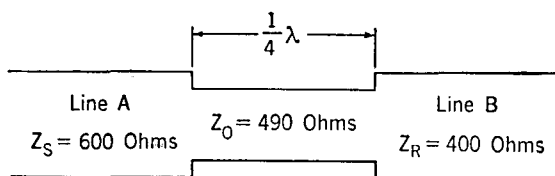


FIGURE 22-5

As an example, assume that it is desired to connect two non-resonant lines A and B having impedances of 600 ohms and 400 ohms respectively for through transmission at a certain high frequency. A quarter-wave line is inserted between them as indicated in Figure 22-5. Then if the quarter-wave line is designed so that its characteristic impedance  $Z_0 = \sqrt{600 \times 400}$  or approximately 490 ohms, the impedance looking from line A into the quarter-wave line will be 600 ohms, and the impedance looking from line B into the quarter-wave line will be 400 ohms.

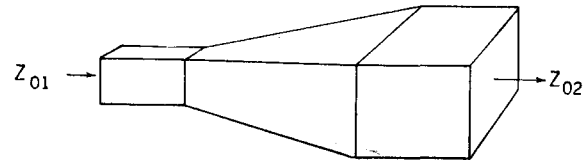


FIG. 22-6 TAPERED IMPEDANCE-MATCHING SECTION FOR WAVEGUIDE

Because of this useful ability to match unequal impedances, the resonant line when so used is commonly known as a **quarter-wave transformer**. It is also interesting to note that a resonant half-wavelength line acts like a transformer of unity ratio. The impedance looking into one end of such a line is equal to the impedance connected to the other end regardless of the value of the characteristic impedance of the half-wave line itself.

Quarter-wave sections or "lines" may also be used to match waveguides of different impedances. Here the coupling section is a piece of waveguide one quarter-wavelength long with cross-sectional dimensions such that its characteristic impedance is equal to the square root of the product of the impedances of the two guides which it couples. Generally, however, matching of waveguides may be more easily accomplished by means of a simple tapered section of guide as shown in Figure 22-6. Such a section, however, must be several wavelengths long if reflection losses are to be completely avoided.

### 22.3 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.

Attenuation equalizers are networks consisting of inductors, capacitors, and resistors, 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

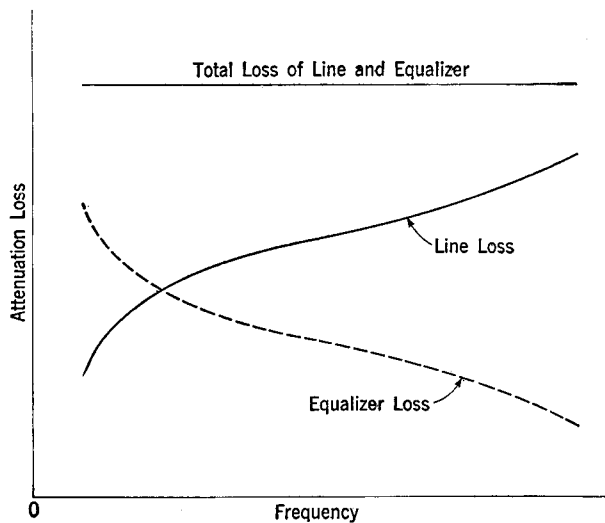


FIG. 22-7 PRINCIPLE OF ATTENUATION EQUALIZER

produced by the equalizer will be substantially the same for all frequencies in the transmitted band. This principle is shown in Figure 22-7.

One of the simplest types of equalizers, shown schematically in Figure 22-8, 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.

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. Instead of changing the net loss of the line, however, this arrangement changes the overall gain-frequency

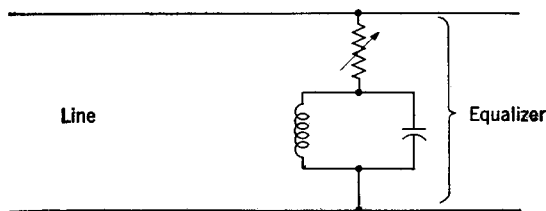
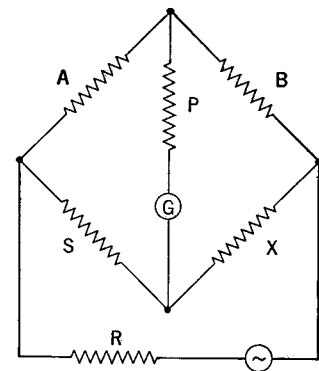


FIG. 22-8. SIMPLE BRIDGED EQUALIZER

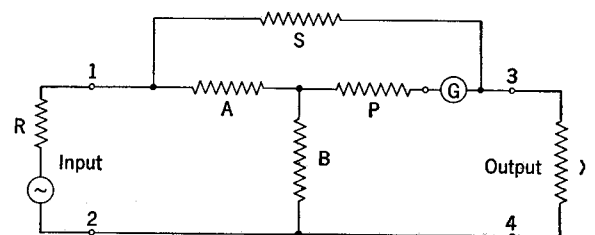
characteristic of the repeater to match reasonably closely the loss-frequency characteristic of the line. That is, for the frequencies where the line loss is high the repeater gain is also high and vice versa. The overall loss-frequency characteristic of the line and repeater together is then reasonably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

## 22.4 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.



A. WHEATSTONE BRIDGE CIRCUIT



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

FIGURE 22-9

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

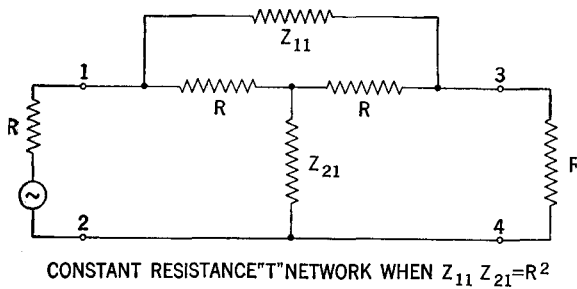


FIG. 22-10. BRIDGED T-EQUALIZER

arrangement, and the principle of its operation may be best grasped by first referring to the ordinary Wheatstone bridge circuit illustrated in Figure 22-9(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:

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

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 22-9(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 22-10, we still have the same bridge which was balanced when  $SB = XA$ , and is now balanced when—

$$Z_{11}Z_{21} = R \times R \quad \text{or} \quad R^2 \quad (22:11)$$

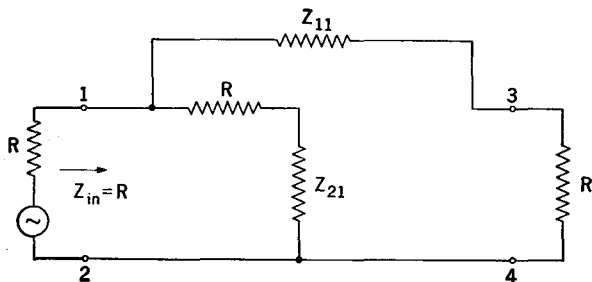


FIGURE 22-11

When the bridge is balanced no current flows in the impedance,  $P$  (Figure 22-9(A)), and for purposes of analysis we may therefore simplify the network by removing the  $R$  resistance which replaced  $P$  in Figure 22-10 giving us the network of Figure 22-11. 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—

$$\begin{aligned} Z_{in} &= \frac{(R + Z_{11})(R + Z_{21})}{R + Z_{11} + R + Z_{21}} \\ &= \frac{R^2 + RZ_{11} + RZ_{21} + Z_{11}Z_{21}}{2R + Z_{11} + Z_{21}} \end{aligned}$$

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

$$Z_{in} = \frac{R(2R + Z_{11} + Z_{21})}{2R + Z_{11} + Z_{21}} = R \quad (22:12)$$

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 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—

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

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 22-11), 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

$$\frac{I_b}{I_a} = \frac{2R + Z_{11} + Z_{21}}{R + Z_{21}} \quad (22:13)$$

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 (22:13), we get—

$$\begin{aligned} \frac{I_b}{I_a} &= \frac{R^2 + 2RZ_{11} + Z_{11}^2}{R^2 + RZ_{11}} \\ &= \frac{R + Z_{11}}{R} = 1 + \frac{Z_{11}}{R} \end{aligned} \quad (22:14)$$

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 22-11 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:

<i>When <math>Z_{11}</math> is:</i>	<i><math>Z_{21}</math> becomes:</i>
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.

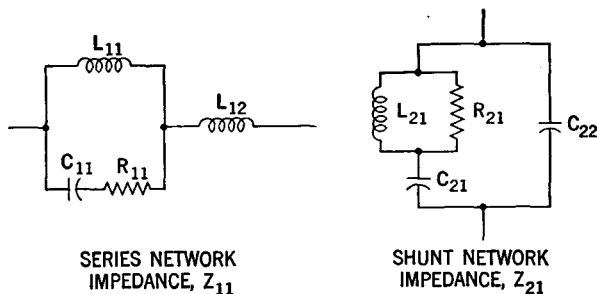


FIG. 22-12 INVERSE SERIES AND SHUNT NETWORKS

This inverse relationship is further illustrated in Figure 22-12 where the series network,  $Z_{11}$ , and its inverse shunt network,  $Z_{21}$ , are shown at 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 22-12  $C_{21}$  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  $R_{11}$ .

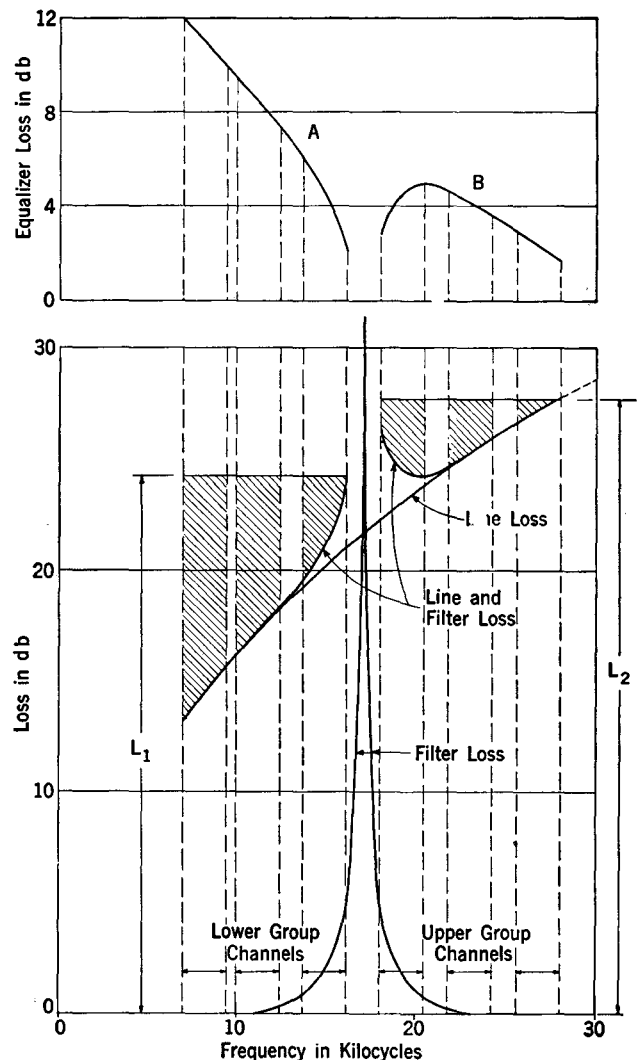


FIG. 22-13 ATTENUATION EQUALIZATION FOR TYPE-C CARRIER SYSTEM

In designing a bridged T-equalizer for a specific use, the attenuation-frequency characteristic of the  $Z_{11}$  network must be complementary to the at-

tenation-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 cutoff 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 22-13. 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

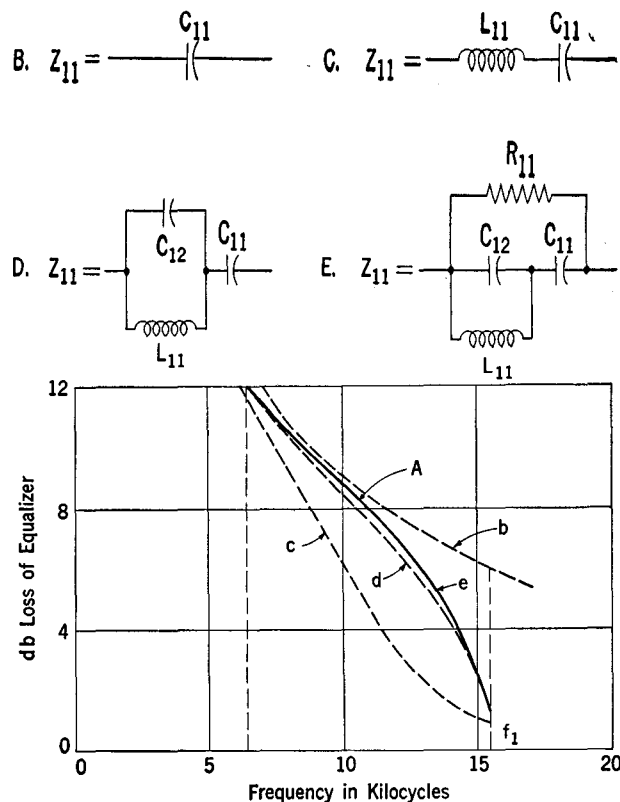


FIGURE 22-14

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 22-13, or with the solid line curve of Figure 22-14, which is the same. As a first approach, a  $Z_{11}$  circuit made up of a single series capacitor, 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 capacitor 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 capacitor 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 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 capacitance 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*.

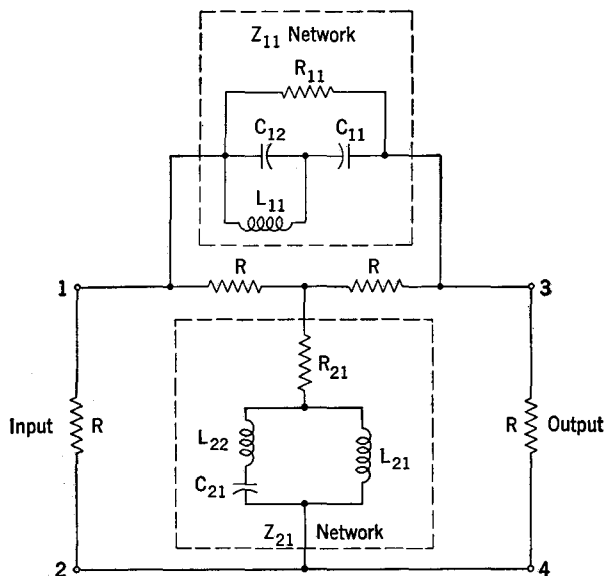


FIGURE 22-15

The final series network,  $Z_{11}$ , and its inverse,  $Z_{21}$ , then take the form illustrated in Figure 22-15. In the inverse network,  $Z_{21}$ , the shunting resistance  $R_{11}$  becomes a series resistance  $R_{21}$ ; the series capacitance  $C_{11}$  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, capacitors, or resistors it is considered economical to use.

## 22.5 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

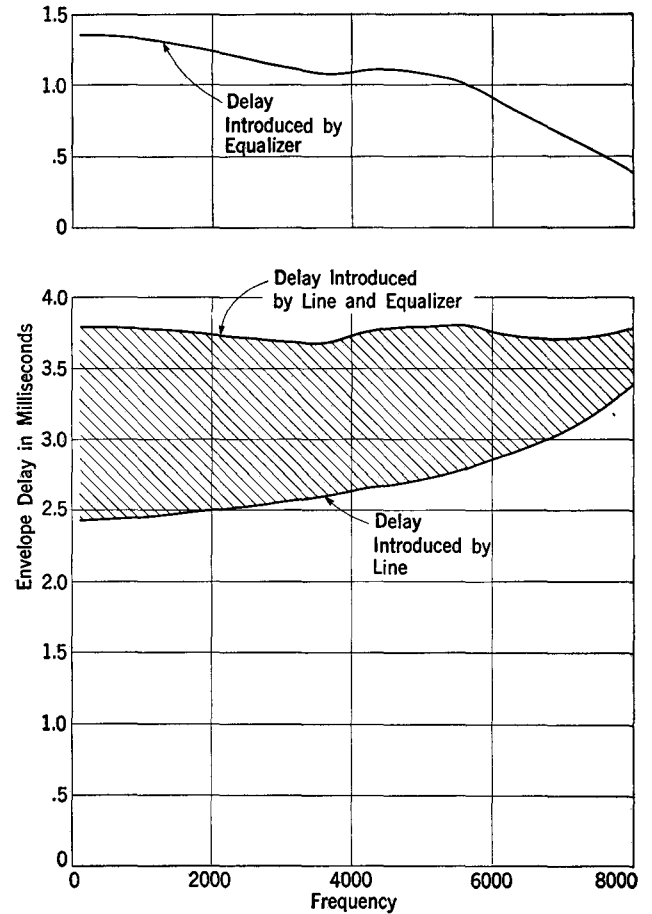


FIG. 22-16 PRINCIPLE OF DELAY OR PHASE EQUALIZING NETWORK

telephone lines vary with frequency. The velocity of propagation,  $W$ , over these lines also may change with frequency.

A pair of wires of zero resistance in free space, separated from all other conductors and without leakage, would transmit electric waves at the

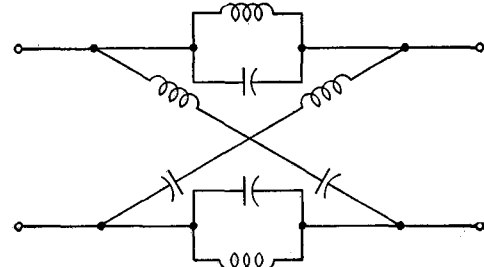


FIG. 22-17 LATTICE-TYPE NETWORK

speed of light. In an open wire circuit, what retardation exists comes largely from the resistance of the wires. In loaded cable circuits, there is a much greater retardation due in part to the larger capacitance between the wires, but much more to the inductance of the loading coils which are inserted to decrease the attenuation. In any case, there is a finite time interval for transmission between the sending end of any circuit and its receiving end.

This delay interval 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 may not be the same for all frequencies in the transmitted band. A changing velocity of propagation at the various frequencies means that the shape of a signal wave at the receiving end of the line will differ to an appreciable degree from the waveshape applied at the sending end. Distortion of this kind is called **envelope distortion**, and its seriousness is measured in terms of **envelope delay** in milliseconds. This delay distortion may be equalized by inserting in series with the circuit a network having the inverse characteristics. The total delay produced by the circuit, added to that of the equalizer, will then be constant over the frequency band transmitted. This is illustrated by Figure 22-16.

Delay distortion, unless excessive, is of little concern in ordinary telephone circuits because the ear seems insensitive to it. For program

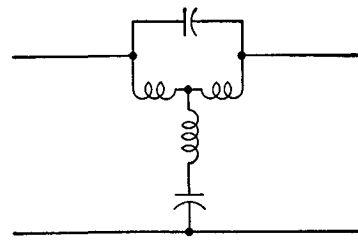


FIG. 22-18 BRIDGED T-NETWORK

transmission, however, where frequency bands extending up to 8,000 or 15,000 cycles may be employed, correction for phase distortion is generally necessary. It is even more necessary in the case of facsimile (telephotograph) transmission and video transmission, where even slight delay deviations may affect the quality of the received pictures.

The design of delay equalizers involves some rather complex network analysis. Two types of networks are generally used. The first known as a balanced lattice-type structure is shown schematically in Figure 22-17. The second is a bridged-T type of structure as illustrated in Figure 22-18. In either case the total equalizer is made up of a number of such sections connected in tandem. The bridged-T type of network is somewhat more economical when balance between the two sides of the line is not necessary, although either type will produce generally similar envelope delay characteristics.

## CHAPTER 23

### FILTERS

#### 23.1 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 **electric wave filters**. As would be expected from our study of resonance in Article 16.4, these filters are essentially networks of inductances and capacitances. While the details of design of some of the more elaborate types of filters are beyond the scope of this text, we may consider here the general principles of the four 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 cutoff 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 cutoff value but readily transmit currents of all higher frequencies; (3) "band-pass" filters which have both an upper and a lower cutoff point, and which permit the transmission of only those frequencies lying between the two cutoff frequencies; (4) "band-elimination" filters which have over-lapping upper and lower cutoff points and thus prevent the transmission of frequencies between the two cutoff 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 designs of filters. Naturally, however, filters are designed to meet specific requirements using the simplest networks practicable.

#### 23.2 Low-and High-Pass Filter Sections

The simplest arrangements of elementary filter networks are the **T** and  $\pi$  sections shown in Figure 23-1, 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.

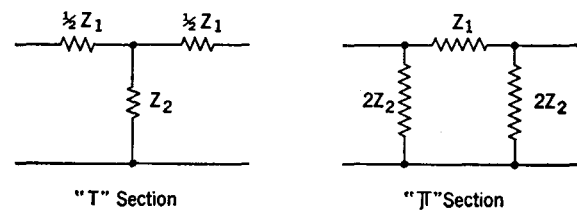


FIGURE 23-1

As covered in Article 18.5, 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 (inductive), or negative (capacitive) reactance. Accordingly, the characteristic impedance must have an angle of  $0^\circ$  or  $\pm 90^\circ$ . When the open- and short-circuited impedances (reactances) have opposite signs—

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

or

$$Z_0 = \sqrt{X_{oc}X_{sc}} \quad (23:1)$$

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

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

or

$$Z_0 = \pm j\sqrt{X_{oc}X_{sc}} \quad (23:2)$$

and the impedance has an angle of  $\pm 90^\circ$  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 electric and magnetic fields, and return it during a later part of the cycle. In other words, the filter will not transmit any energy and therefore may be said to have infinite attenuation.

Over a 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 23-2(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 23-2(D).

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

$$Z_0 = \sqrt{Z_{oc}Z_{sc}} \quad (23:3)$$

but we also know from equation (18:4) that—

$$Z_0 = \sqrt{\frac{Z_1^2}{4} + Z_1Z_2} \quad (23:4)$$

which may also be written as—

$$Z_0 = \sqrt{Z_1\left(\frac{Z_1}{4} + Z_2\right)} \quad (23:5)$$

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. On the other hand, if these two reactances are of opposite signs the characteristic impedance 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 reactance curves for  $Z_1$  and

$\left(\frac{Z_1}{4} + Z_2\right)$  as shown in Figure 23-2(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 23-2(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 (23:5). 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 other side the filter attenuates. This cutoff frequency,  $f_c$ , may be determined from the equation

$$\frac{Z_1}{4} + Z_2 = 0 \quad (23:6)$$

In the low-pass prototype filter section of Figure 23:2(A)—

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

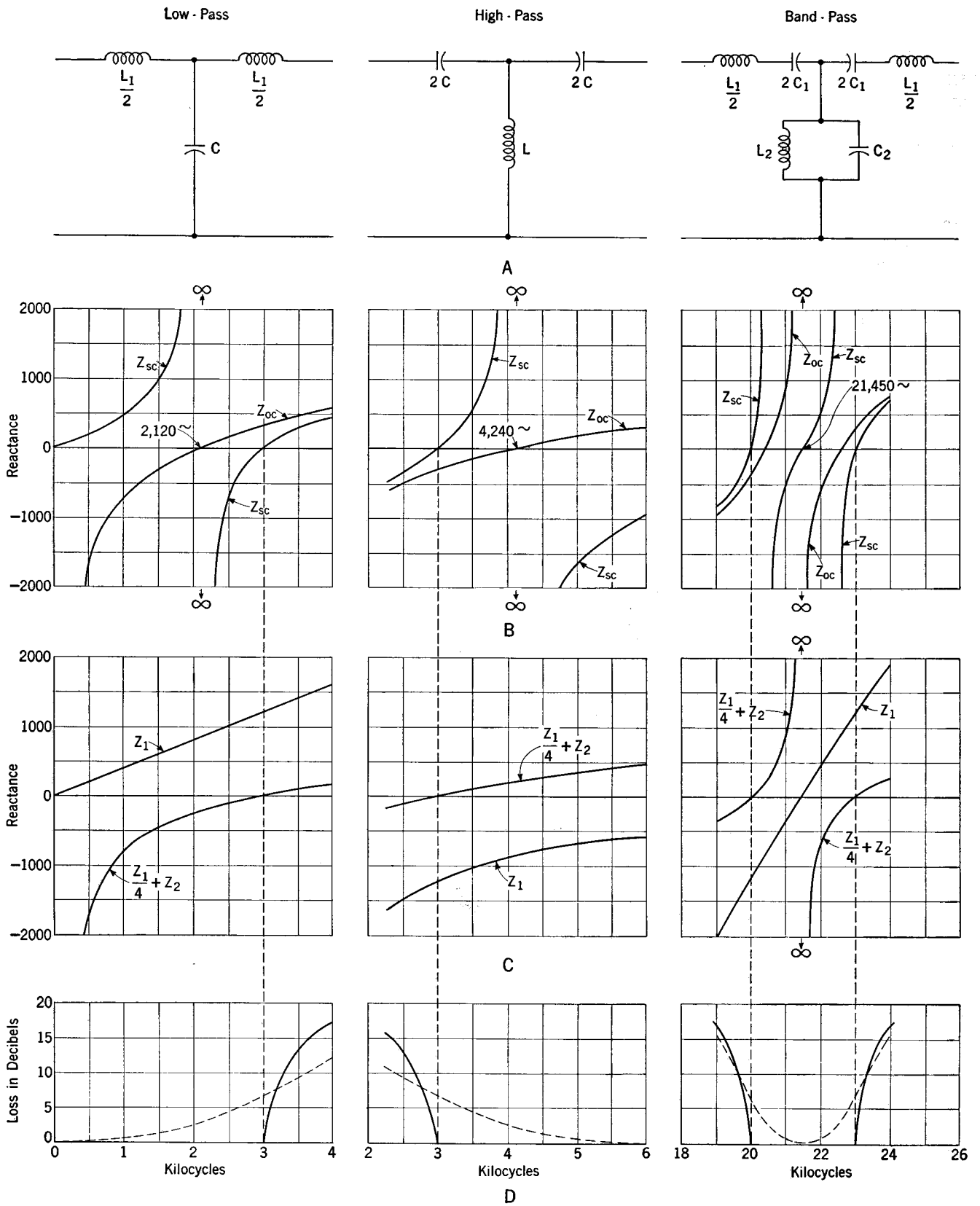


FIG. 23-2 REACTANCE AND LOSS CHARACTERISTICS OF ELEMENTARY PROTOTYPE FILTER SECTIONS

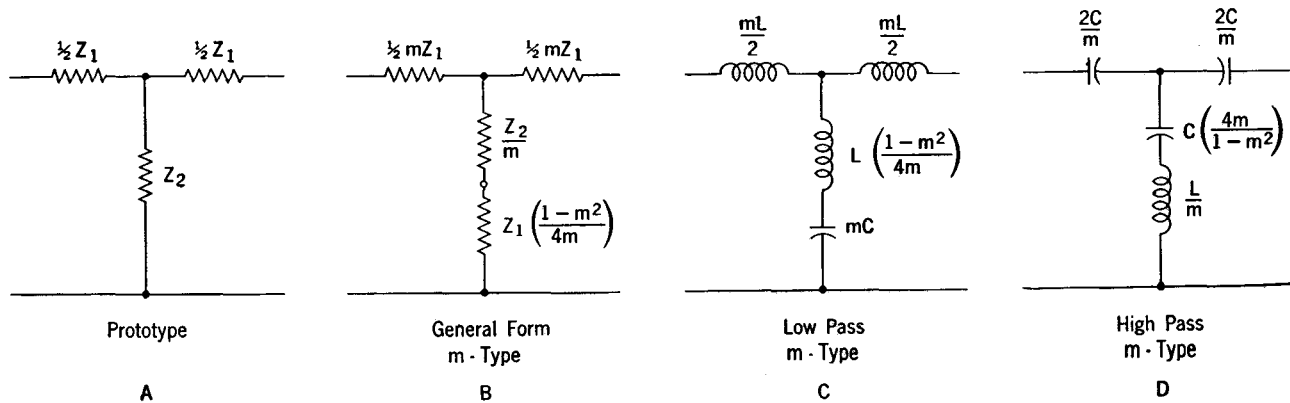


FIG. 23-3 M-TYPE FILTER SECTIONS

at the critical frequency,  $f_c$ . Substituting these values in equation (23:6), we get—

$$f_c = \frac{1}{\pi\sqrt{LC}} \quad (23:7)$$

Similarly, in the high-pass prototype filter section of Figure 23-2 (A)—

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

Then the cutoff frequency is—

$$f_c = \frac{1}{4\pi\sqrt{LC}} \quad (23:8)$$

### 23.3 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 23-2(A) to derive equations for the cutoff 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 23.1. This will be apparent from a glance at the loss-frequency curves shown by the dashed lines in Figure 23-2(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 “insertion” 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 cutoff frequency is anything but sharp.

Some improvement could be obtained by connecting several of these structures in tandem, but the cutoff point would still be not clearly defined and there would be altogether too much attenua-

tion in the pass band. To improve the sharpness of cutoff, what we need is a structure which will produce very much higher losses at frequencies just beyond the cutoff frequency. This objective can be approached by making the shunt impedance of the structure resonant at a frequency a few cycles beyond the cutoff frequency. Such a structure would, of course, have to contain series capacitance 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 23-3(A) where—

$$Z_0 = \sqrt{\frac{Z_1^2}{4} + Z_1 Z_2} \quad (23:4)$$

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—

$$Z'_0 = Z_0 \quad (23:9)$$

and

$$Z'_1 = mZ_1 \quad (23:10)$$

Then using equations (23:4), (23:9) and (23:10), 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—

$$Z'_2 = \frac{Z_2}{m} + Z_1 \frac{(1 - m^2)}{4m} \quad (23:11)$$

The  $m$ -type section then takes the general form shown in Figure 23-3(B). This applies to both low- and high-pass types but Figures 23-3(C) and 23-3(D) show the respective values in terms of  $m$ ,  $L$  and  $C$  for low-pass and high-pass T-sections. It may be noted that the second term of equation (23:11) can be physically realized only if the value of  $m$  lies between zero and +1. By using different values of  $m$  within these limits, 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 cutoff 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—

$$\frac{Z_2}{m} + Z_1 \left( \frac{1 - m^2}{4m} \right) = 0 \quad (23:12)$$

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

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

By substituting these values in equation (23:12) and solving for the frequency of infinite attenuation,  $f_r$ , we get—

$$f_r = \frac{1}{\pi\sqrt{LC(1-m^2)}} = \frac{1}{\pi\sqrt{LC}} \times \frac{1}{\sqrt{1-m^2}} \quad (23:13)$$

Since, as shown in equation (23:7), the cutoff 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—

$$m_{lp} = \sqrt{1 - \left( \frac{f_c}{f_r} \right)^2} \quad (23:14)$$

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 (23:12), and its value found to be—

$$m_{hp} = \sqrt{1 - \left( \frac{f_r}{f_c} \right)^2} \quad (23:15)$$

The closer the values of the two frequencies,  $f_c$  and  $f_r$ , are to each other the lower the value of  $m$ .

Curve B of Figure 23-4 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.

### 23.4 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 cutoff value as we have seen (Figure 23-2). 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.

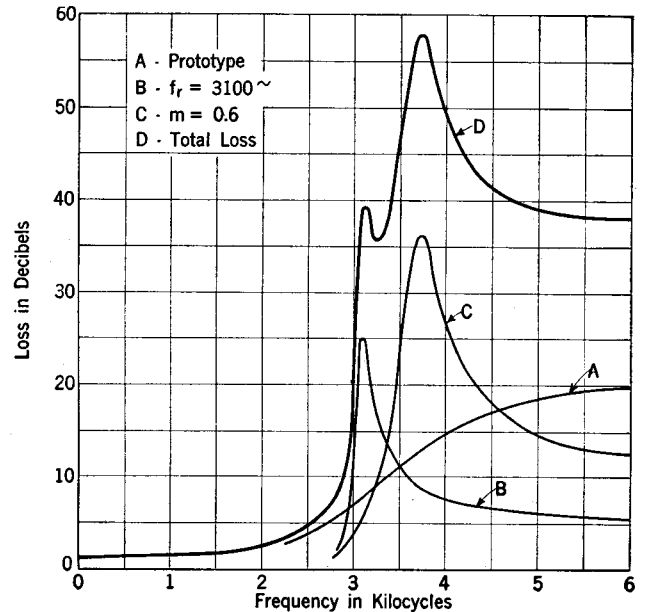


FIG. 23-4 CHARACTERISTICS OF SECTIONS OF COMPOSITE LOW-PASS FILTER SHOWN IN FIG. 23-6

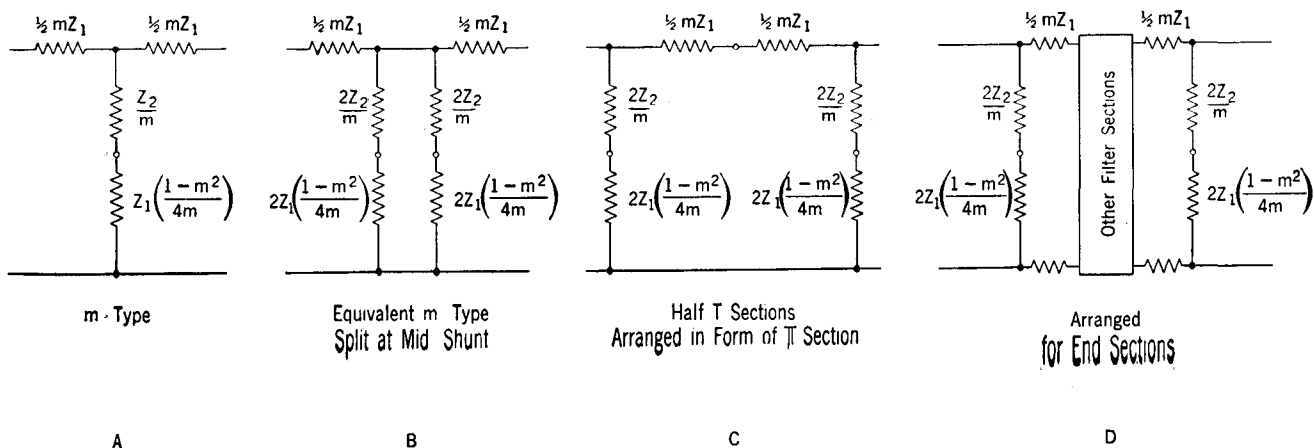


FIG. 23-5 HALF T M-TYPE FILTER END SECTIONS

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. This splitting arrangement is illustrated in Figure 23-5(B). We may then rearrange the series and shunt arms so that the network takes the form of a  $\pi$  as in Figure 23-5(C). It should be noted, however, that we are not converting the T-section into an electrically equivalent  $\pi$ -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, if characteristic impedance is plotted against frequency for various values of  $m$  it will be found that the impedance of such a structure is practically uniform over the major portion of the pass band when  $m$  is equal to 0.6. We can take advantage of this fact by breaking this  $\pi$ -section into two equal parts, each designed with  $m$  equal to 0.6, and using each half as an end section of the composite filter as shown in Figure 23-5(D). The terminal impedances of the complete filter will then be practically constant over the major portion of the pass band.

In designing a filter we must know the frequency band to be passed, the cutoff 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—

$$R = \sqrt{\frac{L}{C}} \quad (23:16)$$

for both the low- and high-pass filters. Combining this relation with the value of the cutoff frequencies as given in equations (23:7) and (23:8), we have for the low-pass filter—

$$L = \frac{R}{\pi f_c} \quad (23:17)$$

and

$$C = \frac{1}{\pi f_c R} \quad (23:18)$$

and for the high-pass filter—

$$L = \frac{R}{4\pi f_c} \quad (23:19)$$

and

$$C = \frac{1}{4\pi f_c R} \quad (23:20)$$

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

**Example:** Design a low-pass filter with a cutoff frequency of 3000 cycles to work into an impedance of 600/0°. To provide a sharp cutoff,  $f_r$  should be 3100 cycles.

**Solutions:** From equations (23:17) and (23:18), the inductance and capacitance for the prototype will be—

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

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

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



From equation (23:14) and Figure 23-3(C), the  $m$ -type section for  $f_r = 3100$  will be—

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

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

$$L\left(\frac{1 - m^2}{4m}\right) = 63.66 \left(\frac{1 - (.252)^2}{4 \times .252}\right) = 59.16 \text{ mil hen.}$$

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

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

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

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

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

The composite filter thus takes the form illustrated in Figure 23-6. The attenuation-frequency characteristics for this composite filter and each of its component structures are illustrated in Figure 23-4.

### 23.5 Band Filters

Band filters may be obtained by connecting a low-pass and a high-pass filter in series, with their cutoff frequencies so arranged as to pass or suppress the desired band. In practice, however,

these filters are designed and built as a single structure having two cutoff frequencies, using the same principles as already discussed for the low- and high-pass types. The band-pass filter shown in Figure 23-2 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 23-7. 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 necessary to use half T-sections for the end terminations with  $m$  equal to 0.6, thus providing uniform terminal impedances over the major portion of the pass band.

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

$$L_1 = \frac{R}{\pi(f_u - f_l)} \quad (23:21)$$

$$C_1 = \frac{(f_u - f_l)}{4\pi f_l f_u R} \quad (23:22)$$

$$L_2 = \frac{R(f_u - f_l)}{4\pi f_l f_u} \quad (23:23)$$

$$C_2 = \frac{1}{\pi(f_u - f_l)R} \quad (23:24)$$

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:

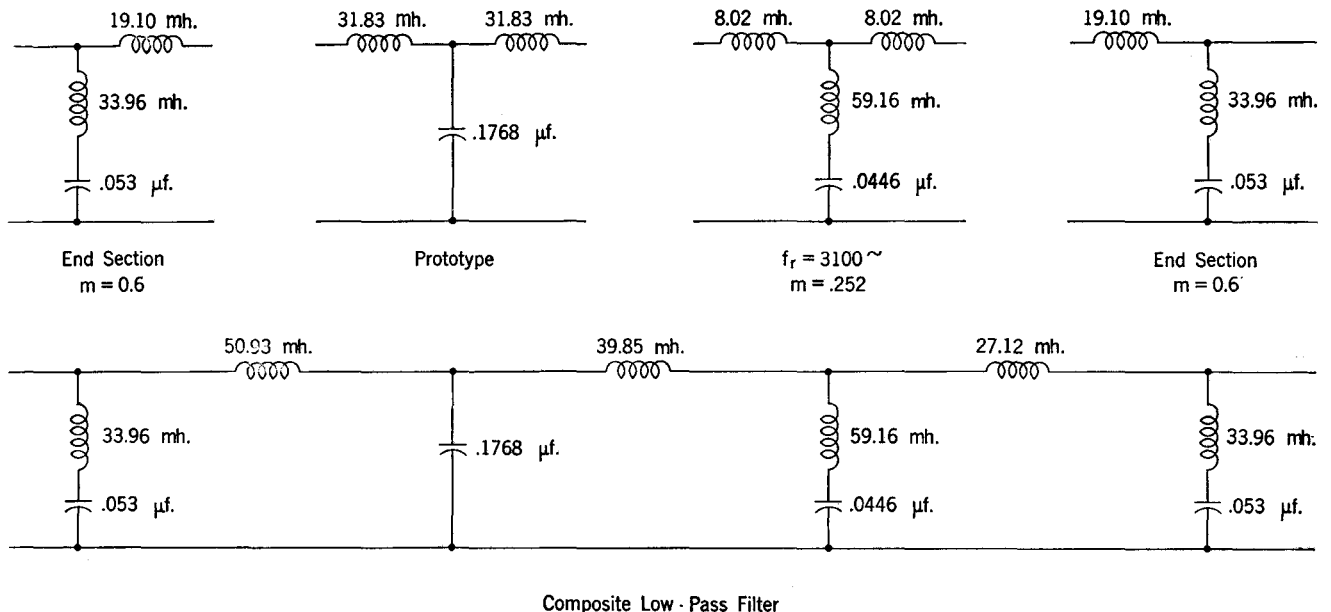


FIG. 23-6. MAKE-UP OF TYPICAL COMPOSITE LOW-PASS FILTER

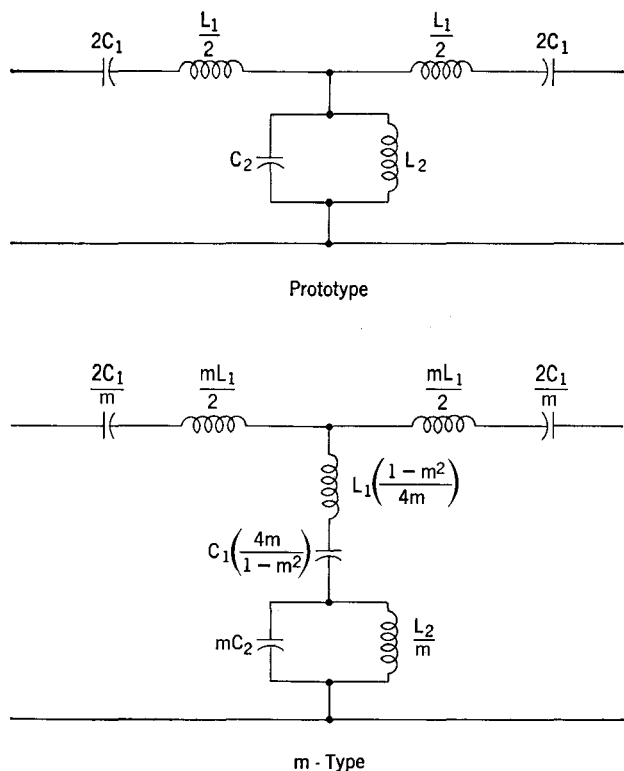


FIG. 23-7 BAND-PASS FILTER SECTIONS

$$m = \sqrt{1 - \left[ \frac{f_r(f_u - f_l)}{f_r^2 - f_u f_l} \right]^2} \quad (23:25)$$

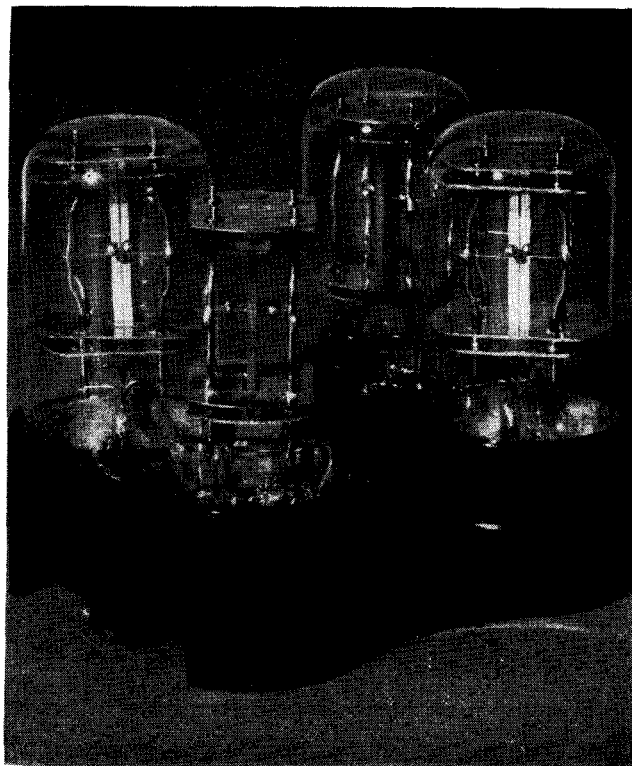
where  $f_r$  is the resonant frequency of infinite attenuation.

Since there are two cutoff frequencies,  $f_u$  and  $f_l$ , there should be two resonant frequencies to provide infinite attenuation both above the upper cutoff frequency,  $f_u$ , and below the lower cutoff frequency,  $f_l$ . It may be noted, however, from an inspection of the  $m$ -type section of Figure 23-7, 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 cutoff 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 cutoff frequency of 23,000 cycles, its value will be 23,460 cycles. By substituting these values in equation (23:25), the value of  $m$  will be found to be 0.627. This same  $m$  (0.627) also establishes an  $f_r$  for the lower cutoff frequency, which is 19,608 cycles or approximately 2% below 20,000 cycles.

## 23.6 Crystal Filters

In our discussion thus far, we have considered filters built with ideal reactance elements having no resistance. This, of course, is not the case in actual practice. The presence of resistance in the inductors 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.

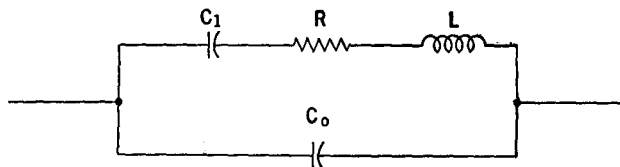
Other things being equal, the sharpness with which a properly designed filter cuts off, or the width of the transition band, is determined by the ratio of the reactance to the resistance of the inductors used in the filter. As we have noted, this ratio of reactance to resistance is generally referred to as  $Q$ . The reactance of a coil of given inductance increases directly with an increase in frequency while its resistance remains fairly con-



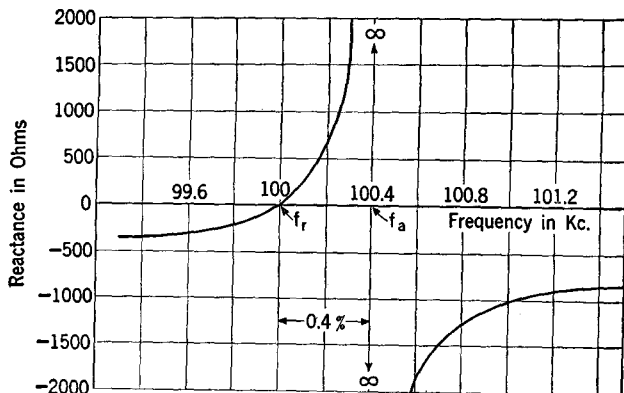
GLASS-SEALED CRYSTAL UNITS

stant 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.

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, one that has been found most satisfactory for these purposes is natural crystalline quartz. Artificially grown crystals of certain chemical compounds are also used effectively. Best known of these are EDT (Ethylene-Diamine-Tartrate) and DKT (Dipotassium-Tartrate). When a piezo-electric crystal is strained mechanically, an electric field is set up in its neighborhood, which will induce electric potential in any conductors in the field. Conversely, when a crystal is placed in an electric field, it is subject to mechanical strain and its shape changes very slightly. Thin slabs of the crystalline material when cut in certain ways will exhibit these same

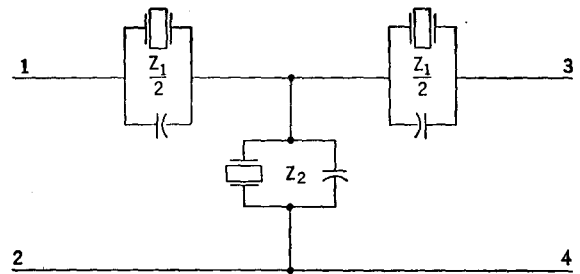


EQUIVALENT ELECTRICAL CIRCUIT OF A QUARTZ CRYSTAL

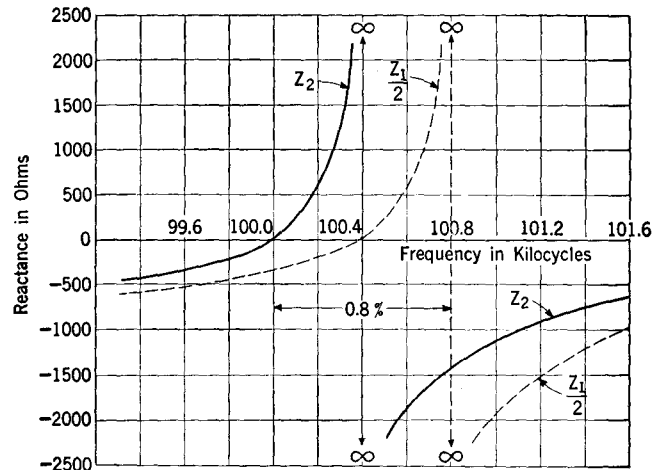


REACTANCE CHARACTERISTIC OF A QUARTZ CRYSTAL

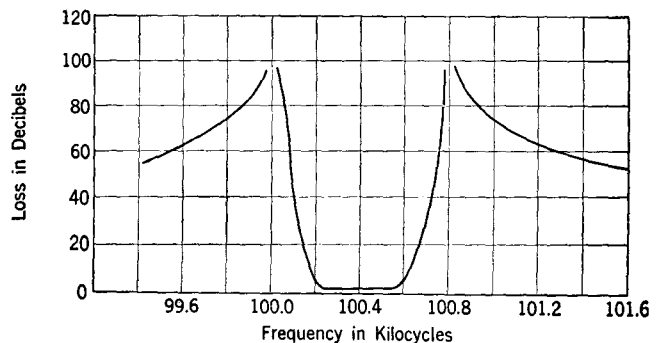
FIGURE 23-8



"T" FILTER SECTION USING CRYSTALS



REACTANCE CHARACTERISTIC CURVES



ATTENUATION-FREQUENCY CHARACTERISTIC

FIGURE 23-9

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 electric circuit such as a filter, a crystal acts like an impedance that can be represented electrically as shown in Figure 23-8, 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).

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 23-8 where the series resonant frequency is  $f_r$ , and the parallel resonant frequency is  $f_a$ . Since there is a fixed ratio of  $C_0$  to  $C_1$  (125 to 1) for quartz, the parallel 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 parallel-resonance, however,  $C_0$  must be taken into consideration and the effective capacitance in the looped circuit becomes  $\frac{125}{126}$  of that for the resonant condition, thereby making the parallel-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.

If an auxiliary capacitor is placed in parallel with the crystal, however, the effective capacitance of  $C_0$  is increased, and as a result the parallel-resonant frequency can be made less than 0.4 per cent above the series resonant frequency. This can be done without distracting from the favorable characteristics of the crystal because the  $Q$  of the added capacitor 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 23-9, where capacitors 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 capacitors 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 parallel-resonant frequency of the shunt arm (infinite reactance), the T-network will have points of maximum attenuation at the parallel-resonant frequency of the series arm (infinite reactance) and the series 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 parallel-resonance to that of

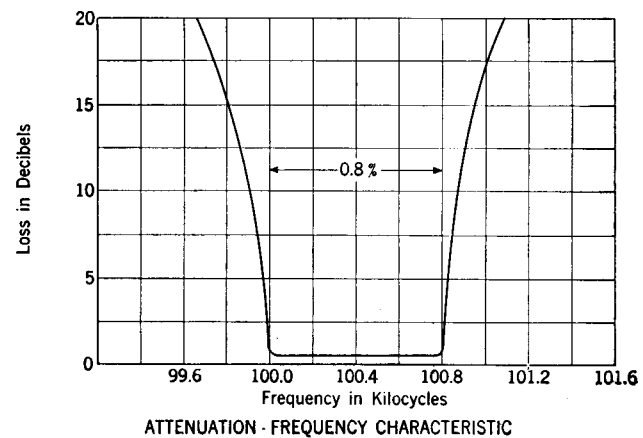
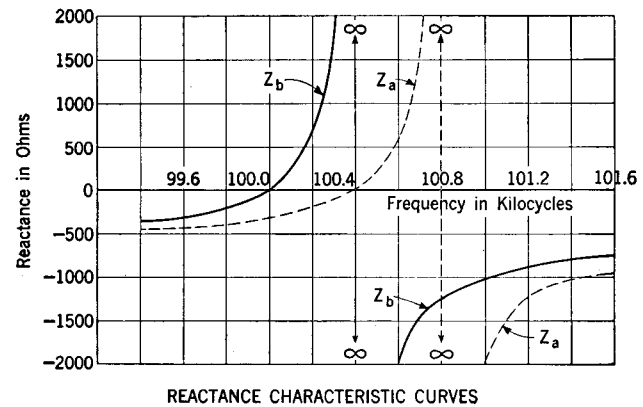
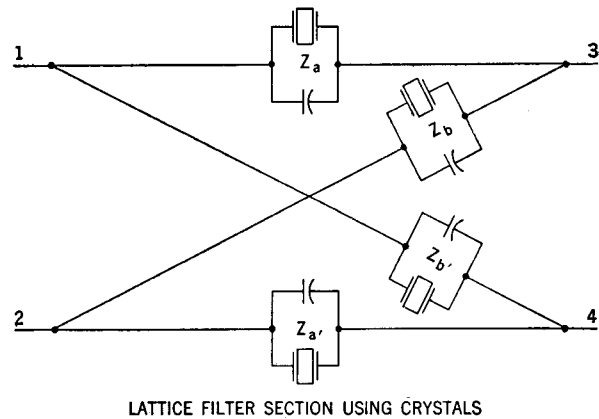
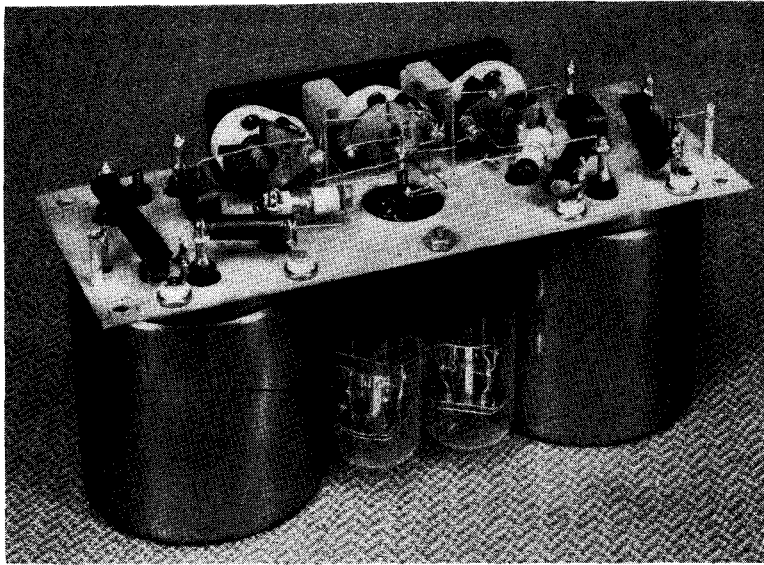


FIGURE 23-10



CRYSTAL CHANNEL FILTER

series resonance, the peaks of maximum attenuation cannot be separated by more than 0.8 per cent of the frequency scale; and with the shunting capacitors, the separation will be less than this amount, depending upon the capacitor values.

### 23.7 Lattice Networks

A more general type of filter is obtained by using a bridge type (lattice) network as illustrated in Figure 23-10, where the two series arms are alike as are the two shunt arms, but the series and shunt arms 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 23-10 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 (assuming no energy dissipated in the crystals).

This may be more easily seen from examination of this network rearranged in the more conven-

tional balanced bridge form of Figure 23-11. Here the series arms,  $A$  and  $A'$ , and the shunt arms,  $B$  and  $B'$ , may be either positive (inductive) or negative (capacitive) 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.

At both the resonant frequency for the shunt arm (zero reactance or short circuit) and the parallel-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 cutoff 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

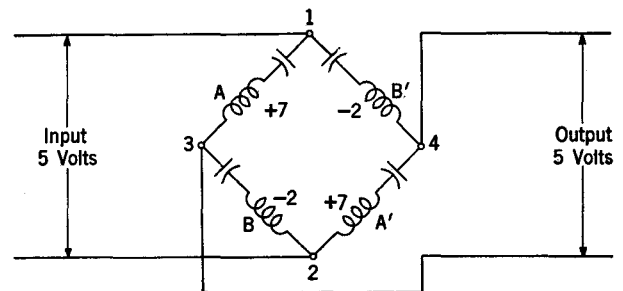


FIG. 23-11 PRINCIPLE OF LATTICE NETWORK

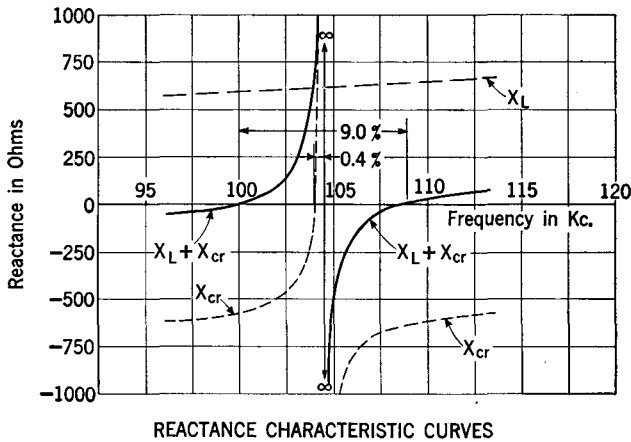
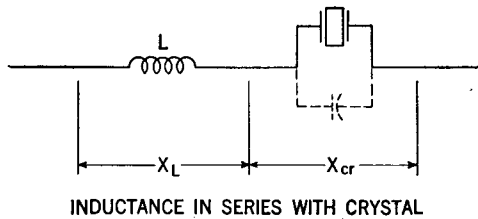


FIGURE 23-12

(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.

If a coil is placed in series with a crystal, the reactance characteristics of the combination are as illustrated in Figure 23-12, where there are now two series 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. It can be shown that when the parallel-resonant frequency is half-way between the two series resonant frequencies, the parallel-resonant frequency is 4.5 per cent greater than the first series resonant frequency. This means the two series resonant frequencies are now separated by 9.0 per cent.

By adding inductances to the lattice network of Figure 23-10 we have the filter illustrated in Figure 23-13, 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  $3 \times 4.5$  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 capacitors in parallel

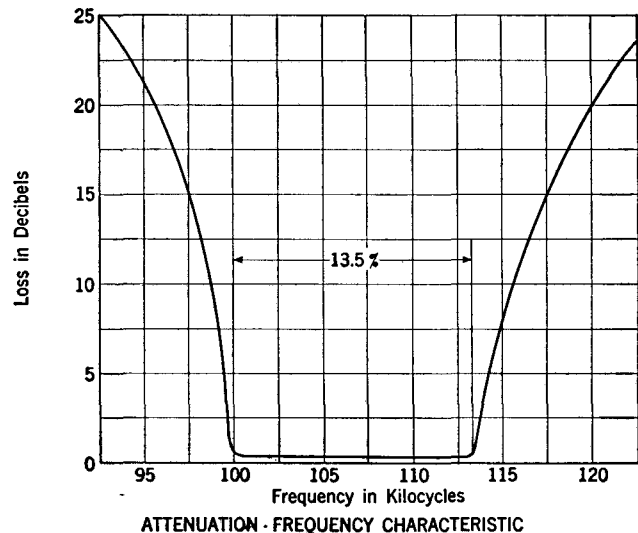
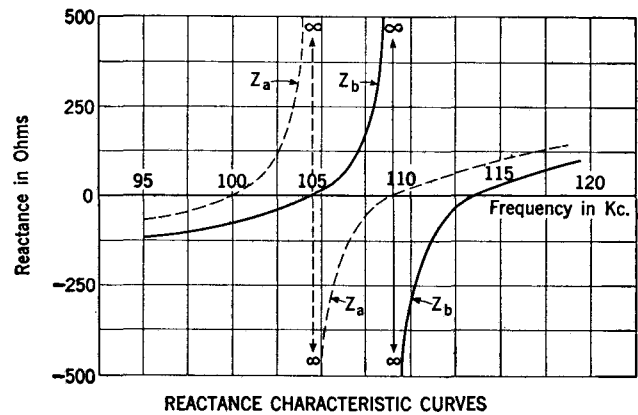
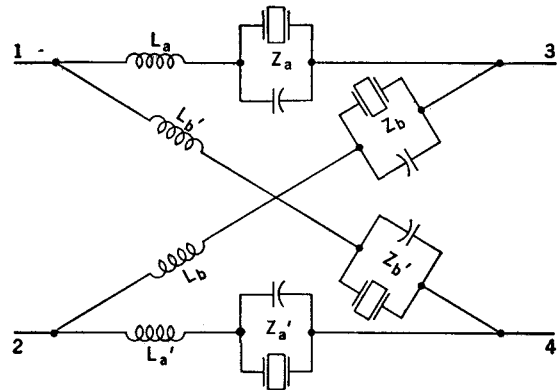


FIGURE 23-13

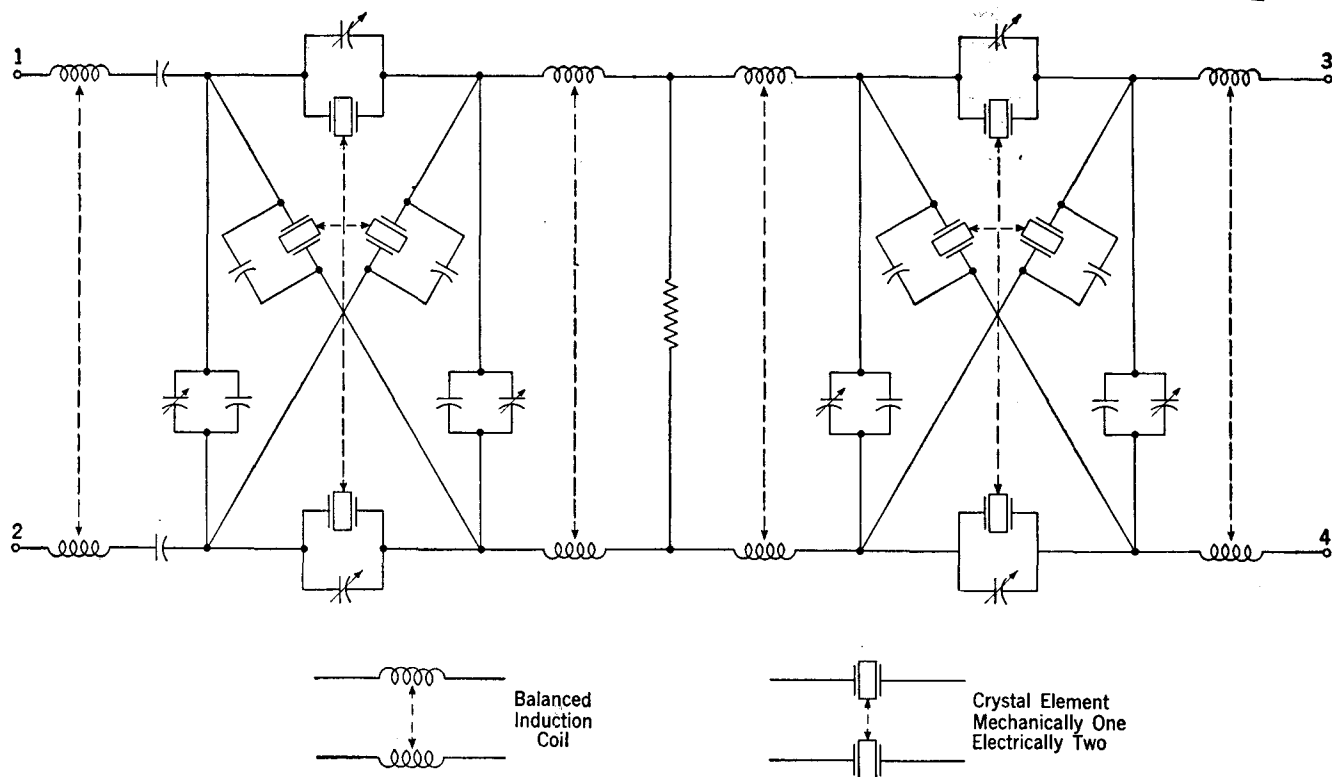


FIG. 23-14. TYPICAL BAND-PASS CRYSTAL FILTER

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 inductors becomes rather large for widths of less than 0.5 per cent. Above 13.5 per cent width, the ordinary coil and capacitor filter is satisfactory, and below 0.5 per cent width, where the losses of the coils becomes 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.

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 23-14. 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 23-14. This is accomplished by plating the surfaces of the crystals used with a thin 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 23-15. The four crystals used in the lowest frequency channel (64 kc) of the Type-K carrier system range in length from about 1.65 to 1.58 inches, while those in the highest frequency channel (108 kc) range from about 0.96 to 0.94 inch. These crystals vary in thickness from about 0.0248 inch

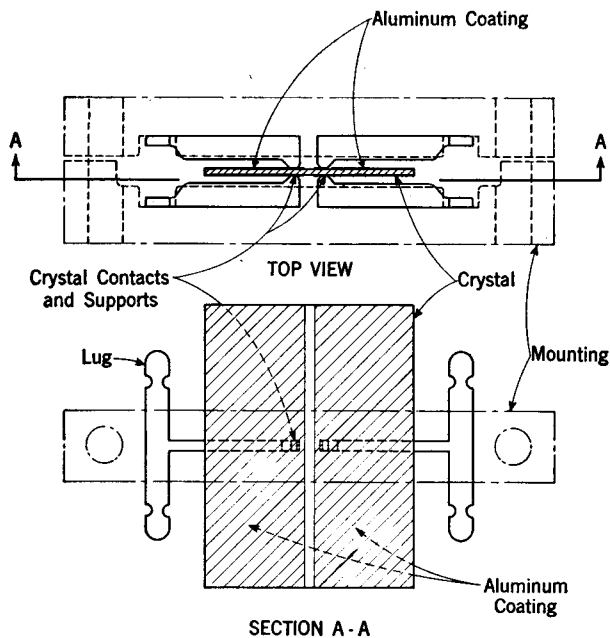


FIG. 23-15. METHOD OF MOUNTING QUARTZ CRYSTAL

to 0.0433 inch, while the widths are in the order of 50 to 80 per cent of their lengths.

Figure 23-16 is a representative loss-frequency characteristic of a two-lattice section crystal band-pass 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 cutoff as occurring at the points where the losses become 10 db greater than the loss at 1000

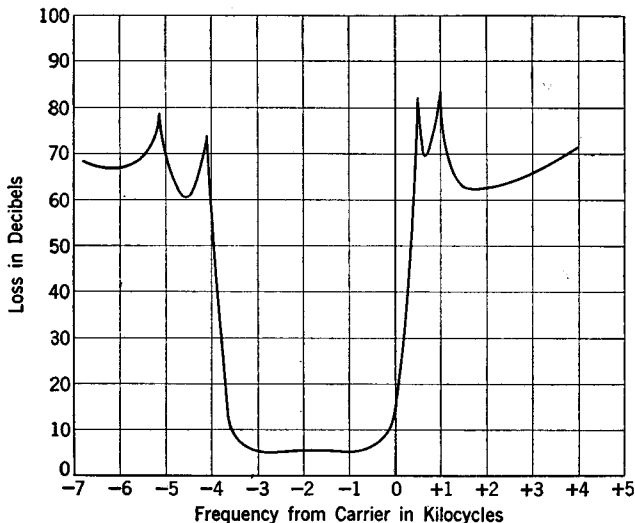


FIG. 23-16. LOSS-FREQUENCY CHARACTERISTIC OF FILTER SHOWN IN FIG. 23-14

cycles, the frequency band passed is about 3600 cycles wide.

### 23.8 Radio-Frequency Filters

At radio frequencies it is customary to speak of devices such as we have been discussing as selecting or tuning circuits, rather than as filters. And generally speaking, such devices at radio frequencies may be considerably simpler in design than many of the complex structures used in low frequency work. Thus the tuning circuits used in ordinary radio receivers are usually simple parallel-resonant **tank circuits** as indicated in Figure 23-17, which may be tuned to a particular frequency by varying the capacitance. In effect, such a circuit is a band-pass filter because, although in theory it is resonant at only one frequency, the resistance which is necessarily included in the circuit broadens the tuning so that appreciable bands of frequencies on each side of the resonant frequency are also selected. This is illustrated in the resonance curves shown in Figure 16-11. Per-

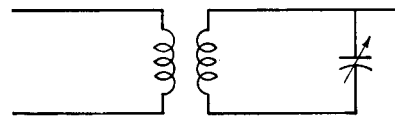


FIG. 23-17 TANK CIRCUIT

centagewise, the tuning may be quite sharp, but at radio frequencies a very small percentage deviation may well be sufficient to include a desired frequency band, such as that required for ordinary voice transmission. Thus, to pass a 10,000 cycle band where the radio carrier frequency is 1,000,000 cycles means a percentage deviation on either side of the carrier frequency of only 1%.

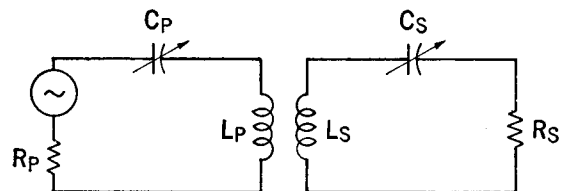


FIG. 23-18 COUPLED CIRCUIT

The selected frequency band may be broadened somewhat if necessary by using a transformer arrangement in which both the primary circuit and the secondary circuit may be separately tuned to resonance. One type of such an arrangement is shown in Figure 23-18. Such a circuit may be designed to serve the purpose of matching impedances as well as acting as an effective band-pass



filter. Here if the primary and the secondary circuits are separately tuned to the same frequency, and the coupling between the primary and secondary windings of the transformer is adjusted so that the current in the secondary is at maximum (critical coupling) the inter-action between the two tuned circuits will be such that the frequency response of the secondary circuit will be broadened appreciably and the current will be constant over a moderately wide band of frequencies. Closer coupling of the transformer windings will result in an even broader frequency response in the secondary, but at the expense of some decrease in the maximum current value.

### 23.9 Waveguide Filters

At super-high frequencies it is impracticable to use tuned circuits, such as we have been discussing, employing coils and capacitors. Analogous principles apply, however, in the design of waveguide filters. Thus the resonant cavity, which we have already considered as being comparable in many respects to the parallel-resonant tank circuit, now becomes the basic element of the waveguide filter. Superficially, at least, if such a cavity is inserted in a waveguide it might be expected to pass a narrow band of frequencies centering about its resonant frequency, and to reject other frequencies.

Figure 23-19(A) is a sketch of such a resonant cavity designed for insertion in a waveguide having the same cross-sectional internal dimensions

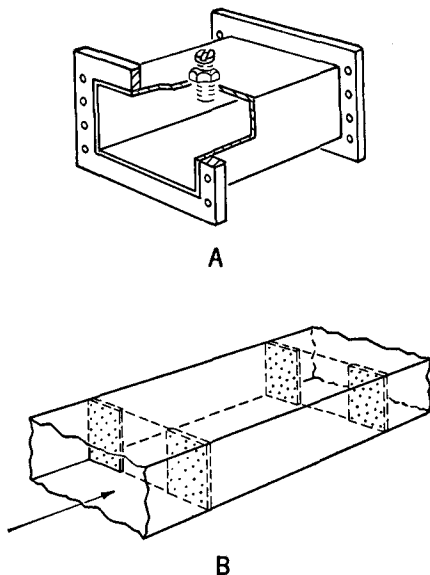


FIG. 23-19 WAVEGUIDE BAND-PASS FILTER SECTION

as the cavity. The screw shown centered in the top wall of the cavity is known as a **trimmer** and is used for making fine adjustments of the resonant frequency. The screw end projecting into the cavity is located in the region of the strongest electric field. Its presence tends to shorten the electric lines of force and thus has an effect similar to increasing the capacitance in an ordinary tank circuit. In other words, inserting the screw farther into the cavity lowers the frequency of resonance of the cavity. This result is effectively the same as if the length of the resonant chamber were increased.

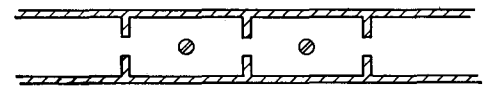


FIG. 23-20 TWO-CHAMBER FILTER IRIS-COUPLED

When the cavity is inserted in the guide its ends are enclosed by thin plates which, however, must contain openings to permit the wave of energy to pass through. The openings in the end plates are usually known as **irises**. In the arrangement shown in Figure 23-19(B), the irises are vertical and the plate assembly has an effect like a shunt inductance because the plates projecting into the region of greatest magnetic field effectively

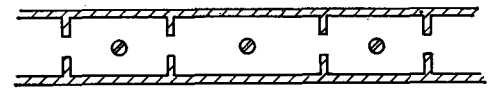


FIG. 23-21 CAVITY-COUPLED FILTER

shorten the lines of magnetic force. Plates, with their irises, of course represent discontinuities in the guide and will therefore tend to cause wave reflections and reflection losses. But, when the cavity section is of proper length (in the order of one-half wavelength) reflections returning into the waveguide line in either direction from the two discontinuities will be so phased as to cancel each other. At the particular frequency at which this is true there will be a substantial standing wave within the cavity but from outside, the line will appear electrically smooth and power will be freely transmitted.

The  $Q$  of cavity resonators, as has been noted earlier, is very high as compared with that of any practicable resonant circuit made up of inductors and capacitors—in the order of 20,000 as compared to a few hundred. The resonance peak of the cavity is therefore relatively very sharp and

for most purposes a single cavity inserted in a waveguide would not permit the passage of a sufficiently wide band of frequencies. This difficulty may be overcome by using two or more cavities in series. Such cavities could be tuned to slightly different frequencies but usually the tandem cavities will be tuned to the same frequency and coupled in such a way as to obtain an increased flat band of frequency transmission on the basis of the principles discussed in the preceding Article. Two types of coupled waveguide filters are in general use. The first, known as iris-coupled, where the resonant cavities are adjacent to each other and coupled by an iris having a considerably smaller opening than those at the outer ends of the cavity, is illustrated by Figure 23-20. The second type, known as a cavity-coupled filter, is illustrated schematically in Figure 23-21. Here the two cavities, tuned for the desired frequency, are coupled by a third cavity which is tuned to a frequency well outside of the range of frequencies with which the particular situation is concerned. Waveguide filters may be built with practically any number of waveguide cavities in tandem, although each cavity causes some loss and it is therefore desirable not to use more than is necessary. In practice as few as three such cavities can be readily arranged to provide a flat pass-band of up to as much as 50 mc in width, with very sharp cutoff on each side.

Band-elimination filters as well as band-pass filters are frequently used in waveguide practice. These are also sometimes known as band-reflection filters. A simple filter structure of this type is shown in Figure 23-22 together with its analogous circuit equivalent. The transverse rod extending partially across the guide has an inductive

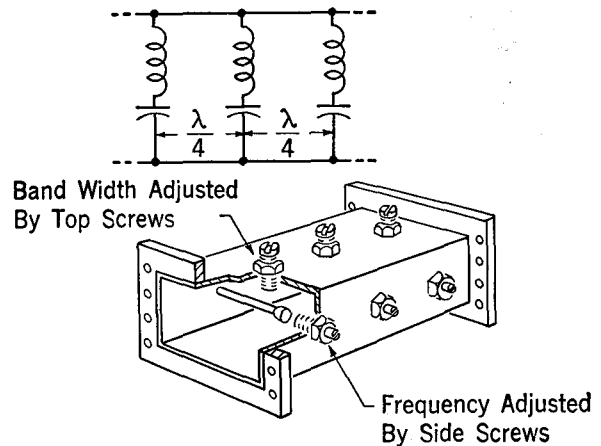


FIG. 23-22 BAND REFLECTION FILTER

effect, as it tends to shorten the lines of the magnetic field. The screw extending into the guide from the top has a capacitive effect, as we have already noted. Together they act like a series resonant shunt across the guide, which can be finely adjusted to present an effective "short-circuit" at a particular frequency. When three of these resonant shunts, spaced approximately a quarter-wavelength apart, are used as shown in the Figure, it is readily possible to adjust them so that the structure will reject quite sharply a band of frequencies as wide as 20 mc, while passing other frequencies freely. It should be noted that the energy of the rejected frequency band is not absorbed in the filter but totally reflected. Furthermore, the design is such that the reflections from the several shunts add in phase. As we shall see in a later Chapter, this behavior is put to very practical use in the channel separating filter arrangement of microwave radio relay systems.

## CHAPTER 24

### ELECTRON TUBES

#### 24.1 Vacuum Tube Elements

Of the several major types of electron tubes used in telephone and telegraph work, the best known and most widely employed is the high-vacuum thermionic tube, generally known simply as a vacuum tube, and familiar to users of radio receivers and comparable devices. This tube consists essentially of a metal or glass envelope from which practically all air or other gases have been evacuated, and into which are sealed two or more electrodes whose terminals may be connected to external circuits. One of these electrodes, known as the emitter or cathode, is the source of the free electrons that are essential to the tube's operation.

Electrons are supplied from the cathode by virtue of thermionic emission. This depends on the fact that certain metals or metallic oxides, when heated to a point of incandescence, tend to release electrons from their surface atoms into the surrounding space. Effectively, electrons "boil out" of the metallic surface in very considerable numbers to form a cloud of free electrons surrounding the electrode. This cloud is of course negatively charged and is known as the "space charge". If other electrodes in the tube are at the same potential as the cathode, the space charge cloud will quickly reach a stage of equilibrium depending on the temperature of the cathode, with electrons in equal numbers being emitted and returned to its surface. However, when a potential positive with respect to the cathode potential, is applied to another electrode, the electrons will be attracted to it and a continuous stream of electrons will flow from the cathode to the positively charged electrode, which is usually known as the anode or plate. This stream of electrons constitutes a flow of electric current (which, however, in accordance with the convention adopted long before electrons were discovered, is considered as flowing from the anode to the cathode.)

Cathodes of thermionic vacuum tubes may be heated directly by a current of electricity supplied from a battery or other appropriate source. In this case, the cathode will be in the form of a filament as in an ordinary electric lamp bulb. Cath-

odes are frequently heated indirectly, in which case the cathode itself usually takes the form of a small oxide-coated cylinder, within which is enclosed an electrically insulated filament. This filament is also heated by an electric current but its electric circuit has no connection with the operating circuit of the tube. In most drawings of vacuum tube circuits, the heating element is for this reason not shown.

#### 24.2 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 or **diode** illustrated in Figure 24-1. Here the cathode is indirectly heated from an external source and the plate battery *B* holds the plate at a positive potential with respect to the cathode. Under these conditions, as we have already noted for the general case, direct current will flow through the plate-cathode-*B* battery circuit and will register on the milliammeter. The value of current flowing for a given value of cathode temperature, and with other operating conditions remaining unchanged, will depend upon the voltage of the *B* battery. Figure 24-2 is a curve showing the relation between this space current and the 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 further with increasing plate voltage is due to the fact that the

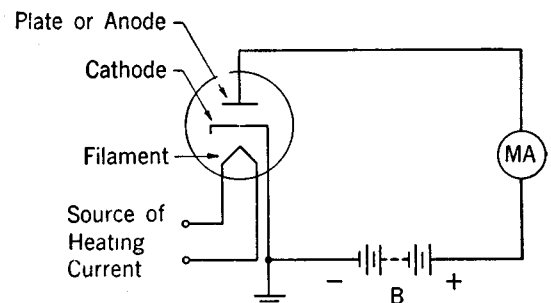


FIG. 24-1 TWO-ELECTRODE VACUUM TUBE

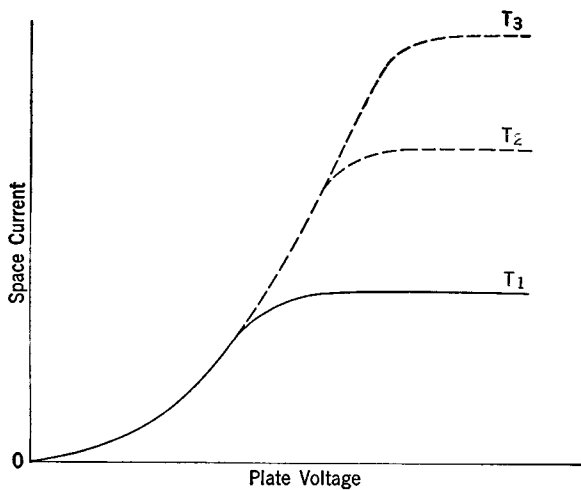


FIG. 24-2 PLATE VOLTAGE VS. PLATE CURRENT CHARACTERISTICS

cathode is emitting the maximum number of electrons possible for that particular temperature. If the cathode temperature is increased, the saturation point will increase correspondingly, as shown by  $T_2$  and  $T_3$  in the Figure. In practice, a vacuum tube is ordinarily operated at such a cathode temperature and plate voltage range that this saturation point will not be reached.

Under these normal operating conditions, the electron flow in the diode is dependent only on the plate voltage. Increases and decreases of plate voltage will cause corresponding increases and decreases of plate current. If the plate voltage is reduced to zero, or is made negative with respect to the cathode, there will be no current flow. The tube permits current flow in only one direction and, therefore, may be used as a rectifier or detector. Thus, if an alternating emf is substituted for the  $B$  battery in Figure 24-1, a space current varying in value for the positive half-cycle of voltage but completely cut off for the negative half of the cycle will result.

### 24.3 Three-Electrode Tubes—Triodes

The type of vacuum tube most widely employed in telephone work, 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 24-3. In this device, the electrons that 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 a typical 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 of the plate to the change in control grid voltage producing an equivalent effect, is called the **amplification factor** of the tube, and is usually designated by the symbol,  $\mu$ . Its value depends entirely upon the mechanical design of the tube.

The utility of the triode 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 or **transfer characteristic** of the particular type of tube. Figure 24-4 illustrates such a curve for a single value of plate voltage. Here the 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 the vertical ordinate shows the corresponding plate current value.

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 reduced, a point will be reached at which the effect of the control grid no longer

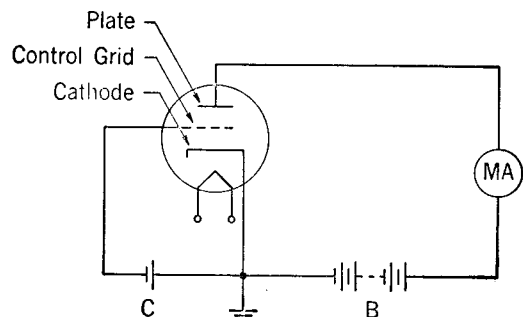


FIG. 24-3 THREE-ELECTRODE VACUUM TUBE

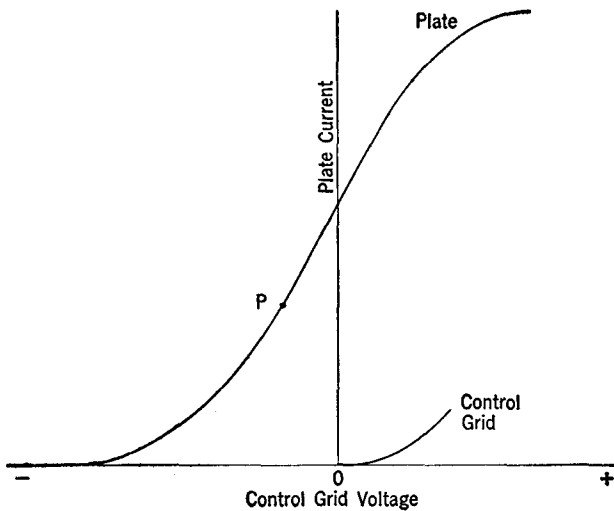


FIG. 24-4 CONTROL GRID VS. PLATE CURRENT CHARACTERISTIC

overpowers that of the plate and a small current flows. No current flows in the circuit of 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 24-4. 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 horizontal 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 24-2.

At low frequencies, the circuit between the control grid and cathode is substantially open when the control grid is negative since no current flows through the space by transfer of electrons. As in the case of any two metallic electrodes separated by an insulating medium, however, there is some capacitance in the cathode-control grid circuit, which will permit some charging current to flow. The capacitance is so small that the charging current will be negligible at low frequencies, but at radio frequencies this capacitance effect may become 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 24-2 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 24-5 for a single value of grid voltage. The potential at which current begins to flow is  $\mu E_g$ , where  $E_g$  is the voltage of the control grid. Above this value the current varies as shown by the curve of Figure 24-5, which is similar to the curve for a two-electrode tube. Such a curve is known as the **plate characteristic** of the tube.

In the design of triodes, one of the major characteristics of the tube to be considered is the amplification factor  $\mu$ . As noted earlier, this is the ratio of a small change in plate voltage to a corresponding small change in grid voltage, with the plate current held constant. This may be represented by the equation—

$$\mu = - \frac{E_{p2} - E_{p1}}{E_{g2} - E_{g1}} \quad (24:1)$$

In more mathematical language,  $\mu$  is the slope at a given plate current value of a curve in which grid voltage is plotted against plate voltage. Its value may therefore be written as—

$$\mu = - \frac{\Delta E_p}{\Delta E_g} \quad (24:2)$$

where  $\Delta$  indicates a very small change in the quantity which it precedes. The minus sign is necessary because the simultaneous changes in plate voltage and grid voltage are in opposite directions.

Another fundamentally important characteristic of a vacuum tube is its output or plate resistance, which is usually designated  $R_p$ . This is somewhat different from a resistance in the ordi-

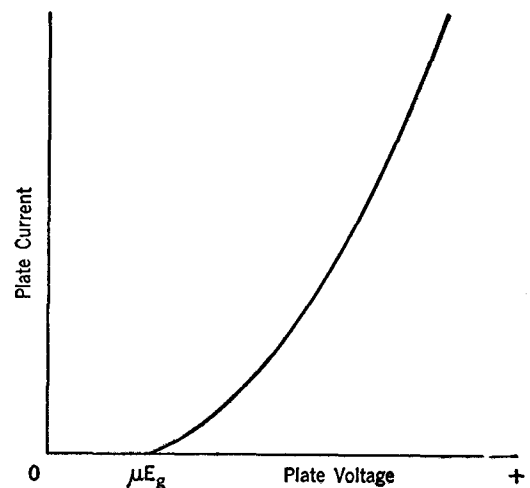


FIG. 24-5 PLATE VOLTAGE VS. PLATE CURRENT CHARACTERISTIC OF THREE-ELECTRODE TUBE

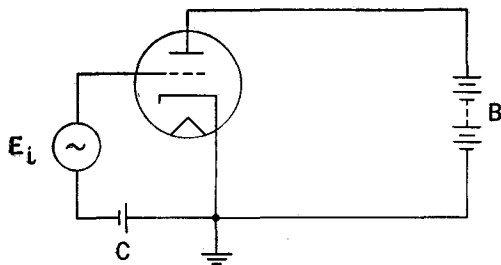


FIG. 24-6 SIMPLE TRIODE CIRCUIT

nary sense because the ratio between plate voltage and plate current is not a simple application of Ohm's Law. Plate resistance of the tube is more in the nature of an "a-c resistance". It is the ratio of a small change in plate voltage to a corresponding small change in plate current, with the grid voltage held constant. In other words, it is the slope of a plate voltage-plate current characteristic curve such as is shown in Figure 24-5, or—

$$R_p = \frac{\Delta E_p}{\Delta I_p} \quad (24:3)$$

A third basic characteristic of a vacuum tube is its **transconductance**, usually designated  $G_m$ . This is defined as the ratio of a small change in plate current to a corresponding small change in grid voltage, with the plate voltage held constant. It is the slope of the grid voltage-plate current characteristic curve of Figure 24-4. That is—

$$G_m = \frac{\Delta I_p}{\Delta E_g} \quad (24:4)$$

where  $G_m$  will be in mhos when  $I_p$  and  $E_g$  are in amperes and volts respectively.

Inspection of equations (24:2), (24:3) and (24:4) will show that the three characteristics are inter-related as follows—

$$\mu = G_m R_p; R_p = \frac{\mu}{G_m}; \text{ and } G_m = \frac{\mu}{R_p} \quad (24:5)$$

#### 24.4 Dynamic Characteristics of Tubes

Curves of the types shown in Figures 24-4 and 24-5 are known as "static" characteristic curves because they are plotted on the basis of fixed d-c potentials at all points. In nearly all practical applications of vacuum tubes, a-c or rapidly varying potentials are superimposed on the fixed d-c potential. Thus in Figure 24-6, the alternating voltage  $E_i$  is shown impressed on the grid in series with the fixed negative potential of the battery C. The outstanding characteristic of this simple circuit is its ability to amplify small voltage changes im-

pressed by  $E_i$  to comparatively large current changes in the plate circuit. The grid voltage-plate current characteristic of the tube, which is of the form shown by Figure 24-4, is redrawn in Figure 24-7. Here the alternating input voltage  $E_i$  is shown as a sine wave impressed on the characteristic curve at the point P, where the distance from P to the vertical zero axis indicates the fixed negative bias maintained by the grid battery. The resulting plate current is shown as a sine wave about the horizontal axis drawn to the right from P. The form of the plate current wave is a faithful reproduction of the a-c wave impressed on the grid as long as the portion of the tube characteristic involved is a straight line. Furthermore, as long as the varying grid voltage remains negative with respect to the cathode, **there is no current flowing to the grid and no power is expended in the input of the circuit.**

The circuit of Figure 24-6 is interesting but it obviously has little practical value unless some means is employed to obtain useful output voltage or power. One method of doing this is to connect a load resistance,  $R_L$ , in the plate circuit as shown in Figure 24-8. In this case, current flowing in the plate circuit will cause a proportional voltage drop across the load resistance  $R_L$ . The output voltage  $E_o$  represented by this voltage drop can then be applied to other circuits. The presence of the load resistance, however, will change the shape of the grid voltage-plate current characteristic of the tube. Instead of the static characteristic curve of Figure 24-4, the new curve will be a "dynamic" characteristic. This is illustrated in Figure 24-9, where A reproduces the static char-

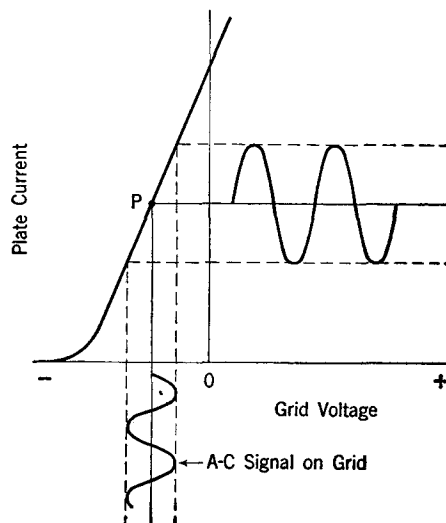


FIG. 24-7 VACUUM TUBE AS AN AMPLIFIER

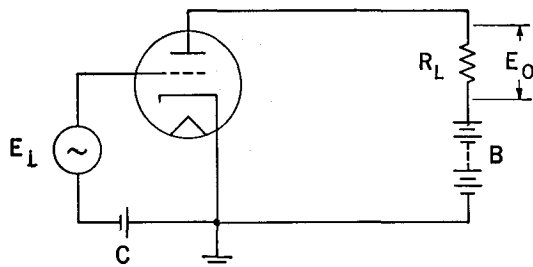


FIG. 24-8 VACUUM TUBE WITH LOAD RESISTANCE

acteristic curve of Figure 24-4, and  $B$  is the dynamic characteristic resulting from the addition of the load resistance  $R_L$ . The difference is due to the fact that the voltage drop across  $R_L$ , being in series with the  $B$  battery, causes the positive potential at the plate to vary with the amount of current flowing in the plate circuit in accordance with the following—

$$E_p = E_B - I_p R_L \quad (24:6)$$

where  $E_B$  is the d-c voltage applied in the plate circuit.

In Figure 24-9,  $P$  is the point on curve  $A$  corresponding to the fixed value of grid bias produced by the  $C$  battery. In order that the dynamic characteristic may pass through this point, it is assumed that the value of the  $B$  battery is increased until the net voltage at the plate is the same as in the static condition with no load resistance present.  $P$  is thus established as a common reference point for the two curves. When the grid voltage increases, the resultant increased plate current causes an increased voltage drop across  $R_L$ . This reduces the plate potential and thus prevents the plate current from increasing to the same extent that it would increase without the load for a like increase in grid voltage. This means that the dynamic characteristic falls below the static characteristic at all points to the right of  $P$ . On the other hand, when the grid voltage decreases below the value at  $P$ , the decrease in plate potential due to the voltage drop in the load resistance is less than the amount by which the  $B$  battery voltage was increased. The dynamic curve, therefore, lies above the static curve at all points to the left of  $P$ . The net result, as indicated in Figure 24-9, is that the dynamic curve is considerably flattened and is more nearly a straight line than the static curve. The difference in slope also shows that the amplification of the circuit with an output load may be expected to be less than that indicated by the amplification factor of the tube.

The voltage amplification of the circuit of Figure 24-8—i.e., the ratio of the output voltage  $E_o$  to the input voltage  $E_i$ , is readily calculated from Figure 24-10. Here the tube and its input has been replaced by an equivalent circuit. This, in accordance with Thevenin's Theorem, consists of the tube output voltage, which is equal to  $\mu E_i$ , in series with its output resistance  $R_p$ , as shown at the left of points 1 and 2. From this simple circuit it is evident that—

$$E_o = \frac{\mu E_i R_L}{R_L + R_p}$$

and from this, the voltage amplification of the circuit is—

$$\frac{E_o}{E_i} = \frac{\mu R_L}{R_L + R_p} \quad (24:7)$$

Inspection of this last equation shows that the voltage amplification or "gain" of the circuit is always less than the amplification factor of the tube, but that it approaches the amplification factor as the load resistance increases.

The voltage amplification can also be determined graphically from a static plate characteristic curve by the method illustrated in Figure 24-11. Here is plotted a family of plate characteristic curves for different values of grid bias voltage. Across these curves a straight line known as a **load line** is drawn. This load line applies to only one specific load resistance value. It merely represents the effect of the voltage drop in the load on the plate potential, and is exactly defined by equation 24:6 ( $E_p = E_b - I_p R_L$ ). The intersection of the load line with the horizontal axis where

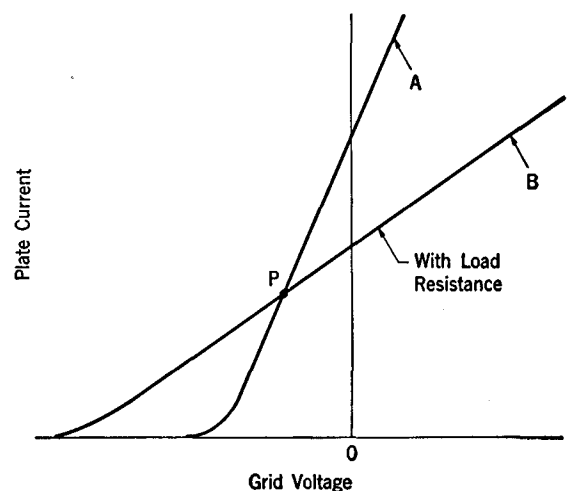


FIG. 24-9 EFFECT OF LOAD RESISTANCE ON PLATE CURRENT-GRID VOLTAGE CHARACTERISTIC CURVE

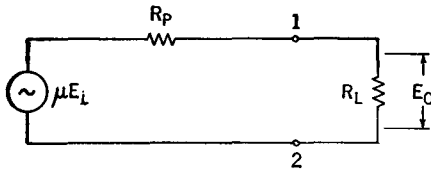


FIG. 24-10 EQUIVALENT CIRCUIT OF FIG. 24-8

$I_p = 0$  is obviously at the point where  $E_p = E_b$ . The intersection with the vertical axis, when  $E_p = 0$ , would be at the point where  $I_p = E_b/R_L$ . The intersection of the load line with any one of the plate voltage-plate current characteristic curves represents the zero input condition of the tube for the particular value of grid bias applying to that curve. When an alternating voltage is applied to the grid, the instantaneous relation between plate current and plate voltage will then be represented by a point swinging back and forth along the load line through the points of intersection.

The application will perhaps be better understood by means of a concrete example. The characteristic curves of Figure 24-11 are drawn for a representative triode connected in a simple amplifying circuit as in Figure 24-8. The plate supply voltage  $E_b$  is assumed at 140 volts, grid biasing voltage  $-4$  volts, and the load line is drawn for a load resistance of 6000 ohms. The static condition of the tube is represented by the intersection of the load line with the  $-4$  volt curve at point  $P$  as indicated. The values of  $E_p$  and  $I_p$  are then 95 volts and 7.5 milliamperes respectively. The corresponding voltage drop across the load resistance is then 45 volts (which is equal to  $140 - 95$  and to  $.0075 \times 6000$ ). Now, if an alternating voltage having a maximum swing of plus

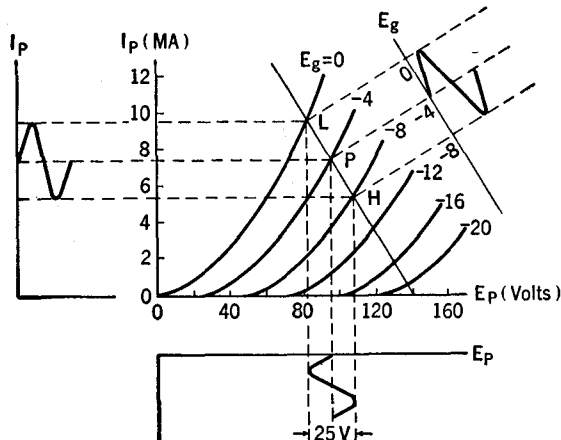


FIG. 24-11 GRAPHICAL ANALYSIS OF TUBE OPERATING CHARACTERISTICS

or minus 4 volts is applied to the grid, the operating point of the tube will swing along the load line between points  $(L)$  where it intersects the  $E_g = 0$  curve, and  $(H)$  where it intersects the  $E_g = -8$  curve. These points establish the maximum and minimum values of the plate voltage as 107 and 82 volts respectively; and the maximum and minimum values of the plate current as 9.6 and 5.4 milliamperes respectively. Thus the diagram shows that an input peak-to-peak voltage variation of 8 volts on the grid results in a variation of 25 volts at the plate. The voltage amplification under these conditions of operation is, therefore,  $25/8$  or 3.1.

### 24.5 Multi-Electrode Tubes

There are many designs of vacuum tubes containing more electrodes than the tubes we have been considering. Most widely used of these are four electrode tubes or **tetrodes** and five electrode tubes or **pentodes**. The basic theory of operation of such tubes is essentially the same as that of the triode. The additional electrodes act to improve the operating characteristics with respect to the amount of amplification to be obtained and may have other desirable effects.

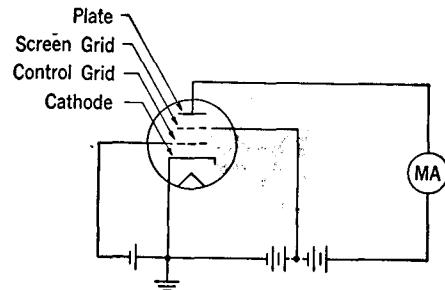


FIG. 24-12 FOUR-ELECTRODE VACUUM TUBE

At relatively low frequencies, the amplification factor  $\mu$  of a triode can be made to have almost any desired value by properly spacing and proportioning the three electrodes. When tubes are used with higher frequencies such as are encountered in radio and other high-frequency systems, the effect of inter-electrode capacitance becomes increasingly important. This is particularly true of the capacitance between the plate and control grid, where its coupling effect may be especially troublesome because it provides a path between the input and output of the tube through which output energy may feed back into the input circuit. This plate-control grid capacitance effect can



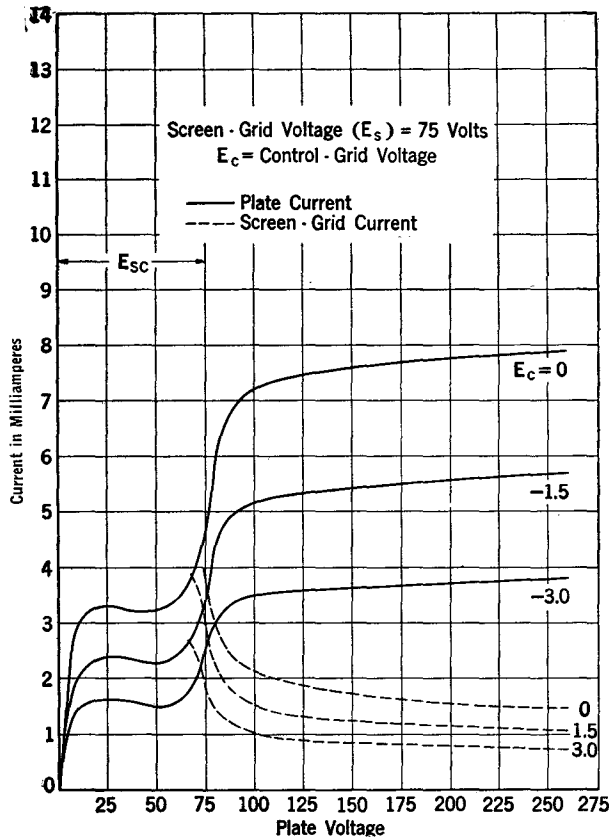


FIG. 24-13 CHARACTERISTICS OF FOUR-ELECTRODE TUBE

be practically eliminated by placing a shielding grid between the control grid and the plate as illustrated in Figure 24-12. This grid is known as a **screen grid**, and the four-electrode tube is then known as a **screen grid tetrode**.

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 by the plate voltage vs. plate and screen grid current curves in Figure 24-13 for a representative screen-grid tetrode.

In the normal working range of the tube, where the characteristic curve is relatively flat, it will be noted that the plate current change is quite small for a considerable change in plate voltage.

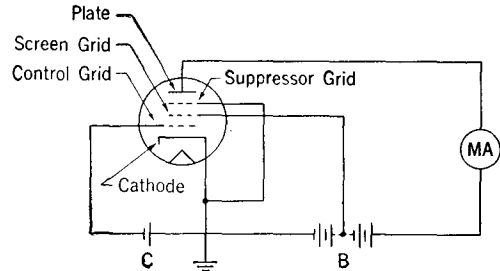


FIG. 24-14 FIVE-ELECTRODE VACUUM TUBE

This means that the output resistance of the tube is very high as compared to the triode. Due to the presence of the screen grid, the variation of plate voltage has relatively little effect on the plate current, but the control grid retains the same control of plate current as in the triode.

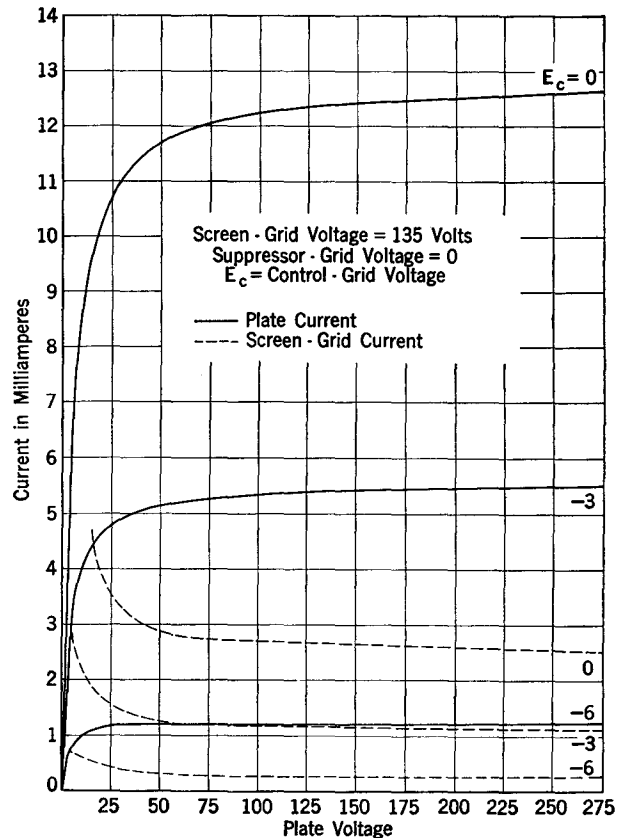


FIG. 24-15 CHARACTERISTICS OF FIVE-ELECTRODE TUBE

The amplification factor is accordingly much higher.

It will be noted, however, that at plate voltages close to or less than the fixed screen grid voltage, the characteristic curves show a pronounced drop in the plate current. This is due in part to the fact that the screen under these conditions is drawing an excessive part of the cathode current because of its relatively high positive potential. More important is the fact that it is now attracting electrons emitted by the plate as a result of **secondary emission**. This emission is caused by the high-speed electrons striking the plate with such force as to knock some of the outer electrons out of the plate material. Under normal operating conditions these secondary electrons will fall back into the plate due to the influence of its positive potential; but if the screen potential is as high or higher than the plate potential, some of them will be attracted to the screen, thus effectively reducing the total flow to the plate and increasing the screen-cathode current. Obviously, operation of the tube in this region would result in marked distortion of the input signal. Generally, therefore, screen grid tubes must be operated with a plate supply voltage sufficiently high so that the maximum negative swing resulting from the control grid input signal will not reduce the instantane-

ous plate potential to a value approaching that of the screen.

While there is always some secondary emission of electrons from the plate of a tube, its effect on the tube's operating characteristics can be practically eliminated by introducing another grid between the plate and the screen grid, as shown in Figure 24-14. This grid, which is maintained at a potential negative with respect to the plate, is called a "suppressor grid" and the tube then becomes a **suppressor-grid pentode**. The suppressor grid is usually connected directly to the cathode, often inside the tube. Its field repels the secondary electrons emitted from the plate, forcing them back to the plate. Figure 24-15 shows characteristic curves for a tube of this type. It will be noted that like the screen grid tetrode, both the amplification factor and output resistance are high

There are many other possible designs of multi-electrode tubes, some practical types of which contain as many as eight electrodes. It is customary also to employ multi-unit tubes in which two or more independent electronic circuits are included in a single envelope. The structure of such tubes is indicated by their designations such, for example, as duplex-diode, duplex-diode-triode, twin-pentode, etc.

## CHAPTER 25

### ELECTRON TUBES—(Continued)

#### 25.1 Gas Tubes

Gas tubes are electron tubes whose envelopes have been evacuated, and then partially refilled with a small carefully regulated amount of an inert gas, such as argon or neon, or a metallic vapor, usually of mercury. Gas tubes are commonly built as diodes or triodes, and sometimes as tetrodes. They are generally similar in appearance to high-vacuum tubes but have considerably different electrical characteristics.

In the thermionic gas diode, which is symbolically represented as shown in Figure 25-1(A), where the black dot indicates the presence of gas in the tube envelope, electrons will be emitted by the heated cathode just as in the case of the high-vacuum tube. Also, if a relatively positive potential is applied to the plate, the electrons will flow from the cathode to the plate. As the plate voltage is increased, the plate current will tend to build up along a characteristic curve similar to that of the high-vacuum tube. But, as the number and velocity of the moving electrons increases, they will collide in increasing numbers with the gas atoms in the intervening space. Some of these collisions will cause ionization of the atoms—i.e., separation of an outer electron from the atom, leaving a positively charged ion. Such action increases the number of electrons moving toward the plate, and further adds to the net current by creating a flow of positive ions moving toward the cathode. The net result is that after a certain critical potential difference is reached, the tube “breaks down” or effectively “arcs over” and becomes a relatively good conductor. The critical voltage at which this occurs is known as the **ionization potential or firing point** of the tube. After ionization has occurred, the value of the current flowing is determined by the resistance in the external cir-

cuit, and it may increase up to the maximum permitted by the emission capabilities of the cathode without any further increase in the plate potential. As a matter of fact, current will continue to flow freely even when the plate potential is reduced considerably below the firing point. There is a minimum sustaining value of plate potential, however, below which the gas will de-ionize and conduction will stop.



FIG. 25-1 GAS-FILLED DIODE AND TRIODE TUBES

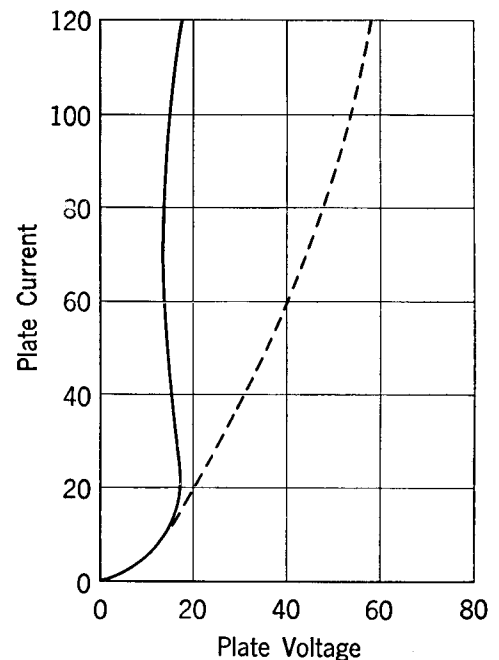


FIG. 25-2 PLATE CURRENT-PLATE VOLTAGE CHARACTERISTIC CURVE OF A GAS-FILLED DIODE

The behavior of the tube is illustrated by the heavy curve of Figure 25-2, which also shows by the dotted curve the comparable behavior of a high-vacuum tube. It may be seen that the tube is effectively an electronic switch, which closes at a certain critical voltage to permit a relatively large current to flow, and opens at some lower voltage to stop the current flow. When the voltage is below the firing point, the tube has an extremely high resistance, but as soon as ionization occurs, its resistance is very low. Thus, it is a switch that can be closed almost instantaneously, and with practically no power, by the mere application of a

suitable voltage to the plate. On opening, however, an appreciable time is required for the gas to de-ionize completely. This time will ordinarily be in the order of a hundred microseconds.

The gas triode is constructed with a third electrode or grid placed between the cathode and anode in much the same way as the control grid of a high-vacuum tube, as shown in Figure 25-1 (B). The operation of the tube is not fundamentally different from the gas diode, except that the firing point may now be controlled by the voltage applied to the grid; and since the grid is closer to the cathode than to the plate, the voltage to be applied to start ionization is smaller. The gas triode cannot act as an amplifier, however, because once ionization is started, the grid has no further control of the current flowing across the tube, unless it is made so negative as to cause de-ionization.

In both gas diodes and triodes employing heated cathodes, the construction is such that the anode is much larger than the cathode. Because of this fact, and the fact that the relatively heavy positive ions naturally travel at a much lower speed than the electrons, the flow of current across the tube is essentially in one direction only. These tubes may, therefore, be used as rectifiers, as discussed briefly in a later Chapter.

Gas tubes in which the cathodes are not heated are known as "cold-cathode" tubes. In this case, there are no electrons "boiling out" of the cathode but there are always a few free electrons floating about in the cathode-anode space. If a relatively positive voltage is applied to the anode of such a tube, these free electrons will be attracted towards the plate. As their velocity is increased, they will ionize atoms with which they collide. If the applied voltage becomes sufficiently high, the tube will break down and conduct a current in the same way as does the hot-cathode tube. As would be expected, the break-down voltage of the cold-cathode tube must be higher; and the total current that can flow in the space, following ionization, is lower than in the hot-cathode tube because the cathode is not supplying electrons by ther-

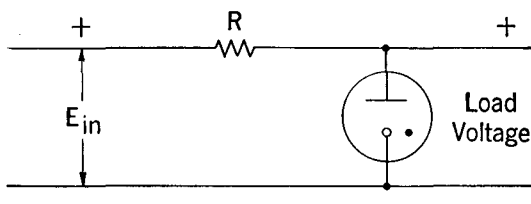


FIG. 25-3 SIMPLE VOLTAGE-REGULATOR CIRCUIT

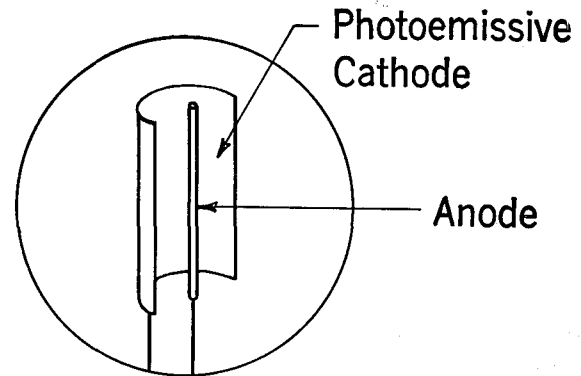


FIG. 25-4 SCHEMATIC OF PHOTOTUBE

mionic action. Within these limitations, however, the cold-cathode tube has the same general characteristics as the heated tube, and has the advantage that it does not require a continuous source of power for its operation. It is, therefore, particularly useful as a switching device where the required operation is highly intermittent. In cold-cathode tubes, the cathode is ordinarily much larger than the anode, with the result that the flow of electricity in the ionized tube may be in both directions. The tube, therefore, has no application as a rectifier.

An important feature of all gas tubes is that after ionization, the voltage drop across the tube (cathode-to-anode) is constant over a considerable range of current variation. This is a characteristic of the arc discharge itself, and has nothing to do with the positioning or shape of the tube elements. The total voltage drop occurs in a thin film surrounding the cathode, and its value depends entirely upon the nature of the gas with which the tube is filled. This characteristic leads to the extensive use of gas tubes as voltage regulators. Thus, if a tube is placed across a circuit as shown in Figure 25-3, and the applied voltage is sufficient to start ionization, the load voltage will take a value depending only on the tube voltage drop characteristic, and will remain constant at that value regardless of variations in the applied voltage.

## 25.2 Phototubes

The phototube is ordinarily a high-vacuum tube having two electrodes. Its action depends upon the property of certain metals, notably sodium, potassium, rubidium, and caesium, to emit electrons when irradiated with visible or ultra-violet light. As indicated in Figure 25-4, the photoemissive cathode is much larger than the anode.

The electrons which are emitted as a result of light falling on the photo-active metal of the cathode, are drawn away to the positively charged anode, thus establishing a small current of electricity which varies in strength with the intensity of the light.

In communications work, phototubes are 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 tubes 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.

### 25.3 The Cathode-Ray Tube

The cathode-ray tube is a highly specialized type of electron tube which has many uses as a testing and measuring instrument, although its current widest application is as the "picture tube" of the ordinary television receiver. Tubes of this type were in use in many electrical laboratories long before the invention of the tubes discussed in the preceding Articles.

Figure 25-5 shows the general structure of a representative "electro-static" cathode-ray tube. The cathode, which is indirectly heated, is a small cylinder, oxide-coated at the end so that it has the ability to emit electrons in large quantities. The control electrode (sometimes called the grid because of its analogy to the corresponding electrode of the ordinary triode) is a metal cylinder with a circular hole in its forward end, which is located close to and partially surrounding the cathode. The control electrode serves the double purpose of controlling the number of electrons emitted by the cathode, and of focusing the emit-

ted electrons so that they will form into a narrow beam emerging through the hole in its end. As in other tubes, the controlling effect of this electrode is affected by the application to it of a voltage negative with respect to the cathode.

In front of the control electrode is an anode, known as the first or focusing anode, which is maintained at a potential positive with respect to the cathode (usually by several hundred volts). This electrode is also in the form of a cylinder, somewhat larger in diameter than the control electrode, having holes in its ends through which the emitted electrons may pass. A second anode, known as the accelerating anode, maintained at a still higher voltage (usually one thousand or more volts) is placed in front of the first anode. This is a still larger cylinder likewise having small holes in the ends. The two anodes act together to attract electrons from the cathode, and to cause their forward movement at high velocity (up to as much as 25,000 miles per second). Some of the emitted electrons will of course strike one of the anodes and be neutralized, but because of their high velocity, and because the arrangement of the two anodes acts as a sort of "lens" to focus the electrons into a narrow beam, most of the emitted electrons will proceed ahead through the holes in the several electrodes to the distant end of the tube. The four electrodes together are commonly called an **electron gun**.

The electrons reaching the distant end of the tube, which is coated on the inside with a fluorescent material, give up their energy of motion causing the fluorescent screen to emit light from the point of impact. The spent electrons then find their way to the conducting coating of aquadag with which the sides of the tube are lined, and are returned to the accelerating anode. Incidentally, it may be noted that the tube is usually operated with the accelerating anode at ground potential, which of course means that the potential of the cathode is highly negative.

The practical usefulness of a cathode-ray tube depends upon deflecting the beam of electrons from the gun in such a way that a pattern will be traced on the screen. In the electrostatic tube shown in Figure 25-5, these deflections of the beam are caused by voltages applied to the two pairs of deflecting plates. Thus, if one of the vertical deflecting plates is made positive with respect to the other, the negative beam of electrons, as it passes between them, will be deflected in the direction of the more positive plate, and the point of light on the screen accordingly will move ver-

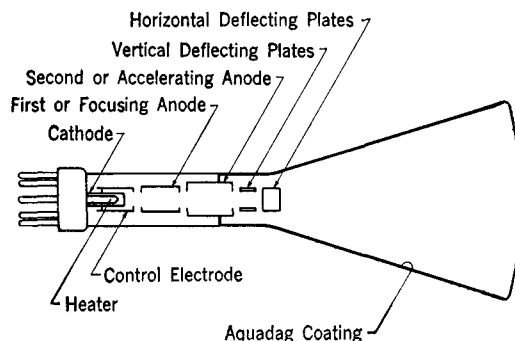
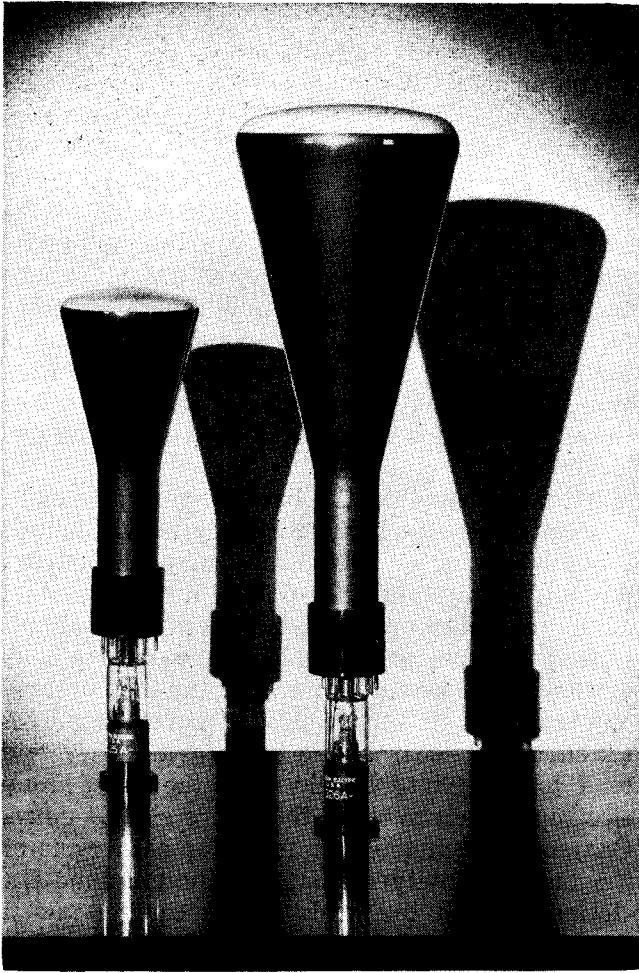


FIG. 25-5 ELECTROSTATIC CATHODE-RAY TUBE



CATHODE-RAY TUBES

tically. Similarly, a difference of potential on the horizontal deflecting plates will cause the light spot to be deflected horizontally. Because of the almost negligible mass of the electrons in the beam, there is practically no inertia, and the beam can be moved about with extreme rapidity. Thus, if a high frequency sine wave voltage is applied to the horizontal deflection plates, the spot of light will sweep back and forth in a horizontal line at a rate equal to the frequency of the applied voltage. Due to the persistency of phosphorescence of the screen material, a horizontal line of light will be seen on the screen. The same voltage applied only to the vertical deflecting plates will similarly produce a vertical line on the screen. It follows from the geometry of the tube that voltages of the same frequency applied simultaneously in phase to the two pairs of plates will produce a straight diagonal line; while the same two voltages applied  $90^\circ$  out of phase will produce a circle.

More significant as a practical matter, is the

fact that the tube may be used to display the pattern of a voltage wave of practically any form or frequency. For this purpose, the voltage being investigated is applied to the vertical deflecting plates. At the same time there is applied to the horizontal deflecting plates a **sweep voltage** having a saw-tooth pattern in which the potential difference increases at a constant rate over a time equal to the period of the voltage applied to the vertical plates, and then returns quickly to its starting point. The pattern of the wave applied to the vertical deflecting plates will then be repeated continuously so that it will appear as a steady picture on the screen.

For the picture tube of a television receiver, both sets of deflecting plates are connected in sweep circuits to move the beam over the surface of the screen in synchronism with the scanning devices of the "television camera". The control electrode varies the intensity of the electron beam (and the light spot) in accordance with the received television picture signal.

Cathode-ray tubes may also employ electromagnetic means for both focusing and deflecting the electron beam. An electromagnetic tube is shown in Figure 25-6. It will be noted that both the focusing and deflecting coils are outside of the tube envelope, which helps to make the tube somewhat more rugged than the electrostatic tube. The focusing coil is in the form of a ring, which fits over the neck of the tube. Its ability to focus the electron stream from the gun into a narrow beam depends on its location with respect to the anode and the amount of current flowing in its windings. The deflection coils are in two pairs, as indicated. The electron beam is of course effectively a current of electricity and like all other electric currents is accompanied by a magnetic field. This field inter-acts with the fields of the deflection coils when they are energized in such a way as to exert pressure on the electron beam,

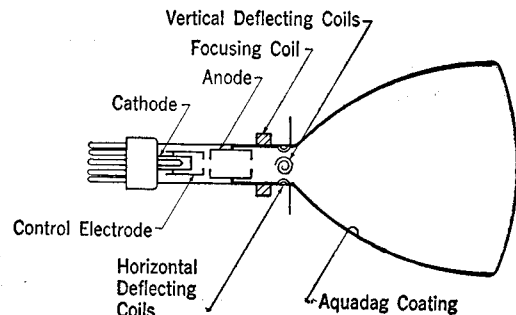


FIG. 25-6 ELECTROMAGNETIC CATHODE-RAY TUBE

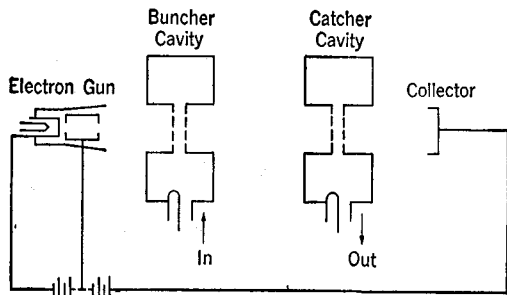


FIG. 25-7 BASIC STRUCTURE OF A KLYSTRON

just as pressure is exerted on a current-carrying wire in the presence of an external magnetic field.

In addition to greater ruggedness, the electromagnetic tube has the advantage that large deflections of the electron beam may be obtained without the employment of the excessively high voltages that would be necessary for like deflections in an electrostatic tube of equal total length. Tubes of this type are, therefore, generally used in television receivers where relatively large pictures are required. The electrostatic type has its chief application in laboratory devices such as the cathode-ray oscilloscope, where it is somewhat more effective than the electromagnetic tube when working with very high frequencies.

#### 25.4 The Klystron

In the operation of electron tubes, it is sometimes necessary to give consideration to the time required for the electrons to travel between cathode and plate. In the usual triode, this "transit time" is normally in the order of  $10^{-8}$  to  $10^{-10}$  seconds, depending on electrode spacing and applied voltage. This is an exceedingly brief interval, but if an alternating voltage as high as 4000 or 5000 mc is applied to the grid of a triode having such a transit time, the polarity of the voltage on the grid will have time to vary from one-half to several cycles while an electron is moving across the inter-electrode space. It is obvious that the behavior of the tube in such circumstances would be seriously affected.

While triodes are built with electrode spacing sufficiently close to overcome the adverse effect of transit time at super-high frequencies (up to approximately 5000 mc), there is an entirely different design of electron tube which makes use of transit time in its operation, instead of allowing it to be a limitation. These tubes are called klystrons,

and are sometimes referred to as **velocity-variation** or **velocity-modulation** tubes.

The principle of operation of the klystron may be followed by referring to Figure 25-7. Included within the evacuated envelope (not shown in the Figure) is an electron gun, two sets of closely spaced grids, and a collector plate. Each of the sets of grids forms part of a resonant cavity, which is tuned to the frequency at which the tube is to operate. The electron gun is generally similar to the gun of the cathode-ray tube described in the preceding Article. It develops a beam of high velocity electrons which travel in a straight line through the grids of the two cavities to the collector plate.

Now, assume that the buncher-cavity is energized at the super-high frequency to which it is tuned by virtue of its size and shape. Electrons passing through its grids during a negative part of the frequency cycle will be slowed down slightly, and electrons passing during the positive half-cycle will be speeded up slightly. As they proceed onward through the so-called "drift space" between the buncher and catcher grids, they will tend to form into bunches—as the speeded up electrons overtake those that were slowed down. One such bunch tends to form for each cycle of the buncher-cavity oscillation. By the time the electrons reach the grids of the catcher-cavity, accordingly, they are traveling in bunches rather than in their normal random spacing.

This electron stream constitutes a current which is effectively varying in amplitude at the frequency of the buncher-cavity. The varying current, in passing through the grids of the catcher-cavity, induces a varying voltage between the two grids just as a varying current in the plate circuit of an ordinary vacuum tube will induce a corresponding varying voltage in a tuned tank circuit to which the plate circuit is coupled. If the catcher-cavity is tuned to the same frequency as the buncher-cavity, the bunched or modulated beam of electrons will transfer energy to the catcher-cavity, which can be delivered to the load as useful output power.

By coupling the output of the catcher to the input of the buncher through an external coaxial connection, the tube becomes an oscillator. If the catcher-cavity has a resonant frequency that is an integral multiple of the resonant frequency of the buncher, the tube will act as a **frequency multiplier**.

## CHAPTER 26

### AMPLIFIER CIRCUITS

#### 26.1 Audio-Frequency Amplifiers

From the discussion of electron tubes in Chapters 24 and 25, it was evident that a major application of such tubes in telephone work—particularly the high-vacuum triodes and multi-grid tubes—would be to amplifying circuits of various types. There are many possible classifications of amplifiers, depending on the purpose for which they are to be used. One broad classification is as voltage amplifiers, or power amplifiers. Another classification depends upon the frequency range in which the amplifier is to operate. This may include audio-frequency amplifiers, broad-band amplifiers, radio-frequency amplifiers, and superhigh radio-frequency amplifiers. Again, amplifiers may be classed according to the band-width of the signal they are required to handle, i.e., whether a relatively narrow band of audio or radio frequencies, or the wide band of frequencies encountered in broad-band carrier systems and in video transmission. Any given amplifier may properly fall into more than one such classification.

Figure 26-1 shows the circuit of a simple voice frequency amplifier. Here the input and output circuits are iron-core transformers, and the grid is biased with a battery. When an alternating signal voltage is applied to the input, the resulting alternating plate current flows through the capacitor  $C_o$  and the primary of the output transformer, but is prevented from flowing through the  $B$  battery circuit by the retardation coil. The capacitor, on the other hand, prevents the direct current of the  $B$  battery from flowing in the output transformer. The overall voltage amplification or gain of this simple circuit depends upon the character-

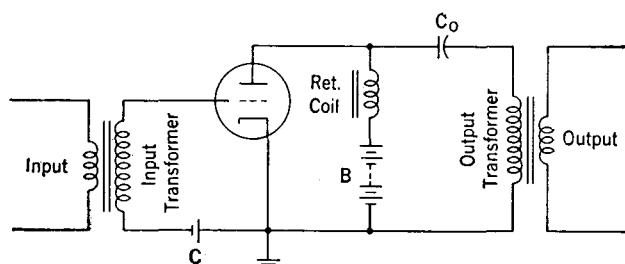


FIG. 26-1 SIMPLE AMPLIFIER CIRCUIT

istics of the tube, and the turn ratio of the input transformer. This gain may be substantially greater than the amplification due to the tube alone when the input transformer has high step-up ratio. The output transformer is used to match the output impedance of the tube to the load, so that maximum power will be transferred. The tube indicated in Figure 26-1 is a triode, although higher gain could be obtained with a tetrode or pentode. However, since the principles involved are the same, it will be convenient in this and the following discussion to consider only triodes.

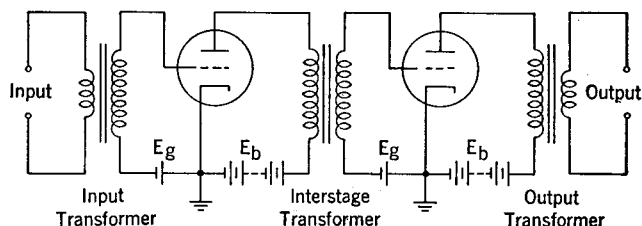


FIG. 26-2 TWO-STAGE TRANSFORMER-COUPLED VOLTAGE AMPLIFIER

There is of course a limit to the total gain that can be obtained from a simple amplifier such as that of Figure 26-1, regardless of the type of tube used or the voltage step-up ratio of the input transformer. Where gains beyond this limit are required, a "cascade" amplifier may be used with the output of the first tube, or stage, working into the input of a second tube, and so on. Figure 26-2 shows a two-stage amplifier in which the output of the first tube is coupled to the input of the second tube through an inter-stage transformer. This transformer, as well as the input transformer, may have a substantial step-up ratio so that it too adds to the overall voltage amplification of the circuit.

The transformer coupled amplifier has limitations with respect to the frequency bandwidth that it can handle on a "flat" basis. This is due to the inductance of the transformer windings, and to their effective shunt capacitance. At frequencies below about 100 cycles per second, the inductive reactance of the primary winding of the



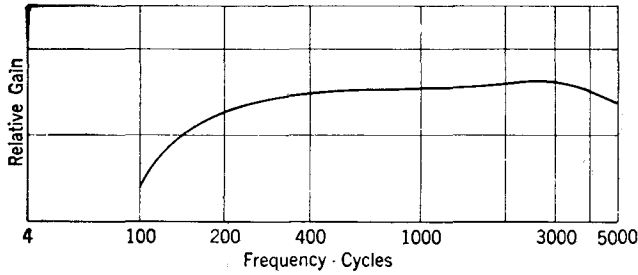


FIG. 26-3 GAIN-FREQUENCY CHARACTERISTIC OF AUDIO AMPLIFIER

inter-stage transformer is low enough in value so that the output resistance of the tube is not negligible in comparison with it. This results in a relative decrease in the voltage across the primary winding, and a consequent reduction in amplification. At frequencies above about 4000 cycles, the shunt capacitance becomes increasingly important. Since capacitive reactance is inversely proportional to frequency, the complex impedance of the transformer input becomes lower at the higher frequencies, with a consequent lower voltage across the primary winding. There is also a tendency to develop a resonance effect between the shunt capacitance and the inductance, which may produce a definite hump in the frequency-gain curve near the higher frequency end. Well constructed amplifiers of this type, however, have a reasonably flat frequency response over a range from a little above 100 to approximately 5000 cycles. This is illustrated by the curve of Figure 26-3.

## 26.2 Resistance-Capacitance Coupled Amplifiers

Where flat frequency response over a greater range than four or five thousand cycles is required, resistance-capacitance coupling is commonly employed. An amplifier circuit of this type is shown schematically in Figure 26-4. Here, the a-c input is through the input capacitor  $C_i$  to a grid resistance  $R_g$ , the drop across which is applied to the grid of the first tube. Grid bias is provided by the drop across the cathode resistor  $R_k$ , through which the d-c component of the plate current flows. The a-c component of the plate current is by-passed by the capacitor  $C_k$  so that it has no effect on the grid. The alternating voltage drop across the resistor  $R_L$  is coupled to the input of the second tube by the capacitor  $C_c$ , which also prevents the plate battery voltage  $E_b$  from being impressed on the grid of the second tube.

An equivalent circuit for one stage of the R-C amplifier is shown in Figure 26-5. Here two

shunting capacitors are indicated, which did not appear in Figure 26-4.  $C_{out}$  represents the inter-electrode capacitance of the first tube—chiefly the plate to cathode capacitance,  $C_{pk}$ —together with such shunt capacitances as may be introduced by the circuit wiring.  $C_{in}$  represents a comparable capacitance  $C_{gk}$  at the input of the second tube. In a mid-frequency range—from about 100 to somewhat more than 10,000 cycles—the effect of the inter-electrode capacitances of the tubes is so small that the shunting capacitors  $C_{out}$  and  $C_{in}$  may be neglected. In this same frequency range, the coupling capacitor  $C_c$  may be considered as an a-c short-circuit because its reactance is negligible. The equivalent circuit then reduces to the parallel combination of resistors  $R_L$  and  $R_g$  across the tube output, as indicated in Figure 26-6. In the low frequency range, however, (below 100 cycles) the coupling capacitor  $C_c$  can no longer be ignored. Its reactance now becomes great enough that the voltage across  $R_L$  is divided between  $C_c$  and  $R_g$  with an increasing amount appearing

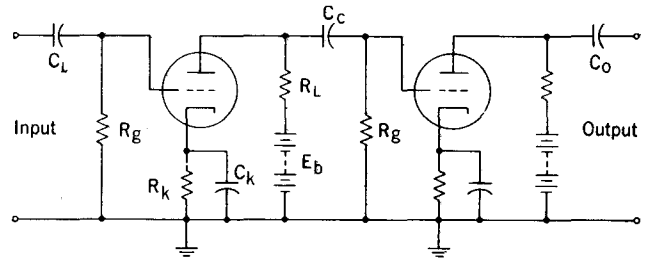


FIG. 26-4 RESISTANCE-CAPACITANCE COUPLED AMPLIFIER

across  $C_o$ , and a decreasing amount appearing across  $R_g$ . The input to the second tube is accordingly decreased, and the gain of the amplifier falls off. In the high frequency range (above 10,000 cycles), on the other hand, the effect of the coupling capacitor  $C_c$  again becomes negligible, but the net reactance of the shunting capacitors,  $C_{out}$  and  $C_{in}$ , then becomes small enough to cause the output voltage to fall off.

It will be clear that the bandwidth of reasonably flat frequency response of the R-C coupled ampli-

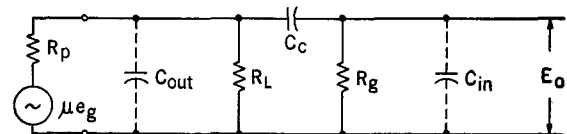


FIG. 26-5 EQUIVALENT CIRCUIT OF SINGLE STAGE OF R-C COUPLED AMPLIFIER

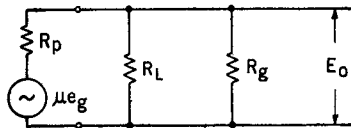


FIGURE 26-6

fier depends mainly upon the values of  $C_c$ ,  $R_L$  and  $R_g$ , and the values of the tube inter-electrode capacitances. Amplifiers with a flat response over a range of 50 to some 15,000 cycles may be readily designed without employing extraordinary methods. Where broader response is required, rather extreme capacitance and resistance values may be required for some of the elements of the interstage coupling networks. Inductors and additional capacitors may also be added to the network in various connections that will help to extend the transmitted band to both lower and higher frequencies. Tubes specially designed to have minimum grid to plate capacitance (or maximum transconductance) may also be necessary. Such tubes will ordinarily be pentodes or tetrodes which have much lower grid-plate capacitance than triodes and much higher transconductance and plate resistance.

The more elaborate interstage networks employed in broad-band amplifiers naturally tend to reduce the gain that can be obtained in each stage and thus may require the use of more stages for a given overall amplification. The range of uniform frequency response can be extended through several million cycles, however, with types of tubes now available and proper design of the coupling networks. In communications work, perhaps the most severe practical requirement occurs in the case of the so-called "video" amplifier, which, ideally, should give a flat response over the total range from zero to about four million cycles.

### 26.3 Radio-Frequency Amplifiers

Voltage amplifiers for most of the applications in radio circuits, such as radio receivers and the low-power stages of radio transmitters, do not have to meet as severe requirements with respect to frequency response as do audio or video amplifiers. This is because the typical radio circuit is designed in theory to handle only the single frequency to which it is tuned. Actually, of course, the tuning is not so sharp that it does not permit the passage of a band of frequencies extending far enough on both sides of the tuned

frequency to carry the complete communication signal. Coupling between the stages of radio amplifiers is commonly accomplished by means of simple air-core transformers. As indicated in Figure 26-7, one or both windings of the coupling transformer are tuned with a paralleling capacitor to the signal frequency. As discussed in Article 17.6, where both primary and secondary are so tuned, a good band of frequency response with sharp cutoff at each end is readily obtained. The transformer itself need have little or no voltage gain because tubes with high amplification factors are used.

### 26.4 Negative-Feedback Amplifiers

For a great majority of amplifier applications in telephone work, it is important not only that the output signal be a faithful reproduction of the input signal, but also that maximum stability of amplifier operation be secured. Both of these objectives can be met to a very large degree by feeding back some of the amplifier output to the input circuit in an inverse phase relationship. An amplifier so connected is called a **negative feedback** or **degenerative amplifier**. Its principle may be understood by referring to Figure 26-8. In this Figure, (A) indicates an amplifier without feedback, having an overall voltage amplification or gain of  $A$ . In Figure 26-8(B), a part of the output voltage is returned to the input,  $180^\circ$  out of phase with the input voltage, through a feedback circuit having a loss  $\beta$ . Without feedback (Figure 26-8(A)) we have—

$$E_o = AE_i = AE_g \quad (26:1)$$

In Figure 26-8(B), on the other hand, the actual input voltage  $E_g$  of the amplifier unit is no longer equal to  $E_i$ , but to the sum of this voltage and the feedback voltage  $\beta E_o$ . That is—

$$E_g = E_i + \beta E_o \quad (26:2)$$

The output voltage therefore is—

$$E_o = AE_g = A(E_i + \beta E_o)$$

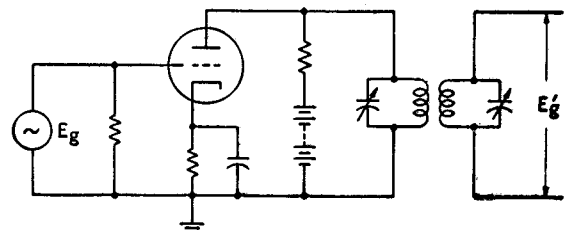


FIG. 26-7 RADIO AMPLIFIER CIRCUIT

Solution of this equation for the overall voltage gain,  $E_o/E_i$ , gives—

$$\frac{E_o}{E_i} = \frac{A}{1 - \beta A} \quad (26:3)$$

Since the feedback is inverse (negative), the value of  $\beta A$  is negative and the denominator of equation (26:3) is greater than unity. Negative feedback accordingly always reduces the net gain, but the reduction can be compensated by the use of an amplifier having as high gain  $A$  as may be required to obtain the desired overall gain. When the product  $\beta A$  is much larger than unity, as is the case in most practical circuits, the overall amplification becomes effectively—

$$\text{Net Amplification} = -\frac{1}{\beta} \quad (26:4)$$

In other words, the effective gain of the circuit depends entirely upon the characteristics of the feedback circuit. This may perhaps be better understood by considering a numerical example. In the circuit of Figure 26-9, the gain  $A$  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). From equations (26:2) and (26:3)—

$$E_o = E_i \left( \frac{1}{1 - \beta A} \right) \quad (26:5)$$

If the applied input voltage  $E_i$  is 1 millivolt, the actual input voltage to the amplifying unit is therefore—

$$\begin{aligned} E_g &= \frac{1}{1 - \frac{10,000}{-1,000}} = \frac{1}{1 - (-10)} = \frac{1}{11} \\ &= .09091 \text{ millivolt} \end{aligned}$$

The output voltage is—

$$A E_g = E_o = 10,000 \times .09091 = 909.1 \text{ millivolts}$$

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 give the actual input voltage to the amplifier, which, therefore, is—

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

This checks the value of  $E_g$  obtained above, which means that the amplifier is stable and as long as

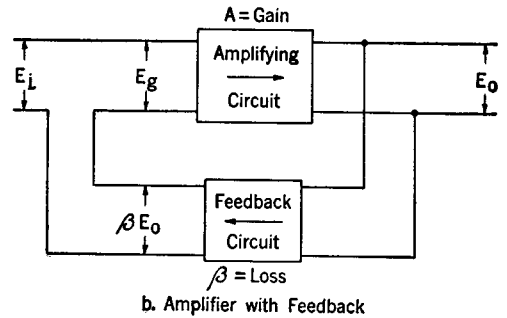
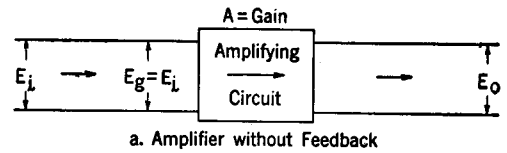


FIG. 26-8 PRINCIPLE OF NEGATIVE FEEDBACK

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—

$$\begin{aligned} 20 \log_{10} \frac{\text{Output voltage}}{\text{Input voltage}} &= 20 \log_{10} \frac{909.1}{1} \\ &= 20 \times 2.9586 = 59.17 \text{ db} \end{aligned}$$

It will be noted that for all practical purposes this gain is the same as the loss  $\beta$  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 100,000)—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 (26:5) again we find the actual input voltage now is—

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

The output voltage,  $E_o$ , accordingly is—

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

and overall gain of the amplifier is—

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

which is again practically equal to the loss in the feedback circuit. This means that **even if the gain  $A$  of the amplifier unit changes due to variations in the battery supply, changing tube characteristics, etc., the overall gain remains the same for all practical purposes.**

Another important feature of the negative feed-

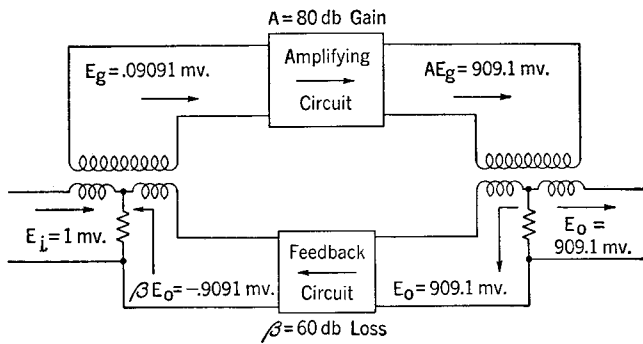


FIG. 26-9 EXAMPLE OF NEGATIVE FEEDBACK CIRCUIT

back amplifier is its ability automatically to reduce to a negligible magnitude any noise or harmonic distortion developed within the amplifier itself. This is true because a part of this noise and distortion appearing in the output is fed back to the input through the feedback circuit where it re-enters the amplifier in such a phase relation that when it is amplified and again appears in the output, it is out of phase with the original noise and distortion, thereby reducing its effect.

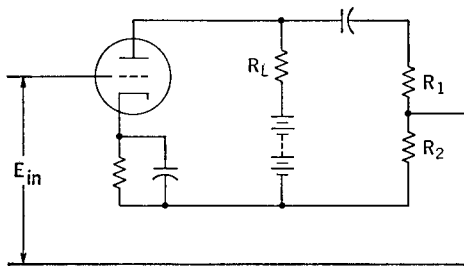


FIG. 26-10 VOLTAGE FEEDBACK CIRCUIT

Feedback circuits may be designed with either voltage feedback or current feedback. Figure 26-10 illustrates a simple voltage feedback arrangement. Here the total resistance ( $R_1 + R_2$ ) of the voltage divider is made large enough so that its shunting effect on the load resistance  $R_L$  is practically negligible. The magnitude of the feedback factor  $\beta$  is—

$$\beta = \frac{R_1}{R_1 + R_2}$$

and the overall gain when  $\beta A$  is large compared to unity is the reciprocal of this—

$$\text{Net gain} = \frac{R_1 + R_2}{R_1} \quad (26:6)$$

The voltage fed back is  $180^\circ$  out of phase with the alternating input voltage because it represents a voltage drop in the plate circuit, and the plate current is in phase with the input voltage.

The simplest type of current feedback arrangement may be obtained by the use of an unby-passed cathode resistor, as indicated in Figure 26-11. Here the alternating plate current  $I_p$  must flow through  $R_k$  as well as through the load  $R_L$ . This causes a voltage drop across the cathode resistor equal to  $I_p R_k$ , and the net input voltage applied to the grid is then—

$$E_g = E_i - I_p R_k$$

This may be rewritten in terms of  $E_o$  as follows—

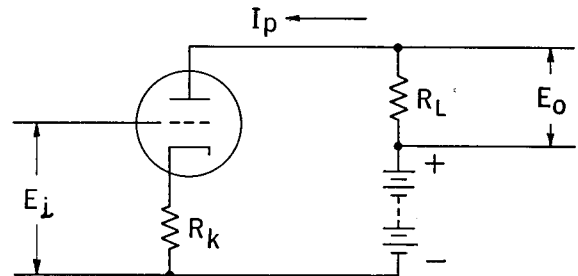


FIG. 26-11 CURRENT FEEDBACK CIRCUIT

$$E_g = E_i + \frac{E_o R_k}{R_L}$$

since  $E_o$  is equal to  $-I_p R_L$ , the drop across the load resistance. Comparison of this equation with (26:2) shows that  $\beta = R_k/R_L$ ; and when  $\beta A$  is large compared to unity—

$$\text{Net gain} = \frac{R_L}{R_k}$$

### 26.5 The Cathode-Follower

An interesting example of maximum application of negative feedback is displayed in the so-called **cathode-follower** circuit, shown in Figure 26-12. Here there is one hundred percent current feedback through the cathode resistor,  $R_k$ —in other words,  $\beta$  is equal to unity. The output is taken across the cathode resistor so that the input and output voltages are necessarily in phase and the

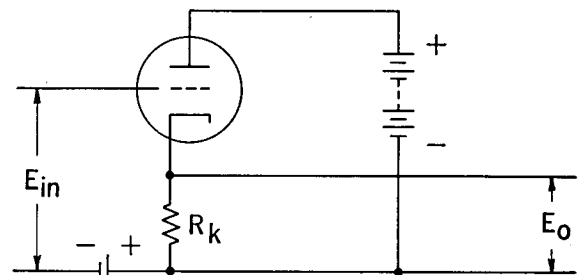


FIG. 26-12 CATHODE FOLLOWER CIRCUIT

a-c cathode to ground voltage varies in the same direction or “follows” the applied grid to ground voltage. The net voltage gain of the circuit is always less than unity because of the hundred percent negative feedback. It may be expressed as—

$$\text{Net gain} = \frac{\mu R_k}{R_p + R_k(1 + \mu)}$$

where  $\mu$  is the amplification factor of the tube and  $R_p$  is its plate resistance.

As an amplifier, the cathode-follower circuit would appear useless since its voltage gain is less than one. However, it is still capable of delivering power to a load without requiring appreciable input power, and with extremely faithful reproduction of the variations in the input voltage. The circuit is very stable and virtually independent of any variation in the tube characteristics. These factors make it useful as a stabilizing coupling circuit between an amplifier and a load. More important is the fact that while the input impedance of the circuit is high, its output impedance is very low for an amplifier. This output impedance consists of  $R_k$  in parallel with an effective plate resistance equal to  $R_p/(1 + \mu)$ . Since the value of  $R_p/(1 + \mu)$  is less than 1000 ohms for most tubes, the value of the net output impedance must be still less. As an impedance-matching device accordingly, the cathode-follower circuit is useful for such purposes as coupling the relatively high impedance output of a video amplifier to the low impedance of a coaxial line.

## 26.6 Power Amplifiers

The classification of amplifier circuits as between voltage amplifiers and power amplifiers is not very definite. Actually, the term “power amplifier” is somewhat misleading because it is perfectly possible for an amplifier to deliver a substantial power output without appreciable power input. The power classification is applied generally in practice to situations where the delivery of a desired amount of power is the controlling criterion, and voltage gain, if any, is of secondary importance. Thus, an amplifier used to drive a load such as a loudspeaker or other device requiring considerable power for its operation, is ordinarily classed as a power amplifier. So is the amplifier that must supply many kilowatts of power to drive the antenna of a radio transmitter.

The power that a vacuum tube amplifier can develop of course depends generally on the maxi-

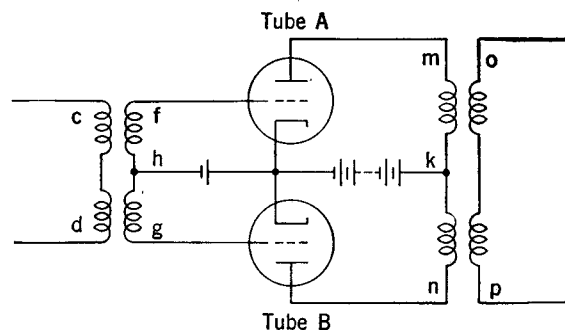


FIG. 26-13 PUSH-PULL AMPLIFIER

imum value of current that may flow in its plate circuit. In most audio amplifier applications, this value is limited by the fact that the tube must operate on the straight line portion of its characteristic curve in order that its output be a faithful reproduction of the input signal. The maximum power that can be delivered, therefore, depends upon the size and characteristics of the tube or tubes used. Where the power requirement is a matter of a very few watts, as for driving the speaker of an ordinary radio receiver, a single triode or pentode may be used. Such a tube is not different in appearance from the tubes used in voltage amplifying circuits, although its design characteristics will generally be such that it will have a lesser amplification factor and a larger plate current.

Where more power than a single tube can deliver is required, together with maximum fidelity of signal reproduction, two tubes may be employed in a “push-pull” circuit, as shown in Figure 26-13. In this circuit, the two tubes, A and B, have identical characteristics. 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 midpoint,  $k$ , 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,  $op$ . The total

output is thus obviously equal to the sum of the outputs of the two tubes.

As a matter of fact, the push-pull circuit will provide a power output appreciably greater than twice the output of an amplifier employing only one tube. This is due to the fact that the tubes of the push-pull amplifier may be given more control grid bias than a single tube without causing distortion in the output. The characteristic curve of a triode is of such shape that its output, when working over a portion of the curve including some curvature, consists principally of the fundamental or desired frequency, and its second harmonic (double the fundamental frequency). Such outputs are illustrated in Figure 26-14(A) and (B) where it will be noted that the net output of each tube, represented by the heavy lines, is considerably distorted. It may also be noted, however, that the second harmonics in the outputs of both tubes become 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 harmonics cancel each other. The net result is indicated in Figure 26-14 (C), where the two output currents are shown to add to produce a sine wave, which is a faithful reproduction of the input signal. Because the push-pull amplifier can thus be operated over a greater range of its tubes' characteristic curves, its output may actually exceed by more than three times the equally distortionless output that could be obtained from a single tube amplifier.

All of the amplifiers that have been considered thus far have operated on a high fidelity basis—i.e., so that the output signal presents a faithful reproduction of the input signal. In radio parlance, such operation is designated “Class A”, a term which indicates, in general, that the amplifier tubes are operating only on a straight line portion of their characteristic curves. All Class A amplifiers have comparatively low “plate efficiency”—that is, the ratio of their useful output power to the total power supplied to the plate circuit by the *B* battery cannot be greater than 50% in theory, and is usually not much higher than 25% in practice. In audio amplifiers, where the output power is in any event not very great, this is not too important. In high-powered radio transmitters, on the other hand, where output power is measured in kilowatts, better efficiency becomes economically significant, and leads to the use

when possible of “Class B” and “Class C” operation. The grids of Class B amplifiers are biased to the cutoff point so that plate current flows only during one-half of the cycles of an applied alternating voltage. In Class C operation, the grid is biased well beyond the cutoff point so that plate current flows during less than one-half of each cycle of applied grid voltage. The effective results are indicated in Figure 26-15(A) and (B) respectively. It is evident that in both cases, the output wave form presents a highly distorted version of the wave form of the input signal. However, plate current flows and draws power from the *B* supply only part of the time as contrasted with the continuous power drain in Class A operation. The plate efficiency of these types of amplifiers is therefore higher, having a theoretical possible maximum of 78% in the case of the Class B

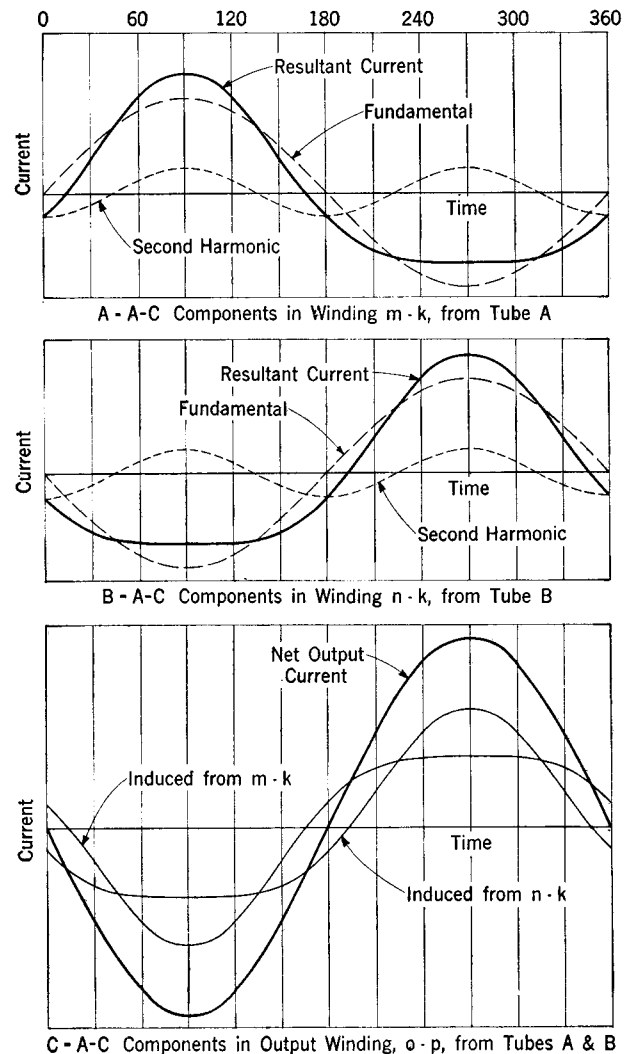


FIG. 26-14 PRINCIPLE OF PUSH-PULL AMPLIFIER

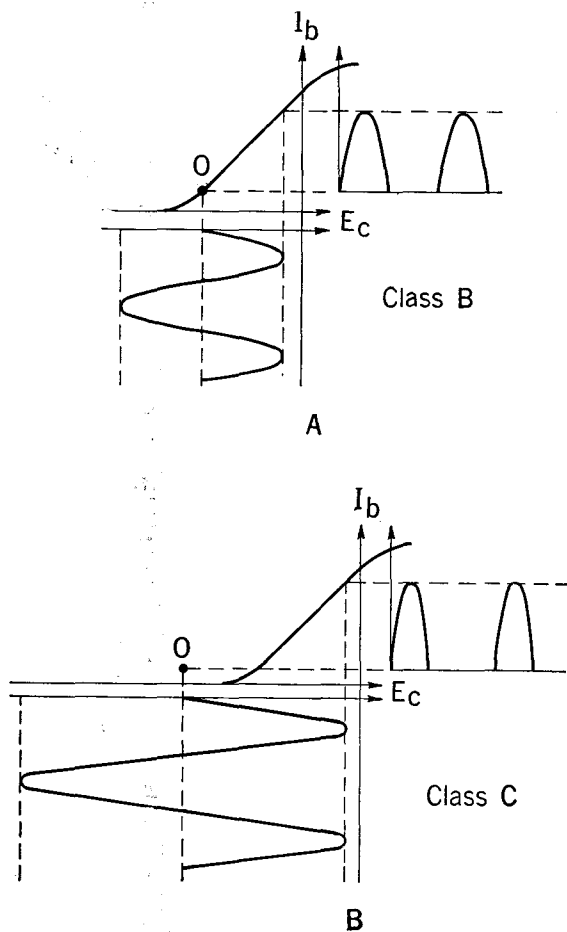


FIG. 26-15 OPERATING CHARACTERISTICS OF CLASS B AND C AMPLIFIERS

operation, and as much as 85% in Class C operation.

Radio power amplifiers ordinarily work into a load impedance which includes a parallel-tuned circuit, as indicated in Figure 26-16. This  $L$ - $C$  tank circuit, when tuned to the operating frequency, acts as a filter to suppress the many harmonics of the fundamental frequency that must obviously be present in the plate circuit of a tube operating Class B or C. The current flowing in the load itself consequently represents only the fundamental input frequency. Its amplitude,

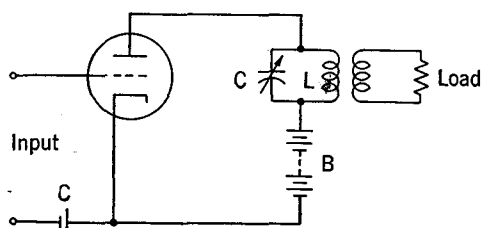


FIG. 26-16 R-F POWER AMPLIFIER

however, can not be expected to be linearly related to the input voltage. In Class B operation, distortion may be held to reasonable proportions by operating two tubes in a push-pull arrangement. In this case, one tube will provide an approximately true reproduction of the positive half of the input voltage wave, while the other tube furnishes a like reproduction of the negative half of the input wave. The net effect is illustrated graphically in Figure 26-17.

### 26.7 The Transistor as an Amplifier

The transistor is a recent development of Bell Telephone Laboratories, which, while working on an entirely different physical principle, is apparently capable of performing most of the functions of an ordinary vacuum tube—including amplification. Figure 26-18 is a cutaway drawing of a type-A transistor. Here two small phosphor-bronze “cat-whiskers” make point contacts very close together to a small block of germanium, the whole being enclosed in an insulated cylinder about one-half inch long and three-sixteenths of an inch in diameter. The cat-whisker electrodes are known respectively as the emitter and collector, while a third connection is made through a conducting base to the germanium block.

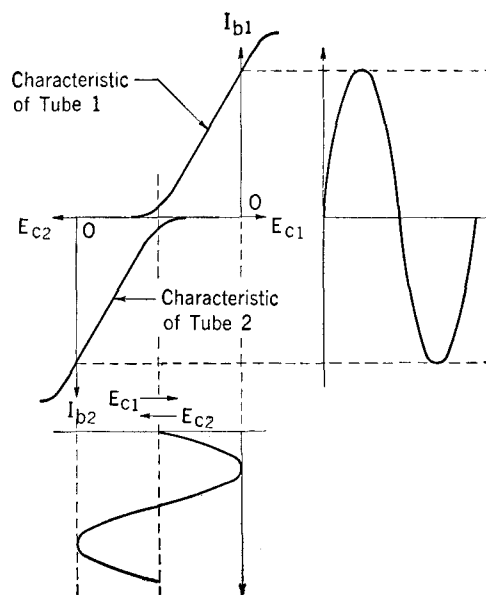


FIG. 26-17 GRAPHICAL REPRESENTATION OF CLASS B PUSH-PULL AMPLIFIER OPERATION

The operation of the transistor depends upon the electrical characteristics of materials known as **semi-conductors**, which include **selenium**, **ger-**

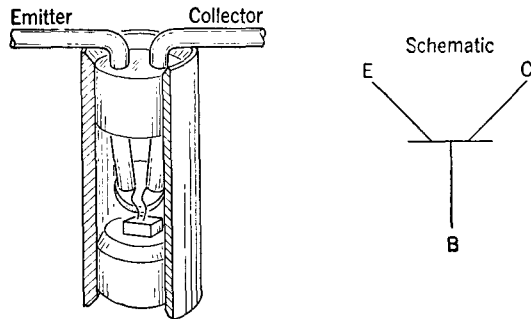


FIG. 26-18 CUTAWAY VIEW OF A-TYPE TRANSISTOR

manium, silicon, copper-oxide and various other oxides. These special properties account for the rectifying action of the copper-oxide and selenium rectifiers that were discussed in Article 7.7. Although a considerable literature has grown up in recent years, the underlying physical theory is still the subject of intensive investigation. It seems well established, however, that the rather extraordinary properties of semi-conductors in electrical circuits are due to their ability to carry current in two distinctly different ways. The usual types of electrical conductors are generally thought of as carrying current by virtue of a drift of free electrons through the atomic structure of the conducting metal. In a semi-conductor, current may be carried in the same way by moving electrons, and also by the movement of "holes" representing the absence of electrons from their normal place in the crystal structure of the semi-conducting material. These holes may be considered as positive charges which can drift like bubbles through the atomic spaces, and thus effectively carry current just as moving electrons do.

Under normal and static conditions, there are practically no free electrons or holes present in a pure semi-conducting material. Such a material, however, can be excited by the application of electric fields so that a limited number of the electrons in its atomic structure are dislodged from their normal fixed positions. These are the so-called

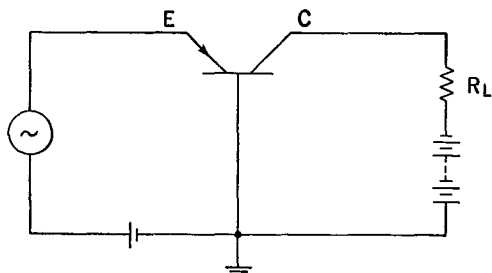
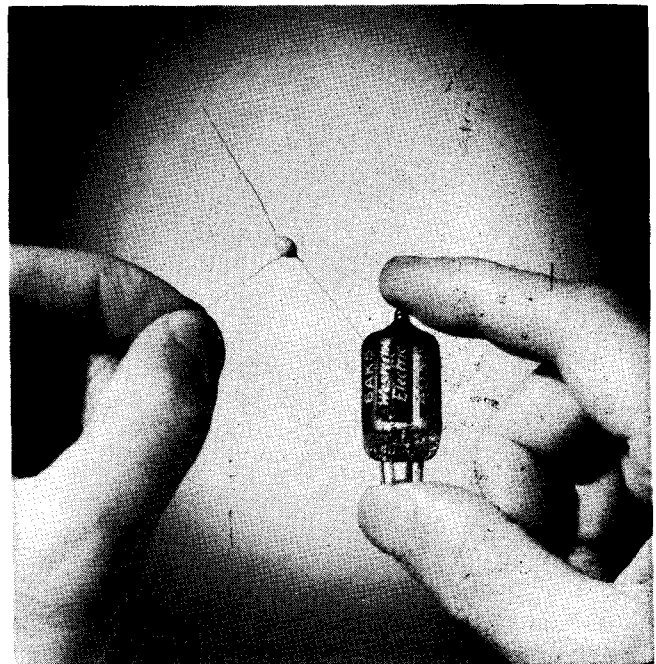


FIG. 26-19 TRANSISTOR AMPLIFYING CIRCUIT

valence electrons, which normally bind together the atoms of the material. The number of such electrons that can be dislodged so that they can move through the material is, in any event, exceedingly small as compared to the number of free electrons in a conductor. The total of possible free electrons can be increased somewhat, however, by adding to the semi-conducting material a limited amount of an impurity whose atoms have one more valence electron than do those of the basic material. The compound is then known as an **n-type** semi-conductor. Similarly, the admixture of an impurity whose atoms have one less valence electron than the base material will tend to cause a deficit of free electrons, or a surplus of holes. This is a **p-type** semi-conductor. Conduction in semi-conductors is apparently confined



A BEADED N-P-N TRANSISTOR (COMPARED WITH SMALL ELECTRON TUBE TO ILLUSTRATE RELATIVE SIZE)

largely to the surface layer of the material, which may be considered to account for their rectifying properties.

Figure 26-19 indicates schematically a transistor connected in one of various possible amplifying arrangements. Here if we assume that the semi-conductor is n-type, and the emitter *E* is positively biased as shown, electrons will tend to flow from the semi-conductor into the emitter, which leaves holes in the material that may be considered as being injected by the emitter. These holes, being equivalent to positive charges, tend to



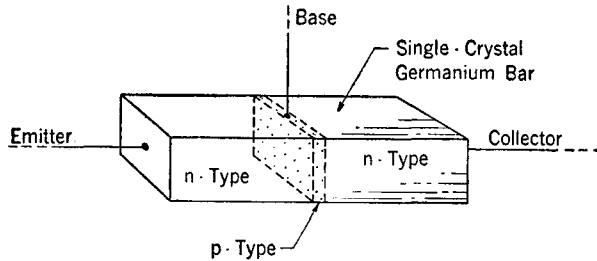


FIG. 26-20 STRUCTURE OF N-P-N TRANSISTOR

flow towards the negatively charged collector. The flow of holes from the emitter appears to "condition" the tiny surface area with which the emitter and collector points are in contact in such a way as to modify the electronic structure of the semi-conducting material. This modification may be great enough to cause a variation in the number of electronic carriers by as much as a thousand-fold or more. The net current in the collector circuit may vary accordingly. Thus, relatively small voltage changes in the emitter circuit may

cause large current changes in the collector circuit—which of course is the basic principle of amplification.

It may be noted that the circuit arrangement of Figure 26-19 has a definite resemblance to a grounded grid vacuum tube amplifier, where the emitter corresponds to the cathode, the collector to the anode, and the base to the grid. The analogy holds good in that both amplifier circuits would have low input impedance, high output impedance, and no phase change of the output signal as compared with the input signal. Care must be taken, however, in carrying this analogy too far.

There are designs of transistors other than the A-type discussed above, which do not employ cat-whisker contacts. One of these is the **n-p-n** type of transistor, which is illustrated schematically in Figure 26-20. The complete transistor may be enclosed in a tiny plastic bead, as shown in the accompanying photograph. Its operating principles are not different from those outlined above for the type-A transistor.

## CHAPTER 27

### AMPLIFIER APPLICATIONS IN TELEPHONE PRACTICE

#### 27.1 Telephone Repeaters

Modern telephone practice requires the use of nearly all types of amplifiers to meet its various needs. In terms of numbers in use, the most important application is the relatively simple voice-frequency amplifier commonly known as the telephone repeater. This device is inserted at regularly spaced intervals in long voice-frequency telephone circuits to offset line attenuation.

Since telephone conversation moves in both directions, each telephone repeater must include two oppositely poled amplifiers. In a two-wire telephone circuit, accordingly, the telephone repeater must be "double-tracked" by means of hybrid

coils or comparable devices, as was discussed in Article 17.7. Figure 27-1 is a schematic of a telephone repeater known as the 22-type. As will be observed, the amplifier units in this arrangement are triodes connected with transformers in both input and output circuits. The drawing also shows the connections of the hybrid coil output transformers with their balancing networks; and the potentiometers in the input circuits for controlling the amplifier gains. Equalizing networks are inserted at the mid-points of the low-impedance sides of the input transformers; and low-pass filters are included in the output circuits to prevent the passage of high frequencies not essential

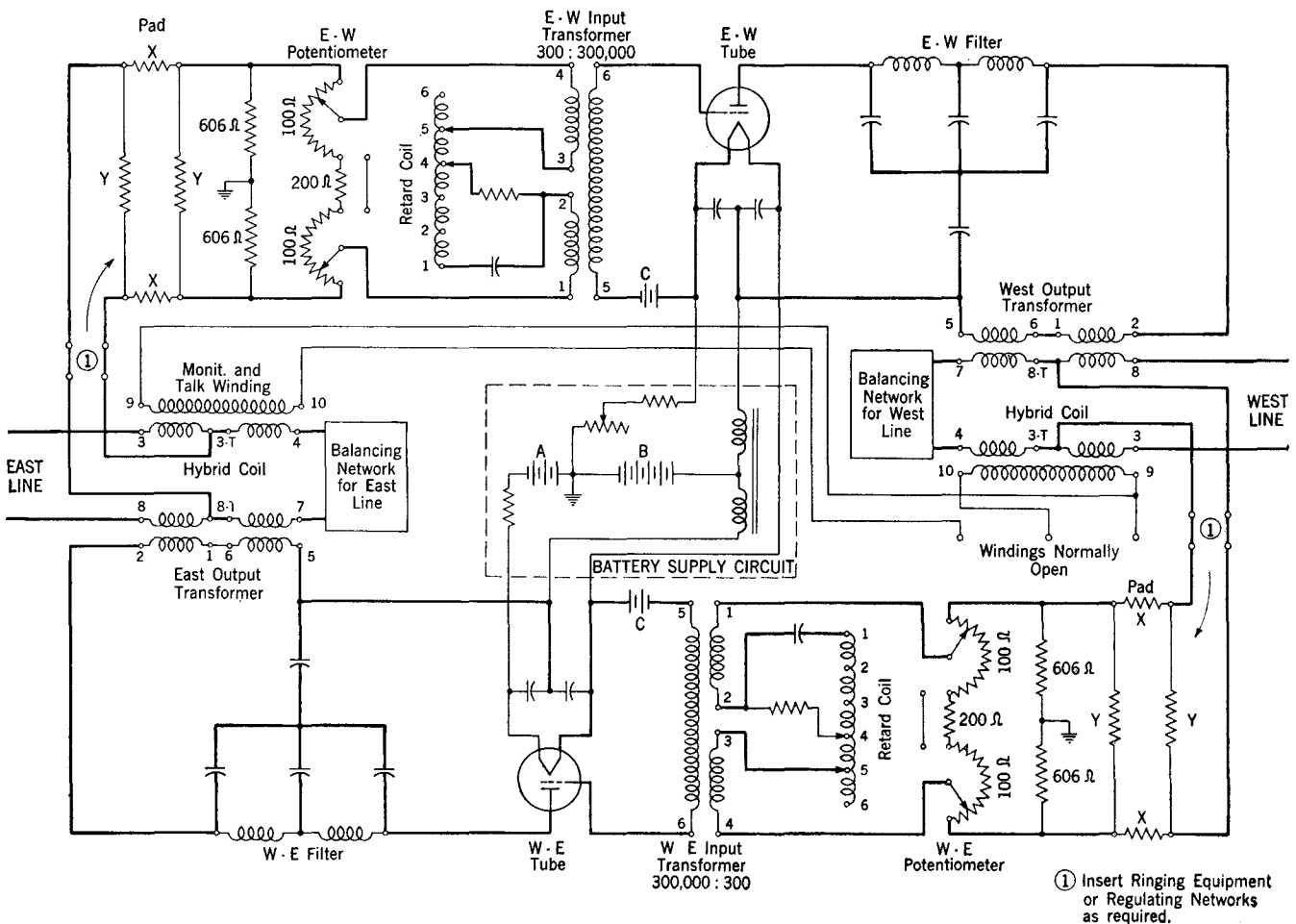


FIG. 27-1 22-TYPE TELEPHONE REPEATER CIRCUIT

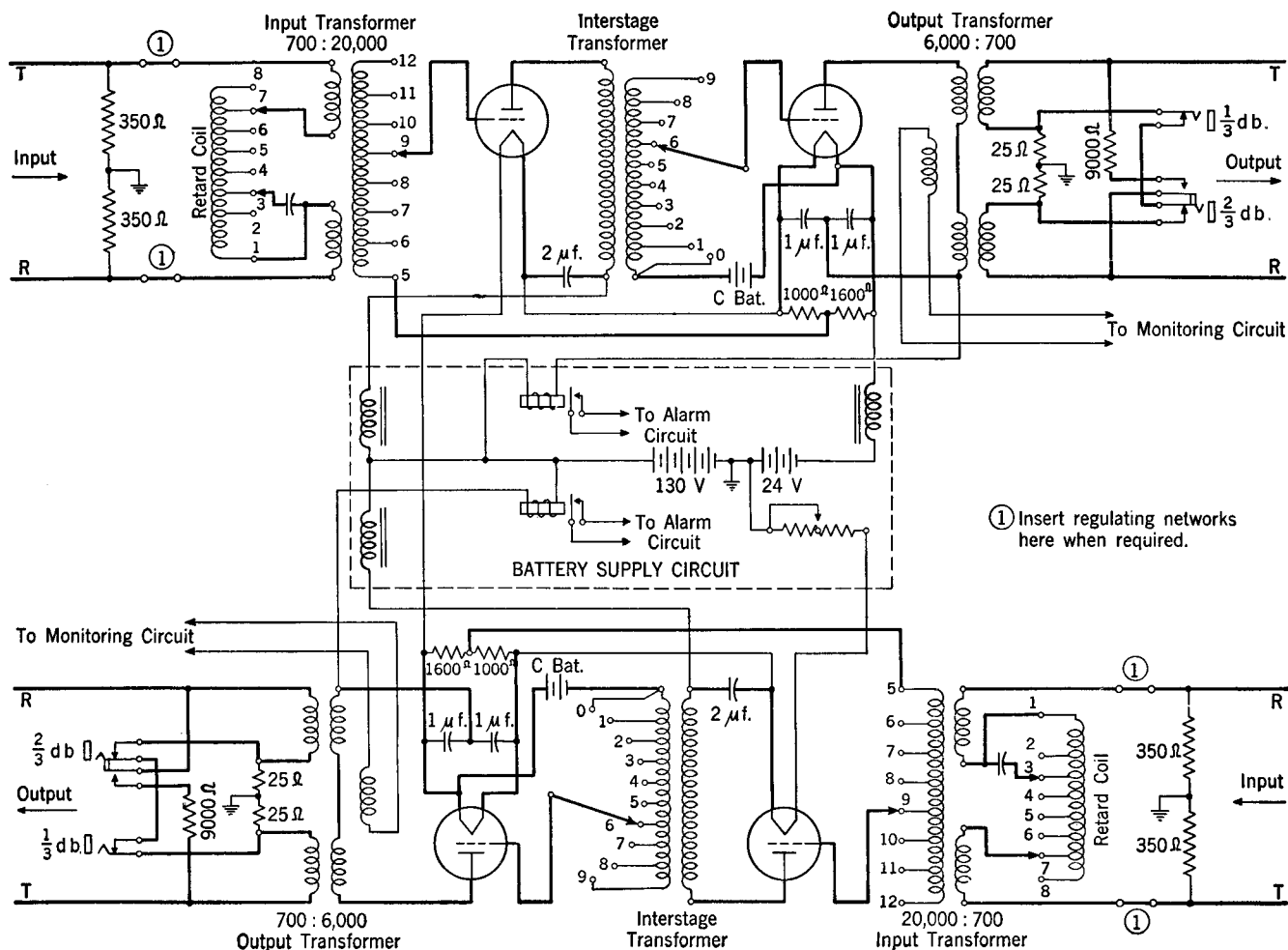


FIG. 27-2 44-TYPE TELEPHONE REPEATER CIRCUIT

for voice transmission. The maximum overall gain of this repeater is approximately 19.5 db when the potentiometers are on top step, but the gain of the amplifying units themselves must be higher than this to overcome the losses in the hybrid coils and other circuit elements. The gain is essentially flat over the frequency range from about 200 to 3000 cycles.

For 4-wire circuits, the repeater corresponding to the 22-type repeater is known as the **44-type**. In this case, the circuit itself is double-tracked so that there is no necessity for using hybrid coils except at the circuit terminals where the 4-wire circuit is converted to 2-wire for connection to the switchboard. There is therefore no need for line balancing networks at repeater points, and little possibility of a "singing" path around the individual repeater. For this reason, 4-wire repeaters may generally be operated at higher gains than 2-wire repeaters.

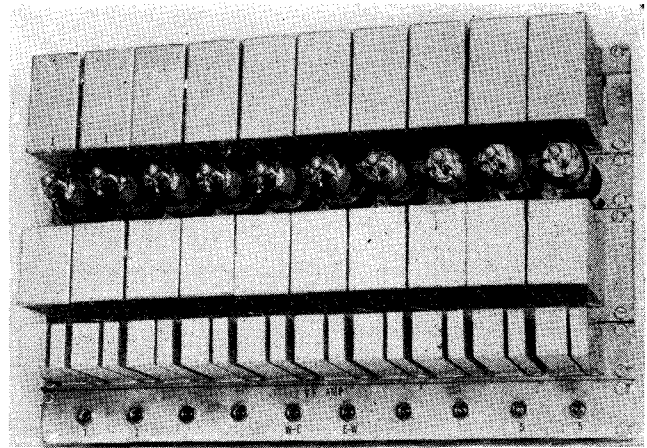
The circuit arrangements of the 44-type re-

peater are illustrated schematically by Figure 27-2. It will be noted that each amplifier has two triode stages, the first tube acting as a voltage amplifier and the second as a power amplifier. Transformers are used for interstage coupling, as well as in the input and output circuits. The gain is controlled by adjustable steps on the secondary windings of both the input and interstage transformers. The shape of the gain-frequency characteristic is controlled by an equalizing network connected in series at the mid-point of the primary of the input transformer. The maximum overall gain of this repeater is 42.7 db and is flat to frequencies well above 3000 cycles.

Where extremely stable amplifier operation is required, as for example in the repeaters of telephotograph circuits, the 44-type repeater may be modified for operation with negative feedback. This reduces the maximum overall gain to about 38 db. Feedback is from an output unit made up of capacitors and resistors inserted between the

second-stage tube and the output transformer, to a similar input unit inserted between the input transformer and the first-stage tube.

Another and more recent design of the voice-frequency telephone repeater is known as the **V-type**. It differs from the 22 and 44-types considerably, both with respect to the amplifiers themselves and the associated equipment arrangements. Hybrid coils, equalizers, filters and regulating networks are associated with the line equipment instead of with the amplifiers, so that the repeater proper consists only of the amplifiers themselves. All repeaters are thus essentially identical and this makes it possible to transfer them freely from one circuit to another, as may be required for maintenance purposes. It also makes possible the use of the same repeaters for either 2-wire or 4-wire operation.



V1 TELEPHONE REPEATERS

as discussed in the preceding paragraph, although about 6 db more negative feedback is provided. This gives somewhat greater stability of gain against tube and battery variation. Overall gain may be controlled by adjustment of a potentiometer connected across the secondary winding of the input transformer, which permits a total net variation from a small loss to about a 36 db gain. The vacuum tube used has about twice the transconductance and substantially the same output power as the tube employed in the earlier V-type amplifier. The entire amplifier unit is of the plug-in type which provides for quick replacement of defective units and facilitates testing and maintenance.

Where long telephone circuits are used for transmitting program material to the transmitters

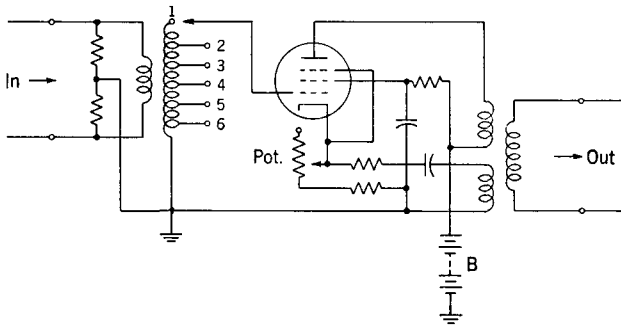
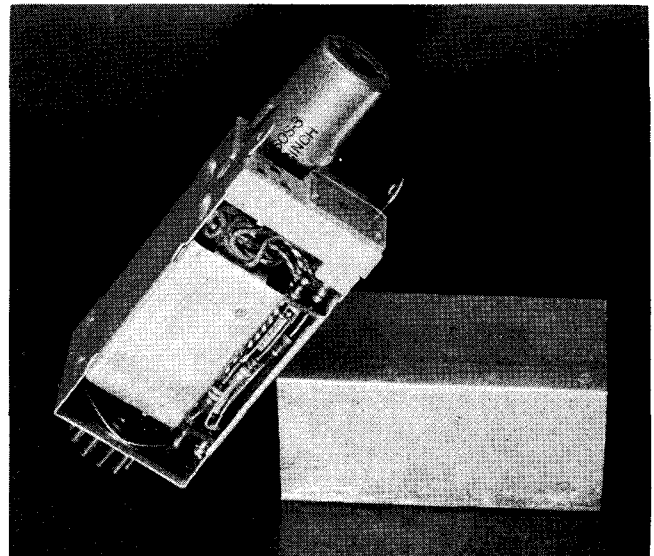


FIG. 27-3 ONE-HALF OF V-TYPE TELEPHONE REPEATER CIRCUIT

A schematic of one repeater of the V-type is given in Figure 27-3 (showing only one-half). It will be noted that the amplifying element is a pentode rather than a triode. This permits a maximum net gain of about 35 db, even though the feedback circuit causes a reduction of approximately 10 db. Feedback is derived from an extra winding on the output transformer and resistances in the cathode circuit. These latter include a potentiometer, which, together with taps on the secondary winding of the input transformer, serves to control the gain. The gain-frequency characteristic of the amplifier varies less than 1 db over a range of about 250 to 4000 cycles.

Another and later V-type repeater (V3) employs "miniature" pentodes, and correspondingly small equipment parts including input and output transformers, capacitors and potentiometers. The miniature technique is used primarily for space saving purposes, and the overall characteristics of the amplifier are approximately the same



DETAIL OF V3 REPEATER AMPLIFYING UNIT

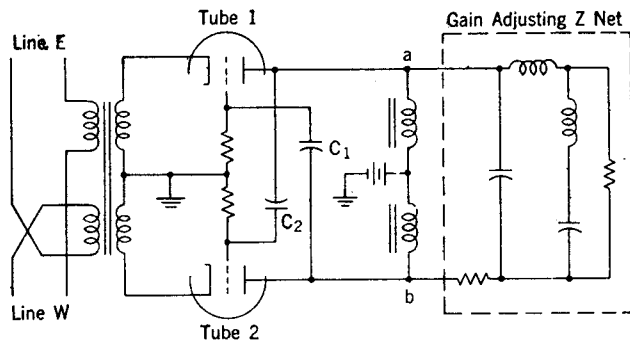


FIG. 27-4 E-TYPE TELEPHONE REPEATER

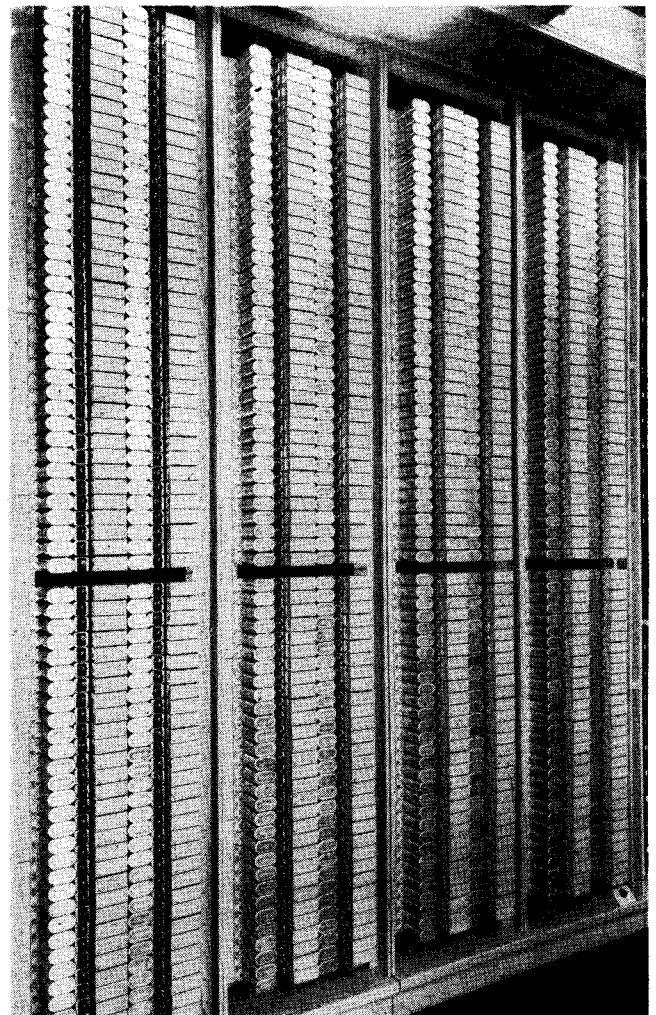
of a radio broadcasting network, it is frequently necessary to employ amplifiers in the lines at terminal or intermediate points. The "program amplifier" need be only a one-way device, but it must handle a frequency band somewhat wider than is required of the ordinary telephone repeater. One widely used design of such an amplifier employs pentodes connected in two transformer-coupled push-pull stages. Negative feedback is through a gain regulating network providing a maximum net gain of about 42 db. The gain-frequency characteristic is flat from 30 to 8000 cycles and is off by only 1 db at 15000 cycles. Other types of program amplifiers, using *R-C* interstage coupling, have comparable characteristics.

An interesting and rather remarkable design of voice-frequency amplifier circuit is illustrated in Figure 27-4. This device is sometimes known as a **negative-impedance repeater** or **converter**, but is coded in the Bell System as an **E-type** telephone repeater. Instead of being inserted in the line as in the case of other types of telephone repeaters, the amplifier is coupled to the line through a transformer without breaking the line continuity. This transformer may be viewed as both an input and output transformer. As indicated in the drawing, the amplifier circuit employs a dual-triode connected in a push-pull arrangement. The grounded grid connection of the tubes results in a very large feedback because the input and output are in a common circuit. The secondary windings of the transformer are included in the output (plate to cathode) circuit as well as in the input circuit and plate to cathode current thus flows through both the output and input circuits. Plate to cathode current changes accordingly tend to set up induced voltages in the primary side of the transformer.

At first glance it might appear that this circuit arrangement would have practically no effect

on the transmission line because current flowing in the line would induce voltages in the transformer secondary, which would be applied across the cathode-grid circuit of the tubes to cause corresponding currents in the plate-cathode circuits that would flow through the transformer secondaries and set up voltages that would counteract the original applied voltages. This would be true if it were not for the capacitors  $C_1$  and  $C_2$ , which couple the plate of each tube to the grid of the other tube. The potential at the grid of Tube 2 is thus determined not alone by the input voltage, but also by the potential at point *a*. This potential depends on the amount of current flowing in the circuit of Tube 1 and the resultant voltage drops across the retardation coil and the elements of the gain adjusting network. Similarly, the grid potential of Tube 1 is controlled by the potential at point *b*.

A careful analysis of the voltages throughout



E-TYPE TELEPHONE REPEATERS

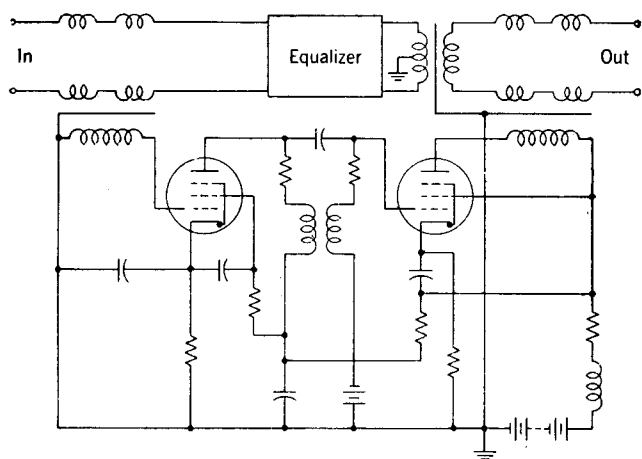
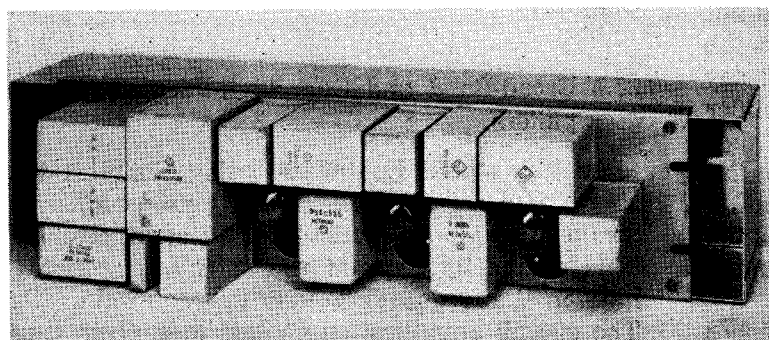


FIG. 27-5 TELEPHONE REPEATER USED IN TYPE-C CARRIER SYSTEM

the circuit when an a-c input signal is applied, will show that amplified voltages are set up in the secondary windings, and that these voltages are of such phase as to induce voltages in the primary windings that add to the line signal voltages so as to increase the current in the line in either direction. The net amount of amplification secured is controlled by the gain adjusting network by virtue of its control over the potentials at points *a* and *b*. In practice, the gain adjusting networks are designed so that the connections of their elements can be adjusted in various specified ways depending upon the characteristics of the line facilities in which the repeater is used. The network connections thus determine not only the overall gain of the repeater, but provide equalization to match the loss-frequency characteristics of the line. Repeaters of this type provide gains up to 8 or 10 db over the voice-frequency range of approximately 300 to 3500 cycles. The application of E repeaters is generally to Exchange telephone plant, where they may be used effectively to improve transmission on long trunks or subscriber lines. They can be applied either at ter-

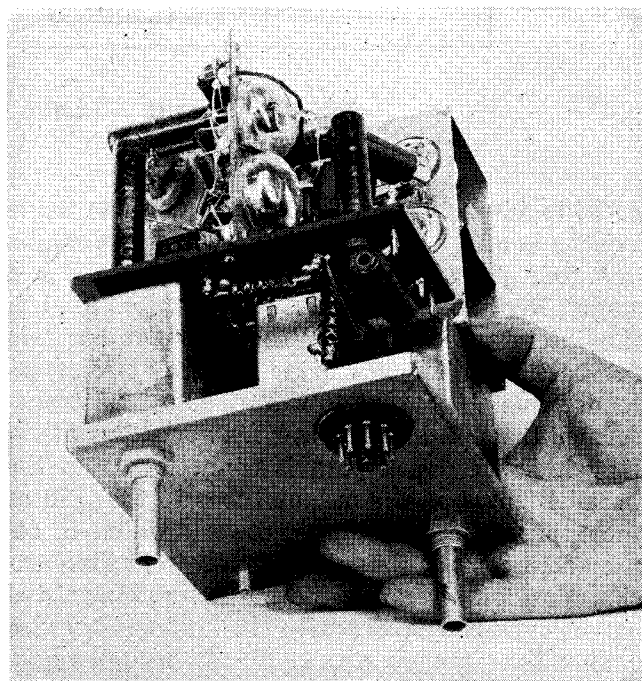


TYPE-K CARRIER LINE AMPLIFIER

minals or intermediate points of such lines or trunks.

## 27.2 Carrier Amplifiers

The several types of carrier systems currently in use in telephone practice employ frequency bands ranging from 4000 cycles up to as high as several million cycles. The amplifiers used in these systems must be designed to handle the entire frequency band of each particular system on a high fidelity basis with a reasonably flat gain over the total frequency range. All such amplifiers are designed with stabilized feedback circuits.



TYPE L1 CARRIER AMPLIFIER

Figure 27-5 is a schematic of an amplifier used in the type-C carrier system which operates over the frequency range from 5 to 35 kc. As indicated, it consists of two transformer-coupled pentodes with hybrid type input and output transformers. The negative feedback circuit is connected between the two hybrids through an equalizer network circuit. This amplifier has a gain of 50 db, flat from 5 to 35 kc.

Amplifiers for the type-J and type-K carrier systems must handle still broader frequency bands. Thus the type-J amplifier must operate through a range

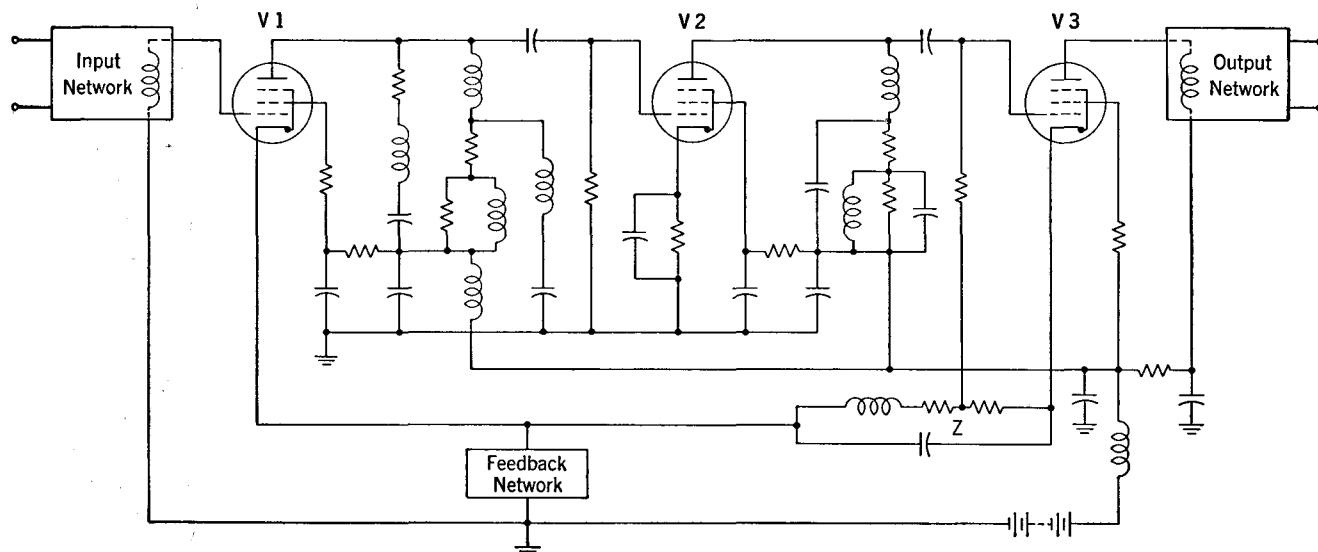


FIG. 27-6 L-CARRIER AMPLIFIER CIRCUIT (PARALLELING TUBES OMITTED)

of 36 to 140 kc. It is a three stage amplifier using voltage amplifying pentodes in the first two stages and four power pentodes in parallel in the last stage. Resistance-capacitance coupling is used in the interstage networks. The amplifier has two feedback circuits. The outer feedback circuit extends from the output to the input hybrid transformer through an equalizer network. An auxiliary inner feedback path is connected from the parallel plates of the output tubes to the grid of the input tube through a network which is designed to control the singing margin at frequencies considerably above the normally transmitted band. Type-K carrier amplifiers (12 to 56 kc) also employ three stages, with feedback through a gain adjusting and equalizing network.

Figure 27-6 shows schematically the circuit of the amplifier used in type-L1 carrier systems, which operate in the range from 60 to 3000 kc. Although the diagram shows only three tubes, these amplifiers are actually built with paralleling tubes in each stage so that a tube failure will not stop the functioning of the amplifier. Because the amplifier must operate over such a wide frequency range, the interstage networks of this amplifier are quite complex. As a matter of fact, neither of the interstage circuits by itself provides a flat gain, but the two interstages in tandem yield an essentially constant gain between the grid of the first tube and the grid of the last tube over the entire carrier range.

The inter-electrode capacitances of the tubes are minimized by the use of special types of min-

ature tubes having relatively high transconductance. Two feedback paths are provided, one around the output tube V3, and the other around the entire circuit. The alternating component of the plate current of tube V3 flows from the cathode of that tube through the impedance  $Z$  and the feedback network to ground; and thence through the primary of the transformer in the output network to the plate of the tube and back to the cathode. The voltage drop to ground across  $Z$  and the feedback network resulting from this current flow is applied to the grid of V3. This local feedback suppresses modulation effects (distortion) developed in the output tube. The alternating voltage developed across the feedback network is applied to the grid of tube V1 through the secondary of the transformer in the input network. This is the main stabilizing feedback of the amplifier circuit. It also provides a means for regulation of the amplifier through appropriate adjustments of the feedback circuit, as is discussed in a later Chapter.

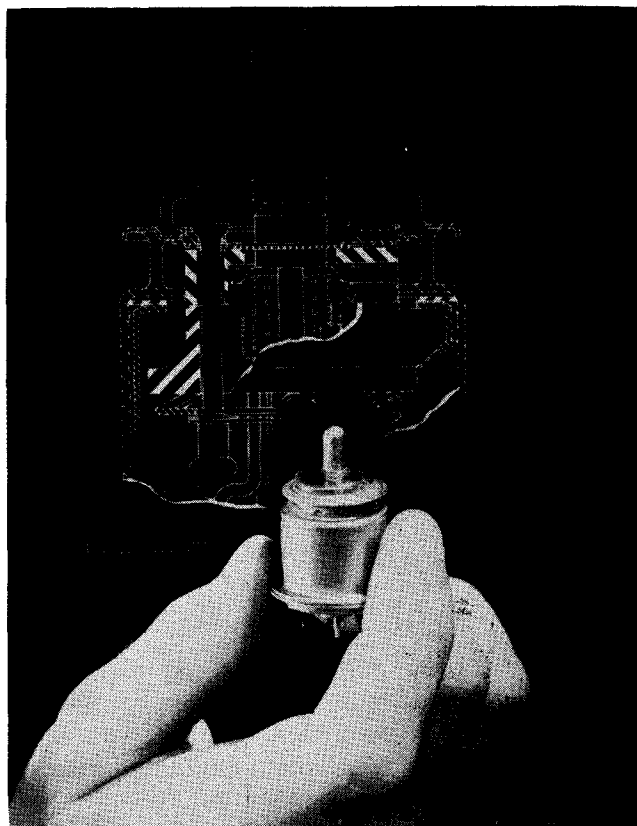
### 27.3 Broad-Band and Superhigh-Frequency Amplifiers

As was stated in the preceding Chapter, the stages of the usual radio amplifier are coupled by means of single- or double-tuned resonant circuits. This ordinarily permits the satisfactory transmission of a frequency band of the required width. The overall amplification desired in a given case is obtained by using as many stages as may be

necessary, with the required output power secured by the use of appropriate power tubes in the final output stage. In most cases, radio amplifiers do not employ stabilized feedback.

In cases where very broad band signals are to be transmitted by radio, such as those of television and L-carrier, reasonably flat amplification over the entire signal range may require the use of a rather large number of stages. For example, the intermediate frequency (IF) main amplifier of the TD-2 radio relay system employs eight stages to give a maximum gain of about 60 db with a bandwidth of about 22 mc. The input circuit, output circuit, and all interstages except the last are double-tuned. The last interstage is triple-tuned. Broad-band IF amplifiers of this type are sometimes "stagger-tuned"; that is, the various interstage networks are tuned to a number of different frequencies within the range of the frequencies to be transmitted.

Amplifiers for use at superhigh-frequencies (3000 to 10,000 mc) present a number of special problems that require rather fundamental design differences. Chief of these perhaps is the fact that the transit time of the electrons in the amplifier tubes becomes a matter of major importance at these extremely high frequencies. As was pointed out in Article 25.4, the adverse effects of transit time likely to be encountered in ordinary vacuum tubes can be overcome by the use of tubes of the klystron type. These tubes, however, are rather



MICROWAVE TRIODE

difficult to maintain. Furthermore, present designs of klystrons do not permit the use of more than a limited number of klystron amplifiers in tandem without cumulative noise and distortion becoming excessive.

For application in very long radio relay systems requiring dozens of amplifiers in tandem, accordingly, a special tube (W.E. 416) was developed by Bell Telephone Laboratories. Because of the wide band of frequencies to be handled (at least 20 mc) this tube had to have very high transconductance. To overcome transit time limitations, the elements or electrodes of the tube also had to have extremely close spacing. Fortunately, these two requirements are compatible. The 416 tube is a triode of the so-called planar type, in which the elements are in parallel planes, with the grid grounded to the frame of the structure. A perspective drawing (greatly enlarged) of the elements of the tube is shown in Figure 27-7. The oxide coating of the cathode is .0005" thick. The cathode-grid spacing is .0006". The grid wires are spaced a thousand to the inch and are .0003" in diameter. The plate-grid spacing is .012". The large number of very fine wires employed in the

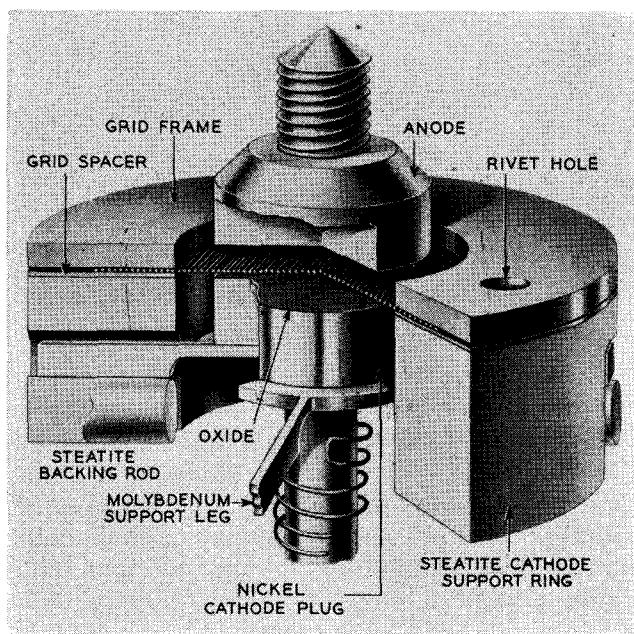


FIGURE 27-7



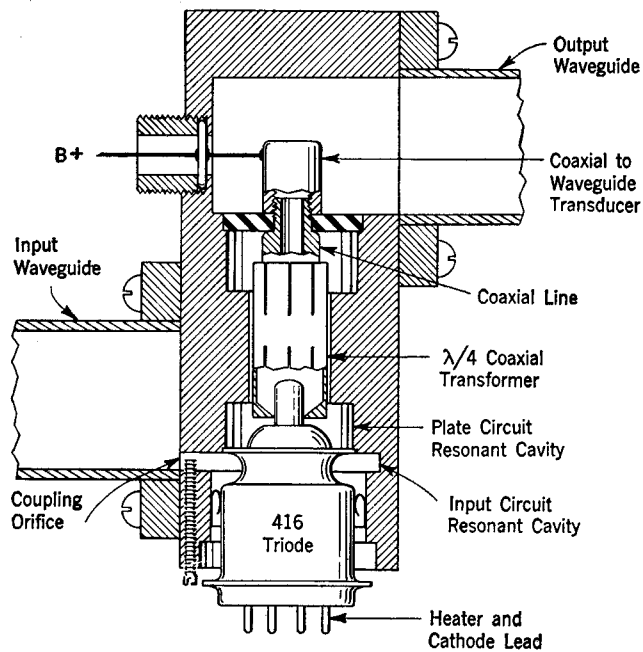


FIGURE 27-8

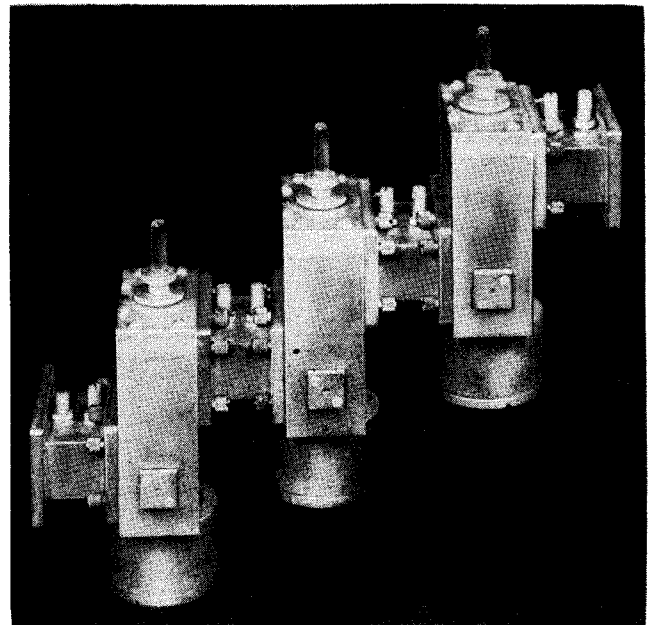
grid structure provides a close approach to a uniform electrostatic shield between cathode and plate, without interfering with the free flow of electrons. The transconductance of the tube is in the order of 50,000  $\mu\text{mhos}$ , the amplification factor is about 350, and the output resistance 7000 ohms.

This triode always operates in a grounded grid circuit arrangement, where the input is applied between the cathode and ground and the output impedance is between the plate and ground. This automatically eliminates coupling between the output and input circuits through the inter-electrode capacitances of the tube, a feature that is especially useful at very high frequencies where neutralization of internal capacitance coupling by conventional methods is difficult. The possible gain, however, is less than that of the more usual grounded cathode arrangement because of the negative feedback inherent in the fact that the cathode to ground input circuit is included in the plate current path.

Figure 27-8 shows one of these tubes connected in a waveguide "circuit". The input waveguide is coupled through an iris to an input cavity which may be tuned to resonance by a trimming screw across the opening. The grid separates the input cavity from a second resonant cavity of the plate or output circuit. This cavity transforms the plate

impedance of the tube to a very low resistance (a fraction of an ohm). A quarter-wavelength coaxial line, which can be adjusted vertically to tune the output cavity, matches this low impedance to the impedance (about 45 ohms) of a short coaxial line leading to the transducer probe which extends into the output waveguide.

Three tubes of this type are used in the transmitter amplifiers of the TD-2 microwave radio relay system. The three stages of the amplifier, which is illustrated in the accompanying photograph, are connected in cascade through waveguide tuners that effectively form double-tuned critically coupled transformers. The overall gain of the amplifier is normally adjusted to 18 db with an output power of 0.5 watt, although a somewhat higher gain is possible. The overall transmission characteristic is flat over about 20 mc between points 0.1 db down. As the output power of the close-spaced triode is increased, its maximum



THREE-STAGE MICROWAVE AMPLIFIER

possible gain decreases. Accordingly, the gains of the three stages of the amplifier are not alike. The first stage has an output of about 80 milliwatts and a gain of about 9 db. The second stage output is about .25 watt with a gain of about 6 db, and the third stage output is about 0.5 watt with a gain of about 3 db.

## CHAPTER 28

### RECTIFIER, DETECTOR, OSCILLATOR AND OTHER ELECTRON TUBE CIRCUITS

#### 28.1 Electron Tube Rectifiers

The operation of most electron tubes requires the application of d-c potentials to certain of the tube electrodes—particularly the anodes. These voltages are sometimes supplied by batteries, but it is generally more economical and practical to make use of ordinary commercial power sources when a considerable number of tubes are involved. Since these sources almost invariably supply only a-c power, it is necessary to convert this to d-c power for tube operation. Devices for accomplishing this result are called rectifiers.

Article 7.5 discussed briefly the “barrier-layer” metallic rectifiers that are quite generally used for developing substantial amounts of d-c power in telephone central office power plants. For more limited requirements, such for example as those of the ordinary radio receiver, electron tubes are used to produce the necessary d-c voltage. Rectifying tubes are diodes, and may be of either the high vacuum or gas type. Figure 28-1 shows a diode connected in a circuit arrangement known as a half-wave rectifier. Here the alternating supply voltage is impressed through a transformer across the plate and cathode of the tube. The tube filament is also heated from this same supply voltage by means of an auxiliary transformer winding as shown. Due to the fundamental characteristics of the tube, current can flow only during the time the plate is positive with respect to the cathode. During the positive half of the impressed voltage cycle, accordingly, the tube conducts and the secondary of the transformer is effectively connected to the load. During the negative half of the cycle, no space current can flow

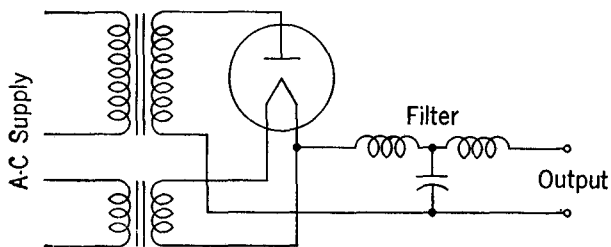
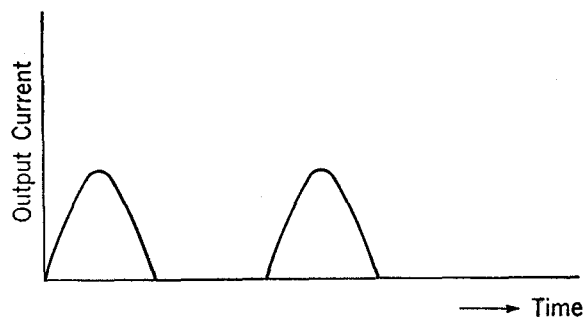
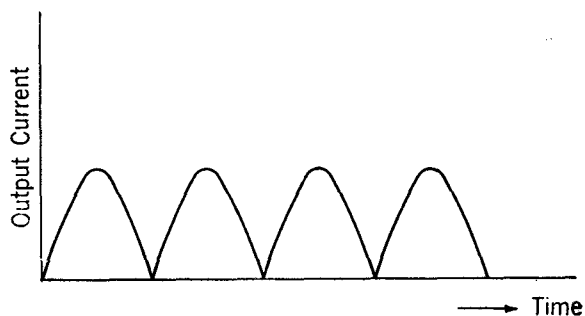


FIG. 28-1 HALF-WAVE RECTIFIER

and the tube effectively opens the circuit to the load. The voltage applied to the filter will therefore consist of a series of pulses, as shown in Figure 28-2(A). If the diode is of the high vacuum type, each pulse will have a waveshape practically identical with that of the positive half of the applied a-c voltage. When the applied volt-



A. Half-Wave Rectifier



B. Full-Wave Rectifier

FIGURE 28-2

age is a sine wave, as is normally the case, the output current will contain a d-c component equal to the average value over each half-cycle pulse plus a-c components at the frequencies of the input voltage and its even harmonics. These a-c components are not only useless but undesirable since they would tend to introduce noise (power hum) into any circuits to which the rectified voltage might be applied. They are accordingly suppressed by a filter which usually consists of series inductors that present high impedance to the a-c

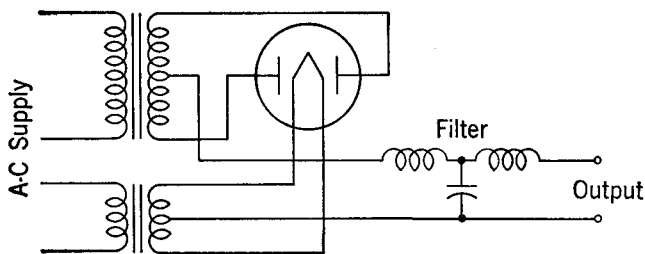


FIG. 28-3 FULL-WAVE RECTIFIER

components, and shunt capacitors whose a-c impedance at the frequencies involved is low.

It is fairly obvious that the ratio of the total values of the a-c components to the d-c component in the output of the half-wave rectifier is rather large. In other words, this rectifier has a high **ripple factor** and is not very efficient. A better arrangement is the full-wave rectifier circuit shown in Figure 28-3, where the tube has two plates so connected that there will always be current between one of the plates and the cathode. Output current will flow during both halves of the impressed cycle to give a series of positive pulses as indicated in Figure 28-2(B). This output can be analyzed into a d-c component which is obviously twice that of the half-wave rectifier output, plus a-c components having twice the input voltage frequency and its harmonics. The ripple factor is substantially reduced and the d-c output power is four times that of the half-wave rectifier since power is proportional to the square of

the doubled current. Because of the lower ripple factor the structure of the smoothing filter may also be somewhat simpler.

Where relatively large amounts of rectified power are needed, as for the operation of amplifiers in radio transmitters or public address systems, gas diodes are frequently employed. Such tubes—especially the mercury-vapor type—are more efficient than high vacuum tubes and will deliver a substantially larger output current for a given size tube. The circuit arrangements for such tubes are no different than those employing high vacuum tubes as shown in Figures 28-1 and 28-3. The output current pulses, however, do not follow the shape of the input so closely because the gas tube does not begin to conduct until the applied voltage has risen to the firing point of the tube.

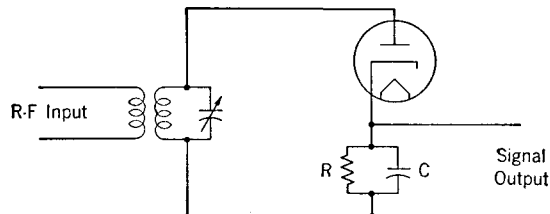
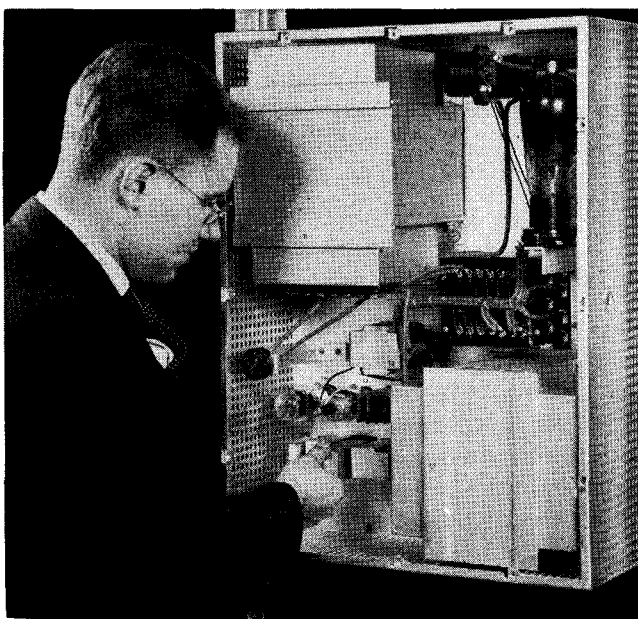


FIG. 28-4 DIODE DETECTOR CIRCUIT



MERCURY-VAPOR TUBE RECTIFIER

Certain special precautions must be taken in operating a gas tube as a rectifier. Most important of these is the requirement that the cathode be heated to its full operating temperature before voltage is applied to the plate. This may take several minutes, depending on the particular type tube. It was pointed out in Article 25.1 that a gas tube when operating normally has a small constant voltage-drop between its electrodes, which in the case of the mercury-vapor tube, is only about 14 volts. If voltage is applied to the plate of the tube before the cathode is sufficiently heated to develop its normal emission of electrons and the accompanying normal voltage-drop, the large applied voltage will cause positive ions to move towards the cathode at high velocities. Their impact with the cathode may quickly break down the emitting coating and thus destroy the usefulness of the tube. Neither gas nor high vacuum tubes should be operated at voltages that will cause the plate current to exceed that determined by the maximum cathode emission as this will cause damage through overheating of the electrodes or disintegration of the active cathode material. Care should also be exercised that the

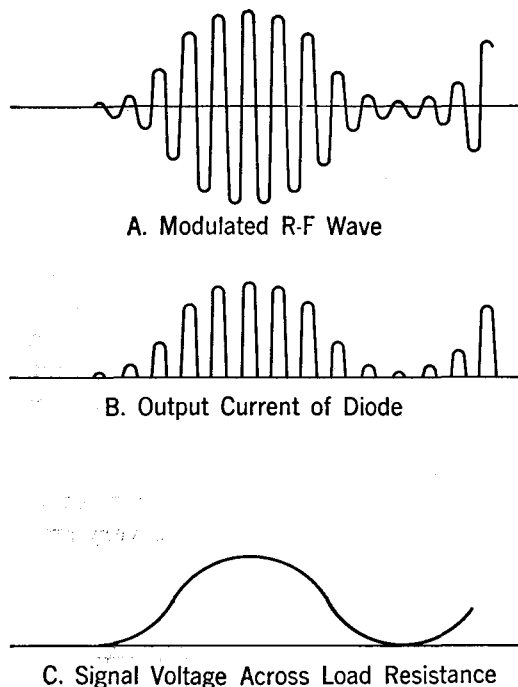


FIGURE 28-5

maximum inverse voltage rating across the tube is not exceeded. This is particularly true in the case of gas tubes which, having higher conductivity, will arc back more easily than the high vacuum tube.

## 28.2 Detectors

**Detector** is the name commonly used in radio work to designate the device employed at the receiving end of the circuit to separate the signal from the high-frequency carrier wave on which it was super-imposed or **modulated** at the transmitting end of the circuit. Detectors, which may also be properly spoken of as demodulators, are essentially rectifiers that normally handle only very small amounts of power.

The high vacuum diode is an effective detector except at very low levels of radio-frequency input voltage. A simple detector circuit employing a diode is shown in Figure 28-4, and its operation in detecting or separating the signal voltage from the radio-frequency input may be understood by referring to Figure 28-5. Figure 28-5(A) indicates a radio-frequency carrier wave that has been modulated by a lower frequency signal. The amplitudes of the pulses of the radio-frequency wave vary in accordance with the signal which the wave is carrying (see Chapter 29). The tube of course will permit the passage of only the posi-

tive half of the cycles, as indicated in Figure 28-5(B). Each positive pulse produces a voltage-drop across the resistor-capacitor circuit  $R-C$  proportional to the value of the plate current. Each pulse tends to charge the capacitor, and between pulses some fraction of this charge will leak off through the resistor  $R$ .  $R$  is of such value, however, that only a very small portion of the charge leaks off before the arrival of the next radio-frequency pulse. As a result, a potential will be developed across the  $R-C$  combination which is proportional to the peak values of the r-f pulses, with only a slight r-f ripple component. On the other hand, the time constant of the  $R-C$  circuit must be low enough so that the charge on  $C$  will be completely dissipated through  $R$  between the cycles of the signal wave. Under these conditions, the net voltage-drop across the  $R-C$  combination will have the shape shown in Figure 28-5(C), which should be identical with the signal applied to the transmitting circuit. These conditions will be realized in a typical detector circuit when the value of  $C$  is in the order of  $100 \mu\mu\text{f}$ , and  $R$  is several hundred thousand ohms.

High vacuum triodes, or multi-electrode tubes, may also be used as detectors. A so-called **plate detector** circuit employing a triode is shown in Figure 28-6. The grid of the tube is biased to the cutoff point so that the tube acts as a rectifier in much the same way as does the diode. The amplitudes of the positive r-f current pulses appearing in the plate circuit follow the amplitude of the signal impressed on the input voltage. Radio frequency is eliminated from the output by means of an appropriate  $R-C$  circuit in the same way as previously discussed.

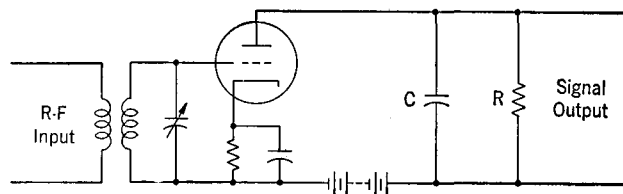


FIG. 28-6 PLATE DETECTOR CIRCUIT

Another circuit employing a triode is commonly known as a **grid-leak detector**. This circuit is illustrated in Figure 28-7. It will be noted that the grid of the tube in this circuit is not biased in the usual fashion, and grid current will tend to flow during the positive portion of each r-f cycle. This current, however, is limited by the charge building up in capacitor  $C_g$ , which sets up a negative

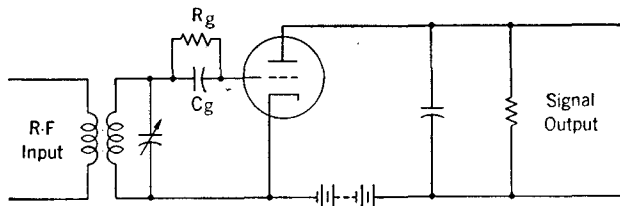


FIG. 28-7 GRID-LEAK DETECTOR

potential on the grid that will vary with the amplitude of the input voltage. The value of  $R_g$  is such that only a negligible portion of the charge on the capacitor can leak off between r-f pulses, but the capacitor will be completely discharged in the time interval between signal pulses. In other words, the grid-cathode circuit acts exactly like the diode detector circuit previously discussed. The net result is that a potential appears on the grid, which varies in magnitude in accordance with the impressed signal voltage. These effective grid bias variations produce corresponding variations in the plate current of the triode giving an output voltage across the load resistance which follows the signal. The circuit thus acts both as a detector and an amplifier.

Both diode detectors and grid-leak detectors draw some current from the input circuit, which means that an effective resistance component is shunted across the tuned tank circuit. This effective resistance is high, but it has some tendency to decrease the sharpness of selectivity of the tank circuit. The plate detector circuit does not affect selectivity because the grid never goes positive with respect to the cathode. On the other hand, it operates over a somewhat curved portion of the grid voltage-plate current characteristic, and therefore causes some distortion of the signal.

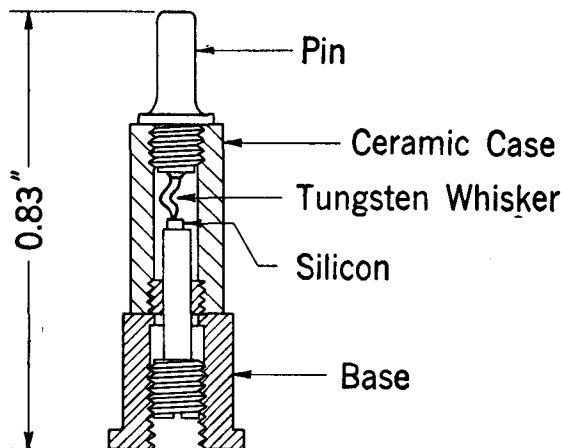


FIG. 28-8 POINT-CONTACT RECTIFIER

At superhigh-frequencies, the usefulness of vacuum tubes as detectors becomes limited because of their inter-electrode capacitances. Detection at these high frequency ranges is usually accomplished by means of **point-contact rectifiers**, commonly known as "crystal detectors". Detectors of this type were widely used in the earliest days of radio before the invention of the electron tube, but they were at that time rather unreliable in their operation. Relatively recent research in the field of semi-conductors and "varistors" has led to the development of a point-contact rectifier that is highly reliable, long-lived and rugged. As illustrated in Figure 28-8, it consists simply of a tiny tungsten cat-whisker held in point contact with a crystal of silicon. For most applications, the silicon is treated with a very small per-

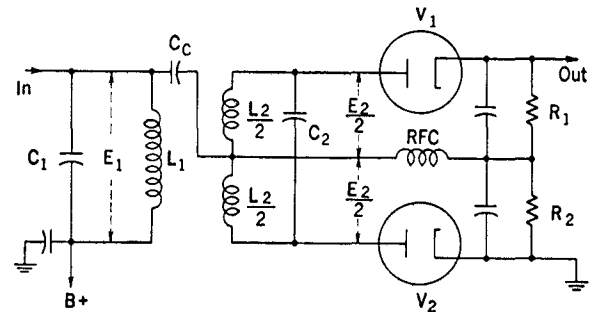


FIG. 28-9 TYPICAL DISCRIMINATOR CIRCUIT

centage of an appropriate impurity to make it an n-type semi-conductor (see Article 26.7). In this case, the rectifier permits current flow from the point to the crystal, but limits current flow in the opposite direction to relatively insignificant values. In the direction of ready current flow, the curve of current vs. applied voltage has a shape quite similar to the corresponding characteristic curve of a vacuum tube diode, and the crystal detector will accordingly perform corresponding functions when connected in the appropriate circuit arrangement. Because of their similarities of function, point-contact rectifiers are frequently referred to as diodes in the current literature.

### 28.3 Discriminators

The detection of the signal carried by a frequency-modulated radio wave requires the use of a circuit that will convert frequency variations into amplitude variations. One of the more common circuit arrangements used for this purpose is known as a **discriminator**. Such a device is also

fundamental to the operation of automatic frequency control circuits.

Figure 28-9 is a simplified schematic of a typical discriminator circuit. As will be noted, this circuit employs two diodes connected in a balanced arrangement. The primary and secondary tank circuits  $L_1C_1$  and  $L_2C_2$  are both tuned to the carrier or mean frequency of the r-f input voltage.  $L_1$  and  $L_2$  are also the primary and secondary windings of a transformer. The input voltage across  $L_1$  accordingly induces a voltage in the secondary winding  $L_2$ , each end of which is connected to the plate of one of the diodes. With reference to the center tap of the secondary winding, the induced voltages in the two halves of the winding are always equal and opposite in sign; and at the carrier frequency, when the tank circuit impedance is wholly resistive, are  $90^\circ$  out of phase with the input voltage  $E_1$ .

In the absence of the coupling capacitor,  $C_c$ , the circuit would always apply equal and opposite voltages  $E_2/2$  to the diodes, which would result in the development of equal and opposite voltage drops across the resistances  $R_1$  and  $R_2$ , and consequent zero output. The capacitor  $C_c$ , however, is relatively large so that the full r-f input voltage  $E_1$  is applied directly to the center tap of  $L_2$ . The net voltages applied to the plates of the diodes are accordingly equal to the sum or difference of this voltage and the voltages induced in  $L_2$  by the current flowing in  $L_1$ . In other words, the voltage at the plate of  $V_1$  will be  $E_1 + E_2/2$ , while the corresponding voltage at the plate of  $V_2$  is  $E_1 - E_2/2$ . When the input voltage is at the carrier or mean frequency to which the tank circuits are tuned, these voltages,  $E_{v1}$  and  $E_{v2}$ , will be equal as illustrated by the vector diagram in Figure 28-10(A). If the frequency of the input voltage becomes higher than the resonant frequency of the tuned circuits, the inductive reactance of the tank cir-

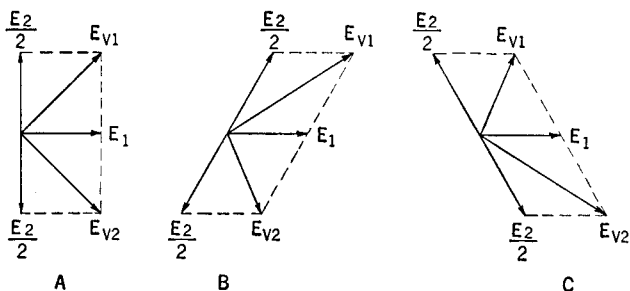


FIG. 28-10 VECTOR DIAGRAM OF VOLTAGES IN DISCRIMINATOR CIRCUIT

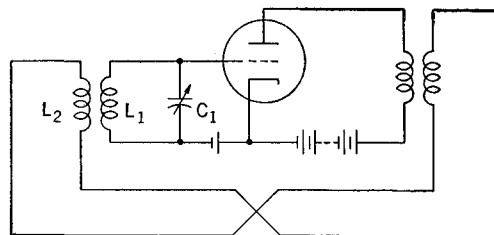


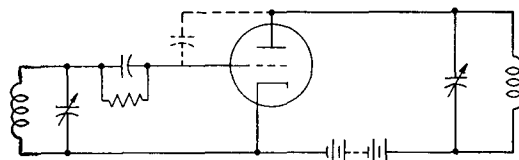
FIG. 28-11 INDUCTIVELY COUPLED OSCILLATOR

cuits predominates and the phase relationship between  $E_1$  and  $E_2$  is no longer  $90^\circ$ . Then, as illustrated by Figure 28-10(B), the voltages  $E_2/2$  add vectorially with  $E_1$  to give unequal resultant values. This results in a larger voltage drop across  $R_1$  than across  $R_2$  and a consequent positive output voltage with respect to ground. Similarly, if the input frequency is lower than the carrier frequency, the capacitive reactance of the tank circuits predominates and the vector diagram takes the form shown in Figure 28-10(C), resulting in the appearance of a negative voltage in the output circuit. Thus deviations of the input frequency from the mean or carrier frequency produce output voltages corresponding in sign to the direction of the variation and in magnitude to the extent of the deviation.

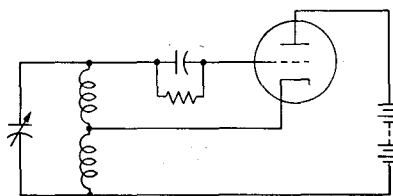
## 28.4 Oscillators

In communications work, the term **oscillator** is usually applied to electron tube devices which act as generators of a-c sine wave voltages. Practically any vacuum tube amplifier circuit will function as an oscillator if some part of the output energy is returned or fed back in phase to the input. The minimum requirement for sustained oscillation is that the energy so fed back must be at least as much as the reciprocal of the total amplification. That is to say, for example, if the energy amplification of the circuit is 100 times, at least  $1/100$  of the output energy must be fed back. This is a condition which it is not at all difficult to obtain. In fact, it is usually necessary in the design of any vacuum tube amplifying circuit to take special precautions to avoid the development of an oscillating condition.

Figure 28-11 represents a simple type of inductively coupled oscillator circuit. Here the amount of energy fed back into the input is determined by the coupling between the coils  $L_1$  and  $L_2$ , and the frequency of oscillation is controlled by the values of  $L_1$  and  $C_1$  in the resonant tank circuit. Many other designs of oscillator circuits are pos-



A. Tuned Grid Tuned Plate Oscillator



B. Hartley Oscillator

FIGURE 28-12

sible. Figure 28-12(A) shows an oscillator circuit in which both the grid and plate circuits include tuned tank circuits. Here the feedback or coupling between the plate and grid is assumed through the inter-electrode capacitance of the tube itself. If this is insufficient, it may be effected by the inclusion of a coupling capacitor as indicated by the dotted lines.

Figure 28-12(B) shows the principle of the well-known Hartley oscillator in which the tuned network is connected between the grid and plate of the tube, and the cathode is connected to a tap-point of the inductor. Thus the current flowing in the plate circuit produces a voltage between grid and cathode whose value depends upon the ratio of the inductive reactances on either side of the tap-point. It may be noted that grid bias is obtained in both circuits of Figure 28-12 by the use of a grid-leak resistor and capacitor. This arrangement has two advantages. It insures that the oscillator will be self-starting, since at the instant that voltage is first applied to the plate the grid will be unbiased and the tube will be working on a high point of its characteristic curve. This will permit an initial surge of current which will begin to charge the grid-leak capacitor and supply sufficient energy to the tank circuit to start it into oscillation. The first few oscillations will continue to build up the charge on the grid-leak capacitor and drive the grid increasingly negative until a steady-state operating condition is reached in which the energy supplied to the tank circuit is just sufficient to overcome its losses and thus maintain oscillations of a constant magnitude. The grid-leak biasing arrangement also tends to make the oscillator self-regulating because the grid bias will automatically change

in accordance with any change in the plate current that may be caused by variations in the load.

The circuits shown in Figures 28-11 and 28-12 do not indicate any load connection, but in any practical case some portion of the plate circuit energy would, of course, be drawn off for application to some other circuit. This can be accomplished by connecting the load directly into the plate circuit, or by connecting it inductively to the coil in the plate tank circuit. It will be evident, however, that such a load connection may affect the constants of the oscillating circuit somewhat, with a possible consequent effect on the oscillating frequency. This may be minimized by inserting a "buffer amplifier" between the oscillator and the load. It may also be avoided by the use of the so-called electron-coupled oscillator shown in Figure 28-13. This circuit employs a tetrode in which the screen grid acts as the plate of an oscillating circuit of the Hartley type. The plate circuit couples the oscillating circuit to the load by means of the varying stream of electrons passing through the screen to reach the plate. Changes in the load impedance thus cannot affect the constants of the oscillating circuit itself.

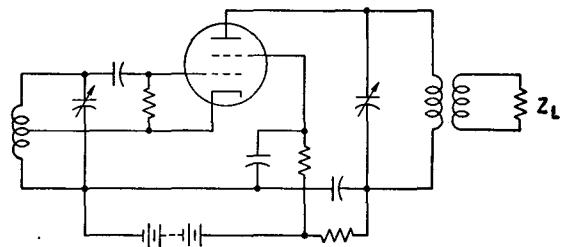


FIG. 28-13 ELECTRON-COUPLED OSCILLATOR

Many applications of oscillators in radio and carrier systems require greater stability of frequency than can be readily obtained with the circuits discussed above. The high degree of stability needed in such cases is usually obtained by employing a piezo-electric crystal in place of the ordinary tank circuit, as indicated in Figure 28-14. As discussed in Chapter 23, such crystals can be cut to precise values of mechanical resonance. When employed in an oscillating circuit as shown, the output frequency will correspond exactly to the resonant frequency of the crystal regardless of variations that may occur in other reactive components of the circuit, or in the characteristics of the tube. If necessary, even greater stability may be secured by enclosing the crystal in a constant temperature oven to preclude any

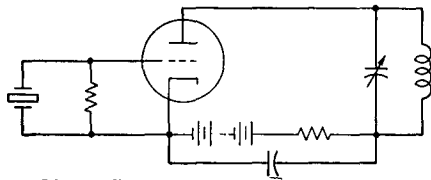


FIG. 28-14 CRYSTAL CONTROLLED OSCILLATOR

changes in the crystal itself that might result from changes in the ambient temperature.

Crystals can be cut to have natural frequencies ranging from a few kilocycles up to about 15 megacycles. Frequencies much higher than this would require making the crystal too thin for practical use. Stable frequencies above this limiting value may be obtained, however, by using **frequency multipliers** in tandem with a crystal source. These consist essentially of vacuum tube amplifiers operated on a non-linear basis so that their output contains substantial harmonics of the fundamental frequency. In a frequency doubler, the second harmonic appearing in the output of the amplifier is selected by an appropriate tuned circuit, while a frequency tripler would select the third harmonic. Because of their relative weakness, higher harmonics than these are not ordinarily used, but any desired multiplying factor can be obtained by employing as many doubling or tripling stages in tandem as may be necessary. The stability of the end frequency remains as great as that of the originating crystal because the multiplying factor is always a fixed integral number.

Alternating voltages at superhigh-frequencies can be generated directly by certain types of electron tubes. In Article 25.4 it was noted that the klystron would act as an oscillator if a coaxial feedback connection was made between the catcher cavity and buncher cavity, with both cavi-

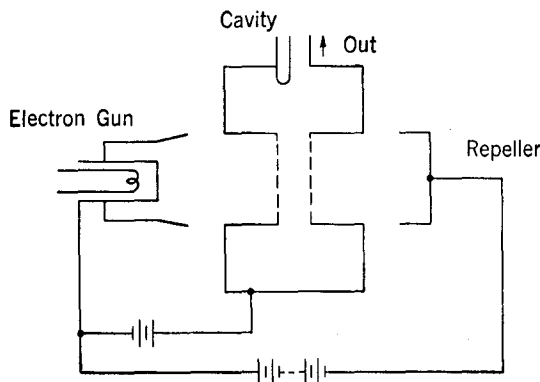
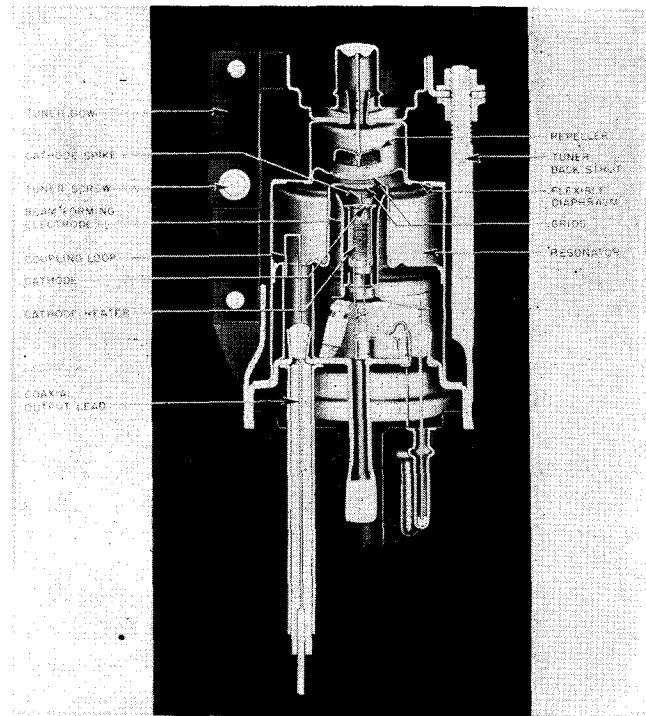


FIG. 28-15 REFLEX KLYSTRON

ties tuned to the same frequency. A somewhat simpler device which will serve the same purpose is the **reflex-klystron** whose basic structure is illustrated in Figure 28-15. This device includes only one pair of grids and one resonant cavity. The cavity acts as both a buncher and catcher because the electrons from the gun pass through its grids twice due to the influence of the "repeller" electrode which carries a high negative charge. As in the ordinary klystron, the electrons proceeding from the gun are velocity modulated in their first passage through the cavity grids. They proceed toward the repeller but are turned back by its negative charge to pass again through the cavity grids in the opposite direction and are finally collected at the accelerating anode of the



REFLEX-KLYSTRON OSCILLATOR TUBE

gun. In the drift interval between their two passages through the cavity grids, the electrons tend to become bunched in accordance with the resonant frequency of the cavity, and when the repeller voltage is appropriately adjusted, the returning electron bunches will pass through the cavity grids in such phase as to give up energy to the cavity.

In addition to its somewhat simpler structure, the reflex-klystron has the advantage over the regular klystron that its output frequency can be



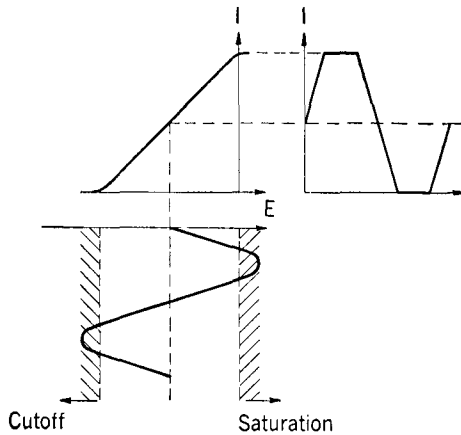


FIGURE 28-16

varied within limits above or below the natural frequency of the cavity by varying the voltage applied to the repeller. This characteristic is extremely useful in certain frequency modulation systems.

### 28.5 Wave-Shaping Circuits

There are numerous situations in modern communications practice where it is necessary to alter the shape of voltage waves under controlled conditions. One such situation frequently encountered is the need to limit or clip the peaks of a sine wave. This requirement may apply to either the positive or negative peaks or to both.

Perhaps the simplest method for limiting or clipping both halves of a wave simultaneously is the use of an over-driven amplifier. If the sine wave voltage applied to the grid of a triode is of sufficient magnitude to drive the grid beyond its cutoff point on the negative swing, and above the saturation point of the plate circuit on the positive swing, the plate current curve will be flattened at its extremities, as illustrated in Figure 28-16. The plate current may thus be made to approach a true square wave as the extent to which the amplifier is over-driven (i.e., as the amplitude of the voltage wave applied to the grid) is increased. Obviously, this same method could be used to limit only the positive or negative pulses by biasing the amplifier grid so that it would be driven above saturation but not below cutoff in one case, and below cutoff but not above saturation in the other case.

In most practical circuits, clipping of the positive half of a wave is accomplished in the grid circuit rather than by driving the plate beyond saturation. If a large resistance is inserted in

series with the grid, the applied positive voltage peak will be able to drive the grid only slightly positive because of the large drop across the resistance when grid current starts to flow. The value of the grid voltage is accordingly held close to zero value during the period when the positive input pulse is attempting to drive it positive, and the plate current will therefore have a constant flat value during this same period.

It will be apparent that a square wave generated by the process outlined above will be of much less amplitude than the sine wave from which it was developed. Generally, therefore, the output square wave will have to be amplified if it is to serve any useful purpose. This requires the use of a very broad band video type amplifier circuit, since a square wave contains an infinite number of frequencies. In practice, the band-width of such amplifiers may range from 2 to 8 megacycles, depending upon the amount of distortion that can be permitted.

There are various other methods of developing waves having approximately square shapes. One interesting circuit that may be used is the **multivibrator** illustrated in Figure 28-17. This consists essentially of a two-stage  $R-C$  coupled amplifier, with the output of each tube fed back to the grid of the other tube in such phase as to encourage oscillation. The net result is that the tubes become alternately conducting and non-conducting at a frequency rate dependent on the constants of the circuit.

The circuit operation may be followed by assuming an instant when the plate current of one tube is increasing slightly due to some random effect. If this occurs in tube B, for example, it will result in a corresponding decrease of its plate potential. Since this plate is coupled by capacitor  $C_1$  to the grid of tube A, the grid voltage of that tube will also be decreased. This will reduce the current across tube A, which will increase the potential at its plate. The plate of tube A is in

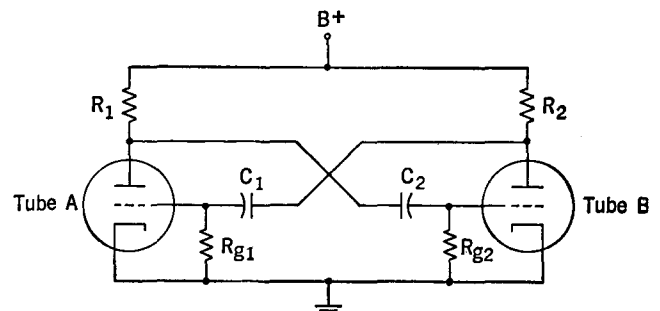
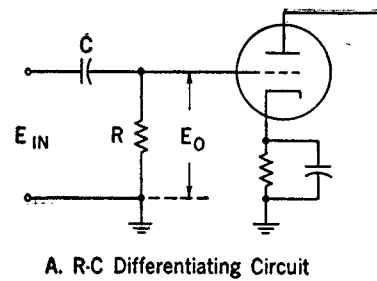


FIG. 28-17 MULTIVIBRATOR CIRCUIT

turn coupled through capacitor  $C_2$  to the grid of tube B so that this grid's potential is correspondingly increased, resulting in a further increase in the current across tube B. This process continues cumulatively until the current across tube B reaches a maximum value, and the grid voltage of tube A falls below the cutoff value. Tube A will then cease to conduct until enough of the charge on capacitor  $C_1$  has leaked off through the resistor  $R_{g1}$  to permit the grid voltage to rise slightly above the cutoff value. The interval during which tube A is non-conducting is of course determined by the  $R_{g1}C_1$  time constant.

As soon as the grid voltage of tube A rises above cutoff, this tube begins to conduct and a cumulative process in the reverse direction proceeds until the current across tube A is at maximum value, and tube B is non-conducting. Although the description of the circuit's action is necessarily somewhat lengthy, the reversing action actually occurs almost instantaneously. The shape of a voltage wave picked off from either of the output resistors  $R_1$  or  $R_2$  is accordingly substantially square. The basic multivibrator circuit discussed is of the so-called "free-running" type, the fundamental frequency of its output wave being determined entirely by its circuit constants. To insure greater stability of operation in practice, an external sine-wave synchronizing voltage of fixed frequency is usually applied to the grid of one of the tubes.

Square waves, as such, have rather limited direct applications in communications systems. They are often used, however, to derive waves of other shapes. Various circuit systems, for example, require uniformly spaced rectangular voltage pulses for their operation. Such pulses may be derived from a square wave by means of a "peaking circuit" known as an **R-C differentiating circuit**. As shown in Figure 28-18(A), this consists of a simple resistor-capacitor combination working into an amplifier. The circuit derives its name from the fact that the output voltage applied to the grid of the tube is proportional to the time rate of change of the input voltage. Assume that a square wave voltage, as indicated in the upper diagram of Figure 28-18(B), is applied to the input. At the instant  $t_0$  the full voltage  $E$  is applied across the  $R$ - $C$  combination. Since it is impossible for the capacitor to charge instantaneously, the full value of the applied voltage will immediately appear across  $R$ , as indicated in the lower diagram of Figure 28-18(B). The capacitor will then charge in accordance with the time con-



A. R-C Differentiating Circuit

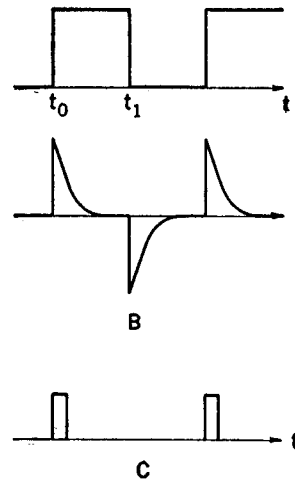
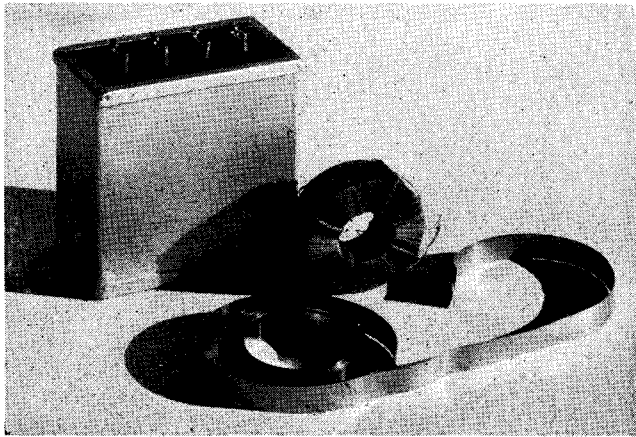


FIGURE 28-18

stant  $RC$ , which is made short relative to the duration of the applied square pulse. As the capacitor charges, the voltage across  $R$  decreases correspondingly until a time equal to about  $5RC$  when the capacitor becomes fully charged and the voltage across the resistor is accordingly reduced to zero. At time  $t_1$  the input voltage suddenly drops to zero and the full voltage of the charged capacitor is impressed across  $R$  in the opposite direction. As the capacitor discharges to ground, the voltage across  $R$  gradually reduces to zero, as shown.

In most applications of peaked waves only pulses of one polarity are desired. In this event, the output wave, indicated in the lower diagram of Figure 28-18(B), may be applied to a clipping circuit that will remove the negative pulses entirely. The sharp tops of the positive peaks may also be clipped as required to leave a series of narrow rectangular pulses, as shown in Figure 28-18(C).

At relatively low frequencies, peaked wave-shapes may be developed by entirely different methods employing devices known as **saturable inductors**. These are small coils wound on a core of permalloy ribbon as illustrated in the accom-



HARMONIC PRODUCING COIL

panying photograph. As indicated by the B-H curve of Figure 28-19, the core material of the coil is of such nature that the magnetic field passes from negative to positive saturation with a comparatively small change in the field intensity. In other words, the coil becomes saturated very quickly and with relatively little current in its windings. This means that the inductance of the coil is high at low current values, but nearly zero as soon as the saturation point is reached. For use in a peaking circuit, the coil is connected at  $L_2$  as shown in Figure 28-20, where the series circuit consisting of  $C_1, L_1$  and  $L_2$  is made approximately resonant.

The behavior of the circuit may be understood by analyzing what happens during a single cycle

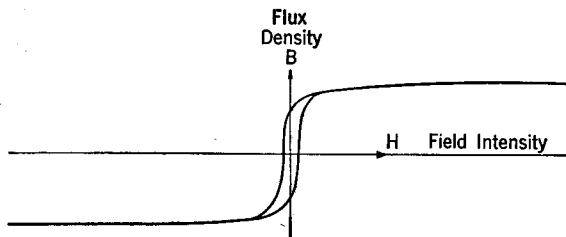


FIG. 28-19 B-H CURVE OF CORE OF SATURABLE INDUCTOR

of an applied sine wave voltage  $E_0$ . This applied voltage is shown in the usual manner in Figure 28-21(A). As this voltage 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 capacitor and the load  $R_2$ . This current is pictured by the small section  $ab$  of the curve of Figure 28-21(B). When the applied voltage increases to the critical value,  $X$ , however, the core of the coil becomes saturated and the in-

ductance 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 capacitor  $C_2$  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 capacitor remains discharged, and no current flows in the load. When the applied voltage reverses in direction, the coil again presents a high inductance to the low values of negative voltage applied and a small negative current,  $de$ , flows into the capacitor and the load. But again, as soon as the coil becomes saturated,

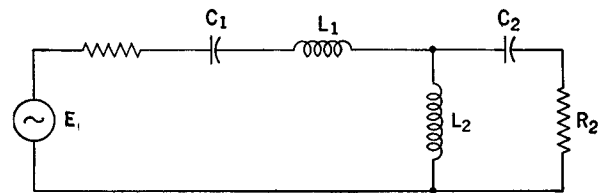


FIGURE 28-20

the capacitor discharges to cause the sharp positive peak of current  $ef$ .

Another wave shape that is used extensively in the operation of cathode-ray oscilloscopes and television cameras and receivers is the so-called saw-tooth wave. This, ideally, is a wave that rises linearly to a specified maximum value in a definite time interval and then returns very rapidly to its

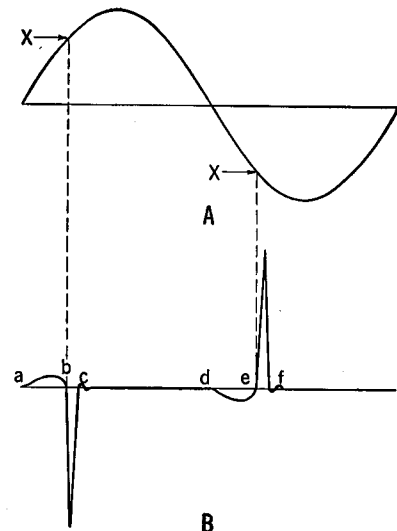


FIGURE 28-21

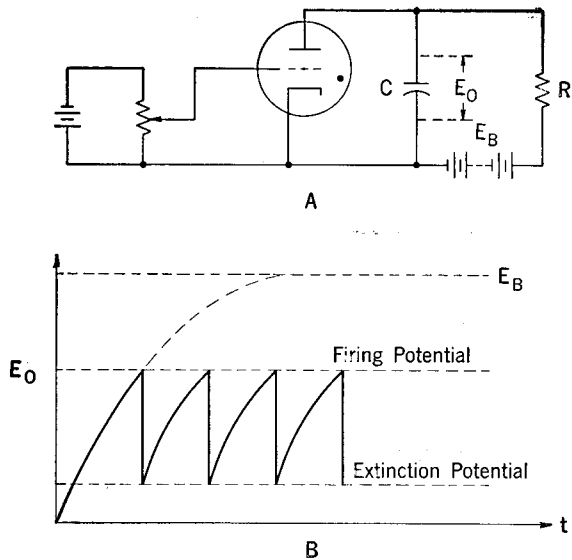


FIG. 28-22 GAS TRIODE SAW-TOOTH GENERATOR

original value. One widely used circuit arrangement for generating such a wave employs a gas triode connected as indicated in Figure 28-22 (A). As the  $B$  battery circuit is closed, the capacitor  $C$  charges through the resistance  $R$  at a rate determined by the  $RC$  time constant. Until the firing potential of the tube is reached, the voltage  $E_o$  across the capacitor will build up in accordance with the solid line curve of Figure 28-22 (B). At the firing potential, the value of which is controlled by the grid bias voltage, the tube suddenly becomes conducting and establishes a low resistance shunt path across the capacitor. The capacitor then discharges quickly through the tube until its voltage drops to the extinction potential when the process repeats itself. It will be seen that the time length of each saw-tooth pulse is determined by the  $RC$  time constant and the firing potential of the tube. The frequency rate can be stabilized by applying to the grid from an external source a small sinusoidal synchronizing voltage having a frequency slightly higher than the natural period of the tube circuit. This causes the grid bias to vary in accordance with the applied synchronizing voltage so that the tube fires at the same specific point in each sine-wave cycle.

It may be noted from Figure 28-22 (B), that the voltage rise across the capacitor  $C$  is not precisely linear since a capacitor charges according to an

exponential curve rather than a straight line. However, if the  $B$  battery value is made large with respect to the tube firing potential so that the tube will break down when the voltage across the capacitor has risen to only about 5% of the  $B$  battery voltage, the rising saw-tooth voltage curve will be sufficiently linear for practical purposes.

Saw-tooth voltage waves may also be developed from a circuit employing a high vacuum triode, to the grid of which is applied a uniform series of negative rectangular pulses as indicated in Figure 28-23. In this arrangement, the pulses applied to the grid have sufficient amplitude to drive the grid well beyond cutoff. During the cutoff period, which is determined by the width of the applied pulse, the capacitor  $C$  charges to a maximum value controlled by the  $RC$  time constant. When the grid bias is reduced to zero by the trailing edge of the pulse, the tube becomes conductive and the capacitor is quickly discharged. The output voltage  $E_o$  accordingly drops to the normal value established by the  $B$  battery at zero grid bias and remains there until the next input pulse is applied to the grid.

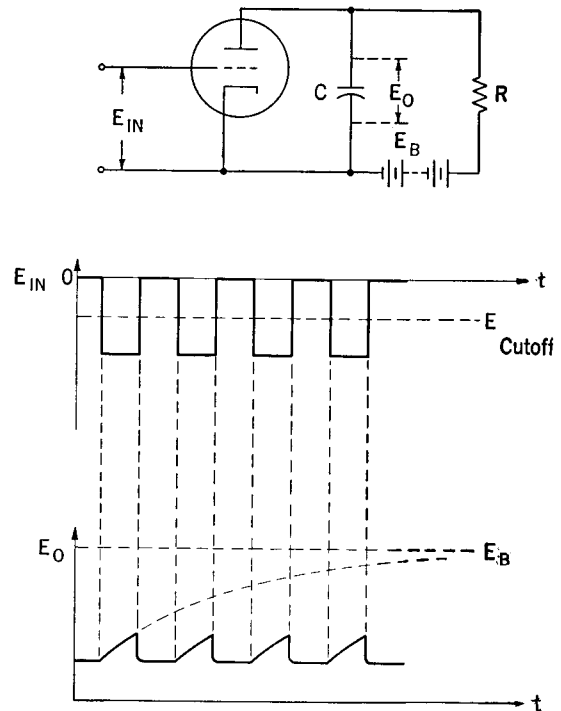


FIG. 28-23 TRIODE SAW-TOOTH GENERATOR

## CHAPTER 29 MODULATION

### 29.1 The Carrier Principle

In radio communication systems, and in the various types of carrier systems, transmission of signals is effected by impressing the signal voltage on a **carrier wave** having a relatively high frequency. The signal is thus transmitted to its destination by electrical waves whose frequencies are normally more nearly comparable in value to the carrier frequency than to the signal fre-

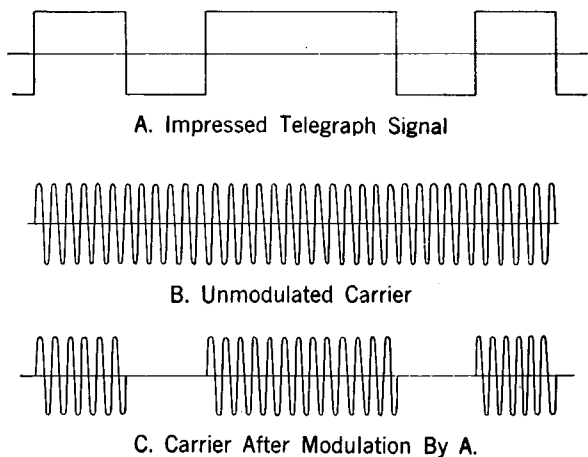


FIG. 29-1 MODULATION IN TELEGRAPH SYSTEMS

quency. The basic reason for this procedure in both carrier and radio systems is to make possible the transmission of a number of different signals over the same transmitting medium without mutual interference by placing each signal in a different portion of the frequency spectrum. In radio transmission, such a procedure is also necessitated by the fact that efficient electromagnetic radiation in space can only be attained at high frequencies. It should be noted, however, that the total width of the transmitted carrier frequency band cannot be less than the sum of the bandwidths of all the signals carried—whether the signals are only a few cycles wide as in telegraph, or millions of cycles wide as in television.

The process of impressing the signal on a carrier is known generally as **modulation**. The inverse process, whereby the signal is retrieved from the modulated carrier-wave, is usually called demodu-

lation in carrier systems and detection in radio systems. There are various methods of modulating carrier waves so that they will effectively transmit signals. The most commonly used of these at the present time are amplitude modulation and frequency modulation.

### 29.2 Amplitude Modulation

In amplitude modulation, the amplitude of the carrier wave is varied in accordance with the variations of the signal wave. The degree of difficulty involved in modulation depends upon the nature of the signal. For a telegraph signal such as that shown in Figure 29-1(A), the method is very simple and consists merely in interrupting the supply of carrier frequency to the line during negative pulses of the telegraph signal and permitting it to flow during positive pulses. The result is a series of "spurts" of current at the frequency of the particular carrier channel, as indicated in Figure 29-1(C).

In telephony, since the variations in voice current are much more complex than telegraph cur-

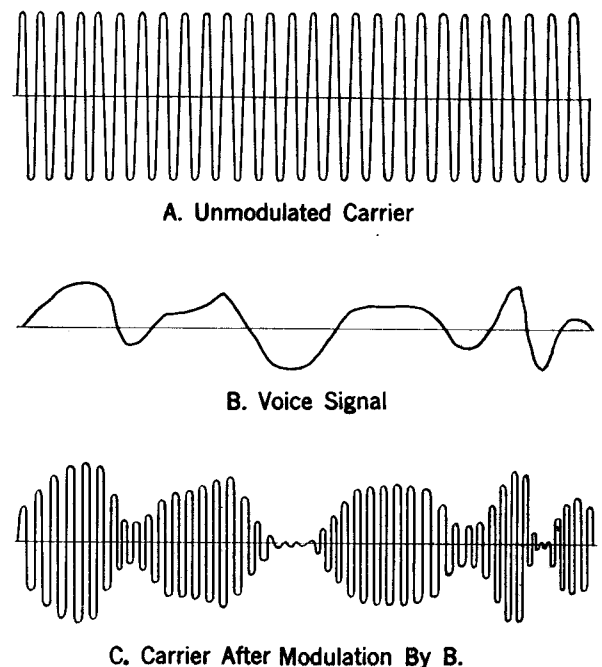


FIG. 29-2 MODULATION IN TELEPHONE SYSTEMS

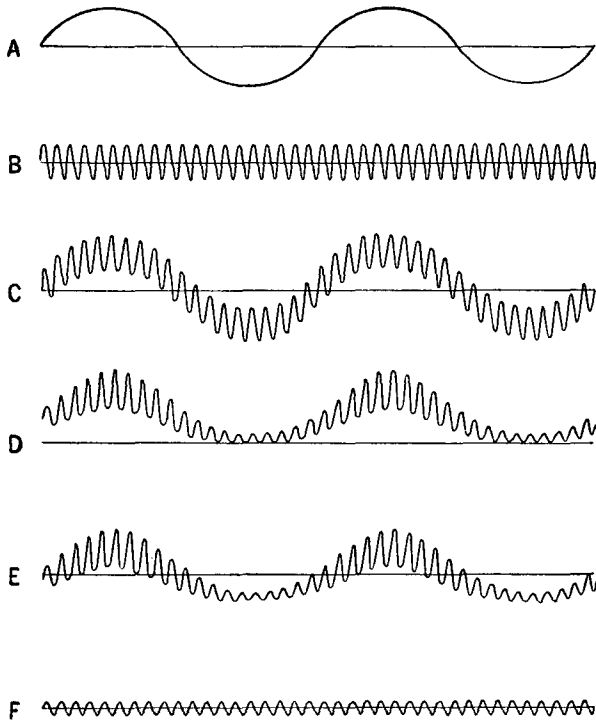
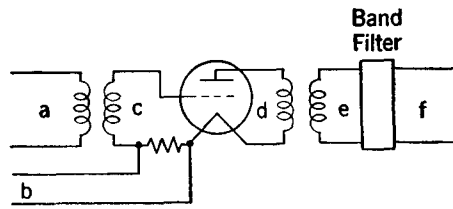


FIG. 29-3 CURRENTS IN MODULATOR CIRCUIT

rent, the process is more involved. This is indicated by Figure 29-2 where *A* is a representation of the unmodulated carrier, *B* is a representative voice signal, and *C* is the modulated carrier. It will be noted that the outline or "envelope" of the modulated wave has the form of the voice signal wave. 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.

The wave of Figure 29-2(C) could be obtained by connecting a transmitter in series with the carrier 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 for this purpose.

Where triodes are used as modulators in carrier telephone systems, the modulating effect is usually obtained by applying a biasing voltage to the grid of the triode of such magnitude that the tube operates on a definitely curved portion of its grid voltage-plate current characteristic. Under these conditions, the amplification supplied by the tube will not be constant but will vary with the value of any alternating voltage applied to the grid.

In the simple circuit of Figure 29-3, 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 the voice voltage is here assumed to be sinusoidal in form although this, of course, is not generally the case. The two voltages, being in series, add 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 29-4, 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 29-3(D). After passing through the output transformer, the current curve will be as pic-

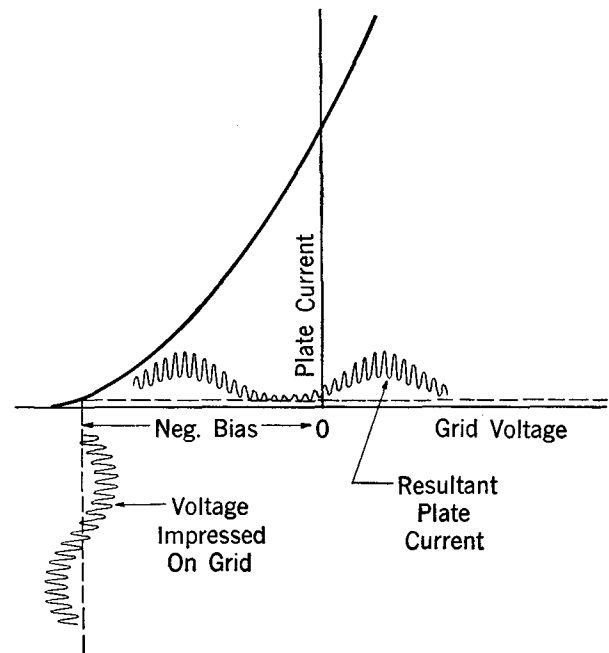


FIG. 29-4 VACUUM TUBE AS MODULATOR

tured in Figure 29-3(E). Analysis of this curve shows the principle frequencies present in terms of the voice and carrier frequencies to be:

- $V$ —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.

These same results may be reached mathematically 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:

$$i_b = K(E_b + \mu E_c + \mu e)^2 \quad (29:1)$$

where—

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

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,

$$i_b = K(E_b + \mu E_c)^2 + a_1 e + a_2 e^2 \quad (29:2)$$

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

and  $a_2 = K \mu^2$ .

Since the impressed voice and carrier currents represented in Figures 29-3(A) and (B) are both sinusoidal in form they 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—

$$e = A \sin Vt + B \sin Ct. \quad (29:3)$$

Substituting (29:3) in equation (29:2), 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—

$$i_b = K(E_b + \mu E_c)^2 + a_1 A \sin Vt + a_1 B \sin Ct + a_2 A^2 \sin^2 Vt + 2a_2 AB \sin Ct \sin Vt + a_2 B^2 \sin^2 Ct. \quad (29:4)$$

Making use of the trigonometric relationships—

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

and

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

we may expand further to obtain—

$$i_b = K(E_b + \mu E_c)^2 + a_1 A \sin Vt + a_1 B \sin Ct + \frac{1}{2} a_2 A^2 - \frac{1}{2} a_2 A^2 \cos 2Vt + a_2 AB \cos (C - V)t - a_2 AB \cos (C + V)t + \frac{1}{2} a_2 B^2 - \frac{1}{2} a_2 B^2 \cos 2Ct = K(E_b + \mu E_c)^2 + \frac{1}{2} a_2 (A^2 + B^2) + a_1 A \sin Vt + a_1 B \sin Ct - \frac{1}{2} a_2 A^2 \cos 2Vt - \frac{1}{2} a_2 B^2 \cos 2Ct + a_2 AB \cos (C - V)t - a_2 AB \cos (C + V)t. \quad (29:5)$$

The first and second terms of this equation are 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 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 (29:5) 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 signal current to the receiving end of the circuit. In carrier telephone practice, it is customary in many cases to suppress by means of filters or otherwise, all of the frequencies in the output of the modulator except one side-band for transmission over the line. Thus in Figure 29-3(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 relative 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 (29:5). 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 carrier systems employing vacuum tube modulator circuits, 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. Such a modulator circuit is shown schematically in Figure 29-5. 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.

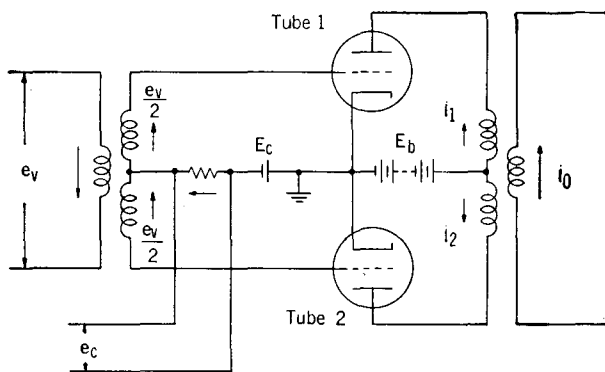


FIG. 29-5 BALANCED TUBE MODULATOR CIRCUIT

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 with respect to ground 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:

$$e_1 = e_c + \frac{e_v}{2}$$

$$e_2 = e_c - \frac{e_v}{2}$$

These values of input voltage may be substituted in equation (29:2) 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, their difference determines the value of the current in the secondary. Since no direct current passes through the transformer, the secondary current may accordingly be written as—

$$\begin{aligned} i_0 &= i_1 - i_2 = a_1 e_c + \frac{a_1}{2} e_v + a_2 e_c^2 + a_2 e_c e_v + \frac{a_2}{4} e_v^2 \\ &\quad - a_1 e_c + \frac{a_1}{2} e_v - a_2 e_c^2 + a_2 e_c e_v - \frac{a_2}{4} e_v^2 \\ &= a_1 e_v + 2a_2 e_c e_v \end{aligned} \quad (29:6)$$

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

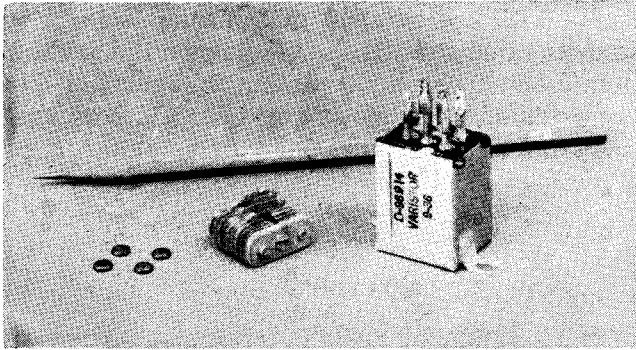
$$i_0 = a_1 A \sin Vt + 2a_2 A B \sin Ct \sin Vt \quad (29:7)$$

By the second of the trigonometric relationships cited above, equation (29:7) may be converted to read—

$$\begin{aligned} i_0 &= a_1 A \sin Vt + a_2 A B \cos (C - V)t \\ &\quad - a_2 A B \cos (C + V)t \end{aligned} \quad (29:8)$$

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 car-





COPPER-OXIDE VARISTOR USED AS MODULATOR AND DEMODULATOR

rier voltage much larger than that of the applied voice voltage.

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, a single frequency has been assumed for the applied signal voltage in all cases. Actually, a voice signal normally includes several different frequencies which are 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 (29:5) and (29:8) 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.

Demodulation in carrier systems is effected by circuits that are essentially the same as the circuits used for modulation and operate on the same principle. This is illustrated by Figure 29-6. Here carrier-frequency  $B$ , identical in frequency to the carrier frequency employed at the sending end of the line, adds to the incoming 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 im-

pressed voltage then is equal to  $(C + V)$  plus  $C$ . If these values are substituted for  $e$  in equation (29:2) and expanded, the resultant output currents will be found to be—

$C$ —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 29-6 (D); and Figures 29-6(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 cir-

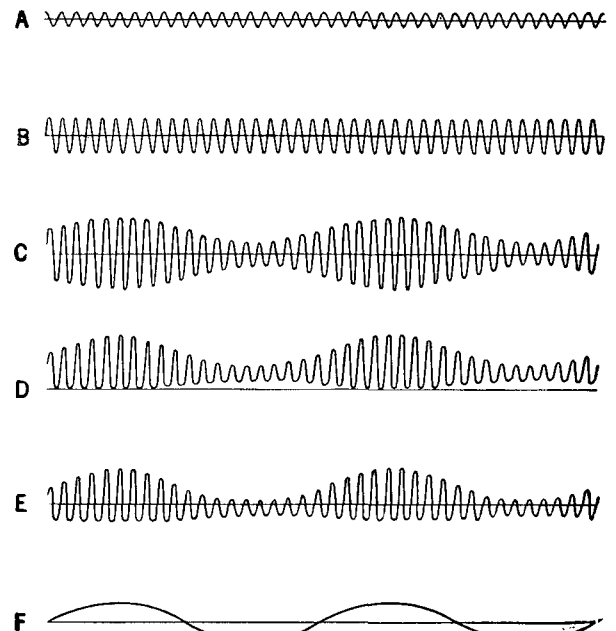
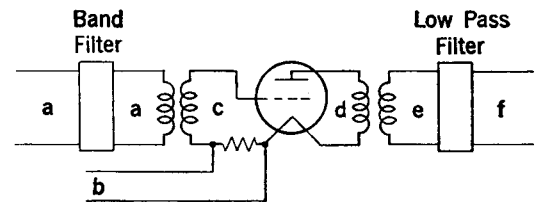


FIG. 29-6 CURRENTS IN DEMODULATOR CIRCUIT

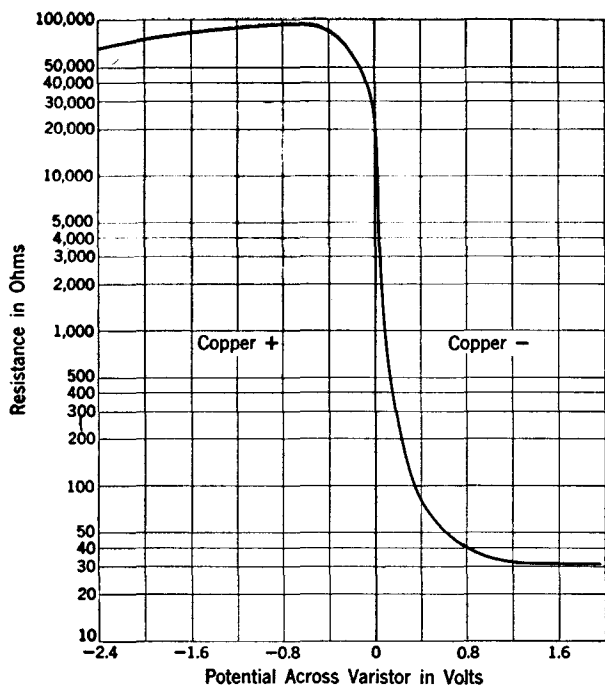


FIG. 29-7 RESISTANCE CHARACTERISTIC OF COPPER-OXIDE VARISTOR

circuit employed as a demodulator, if the upper sideband ( $C + V$ ) is applied, the output frequencies will be  $V$ ,  $C + V$ , and  $2C + V$ . This will be apparent from equation (29:8) where  $(C + V)$  may be substituted for  $V$  to determine the demodulation products.

Many types of carrier telephone systems employ copper-oxide varistors instead of vacuum tubes in their modulating and demodulating circuits. These devices are capable of accomplishing essentially the same results.

The principle of the copper-oxide varistor as a rectifier for converting alternating to direct currents in power supply circuits was discussed briefly in Article 7.5. Its essential characteristic for the present purpose is that, as shown in Figure 29-7, 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 systems, four of these tiny copper-oxide units are mounted in a sealed container having a maximum dimension of less than one inch. The char-

acteristics of such units are very stable and their useful life is apparently indefinite under normal operation.

In the channel modulator and demodulator circuits of most carrier systems, the varistor units are connected in the Wheatstone bridge arrangement illustrated in Figure 29-8. (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 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, the network of varistors will act to virtually short-circuit the line during the positive halves of the carrier voltage

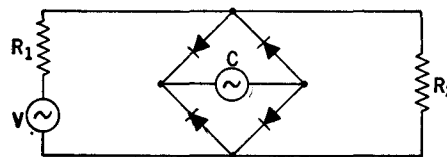


FIG. 29-8 BALANCED BRIDGE MODULATOR CIRCUIT

cycle; and to present an open circuit across the line during the negative halves of the carrier voltage cycles. This is illustrated by the two diagrams of Figure 29-9 where the varistors are 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. The varistors thus act effectively like a

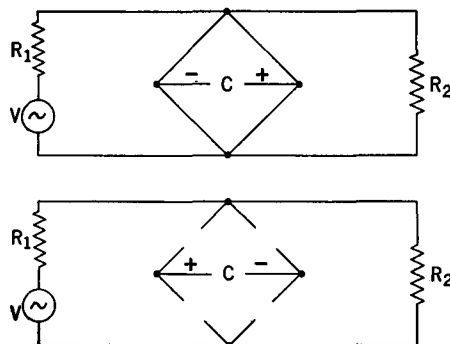


FIG. 29-9 OPERATING PRINCIPLE OF CIRCUIT OF FIG. 29-8

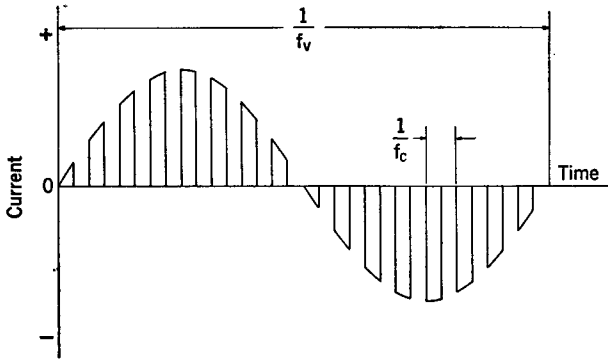


FIG. 29-10 OUTPUT CURRENT OF BALANCED BRIDGE MODULATOR

switch, opening and closing at the frequency of the carrier voltage. The resultant output current is shown in Figure 29-10.

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  $2\pi$  times the signal frequency, an approximate equation for the output current represented by Figure 29-10 may be written as follows:

$$I = \frac{A \sin Vt}{2(R_1 + R_2)} + \frac{2A}{\pi(R_1 + R_2)} [\sin Vt \sin Ct + \frac{1}{3} \sin Vt \sin 3Ct + \frac{1}{5} \sin Vt \sin 5Ct + \dots] \quad (29:9)$$

Here  $R_1$  and  $R_2$  are respectively the input and output resistances as indicated in Figure 29-8 and  $C$  is  $2\pi$  times the carrier frequency.

Making use of the trigonometric relationship—

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

the above equation may be rewritten as—

$$I = \frac{A \sin Vt}{2(R_1 + R_2)} + \frac{A}{\pi(R_1 + R_2)} [\cos (C - V)t - \cos (C + V)t + \frac{1}{3} \cos (3C - V)t - \frac{1}{3} \cos (3C + V)t + \frac{1}{5} \cos (5C - V)t - \frac{1}{5} \cos (5C + V)t + \dots] \quad (29:10)$$

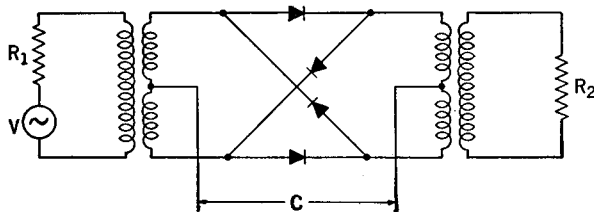


FIG. 29-11 LATTICE MODULATOR CIRCUIT

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, only one of the side-bands of the carrier frequency is made use of in most cases 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 29-8, is used at the receiving end of the carrier line to restore the original signal frequency. In this case,

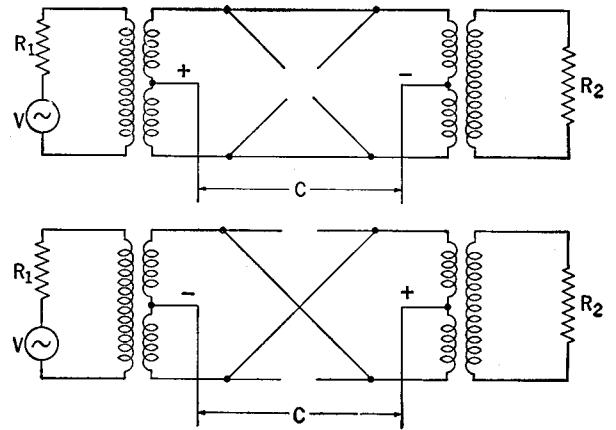


FIG. 29-12 OPERATING PRINCIPLE OF CIRCUIT OF FIG. 29-11

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 (29:9) in place of  $A \sin Vt$ , the first term inside the brackets in equation (29:10) 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

broad-band carrier systems, a somewhat different arrangement of the varistor units is frequently employed. This is illustrated in Figure 29-11. It is also a balanced bridge arrangement but the circuit connections and the configuration of the varistors are such that, as indicated in Figure 29-12, 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 at the carrier frequency and results, in the ideal case, in the output current wave shown in Figure 29-13.

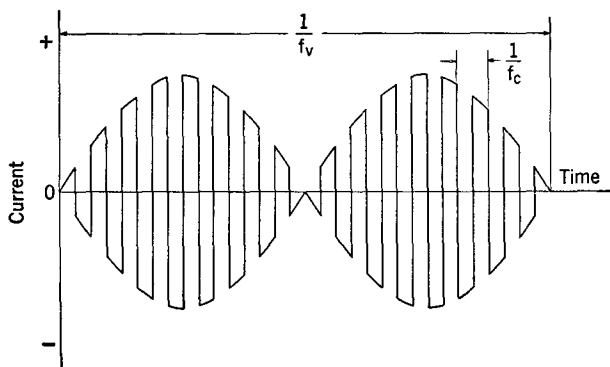


FIG. 29-13 OUTPUT CURRENT OF LATTICE MODULATOR

Using the same terminology as in the preceding discussion, the approximate equation for the curve of Figure 29-13 is—

$$I = \frac{2A}{\pi(R_1 + R_2)} [\cos(C - V)t - \cos(C + V)t + \frac{1}{3} \cos(3C - V)t - \frac{1}{3} \cos(3C + V)t + \frac{1}{5} \cos(5C - V)t - \frac{1}{5} \cos(5C + V)t + \dots] \quad (29:11)$$

Comparing this equation with (29:10), 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 arrange-

ment 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 29-7, 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 (29:10) and (29:11), 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 multiple or series-multiple.

Modulation in radio systems ordinarily involves substantially larger amounts of power than the carrier systems discussed above. Considerations of power efficiency, therefore, are of much greater importance. Also, because of the much larger power requirements, vacuum tubes must always be used in the modulating circuits since any practical varistor arrangement would be inadequate to handle the required power. Power efficiency, in this case, also demands that relatively more attention be given to the percentage or extent of the modulation. This is measured by the ratio of the peak signal voltage to the peak carrier voltage, which is known as the **modulation factor**. This factor normally ranges in value from 0 to 1. At zero no modulation exists and the modulator out-

put is simply the carrier wave. A modulation factor of 1 corresponds to 100% modulation, meaning that the modulated wave varies in amplitude at the peaks from twice the amplitude of the carrier wave to zero amplitude. In other words, the maximum signal amplitude is equal to the carrier amplitude. Greater modulation than this, a condition known as over-modulation, would of course indicate a modulation factor greater than 1 and would result in distortion because the modulated wave would not be continuous.

The total power of an amplitude-modulated carrier wave is equal to that of the carrier plus that of the two side-bands. It can be shown that the power in the side-bands is proportional to the square of the modulation factor, and with 100% modulation, the total power of the modulated wave is 50% greater than that of the carrier wave; i.e., one third of the total output power is contained in the side-bands. With lesser modulation, the side-band power drops off rapidly. Since it is only the power in the side-bands that carries the signal, it is clear that for efficient transmission, the modulation ratio should be held as close to 100% as possible.

The general type of vacuum tube modulator that was discussed earlier in connection with carrier systems, in which the carrier and signal wave were applied to the grid circuit, is sometimes used in relatively low power radio transmitters. In radio terminology, such a circuit would usually be known as a grid-modulated Class C amplifier. More generally used in radio work is the plate-modulated Class C amplifier illustrated in Figure 29-14. In this circuit, the carrier voltage is applied to the grid as shown. When no signal voltage is applied, the plate current consists of a series of positive pulses, as indicated in (A) of Figure 29-15, because of the large negative grid bias. When a signal voltage is applied to the plate circuit in series with the  $B$  battery as shown, the total plate voltage varies above and below the fixed  $B$  battery voltage in accordance with the signal. This is illustrated in (B) of Figure 29-15. The varying plate voltage causes the plate current

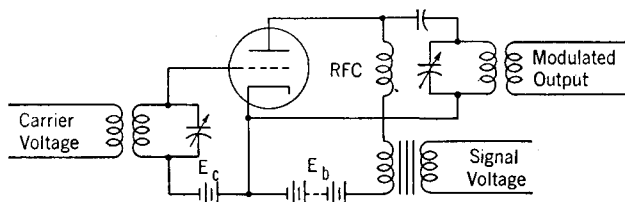


FIG. 29-14 PLATE-MODULATED AMPLIFIER

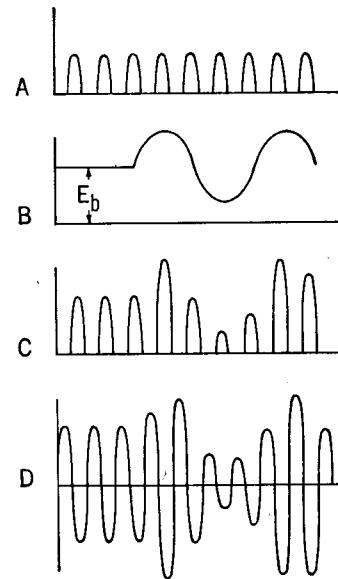


FIGURE 29-15

to vary as indicated in (C) of Figure 29-15, where the envelope of the wave has the form of the signal. The current pulses produce an oscillatory current flow in the tank circuit, which results in the output voltage wave shown at (D).

Examination of Figure 29-15 will show that for 100% modulation (modulation factor of 1), the maximum amplitude of the applied signal voltage must be equal to the value of the d-c plate supply voltage. This means that 50% of the total r-f output power must be furnished by the signal source at 100% modulation. The requirement that this large portion of the power must come from the modulating source has certain disadvantages in comparison with a grid-modulated transmitter in which all the output power is furnished by the d-c plate-supply.

Demodulation in radio systems involves the same principles as were outlined above in connection with carrier systems. But the particular devices used in this case are generally known as detectors. Such devices were discussed in Chapter 28.

### 29.3 Frequency Modulation

The use of frequency modulation is confined entirely to radio systems operating in the very high frequency band or above, where it has certain definite advantages over amplitude modulation in minimizing interference from "static" and extraneous signals. It depends upon varying the frequency of a carrier wave of fixed amplitude above

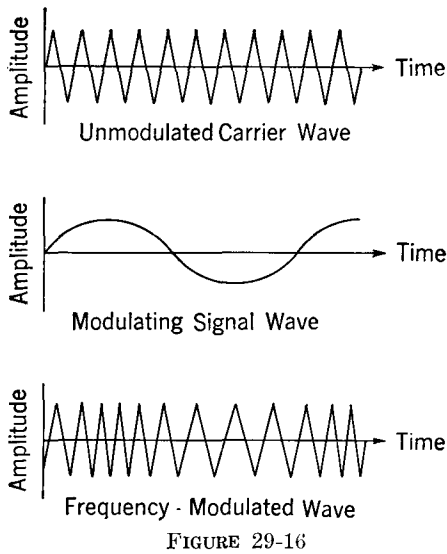


FIGURE 29-16

and below a central or normal frequency in accordance with the amplitude variations of an applied signal voltage. The process is roughly illustrated by the three wave diagrams of Figure 29-16. The amount of frequency change that is produced by the signal is called the **frequency deviation** and, ideally, this should be as high as possible in order to obtain the maximum signal to noise ratio. However, since it is obvious that the total bandwidth of the modulated wave to be transmitted will increase with increases in the maximum frequency deviations on both sides of the unmodulated carrier frequency, it is necessary as a practical matter to arbitrarily limit the maximum permissible deviations to values that will keep the total bandwidth that must be assigned in the radio spectrum to each FM channel as narrow as feasible. The maximum permissible deviation has been specified by the Federal Communications Commission at 75 kc for FM broadcasting, and at 15 kc for such applications as mobile radio service.

As in amplitude modulation, frequency modulation results in a modulated wave containing the carrier frequency and other frequencies above and below the carrier frequency. If, for purposes of analysis, we assume a sinusoidal signal wave of frequency  $f_s$  impressed on a carrier of frequency  $f_c$ , it may be shown that the instantaneous value of the modulated resultant wave may be written as—

$$e = E_m \sin \left( \omega_c t + \frac{\Delta f}{f_s} \sin \omega_s t \right) \quad (29:12)$$

where  $E_m$  represents the constant voltage ampli-

tude of the carrier wave,  $\Delta f$  the frequency deviation, and  $\omega_c$  and  $\omega_s$  are  $2\pi f_c$  and  $2\pi f_s$  respectively. This equation can be expanded into the following infinite series—

$$e = E_m \{ A_0 \sin \omega_c t + A_1 [\sin (\omega_c + \omega_s)t - \sin (\omega_c - \omega_s)t] + A_2 [\sin (\omega_c + 2\omega_s)t + \sin (\omega_c - 2\omega_s)t] + \dots \} \quad (29:13)$$

This equation shows that in addition to the carrier frequency itself, the modulated wave includes an infinite series of side frequencies having values equal to the carrier frequency plus and minus the signal frequency and all of its integral multiples. The relative magnitude of the carrier frequency and the side frequencies depends upon the ratio of the frequency deviation to the signal frequency. This ratio is commonly designated  $m_f$ , and is known as the **modulation index**.

In general, the number of side frequency pairs that must be transmitted to carry a signal with reasonable fidelity increases with the value of the modulation index. While the number of side-pairs is theoretically infinite, only side frequencies containing 1% or more of the total energy are considered significant in practice. This limits the total number of frequencies necessary for satisfactory transmission within reasonable bounds. For example, Figure 29-17 shows the relative amplitudes and positions in the spectrum of a representative FM wave where an 80 mc carrier is modulated with a 4 kc signal when the maximum frequency deviation is 40 kc. It will be noted that the amplitudes of the carrier and side frequencies vary according to an irregular pattern, but tend to decrease in value rapidly at the extremities. The values of these amplitudes of course correspond to the constants  $A_0, A_1, A_2 \dots$  of equation (29:13). Their calculation involves the use

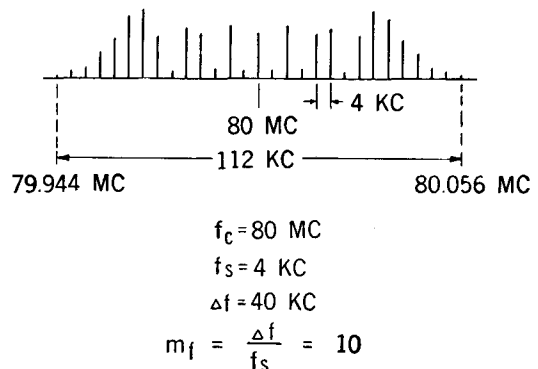


FIG. 29-17 INSTANTANEOUS FREQUENCY COMPONENTS OF F-M WAVE

of certain complex mathematical functions, but tables are to be found in the handbooks showing their values for various values of modulation index,  $m_f$ . The number of significant side frequencies varies inversely with the value of the modulation index.

It will be observed that the frequency deviation in this example is only 40 kc, although a maximum deviation of 75 kc is permitted. This is the more usual situation because the extent of deviation is determined by the signal amplitude and will reach maximum values only at peak signal strength. The total bandwidth required by the frequency-modulated carrier at any time thus depends both upon the intensity of the applied modulating signal and its frequency, since  $m_f$  is equal to  $\Delta f/f_s$ . In the example, the total swing is shown as 112 kc, which, it will be seen, is appreciably more than twice the deviation of 40 kc. At higher values of  $m_f$ , however, the total required bandwidth approaches closer and closer to  $2\Delta f$ .

At lower frequencies in the voice range, the modulation index will be high and the number of significant side frequencies will be correspondingly large. But the separation of these frequencies in the spectrum is equal to the signal frequency so that the required bandwidth is generally no greater than 150 kc, twice the maximum permissible deviation. At an extreme high voice-frequency—say 15 kc—a maximum-intensity signal would result in a modulation index of 75/15 or 5. In this case, the theoretical required bandwidth would be 240 kc. This is an extremely unlikely situation, however, because the amplitude of any normal high-frequency signal is substantially less than that of the low-frequency components, and maximum frequency deviation at 15 kc would not be expected to occur. Practice has shown that a total bandwidth of 200 kc is suffi-

cient for high fidelity transmission of audio signals, and this is the bandwidth designated by the F.C.C. for FM radio broadcasting channels. Where the permissible deviation is less than 75 kc, as in mobile radio service, the permissible channel bandwidth is of course reduced proportionately.

Various circuits have been devised for modulating the frequency of a carrier wave in accordance with the amplitude variations of a signal voltage. One basic arrangement for accomplishing this purpose makes use of what is known as a "reactance tube" circuit. The principle is illustrated schematically in Figure 29-18. The oscillator circuit at the right is of the Hartley type, and its output frequency, when the circuit is open at points 1 and 2, is determined by the constants  $L_1$  and  $C_1$  of the tuned tank circuit. When connected to the reactance tube circuit, the oscillator circuit presents a purely resistive load to the output of the reactance tube circuit at the operating frequency. The reactance tube is bridged across points 1 and 2 as is also the voltage divider,  $C-R$ . The total impedance of  $C-R$  is high, and the value of  $C$  is so selected that its reactance in the carrier frequency range is much higher than the resistance of  $R$ . The oscillator sets up a small r-f current through  $C-R$ , which, because of the dominating capacitive reactance, will lead the voltage across 1 and 2 by almost  $90^\circ$ . The voltage across  $R$ , which is applied to the grid of the reactance tube, will also lead  $E_{1,2}$  by the same angle. Since the plate current of a vacuum tube is in phase with the grid voltage, the output current of the reactance tube must also lead  $E_{1,2}$  by  $90^\circ$ . The reactance tube circuit accordingly acts effectively like a pure capacitive reactance connected across points 1,2. Since this reactance is in parallel with the capacitance  $C_1$  of the oscillator tank circuit, it will affect the frequency of the oscillator. The magnitude of the bridged capacitive reactance is determined by the value of the plate current in the reactance tube, which is in turn controlled by the value of the voltage applied to its grid. When the signal voltage is connected across  $R$  as shown, therefore, the net reactive capacitance of the reactance tube circuit will vary as the signal voltage amplitude varies, and the frequency of the oscillator will vary accordingly.

Because it is controlled by simple inductive and capacitive elements, the frequency of the oscillator of Figure 29-18 would ordinarily not be sufficiently stable to meet the exacting requirements of radio transmission. This difficulty is overcome

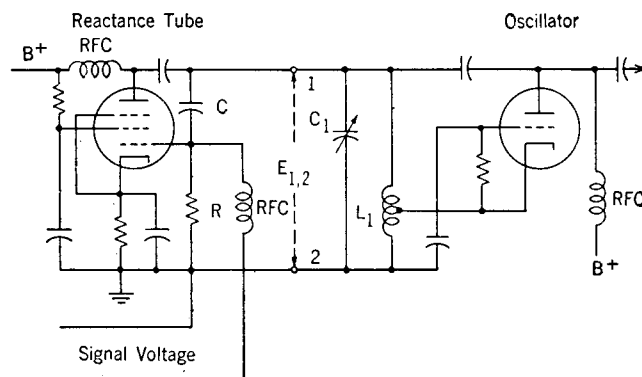


FIG. 29-18 FREQUENCY MODULATION WITH REACTANCE TUBE CIRCUIT

in one type of FM transmitter by means of an automatic frequency control circuit employing a discriminator and a fixed crystal-controlled auxiliary oscillator. In the arrangement most widely used in FM broadcasting transmitters, the basic oscillator frequency is controlled by a crystal. This makes it impossible for the signal voltage to cause the oscillator frequency to vary. It can, however, cause variations in the phase angle of the oscillator output corresponding to the amplitude variations of the signal. By the use of appropriate phase-shifting networks, the "phase-modulated" output is then translated into an equivalent frequency modulation.

In practice, the basic oscillator is operated at a much lower frequency than the actual carrier frequency radiated from the transmitting antenna, but is brought up to the desired value by several stages of frequency multiplication. The multiplication process naturally increases the deviation frequencies in the same way that it increases the mean carrier frequency.

#### 29.4 Pulse Modulation

Although applications have been limited in telephone practice up to this time, it is possible to transmit telephone signals by means of a series of pulses of energy. The conversion of the telephone signal into energy pulses from which it can be reproduced at the receiving end is known as **pulse-modulation**. The method depends on the fact that any signal waveshape can be completely defined by taking very short (essentially instantaneous) samples of its amplitude at regularly spaced successive time intervals, such that at least two samplings are taken for each cycle of the highest frequency included in the signal wave. For example, if a voice-wave has been passed through a low-pass filter having a cutoff frequency of 4000 cycles, all of the information required for its reproduction will be contained in a regular series of samples taken at the rate of 8000 per second.

The transmission of information by means of discrete energy pulses has certain definite advantages in radio systems particularly. It permits the utilization of amplitude limiting techniques by means of which noise and other interference, including fading, may be minimized. It also tends to reduce power consumption at the radio transmitter because the energy output is not continuous. Moreover, it makes possible the transmission of a number of separate signals over a single car-

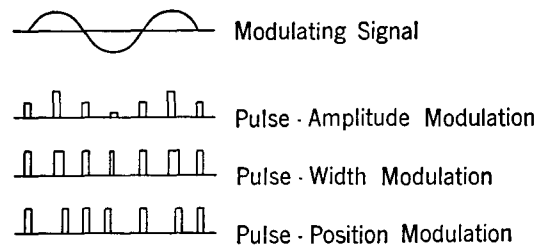


FIG. 29-19 PULSE MODULATION METHODS

rier by means of **time-division multiplex**. That is, the intervals between the successive pulses of a given signal can be employed to transmit comparable sampling pulses of other signals.

There are several possible ways in which the signal sampling information can be incorporated in the energy pulses. Three of these are illustrated in Figure 29-19. In pulse-amplitude modulation, the sampling pulses are of uniform time duration, but vary in amplitude in accordance with the signal amplitude at the sampling instants. In pulse-width modulation and pulse-position modulation, the pulses are of constant amplitude but vary with the signal amplitude in duration or width in the first case, and in position with respect to a uniform time scale in the second case. Because of the constant amplitude, the latter two types of pulse modulation have the advantage that limiting circuits can be used at the receiving end to practically eliminate noise, and the amplifiers in the system do not need to have linear characteristics.

It may be evident that transmission of information by pulse systems has much in common with ordinary telegraph systems, except that the repetition rate of the pulses is very much higher. When modulated pulses of constant amplitude are superimposed on a high frequency carrier by ordinary amplitude methods, the transmitted wave consists merely of a series of spurts of the carrier frequency just as in simple carrier telegraph systems. Transmission of amplitude-modulated pulses would of course involve variations in the amplitudes of the carrier frequency spurts. This technique may nevertheless have its advantages by permitting the use of simpler and less costly transmitting and receiving apparatus in certain cases.

In a further elaboration of the pulse modulation technique that has recently been developed by Bell Telephone Laboratories, information regarding the amplitude of the signal samples is coded into a rapid sequence of pulses of uniform amplitude similar to the "off or on" code pulses that



represent letters in teletypewriter systems. This technique is called **pulse-code modulation**. The amplitude of a signal may of course vary over a continuous range and may thus have an infinite number of values. To keep the total number of codes within reasonable limits, it is accordingly necessary for this purpose to divide the signal amplitudes into a limited number of finite units. For example, one code might represent any value of signal amplitude between  $-0.5$  and  $+0.5$ , and so on. This technique is called "quantizing" and its use necessarily results in some "granularity" in the final transmitted signal. It has been found, however, that by making the quantizing steps vary logarithmically and by using as many as 64 codes, speech signals may be reproduced with a high degree of fidelity. This number of separate codes requires seven positions per code as distinguished from the five positions of teletypewriter operation where only 32 different codes are necessary.

Although the basic principles involved are very similar to those applying to teletypewriter systems, it will be clear that the time factors are widely different. The extremely rapid generation of the unit code signals of the PCM system is accomplished entirely by electronic rather than mechanical methods. A simple arithmetical calculation will show that even the transmission of a single voice signal by this method requires a very broad frequency band. On the other hand, pulse modulation methods may have distinct advantages in eliminating noise interference. Moreover, coded pulse signals may be perfectly regenerated at repeater points, in a manner generally similar to the way in which teletypewriter signals are completely restored with regenerative repeaters. These advantages may well lead to the useful application of pulse modulation systems for radio transmission in the superhigh-frequency ranges where extremely broad frequency bands are available.

## CHAPTER 30

### LONG DISTANCE TRANSMISSION SYSTEMS VOICE-FREQUENCY TELEPHONE CIRCUITS

#### 30.1 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 most 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. If they are of 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 coaxial conductors, carrier circuits employ the same or similar kinds of wire facilities for transmission as do voice-frequency circuits. They also require the use of amplifiers or repeaters at regular intervals along the line. This latter statement also applies to superhigh-frequency overland radio circuits where relay stations are necessary at regularly spaced intervals along the route.

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, 2-wire circuits are commonly used for relatively short distances—not more than a few hundred miles maximum. Four-wire cable circuits are used for somewhat longer distances when equipped with “echo suppressors”. Carrier or radio circuits are generally used for the longest distances, although their use is not limited to such application.

In this and following Chapters, some of the more essential problems involved in the design and operation of long voice-frequency, carrier, and radio circuits are considered.

#### 30.2 Repeater Spacing

The different types of circuit facilities, whose principal characteristics are discussed in Chapter 21, 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 helped to make 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 capacitance 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 wire conductors are usually of 16 or 19-gage and even though loading is used, repeaters must be inserted at 50 to 60-mile intervals. 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, 165 wire will generally have the best electrical and mechanical characteristics, with 128 wire circuits next, and 104 wire circuits last. By increasing the repeater spacing when larger wires are used, these three types of circuits can 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 strength 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 usually been found that 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.

Long toll cables usually employ 19 and 16-gage conductors for voice-frequency transmission. The former gage is used for both 2 and 4-wire circuits, while the latter is frequently 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 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 and buried 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.

### 30.3 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 practicable level in order to minimize 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. This can be overcome by keeping the transmission level of the message currents high with respect to the level of the induced noise currents. However, it should not be forgotten that there is a limit to the amount of energy 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 may ordinarily be operated to deliver a volume of transmission not exceeding the zero level by more than 6 db, and with a maximum net gain of not more than 18 db. The output of V-type repeaters presented to the input of the hybrid coil (which in this case is not a part of the repeater) may be as high as 10 db above zero level and the total gain of the repeater may be permitted to approach 25 db.

In repeaters used in 4-wire cable circuits, much higher gains are possible. However, crosstalk considerations usually prevent the attainment in practice of the maximum possible gain. It is permissible to operate these 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 an incoming cable pair is adjacent to an outgoing cable pair of another 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 30-1 gives such a diagram for a typical 2-wire cable circuit. 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 and equipment 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 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.

It will be noted that in the case of the 22-type repeaters, a single vertical line is used to represent the net gain, which is made up of the difference between the gain of the amplifier itself and the losses of the hybrid coils that are part of the repeater. In the case of the V-type repeater, on the other hand, the hybrid coil losses are indicated separately from the amplifier gain. This is desirable as a practical matter because the amplifiers of the V-type repeaters and their gain adjusting controls are separated physically from the hybrid coils.

### 30.4 Return Loss and Singing Points

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 2-wire 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 pointed out in Article 17.7, 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. The total loss across the hybrid between the series winding and the bridge taps, which may be designated  $L$ , is made up of the normal hybrid losses caused by the divi-

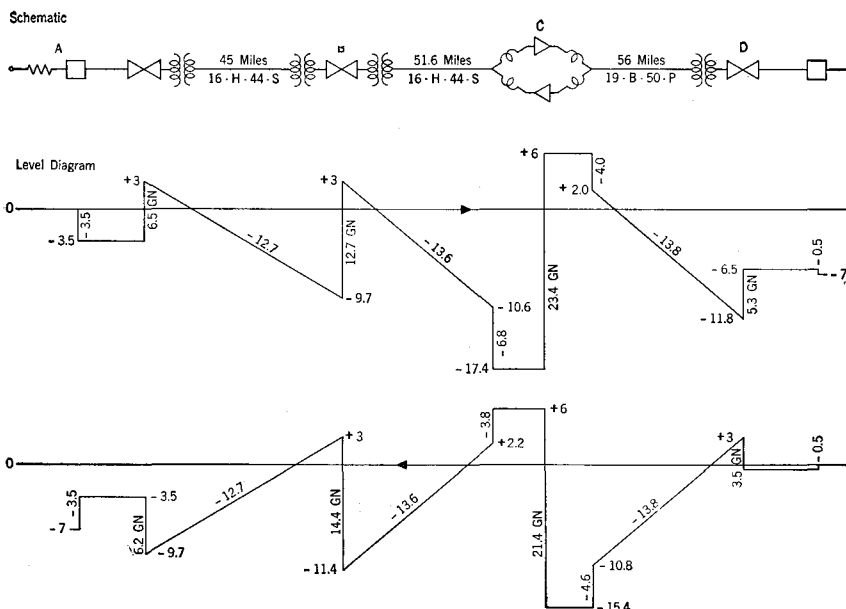


FIG. 30-1 ENERGY LEVEL DIAGRAM FOR TWO-WIRE CABLE CIRCUIT

sion of energy between the balanced circuits and a loss that is dependent upon the degree of balance between the line and network which is known as the **return loss** and designated *R.L.* When the line impedance differs from the network impedance, a portion of the power on the line is 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 return loss. 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 between the bridge taps and the series winding, thus incurring another hybrid loss. The total trans-hybrid loss then equals the *R.L.* plus the normal loss from the series winding to the line, plus the normal hybrid loss from the line to the bridge taps. In the ordinary 22-type repeater circuit, the normal hybrid losses are about 3.25 db making the total trans-hybrid loss *L* equal to *R.L.* + 3.25 + 3.25. The return loss at a given frequency is thus the measured transmission loss across the hybrid coil at that frequency less the fixed losses due to the characteristics of the coil itself.

The value of the return loss is 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 equation—

$$R.L. = 20 \log_{10} \frac{Z_N + Z_L}{Z_N - Z_L} \quad (30:1)$$

where  $Z_N$  is the impedance of the network and  $Z_L$  is the impedance of the line. If the network perfectly balances the line, that is, if  $Z_N = Z_L$ , equation (30:1) shows that the return loss is infinite. When an unbalance exists, the loss takes a finite value. Assuming as an example that  $Z_N = 600$  ohms, and  $Z_L = 400$  ohms, then—

$$\begin{aligned} R.L. &= 20 \log \frac{600 + 400}{600 - 400} \\ &= 20 \log 5 = 20 \times .7 = 14 \text{ db} \end{aligned}$$

A return loss measurement or computation is made in terms of a single frequency, and the gains and losses in the measuring circuit must be expressed for the particular frequency used since both  $Z_N$  and  $Z_L$  may vary with frequency. 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 poorest.

In singing point tests, however, the repeater automatically selects the frequency at which singing is most likely to occur, and thus in a single measurement gives the approximate balance condition at the worst frequency. 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. Phase relationships, also, 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. They provide a ready means of ascertaining what is the maximum safe working gain of a repeater when connected to a given 2-wire 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.

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 suitably terminated in resistors so as to present its nominal impedance to the line. 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**.

### 30.5 Balancing Networks

From the preceding discussion it will be evident that the satisfactory operation of 2-wire repeated circuits depends upon the impedance of the line as well as 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 30-2. Here we have the  $R$  and  $X$  components of the characteristic impedance,  $Z_0$ , 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.

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 a network used to balance this type of line.

The essential elements of a basic network designed to balance an open wire circuit are illustrated by Figure 30-4(A). This simple arrangement, with proper values of resistance and capacitance, will closely approximate the impedance

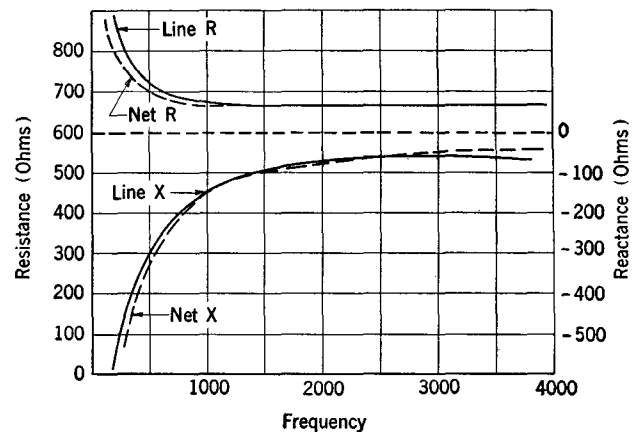


FIG. 30-2 IMPEDANCE CHARACTERISTICS OF OPEN WIRE CIRCUIT AND ITS BALANCING NETWORK

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 30-3(A) shows the resistance components of the impedance of an ideal loaded line for various terminations, 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 30-3(B) shows the corresponding reactance components.

An inspection of Figure 30-3(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

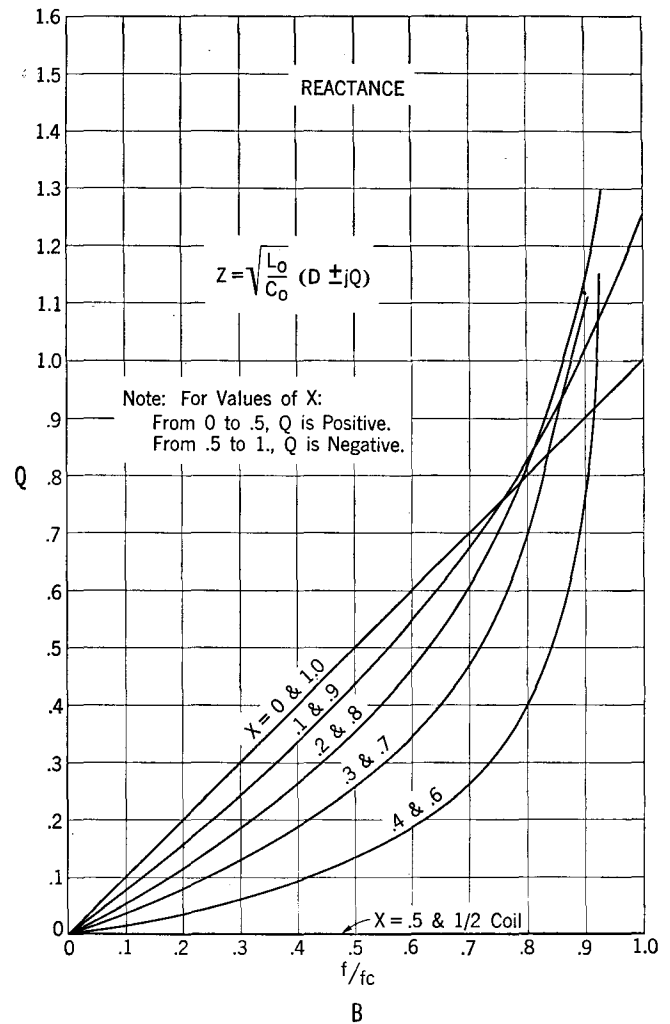
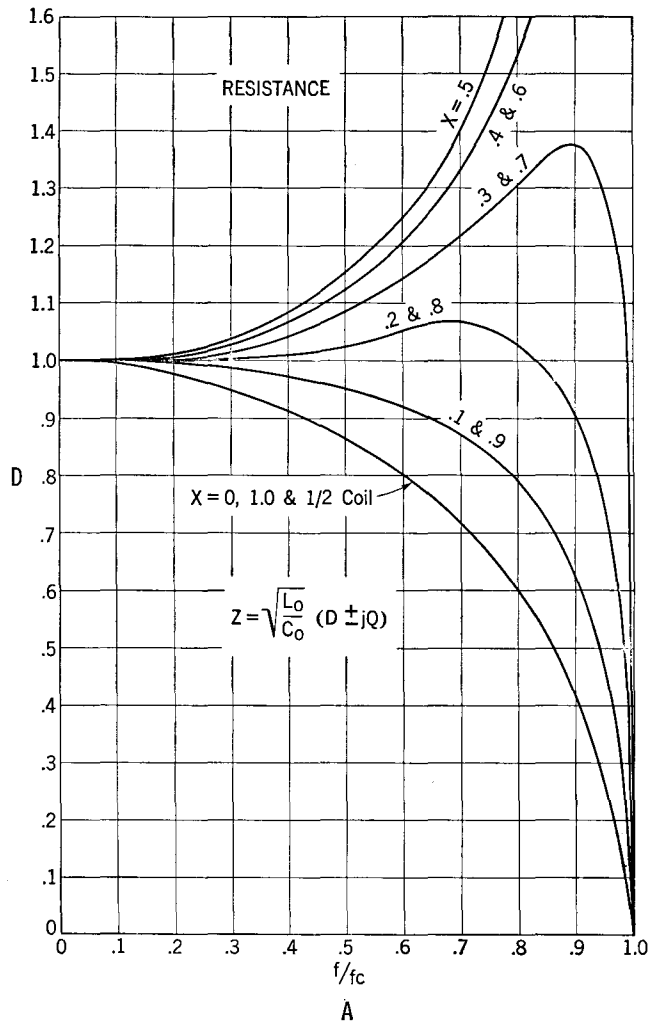


FIG. 30-3 IMPEDANCE COMPONENTS OF IDEAL LOADED CIRCUIT AT VARIOUS END SECTIONS

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 capacitance that will approximate the corresponding reactance component shown in Figure 30-3 (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 capacitance value in parallel with an inductance value. The essential elements of a simple basic network for a loaded circuit at .2 section termination are shown in Figure 30-4 (B).

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 an approximate .2 section termination sending end impedance for a smooth line of infinite length in the case of loaded 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. This may be due to the use of toll entrance cable in the case of open wire circuits, or to the fact that the termination may not be at the .2 section point in the case of loaded cable circuits.

It is the practice to make adjustments on the network sides of the telephone repeater's hybrid coils to take care of these irregularities by means of "building-out sections". That is, if an open wire circuit has a short section of non-loaded toll entrance cable, a capacitance value equal to the capacitance of this section may be bridged directly across the basic network as illustrated by Figure 30-4 (C). If the toll entrance cable is long, it may be necessary to compensate for its resistance as well as its capacitance and the building-out

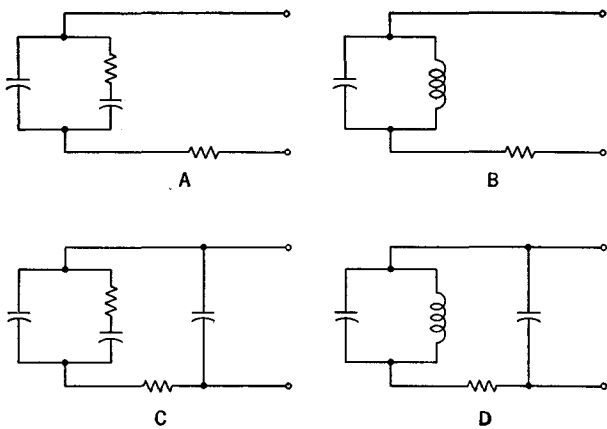


FIG. 30-4 BASIC BALANCING NETWORK CIRCUITS

section may then include a series resistance as well as the bridged capacitance.

Similarly, in the case of the loaded cable circuit, if the capacitance 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 capacitance, as shown in Figure 30-4(D). If, on the other hand, the circuit should be so terminated that the capacitance from the office side of the last loading coil was less than that of .2 loading section, it would be necessary to add bridged capacitance to the line of such value as to make the termination equivalent to .2 of a section.

In order to keep at a minimum the number of different basic network designs, it is current practice to employ networks whose characteristics can be varied somewhat by an appropriate strapping of certain terminals. Thus Figure 30-5(A) shows the elements of a network designed for balancing 104 open wire side circuits. The additional terminals 3 and 4 may be strapped to terminal 2 so as to short out either resistance  $R_1$  or  $R_2$ , or both. This makes it possible to adjust the network for different wire spacings of this particular type of facility. Similarly, the network of Figure 30-5(B) represents the basic design for a 19-gage H-44-S cable circuit, and the strapping possibilities permit adjustments for deviations of the circuit capacitance from its nominal value. Both networks are made up with built-in building-out capacitors which can be adjusted to the desired values. Networks of like designs are available for other standard types of circuit facilities.

### 30.6 Transmission Regulation

The continued satisfactory operation of a long telephone circuit 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, 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 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 XIII 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

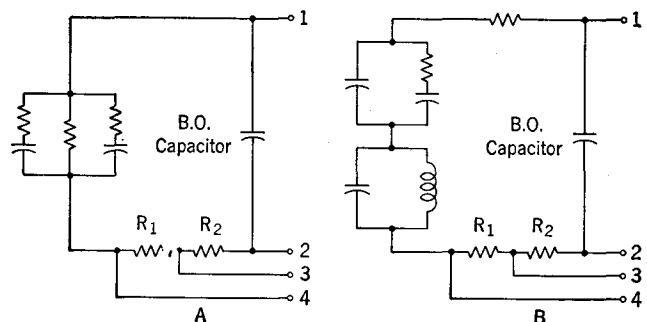


FIG. 30-5 ADJUSTABLE BALANCING NETWORKS



**TABLE XIII**  
**TRANSMISSION EQUIVALENTS IN DECIBEL PER MILE OF 16 AWG & 19 AWG CABLE CIRCUITS AT 55° F.**  
**SHOWING YEARLY VARIATIONS IN EQUIVALENTS DUE TO 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.

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 master relays which in turn control the operation of relays in all the regulating repeaters, causing the gains 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 27-2). 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 27-1).

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. 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.

The operation of the regulating system is illustrated by Figure 30-6. 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<sub>1</sub> and S<sub>2</sub>) together with a part of the slide-wire resistor. In the fourth arm, C, is a fixed resistance, K, and the remainder of the slide-wire resistor. 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 contact arm will be at mid-position (step 0) when the cable is at this temperature. 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 becomes sufficiently large to cause movement of a mechanism to take place in such a way that the shaft on which the slide-wire arm is mounted is caused to turn in the proper direction

to restore balance. When the arm 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 30-6). 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 30-6, 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 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

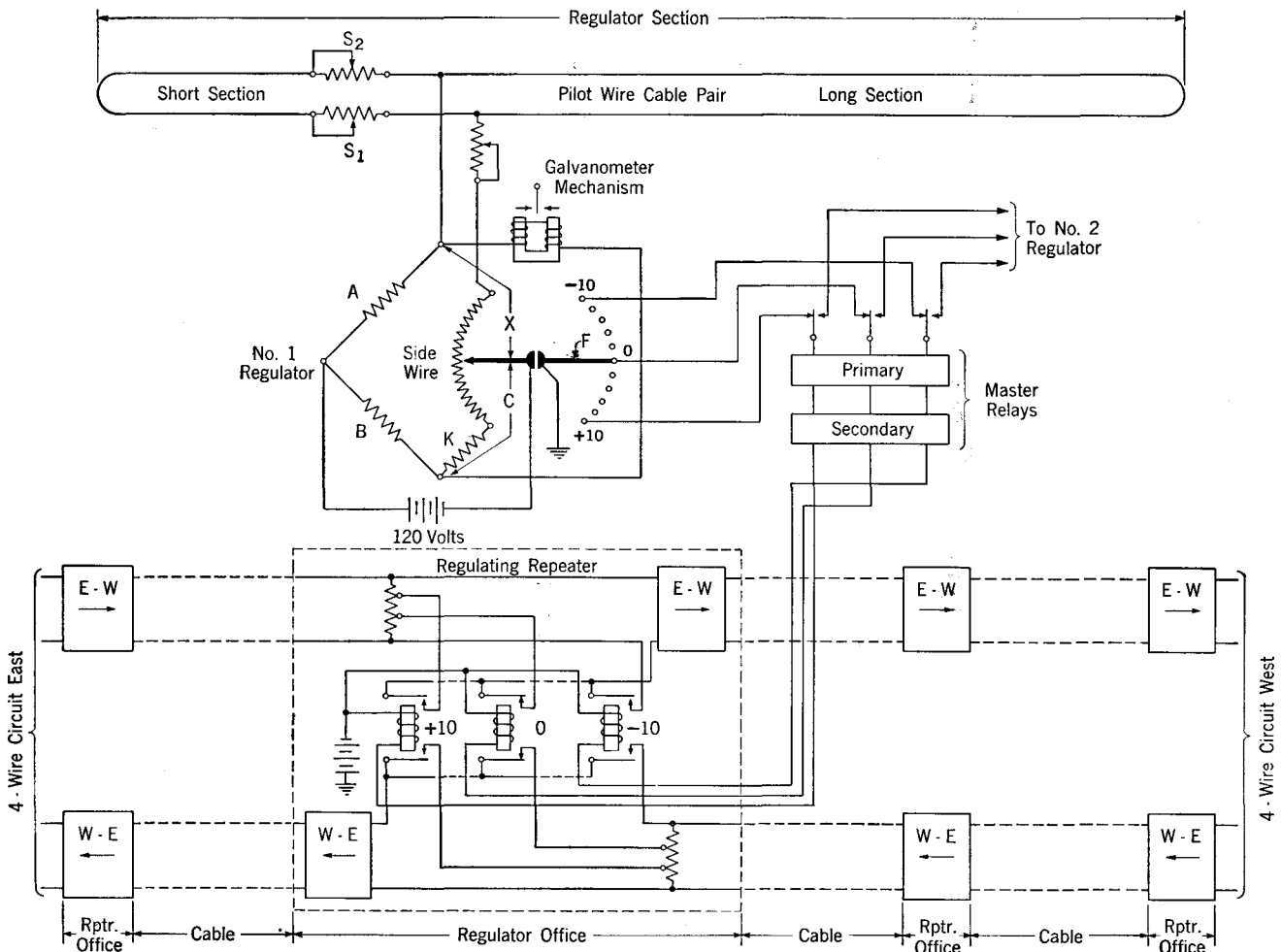


FIG. 30-6 PILOT WIRE TRANSMISSION REGULATOR CIRCUIT

the same for all the circuits under the control of a single master regulator. In order to use the same master regulator under these conditions, the systems are designed so that a given movement of the master regulator produces different changes in the gains of the regulating repeaters on the different types of facilities. This is the purpose of the two sets of master relays, primary and secondary. A separate chain of secondary relays is provided for each of the different types of regulating networks. The primary relays are operated directly from the dial switch on the master regulator. These relays are numbered from +10 to zero to -10, corresponding to the stud from which they are operated. The contacts on the primary relays are used only to operate the secondary relays, and do not control directly any of the regulating repeaters. Each of the different groups of secondary relays then controls directly the relays of all of the like regulating repeaters with which it is associated.

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.

Figure 30-7 is a transmission level diagram of a 4-wire circuit made up of three 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 time. 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 midpoint of the circuit units, although this is not strictly necessary in every case.

### 30.7 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 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 30-8(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. 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 30-8(B) marked "direct transmission". With perfect balance between the network and

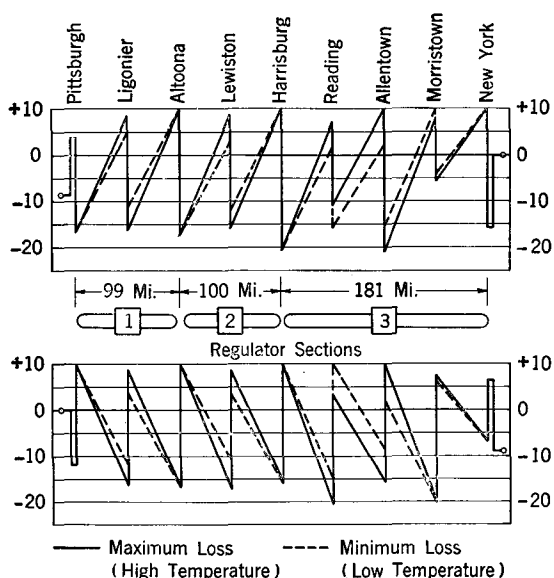
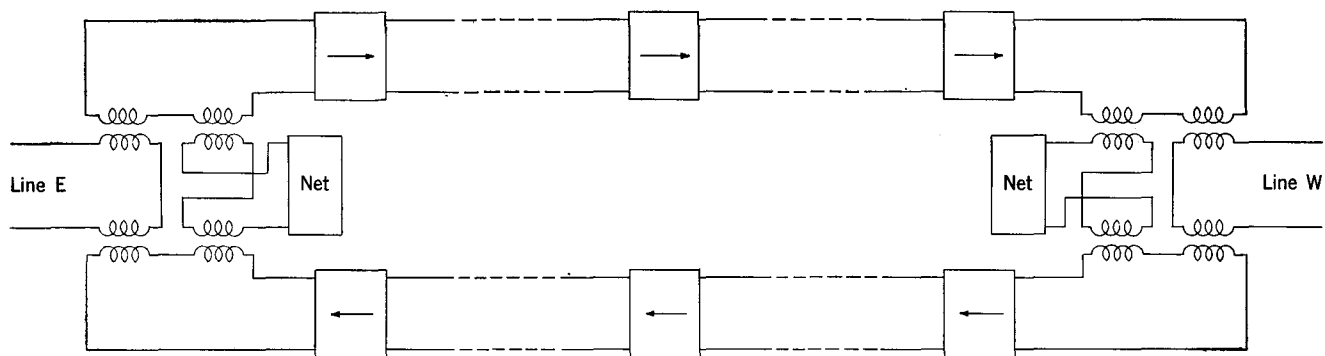
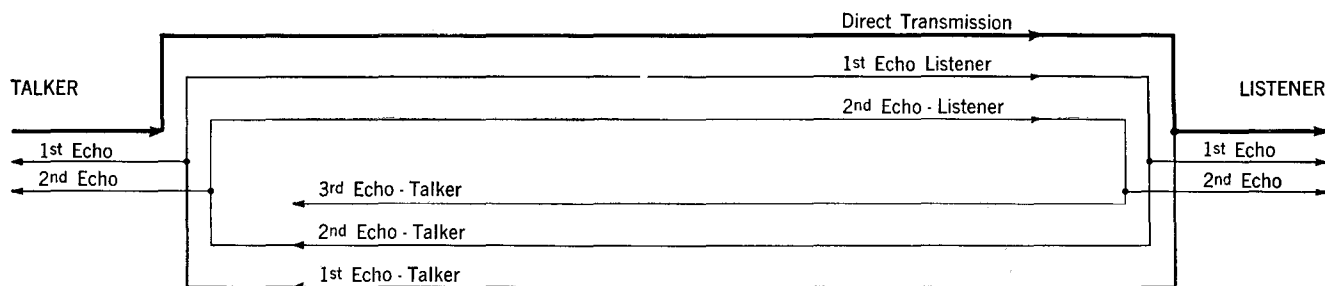


FIG. 30-7 ENERGY LEVEL DIAGRAM FOR FOUR-WIRE CABLE CIRCUIT



A



B

FIG. 30-8 ECHO PATHS IN FOUR-WIRE CIRCUIT

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 30-8(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 sets, 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 re-enforce 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 X, the velocity of propagation of the various types of facilities most commonly employed for voice-frequency 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. 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.

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.

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.

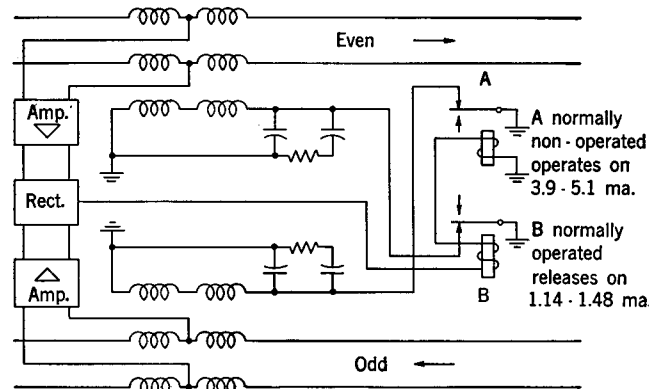


FIG. 30-9 PRINCIPLE OF ECHO SUPPRESSOR

The principle of operation of the echo-suppressor may be understood by referring to the simplified schematic of Figure 30-9. As there indicated, hybrid coils are inserted in each side of a 4-wire circuit or in each side of one of the repeaters of a 2-wire circuit. The bridge points of the hybrid coils are connected to separate amplifiers, the outputs of which connect to a common rectifier circuit. The rectifier output is fed in series through the windings of relays **B** and **A**, as shown. The rectifier circuit is designed to act differentially. When there is no energy, or only low levels of speech or noise, in both sides of the 4-wire circuit, the d-c rectifier output current will be at its normal value which lies between 3.9 ma., the minimum operating current for relay **A** and 1.48 ma., the maximum release current of relay **B**. Relatively high speech energy in the even side of the circuit will cause the rectified current to increase above normal value; and high speech energy in the odd side causes the rectified current to decrease below normal. In the first case, relay **A** operates while **B** is unaffected and in the second case, relay **B** releases while **A** is unaffected. Operation of relay **A** removes the ground connection which had been short-circuiting the resistance-capacitance network connected to the series winding of the odd hybrid coil. The impedance of this network balances, in the voice-frequency range, the input impedance of the amplifier bridged across the hybrid coil. When the short-circuit is removed from the network, accordingly, the now balanced hybrid coil inserts a large loss in the odd path, thus effectively suppressing energy transmission over that side of the line. Similarly, when the **B** relay releases as a result of high energy in the odd path, the ground connection is removed from the even network and transmission over that side of the line is suppressed.

## CHAPTER 31 CARRIER SYSTEMS

### 31.1 Types of Carrier Systems

The general principles of carrier operation were outlined in Chapter 29 and these principles apply alike to all of the several types of carrier systems used in the telephone 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 voice range, up to systems applied to coaxial conductors where frequency bands millions of cycles in width may be involved. As has been previously noted, radio transmission systems have much in common with carrier systems in that both make use of the frequency-division principle for obtaining numbers of separate transmission paths over a single transmission medium. Radio systems as employed in telephone work, however, are discussed in other Chapters. In general, the term carrier system is here used only with reference to frequency-division multiplex systems employing as a transmission medium physical facilities such as open wires, cable pairs, or coaxial conductors.

Figure 31-1 shows the overall frequency allocations for the types of telephone and telegraph carrier systems currently in use in the Bell System. In most of the carrier systems applied to open wire conductors (Type-B telegraph and Types C, J and O telephone) several different sets of frequency allocations are used within the overall range indicated in Figure 31-1. The purpose of this is to reduce the possibilities for inter-system crosstalk where a number of signals are operated on the same pole line. All but one (Type-G) of the open wire carrier systems use separate channels for transmission in opposite directions, thus in effect operating on a 4-wire basis. The Types K and N systems, which are designed for application to wire cable facilities, employ the same channel frequencies for transmission in both directions but use 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 tele-

phone circuit. Type-L systems, which are applied to coaxial conductors, also operate on a 4-wire basis.

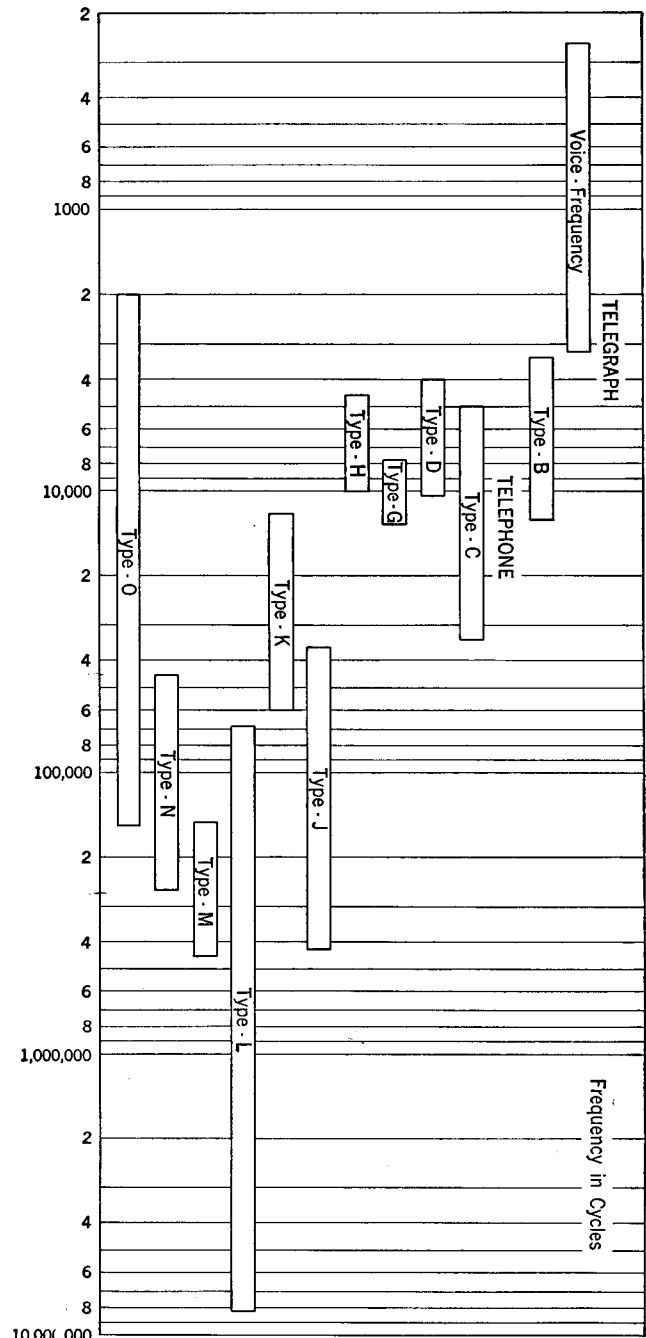


FIG. 31-1 OVERALL FREQUENCY ALLOCATIONS OF STANDARD TYPES OF CARRIER SYSTEMS

The principles of most of the major apparatus units and circuit arrangements that are employed to make up a carrier system such as filters, equalizers, modulators, amplifiers, etc., have been discussed in earlier Chapters. This and the following Chapter will be devoted to a brief survey of the general overall design of the types of systems most widely used in long distance telephone and telegraph work. It will include discussion of the carrier frequencies used to obtain the various channels, the modulation steps required, and the important regulating and equalizing procedures that must be employed for satisfactory overall transmission.

### 31.2 Carrier Telegraph Systems

Two types of carrier telegraph systems are in use, as indicated in Figure 31-1. First in historical order of development, although currently receding in importance, is the Type-B system. This system ordinarily provides for superimposing ten two-way telegraph channels on an open wire telephone circuit, which may also be composited for ordinary grounded telegraph operation. Two frequency allocation groups are used, coded BL and BH. Each group employs twenty carrier fre-

quencies, 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 specific channel frequency allocations are given in the adjacent table.

As may be noted, the separation between carrier frequencies ranges from less than 200 cycles at the lower end of the frequency band, to 750 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 general arrangement of the Type-B system terminal is indicated in the block diagram of Figure 31-2. 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.

There are also two special arrangements of open wire telegraph systems coded BC and BS. The former is designed to operate on the same wires as a Type-C telephone system and provides

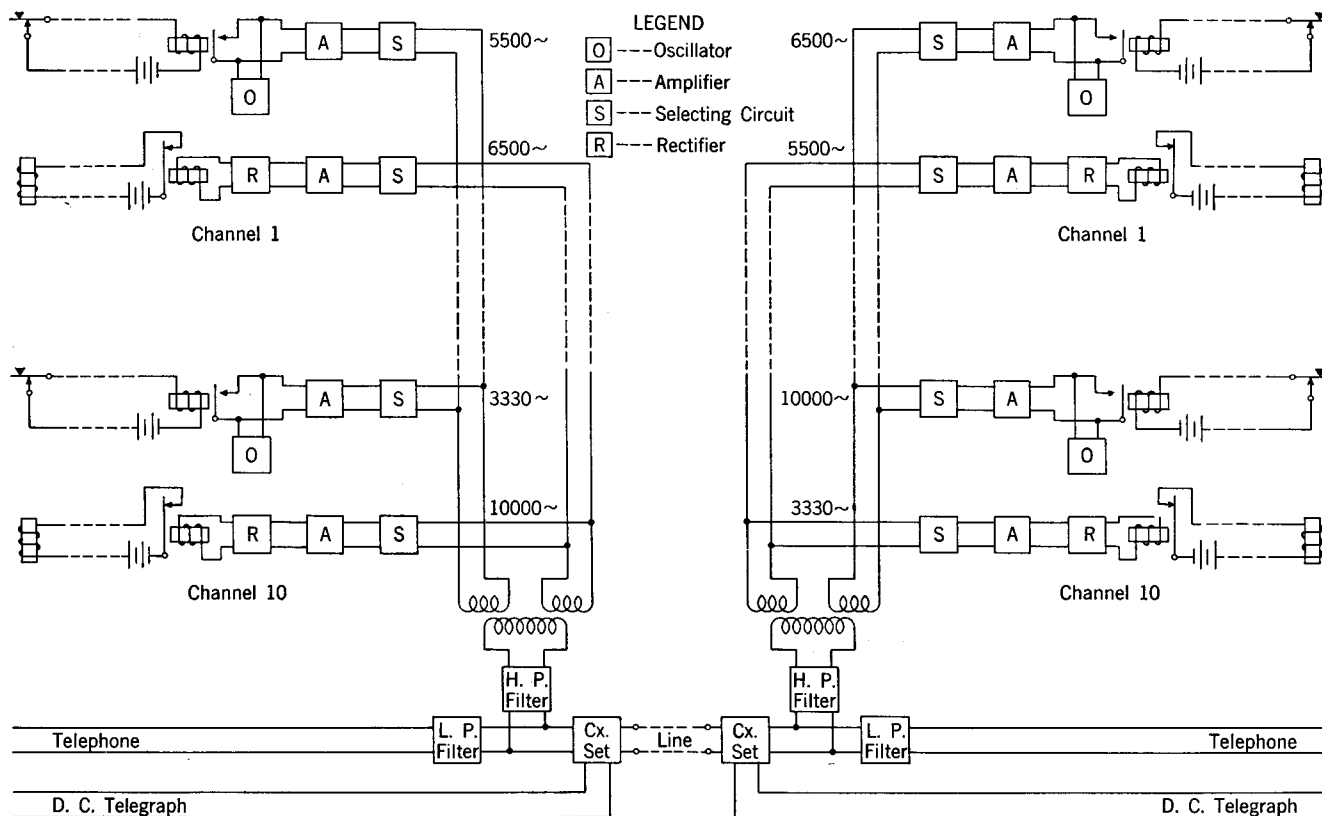


FIG. 31-2 HIGH-FREQUENCY CARRIER TELEGRAPH SYSTEM

B-H FREQUENCY ALLOCATION  
CHANNEL NO. AND FREQUENCY

E-W	W-E
1— 6800	10—3720
2— 7110	9—3910
3— 7440	8—4110
4— 7800	7—4320
5— 8180	6—4540
6— 8590	5—4770
7— 9030	4—5010
8— 9500	3—5250
9—10,000	2—5500
10—10,750	1—5760

B-L FREQUENCY ALLOCATION

E-W	W-E
1— 6500	10—3330
2— 6800	9—3570
3— 7110	8—3810
4— 7440	7—4050
5— 7800	6—4290
6— 8180	5—4530
7— 8590	4—4770
8— 9030	3—5010
9— 9500	2—5250
10—10,000	1—5500

a maximum of three channels in the frequency range between 3.57 kc and 5.01 kc. The BS system is used in certain cases to minimize interference with telephone program circuits. It provides a maximum of seven channels in the frequency range between 6.65 kc and 12.75 kc.

Much more widely used in current practice than the Type-B system is the voice-frequency carrier telegraph system. The principle of this system is not essentially different from that of Type-B but there is considerable variation in the details of the apparatus. The voice-frequency 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. Voice-frequency systems provide up to 18 telegraph circuits. The carrier frequencies employed are odd multiples of 85 cycles beginning with channel 1 at 425 cycles and extend-

ing at 170-cycle intervals through channel 17 at 3145 cycles. Channel 18, when used, is placed below channel 1 at 255 cycles. Because of the generally poor transmission characteristics of line facilities at that frequency, however, this channel is likely to be inferior to the higher frequency channels.

The carrier sending and receiving apparatus is similar in principle 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 d-c telegraph signals. The channel receiving circuit includes a two-stage amplifier, a bridge-type varistor rectifier, an output and control tube, and a receiving telegraph relay, which is operated by the plate current of the control tube and is normally biased to spacing by an auxiliary winding. The first stage of the amplifier is arranged to provide partial automatic gain control. The gain of the tube in this stage is variable, and the circuit is designed so that the amount of gain varies in approximate inverse proportion to the amplitude of the incoming signal voltages. The voltages impressed on the rectifier, and the resultant rectified signal, are therefore nearly constant in magnitude regardless of the input level. The control tube acts effectively to amplify the rectified signals and to "square-up" the signal form. In other words, the tube circuit is so arranged that during spacing intervals, when there is no incoming rectified signal voltage, its control grid is biased well beyond the cutoff point and no current flows in the plate circuit or the winding of the receiving relay to which the plate is connected. During marking intervals, on the other hand, the rectified signal voltage overcomes the bias so that a large plate current flows and causes the relay to operate sharply to its marking contact.

The carrier supply voltages in recent designs of voice-frequency telegraph systems are provided by vacuum tube oscillators in the manner indicated schematically in Figure 31-3. As shown, a base frequency of 85 cycles is developed by a common oscillator, amplified and passed on to a harmonic generator circuit. This includes a non-linear coil, or saturable reactor, connected in an arrangement like that described in Article 28.5. As was there pointed out, the output of such a circuit arrangement is a sharply peaked wave form which contains all the odd harmonics of the



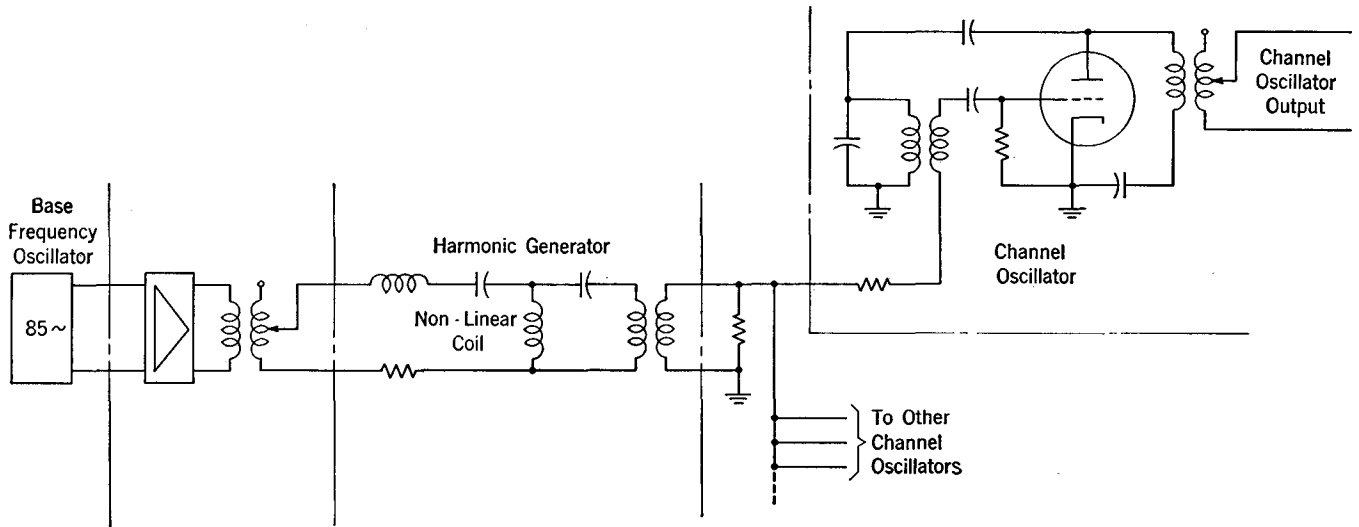


FIG. 31-3 V-F CARRIER GENERATING CIRCUIT

applied sine wave. This wave is introduced into the grid circuits of all the channel-frequency oscillators. Each channel oscillator tuned circuit then acts as a filter to select the particular harmonic to which it is tuned, and the channel oscillator falls into exact step with this frequency.

### 31.3 Type-C Carrier Telephone System

The oldest of the carrier telephone systems still in general use is coded Type-C. 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 31-4. Transmission over each channel consists of a single side-band, the carrier frequency being suppressed in the modulator. Earlier designs of C-systems used vacuum tube modulators and demodulators. Copper-oxide varistors are employed for both purposes in the more recent designs. The required carrier frequency at each end of each channel is supplied by individual vacuum tube oscillators whose stability is such as to maintain satisfactory frequency synchronization at all times.

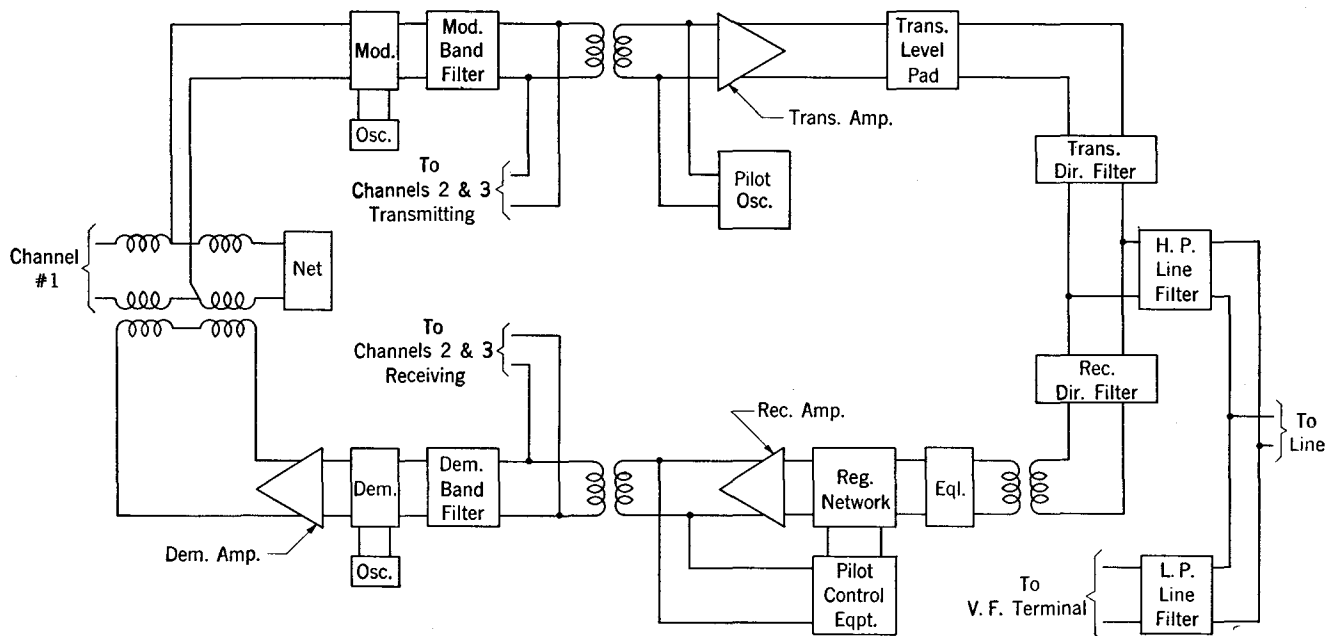
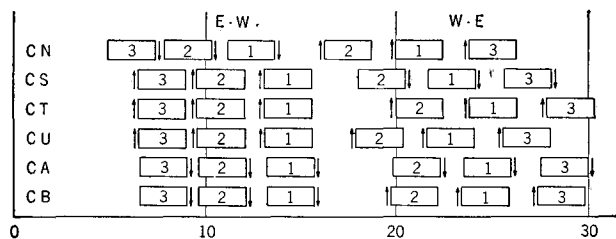


FIG. 31-4 TYPE-C CARRIER TELEPHONE TERMINAL

The individual channel carrier frequencies used in Type-C systems are shown in Figure 31-5. It will be noted that a total of six different frequency allocations are used. All six allocations, however, cannot be used for any one C-system design. Allocations coded CN, CT, CS and CU are suitable for the earlier designs of C-systems, while allocations CS, CU, CA and CB can be used with the more recent design. In all cases, the separation between carriers is at least 3000 cycles, which permits transmission of a side-band about 2500 cycles in width. Separation between the transmission in the two directions at terminal and repeater points



Note Arrows Indicate carrier frequencies

Upper side band Lower side band

	E-W			PILOT
	3	2	1	
CN	7.6	10.6	13.9	10.55
CS	6.3	9.4	12.9	9.45
CT	6.3	9.4	12.9	9.45
CU	6.3	9.4	12.9	9.45
CA	9.3	12.4	15.9	12.35
CB	9.3	12.4	15.9	12.35

	W-E			PILOT
	2	1	3	
CN	16.1	19.8	23.7	19.85
CS	20.7	24.4	28.4	24.35
CT	19.8	23.7	27.7	23.75
CU	17.7	21.4	25.4	21.45
CA	22.5	26.20	30.2	26.15
CB	19.5	23.2	27.2	23.25

FIG. 31-5 TYPE-C CARRIER FREQUENCY ALLOCATIONS

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. High-frequency signaling is employed, the ringing current being transmitted over the system in exactly the same way as the voice currents. C-systems require the use of repeaters at intervals of about 140 to 180 miles, depending upon the transmission characteristics of the conductors. The repeater consists of two amplifiers, one transmitting in each direction, together with directional filters for obtaining the necessary separation.

Regulation to compensate for changes in line

attenuation due to weather and temperature changes is effected automatically by means of automatic gain control circuits associated with the line amplifiers at repeaters and with the receiving amplifiers at terminals. The regulating equipment is controlled by pilot channel frequencies, which are supplied by an oscillator at the transmitting terminal as indicated in Figure 31-4. The values of the pilot channel frequencies for each direction of each system frequency allocation are listed in Figure 31-5. The pilot frequencies in each case have values lying between the carrier frequency and the side-band of the middle channel. Since the pilot frequencies are transmitted along with the regular signal frequencies, they suffer essentially 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.

As shown in Figure 31-4, the pilot channel voltage is applied at the input of the terminal transmitting amplifier. At repeater points, this single-frequency voltage is tapped off at the output of the line amplifier and fed back through a rectifier circuit to a pilot channel control circuit which controls the net gain of a regulating network and amplifier connected into the main transmission path in front of the line amplifier. The same general plan applies at the receiving terminal, except that here the pilot channel voltage is taken off at the output of the receiving amplifier.

The regulating amplifier circuit consists essentially of a variable attenuator in series with a vacuum tube 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 capacitor whose movable plate is rotated under the control of the pilot channel voltage. The rotor of the capacitor is connected to the control grid of the amplifier tube so that the voltage applied to the amplifier depends upon the position of the capacitor plate. When the capacitor rotor is at its extreme left position, the regulating network is effectively out of the transmis-

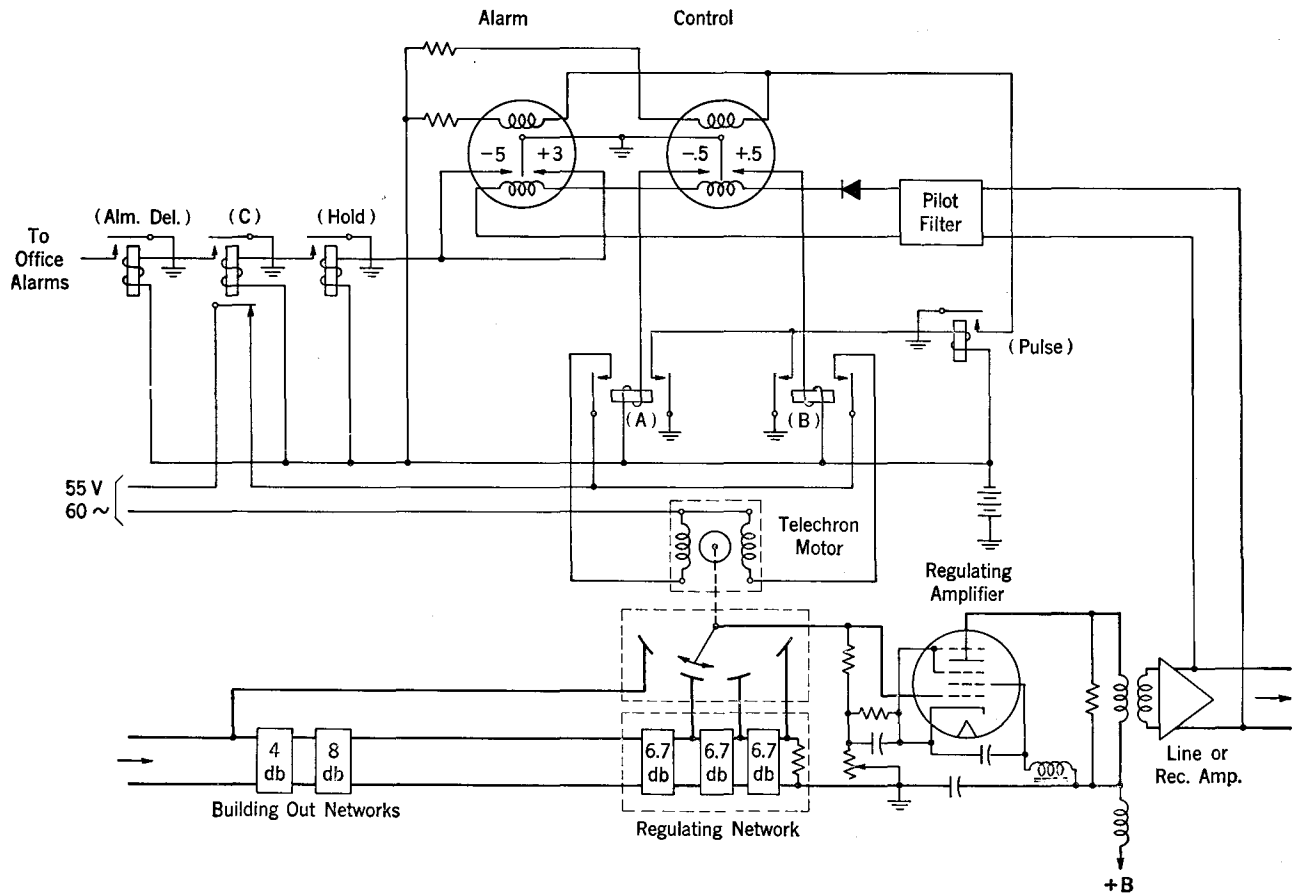


FIG. 31-6 C-CARRIER AUTOMATIC GAIN CONTROL CIRCUIT

sion 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 network. This arrangement provides a smooth control of the net gain of the regulating circuit and avoids the use of sliding contacts or relays in the transmission path.

Figure 31-6 shows how the pilot channel voltage controls the position of the regulating capacitor. As may be seen, the incoming pilot channel voltage is selected by a band filter, rectified, and led through the winding of a control relay. This latter is a highly sensitive type of relay designed to act very positively by 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 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 voltage 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, correcting at the rate of approximately 1/16 db every 4 seconds. Release is effected by means of the "pulse" relay, the winding of which is connected to a second contact of relay *B*. The pulse relay is a special mercury delay 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.

What happens, then, is that when the pilot channel current deviates from normal, the tele-

chron 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 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 31-6 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, as indicated. 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 network circuit. At the end of four seconds, the operation of the pulse relay restores both the control and

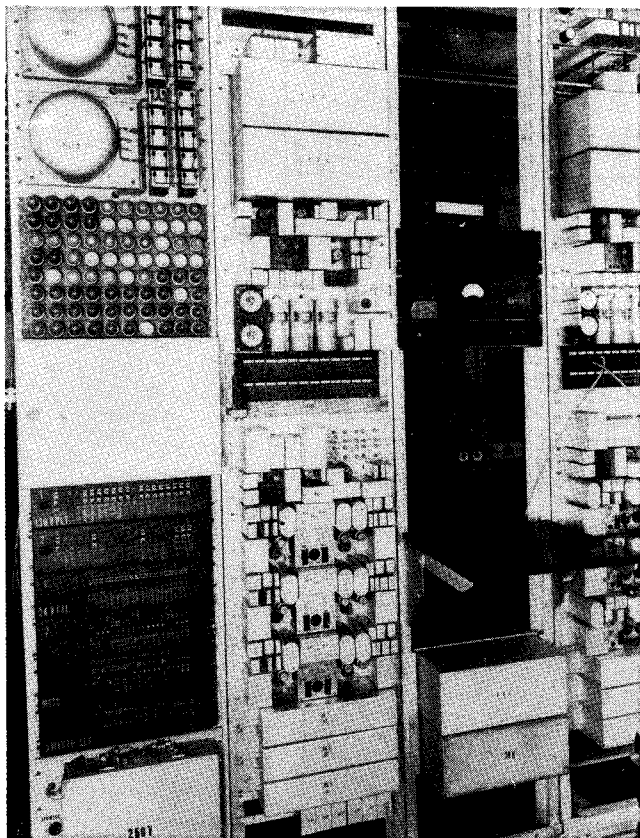
alarm relays to normal. The hold relay has a release time of four seconds and if the level change still persists, no correction will be made because relay *C* remains operated. If the level change was temporary, however, relay *C* will release after four seconds and normal regulation will be resumed. If a level change of sufficient magnitude to operate the alarm relay persists for a considerable time, the second contact on relay *C* will cause the "alarm delay" relay to operate after a period of 25 seconds, which will cause the operation of other relays to give visible and audible alarms that the circuit needs manual attention.

### 31.4 Type-J Carrier Telephone System

The Type-J carrier telephone system provides 12 two-way telephone channels on an open wire pair using a frequency range from about 36 kc to 143 kc. As in the Type-C system, operation is on an equivalent 4-wire basis with West to East transmission employing a frequency band lying between 36 and 84 kc, and East to West a band lying between 92 and 143 kc. Channel carriers are spaced at 4 kc intervals giving an effective channel bandwidth of well over 3 kc. Four frequency allocations coded NA, NB, SA and SB are available for use in limiting crosstalk where more than one J-system is applied to the same line. Since the lowest frequency on the line is 36 kc, it is possible for a pair of wires to carry both a voice-frequency telephone circuit and three Type-C carrier telephone channels in addition to a J-system, thus providing a total of 16 voice circuits.

Figure 31-7 is a block schematic of the essential elements of a West (or South) J-system terminal. It will be noted that the principle of operation does not differ essentially from that employed in Type-C and other carrier 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 desired bands of frequencies. 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 for the basic "channel bank". In the first place, it happens that high-grade crystal filters can be economically



TYPE-C CARRIER TERMINAL

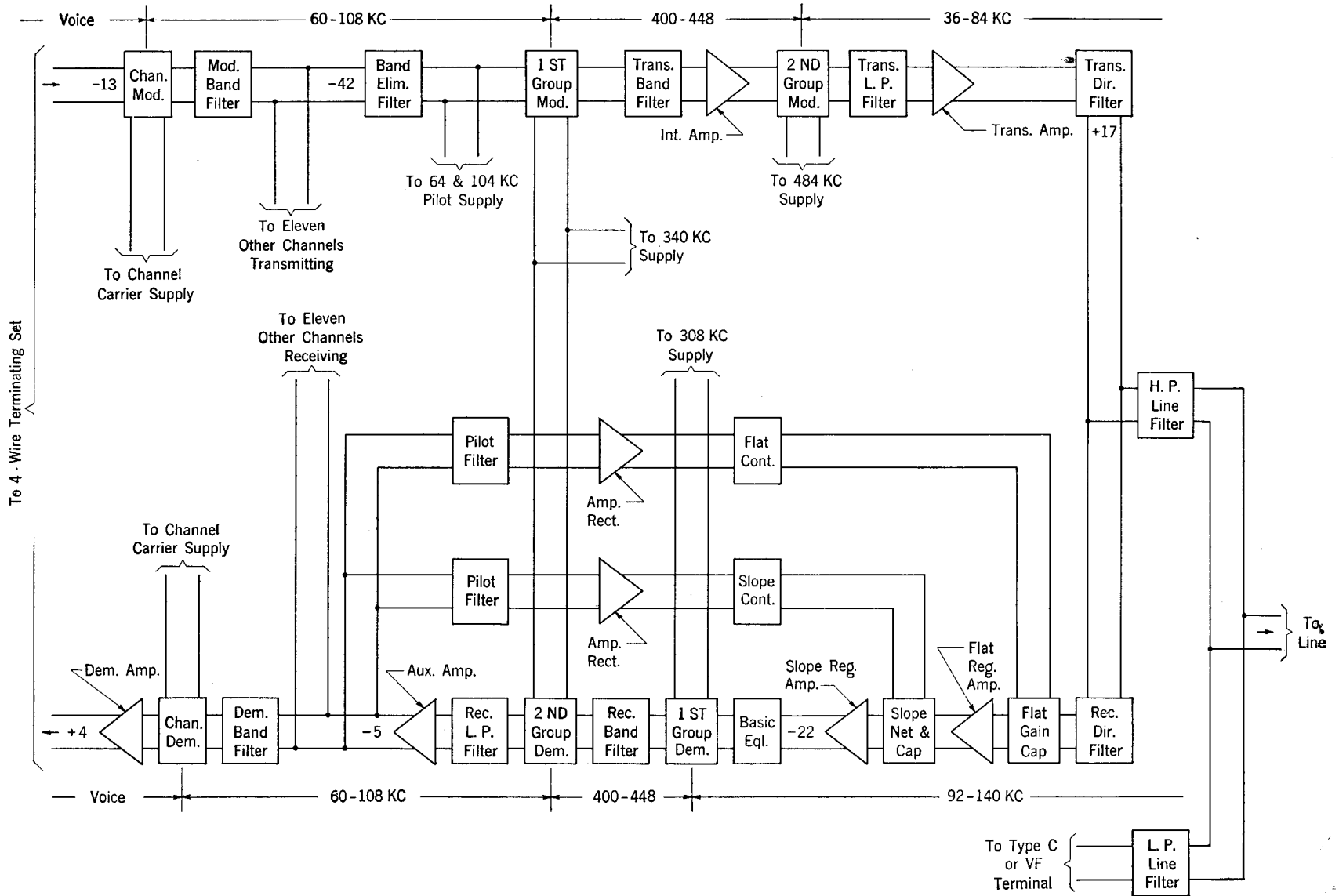


FIG. 31-7 TYPE-J CARRIER TELEPHONE TERMINAL (WEST)

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 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 channel bank for other carrier systems, including Types K, L, and N.

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 NA frequency allocation, 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.

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 31-8. 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 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 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 translations occurring on the receiving side of the circuit are, of course, in the opposite direction and may be followed through in a like

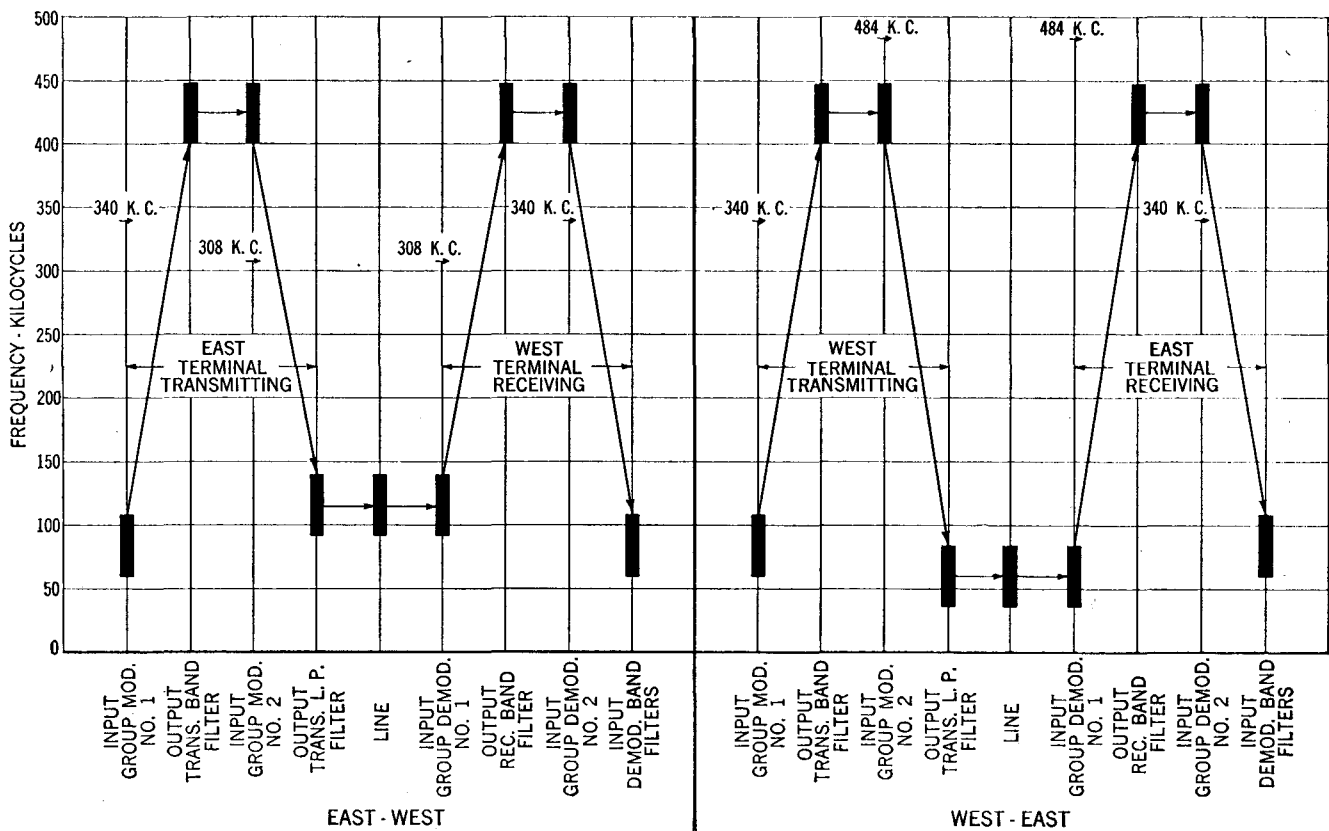


FIG. 31-8 FREQUENCY TRANSLATIONS IN TYPE-J CARRIER SYSTEMS

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 relatively small amounts of energy, so that the transmission levels at both their inputs and outputs are necessarily rather low. As may be seen by referring to Figure 31-7, 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 frequencies employed are relatively 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 applied, the average spacing is about 50 miles. At the repeater points, the transmission in the two directions is separated by directional filters and each repeater includes two line amplifiers, one "pointed" in each direction.

Automatic regulation to compensate for attenuation variations caused by temperature and weather changes is accomplished in the J-system by means of two pilot channels associated with each direction of transmission. One pilot channel provides for "flat gain" regulation and the other

for "slope" regulation. The flat gain pilot channel frequencies on the line are at 80 kc for W-E transmission, and 92 kc for E-W transmission. The corresponding slope pilots are at 40 and 143 kc respectively. It may be noted in Figure 31-7 that the pilot supply frequencies applied at the input of the first group modulator are indicated at 64 and 104 kc for W-E transmission. The two subsequent frequency translations, however, change these values to 80 and 40 kc respectively when they reach the line. On the receiving side, the pilot channel voltages are picked-off at the output of the auxiliary amplifier, separated by pilot filters, amplified and rectified. The rectified flat gain pilot voltage is fed to a flat control circuit of the same type as was shown in Figure 31-6. This circuit controls the position of a variable flat gain capacitor which in turn determines the gain of the flat regulating amplifier. Similarly, the rectified slope pilot voltage drives a slope control circuit which, through another variable capacitor, controls a slope network and an associated slope regulating amplifier. This slope network, with its amplifier, is effectively a variable equalizer. It should be noted that the diagram of Figure 31-7 is for a West terminal and the incoming transmission is therefore in an E-W direction. The incoming pilot frequencies from the line are accordingly 92 kc (flat) and 143 kc (slope), which after the two demodulation stages take values of 60 and 111 kc respectively. The regulating circuit arrangement at an East terminal, and the W-E amplifiers of repeaters are somewhat different in detail design but function according to the same general principles.

For discussion purposes, we will consider only the West terminal and the corresponding E-W repeater amplifiers where slope and flat gain regulation are controlled by line pilot frequencies of 143 kc and 92 kc respectively. To understand the overall operation, we may assume weather conditions where the line is changing from dry to wet. The wet line loss will become higher at all frequencies, but the increase in loss will be greater at the higher than at the lower frequencies. The flat gain regulator will adjust automatically to compensate for the increased loss at 92 kc, but there will remain a deficiency of compensation for the higher frequencies. The slope regulating circuit, which is controlled by the high 143 kc pilot frequency, will operate to compensate for this deficiency by increasing the gain at 143 kc, and at lower frequencies on a gradually decreasing (approximately straight line) basis. Actually the two

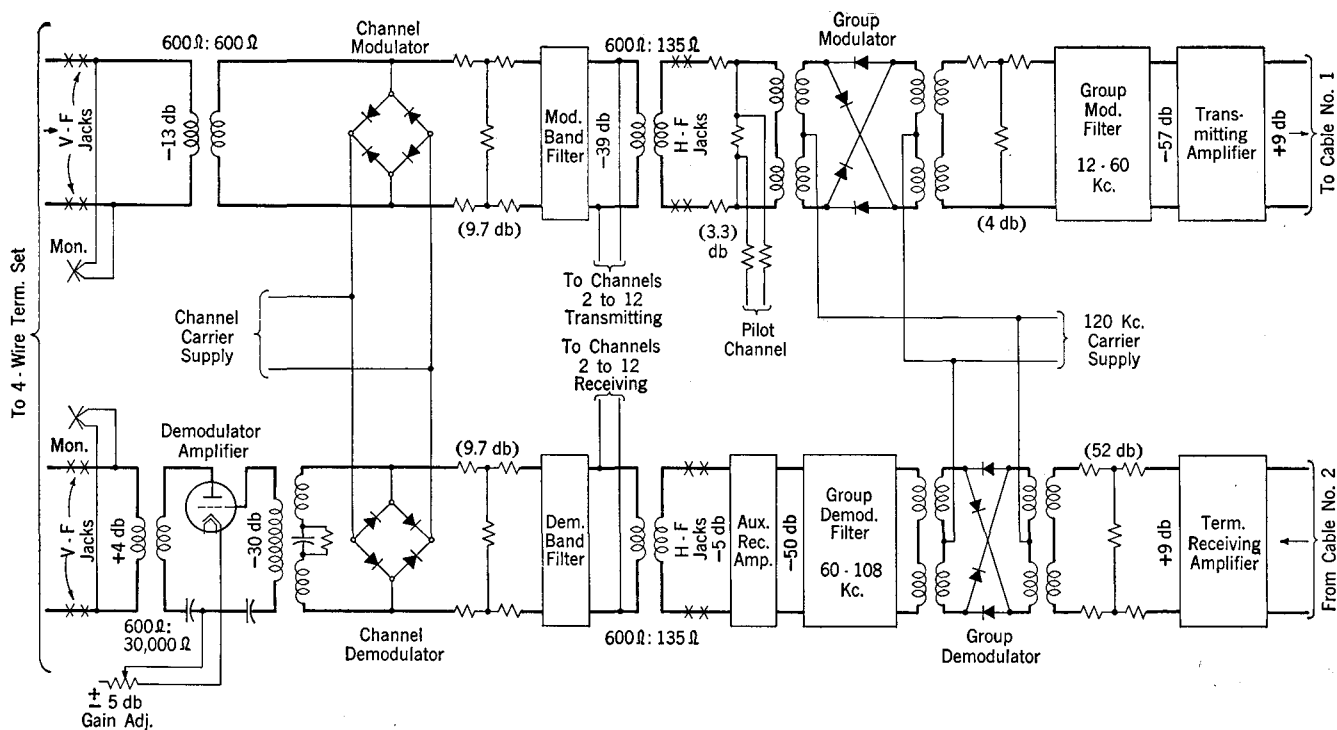


FIG. 31-9 TYPE-K CARRIER TELEPHONE TERMINAL

controls are independent and both actions may take place simultaneously, thus at the same time increasing the total flat gain of the amplifier and changing the slope of the gain-frequency characteristic to compensate for the changed loss-frequency characteristic of the line.

### 31.5 Type-K Carrier Telephone System

Type-K carrier systems provide 12 two-way telephone channels over two 19-gage non-loaded cable pairs. As shown in Figure 31-9, the terminal channel banks are identical in design with those of the Type-J system, spacing the 12 channels in the 60 to 108 kc band with 4 kc intervals between carriers. For transmission over the line, however, the techniques employed in the two systems are necessarily quite different. Because the attenuation of non-loaded cable conductors is high and of course increases with frequency, it is desirable to keep the maximum line frequency at 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 17 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.

Three basic types of amplifiers are used in K-systems for counteracting normal line attenuation, and for gain regulation to compensate for variations in line attenuation. Such variations in cable facilities are due entirely to temperature changes, since the insulation of the cable conductors is not affected by weather conditions. The three types of amplifiers are known as terminal transmitting amplifiers, line amplifiers and twist amplifiers. The latter two types of amplifiers furnish automatic regulation as well as flat gain. The terminal receiving amplifier is a twist amplifier.

Variations in attenuation of cable facilities due to temperature changes may be considered as being composed of two components. The first of



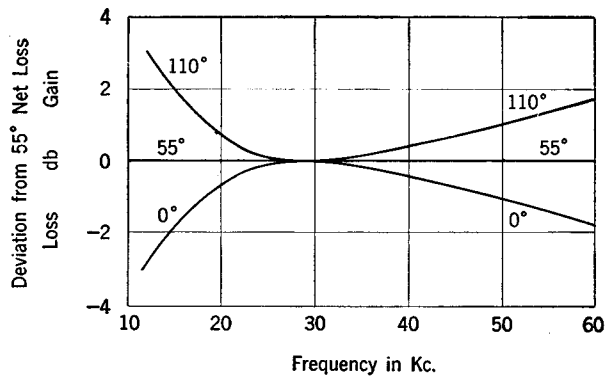


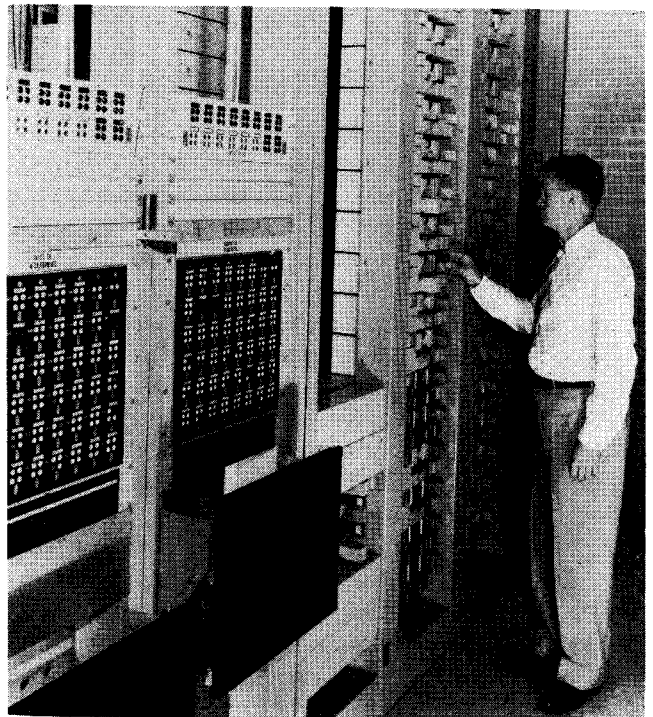
FIG. 31-10 TWIST EFFECT IN 100-MILE AERIAL CABLE CIRCUIT

these is a "flat loss" variation, which is independent of frequency. In addition, changing temperatures cause attenuation variations which are not exactly the same at all frequencies in the transmitted range. In other words, the shape of the attenuation-frequency characteristic of cable facilities may be different at different temperatures. This effect is known as "twist". The curves of Figure 31-10 indicate the general magnitude of the twist effect for a 100-mile aerial cable circuit over the Type-K frequency range, in terms of deviations from the loss at 55°F. Compared in magnitude to the flat loss variation resulting from temperature changes, the twist variations may appear insignificant. Nevertheless, they are too large to be allowed to accumulate over a long circuit. Automatic regulation is accordingly required in practice at intervals of about 100 miles in aerial cable, and 200 miles in underground cable, to take care of the twist effect. Flat gain regulation, in contrast, is required at every repeater point to take care of the flat loss variation of the adjacent 17-mile cable section.

In the original design of K-systems, both flat and twist regulation are provided by means of pilot wires extending through one repeater section for flat gain regulation, and through a twist section (several repeater sections) for twist regulation. The loss variations in these pilot wires control mechanical devices associated with the regulating amplifiers, in a manner similar in principle to the voice-frequency regulating circuit discussed in Article 31.2. In more recent designs, regulation is secured through pilot channels employing frequencies of 12, 28, 56, and 60 kc on the line. The 60 kc channel controls the flat gain of the line amplifiers while the other three channels control the twist amplifiers. Changes in the regulating amplifier gains are effected in all cases through

the use in the feedback circuits of **thermistors** whose resistance is varied by the pilot channel currents.

The thermistor is a circuit element made of a number of semi-conducting metallic oxides, which have a large negative temperature coefficient. In contrast to the usual conducting metals, which have small positive temperature coefficients, the resistance of a typical thermistor decreases in the order of 2.5% per degree Fahrenheit so that its resistance is halved for about every 27°F increase in its temperature. Several types of thermistors are used in K-carrier systems. Physically, these take the form of a tiny "bead" of semi-conducting material not much larger than a pin-head, or of a "disc" of like material about the size of a 5-cent coin. The disc type is used as a thermostatic control device whose resistance is determined by the ambient temperature. The bead thermistor may be either directly or indirectly heated, but in either case is surrounded by a tiny coil of nichrome wire, which is insulated from the bead itself and may be used to vary the temperature of the bead and thus change its resistance. Where slow action of the thermistor is desired, the bead is enclosed in glass to increase the thermal capacity, and a thin film of gold is deposited on the outside of the glass to reduce heat loss by radiation.



TYPE-K CARRIER REPEATERS AT AUXILIARY STATION

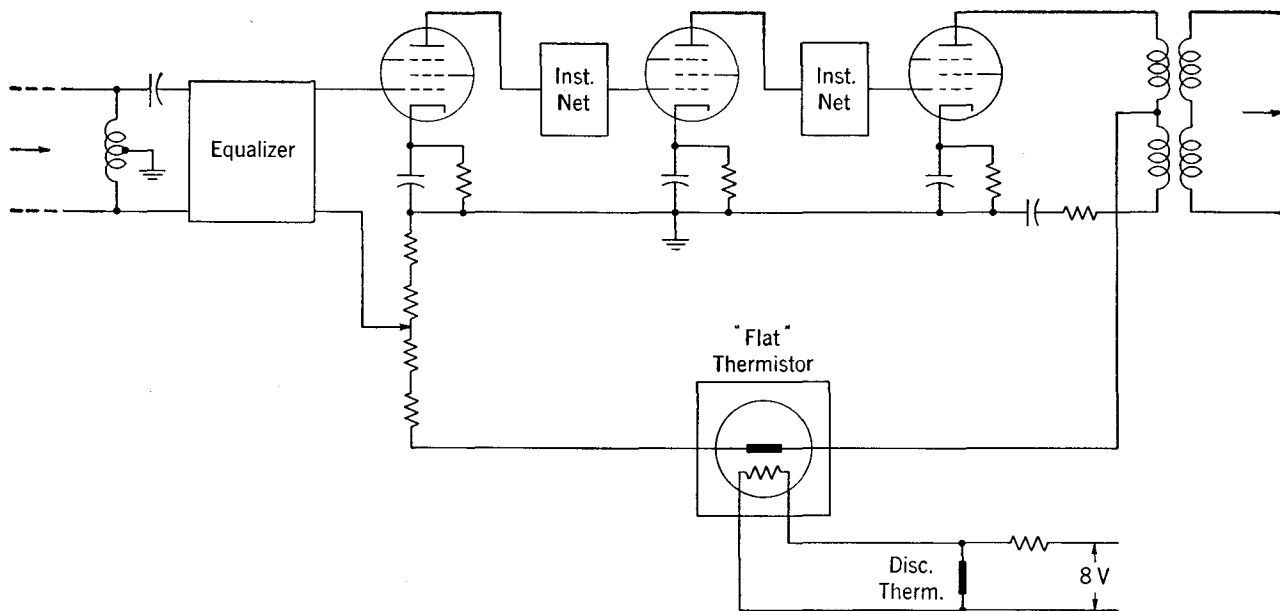


FIG. 31-11 TYPE-K LINE AMPLIFIER-REGULATING

In the line amplifiers of K-systems, a directly heated thermistor bead is inserted in series with the amplifier negative feedback path as indicated in Figure 31-11. A portion of the amplifier output current, whose total value is controlled by the 60 kc pilot channel, is fed back through this thermistor into the amplifier input. If the amplifier output tends to decrease below its normal value as a result of increased attenuation in the preceding line section, the current through the thermistor will tend to decrease accordingly. This causes the thermistor resistance to increase, which reduces the total current fed back into the amplifier input and thus increases the amplifier gain in amount sufficient to compensate for the increased line attenuation.

Figure 31-12 shows how the thermistor resistance varies with the current flowing through it. It will be noted that the normal operating range of this thermistor is between 1100 and 20,000 ohms, which corresponds to current values of 1.7 and .15 ma. In order that the thermistor operation be uniform and within this operating range, the thermistor bead is maintained at a constant reference temperature of 160°F by means of an auxiliary current through its heater winding. This current is supplied from an 8-volt filament supply tap, as indicated by Figure 31-11. To avoid changes in the bead's temperature as a result of ambient temperature changes, a disc-type thermistor is bridged across the heater winding supply. The disc is mounted close to and in the same

container as the bead, but because of its size, its temperature is controlled by the surrounding ambient temperature and not affected by the temperature of the bead itself. Accordingly, if the ambient temperature should increase, the resistance of the disc will decrease thus reducing the heater current by its increased shunting effect. The thermistor circuit is arranged to be relatively

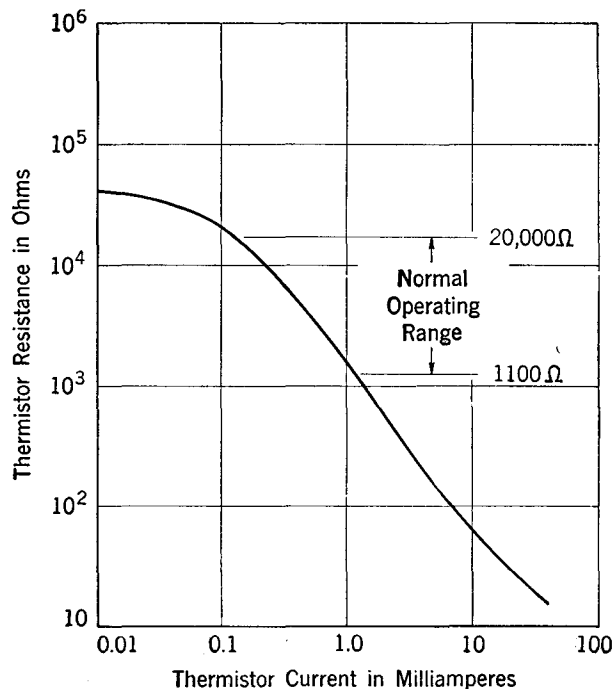


FIG. 31-12 CHARACTERISTIC OF FLAT GAIN REGULATING THERMISTOR IN LINE AMPLIFIER

slow acting (10 to 15 minutes) to correspond to the normally slow variation of the line attenuation.

The design objective of the regulating line amplifier is to maintain practically constant total output power at each amplifier, regardless of possible substantial variations in input power. To accomplish this objective, it is necessary that the same constant power value be supplied to the line by the terminal transmitting amplifier. This total power output consists of the power of the 60 kc pilot channel plus the combined power of the signal channels. Its value in the K-system is +15 dbm (15 decibels above 1 milliwatt). The transmitting amplifier accomplishes the double purpose of raising the signal currents to a proper level for transmission over the line, and of generating the 60 kc pilot channel power. The arrangement is such that when no signals are being transmitted, the 60 kc output power alone will total +15 dbm. When signal currents are present, the amplitude of the 60 kc oscillations automatically decreases to whatever extent is necessary to still maintain the total output power at +15 dbm. This means that the output of the transmitting amplifier, and the gains of the succeeding line amplifiers, do not depend upon whether any or all of the 12 signal channels are in use.

The design of this transmitting amplifier-oscillator is shown in simplified form in Figure 31-13. It will be noted that there are two distinct feedback paths. One of these is a standard negative feedback which includes a gain adjusting pad by which the amplifier gain may be set at its desired

value. The other feedback path, which includes a resistance lamp having a negative temperature coefficient and an oscillator coil, is positive. Oscillations set up in the 60 kc tuned tank circuit will increase in magnitude until the current flowing in the resistance lamp reduces its resistance to a point where the loss in the positive feedback path is just equal to the amplifier gain. Currents flowing in the oscillator coil are limited to frequencies in the immediate neighborhood of 60 kc because of the tuned tank circuit. Any signal currents present in the amplifier output, however, can flow through the resistance lamp, thus adding to the heat produced by the 60 kc current. Because the operating resistance of the lamp must retain a value that will cause the loss in the feedback path to equal the amplifier gain, the value of the 60 kc oscillating current is reduced accordingly. The total output of the amplifier, therefore, is held at a constant value.

The twist amplifiers which, as has already been stated, are located at intervals of 100 to 200 miles along the line and at receiving terminals, perform three separate regulating functions as well as providing the gain required to compensate for the normal loss of the immediately preceding repeater section. The regulating functions are under the control of the 12, 28, and 56 kc pilot channels previously mentioned. The output power of these pilot channels at the amplifiers is very low (-11 dbm) as compared to the +15 dbm power of the full-strength 60 kc pilot channel. The 60 kc pilot channel controls only the line amplifiers and has no effect on the twist amplifiers.

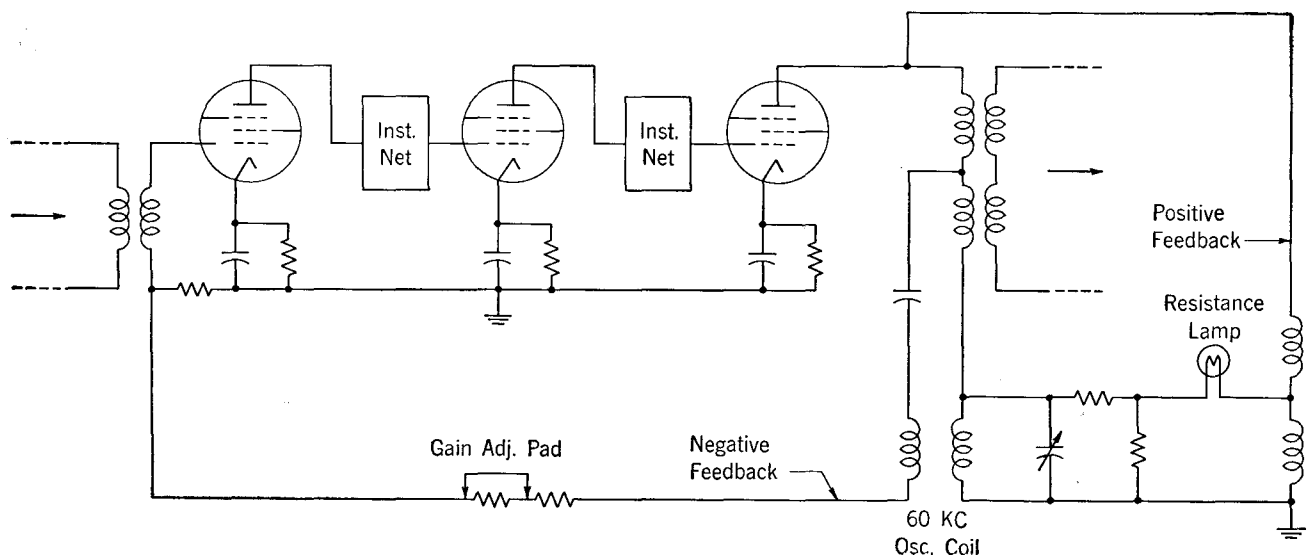


FIG. 31-13 TYPE-K TRANSMITTING AMPLIFIER

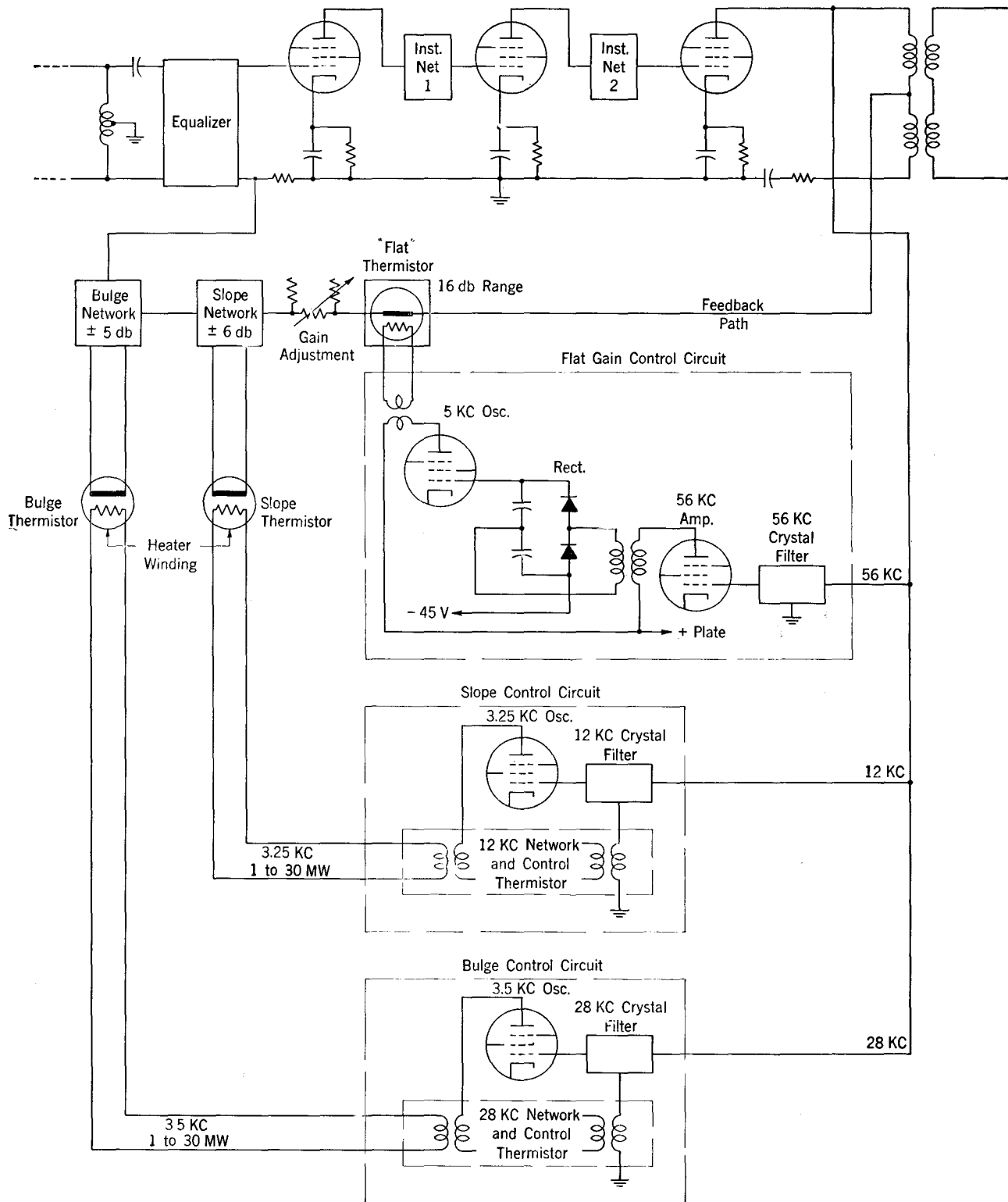


FIG. 31-14 SCHEMATIC OF TWIST AMPLIFIER

The 56 kc pilot channel regulates the twist amplifier gain to correct for temperature changes in the preceding repeater section in the same way that such regulation is effected in line amplifiers

under the control of the 60 kc channel. It also compensates for minor gain variations that may have been picked up in the preceding line amplifiers of the entire twist section.

The 28 and 12 kc pilot channels together control regulation to counteract twist effect. Twist, as previously discussed, is caused by the fact that the slope of the loss-frequency characteristic of a line section varies somewhat with temperature changes, and the further fact that the slope characteristic deviates somewhat from a straight line at temperatures above and below the reference operating temperature of 55°F. This deviation is known as **bulge**. Slope regulation is under the control of the 12 kc pilot and bulge regulation is under the control of the 28 kc pilot. All three pilot controls may function simultaneously and independently.

The normal slope of a repeater section, which is considerable in the total operating range between 12 and 60 kc, is taken care of in the usual manner by equalizers inserted in front of each line and twist amplifier. Slope regulation in the twist amplifier takes care of deviations from the normal 55°F slope, which may accumulate to as much as  $\pm 6$  db over a twist section. Similarly, bulge regulation corrects for as much as  $\pm 5$  db maximum deviation from the straight line slope characteristic, which maximum occurs at the 28 kc point on the curve.

Figure 31-14 shows schematically the general arrangement of the twist amplifier circuit. As

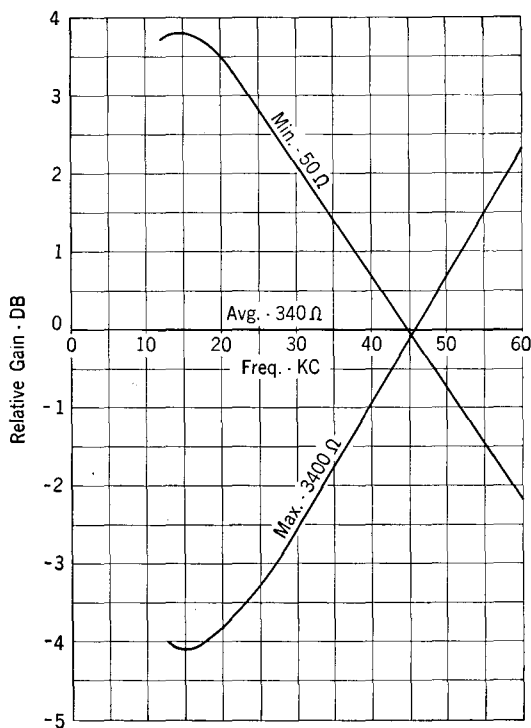


FIG. 31-15 GAIN-FREQUENCY CHARACTERISTICS OF SLOPE NETWORK

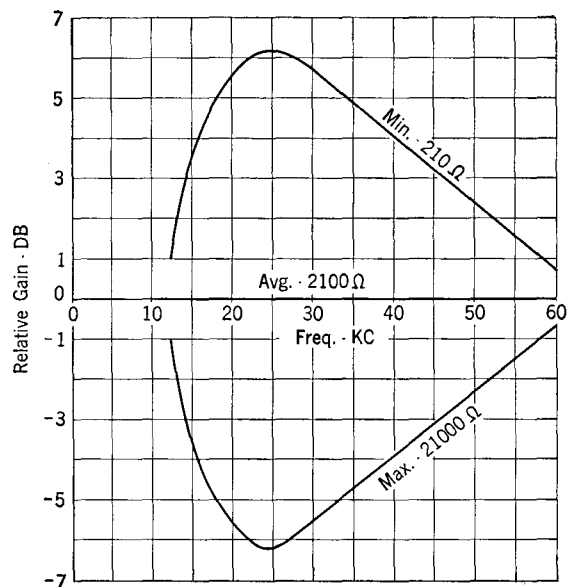


FIG. 31-16 GAIN-FREQUENCY CHARACTERISTICS OF BULGE NETWORK

there indicated, flat gain regulation is obtained by means of a "flat" thermistor inserted in the negative feedback path of the amplifier. This is an indirectly heated thermistor. Although the thermistor bead is in series with the feedback path, the thermistor design is such that the 60 kc pilot and signal currents flowing through the bead do not affect its temperature appreciably as compared with the temperature change caused by the heater winding. The value of the current in the heater winding is determined by the value of the 56 kc pilot voltage by means of the flat gain control circuit. In this circuit, the 56 kc pilot voltage is selected by crystal filters, amplified and rectified. The rectifier output drives a 5 kc oscillator tube, the output of which is fed to the heater winding of the thermistor. The net effect of the control circuit is to produce large changes in the power supplied to the heater winding that are accurately proportional to small changes in the 56 kc pilot power. The normal pilot channel power level is at  $-11.0$  dbm and a change of this level to  $-11.55$  dbm will produce a change in the 5 kc output from 180 milliwatts to 68 milliwatts. This in turn will vary the resistance of the thermistor bead between 2500 and 18000 ohms, which will produce a net change in the amplifier gain of 16 db.

Twist regulation is effected by means of the slope and bulge networks which are also in series with the amplifier feedback path. These networks are actually variable equalizers of rather complex

design. Variation of the network transmission characteristic is controlled in each case by variations in the value of a single resistor which forms a part of the network circuit. These resistors are the beads of the indirectly heated slope and bulge thermistors.

Figure 31-15 shows the maximum range of the slope network characteristic which, of course, is continuously variable between the limits shown. As indicated, the network has zero slope when the resistance of the slope thermistor is 340 ohms. Negative slope is obtained when the resistance of this thermistor is between 50 and 340 ohms, and positive slope when it is between 340 and 3400 ohms. Similarly, Figure 31-16 shows the characteristic of the bulge network between its maximum and minimum correcting values, corresponding to resistance values of its thermistor ranging between 21000 ohms and 210 ohms.

As in the case of the 56 kc channel, the power supplied by the 12 and 28 kc channels is relatively low. It is selected in each case by a crystal filter and used to drive an oscillator tube in a control circuit, the output of which is fed to the heater winding of the thermistor. The slope control os-

cillator output produces a frequency of 3.25 kc, and the bulge control oscillator output is at 3.5 kc. The amount of power delivered by each oscillator tube to the respective thermistor heater windings is controlled by the amount of 12 or 28 kc pilot power at the output of the twist amplifier. The control circuits are so designed, however, that a small change in pilot level produces a large change in the thermistor heating currents supplied by the oscillators. This is effected by means of "control thermistors" which form a part of the control networks, and which act to hold the sum of the pilot power and the oscillator power at a substantially constant value. Thus, if the pilot power tends to decrease, the oscillator power and the resultant heater current increases, and vice versa. The design is such that a variation in pilot output level of about  $\pm 0.2$  db causes the heater power in the thermistors to vary over their entire operating range from 1 to 30 milliwatts. The slope and bulge thermistors are of the slow acting type, requiring many minutes to stabilize. This is in contrast to the flat thermistor associated with the 56 kc gain control circuit which stabilizes in a fraction of a second.

## CHAPTER 32

### CARRIER SYSTEMS—(Continued)

#### 32.1 Type-L Carrier Systems

Type-L carrier telephone systems are designed for application to coaxial conductors. The telephone terminal equipment may also be used on microwave radio systems or other mediums capable of handling an extremely wide range of frequencies. The original system, coded L1, was designed to provide 480 two-way telephone channels over a pair of coaxial conductors, using the frequency range between 68 and 2044 kc. Later development added 120 more channels for relatively short distance transmission, thus extending the frequency range utilized up to 2788 kc. The 600 voice channels, each 4 kc wide, are placed in their line frequency allocations by means of three modulation steps in the manner indicated in Figure 32-1. The first modulation step places 12 voice channels in the 60 to 108 kc range to form a channel bank identical with the channel banks used in the J and K systems, as previously discussed. In a second step of modulation, five channel banks are translated to the frequency band between 312 and 552 kc. This constitutes a **basic supergroup** of 60 voice channels. The final modulation step translates the supergroups to appropriate line frequency positions as shown in Figure 32-1, which also indicates the carrier frequencies used in the group and supergroup modulators. It may be noted that supergroup No. 2 does not require modulation but is placed directly on the line in the 312-552 kc position. In practice, of course, any number of supergroups up to the total of 10 may be used, depending upon the total circuit requirements.

Figure 32-2 shows diagrammatically the essential elements of the L1 terminal. The group and supergroup modulators and demodulators employ copper-oxide disc varistors in a bridge arrangement similar to that of Figure 29-8. Intermediate amplifiers are included as required to compensate for the losses caused by the filters and hybrid coils, which form a necessary part of the total circuit. A group pilot of 92 kc is applied at the input of the modulator for each group, as indicated in Figure 32-2. On the receiving end, this pilot frequency is picked off at the output of the

auxiliary amplifier following the group demodulator to provide a rapid means of trouble location. It may also be used as a means of system lineup and as a convenient check on transmission.

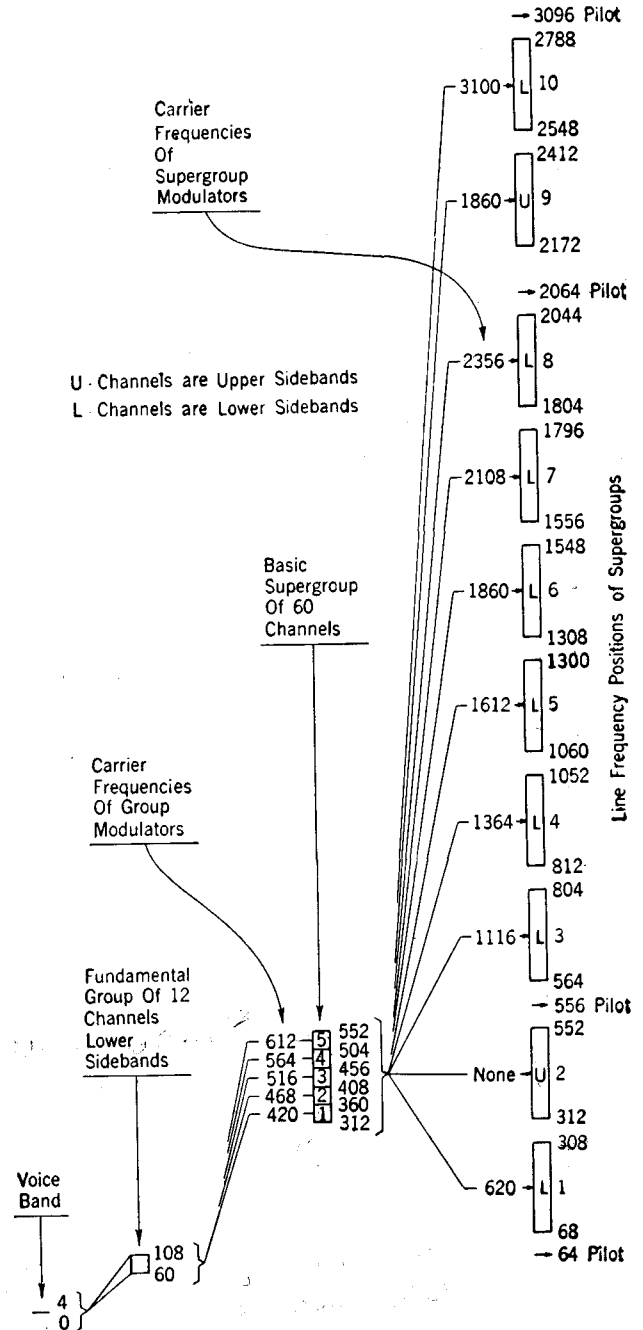


FIG. 32-1 FREQUENCY TRANSLATIONS OF L1 SYSTEM

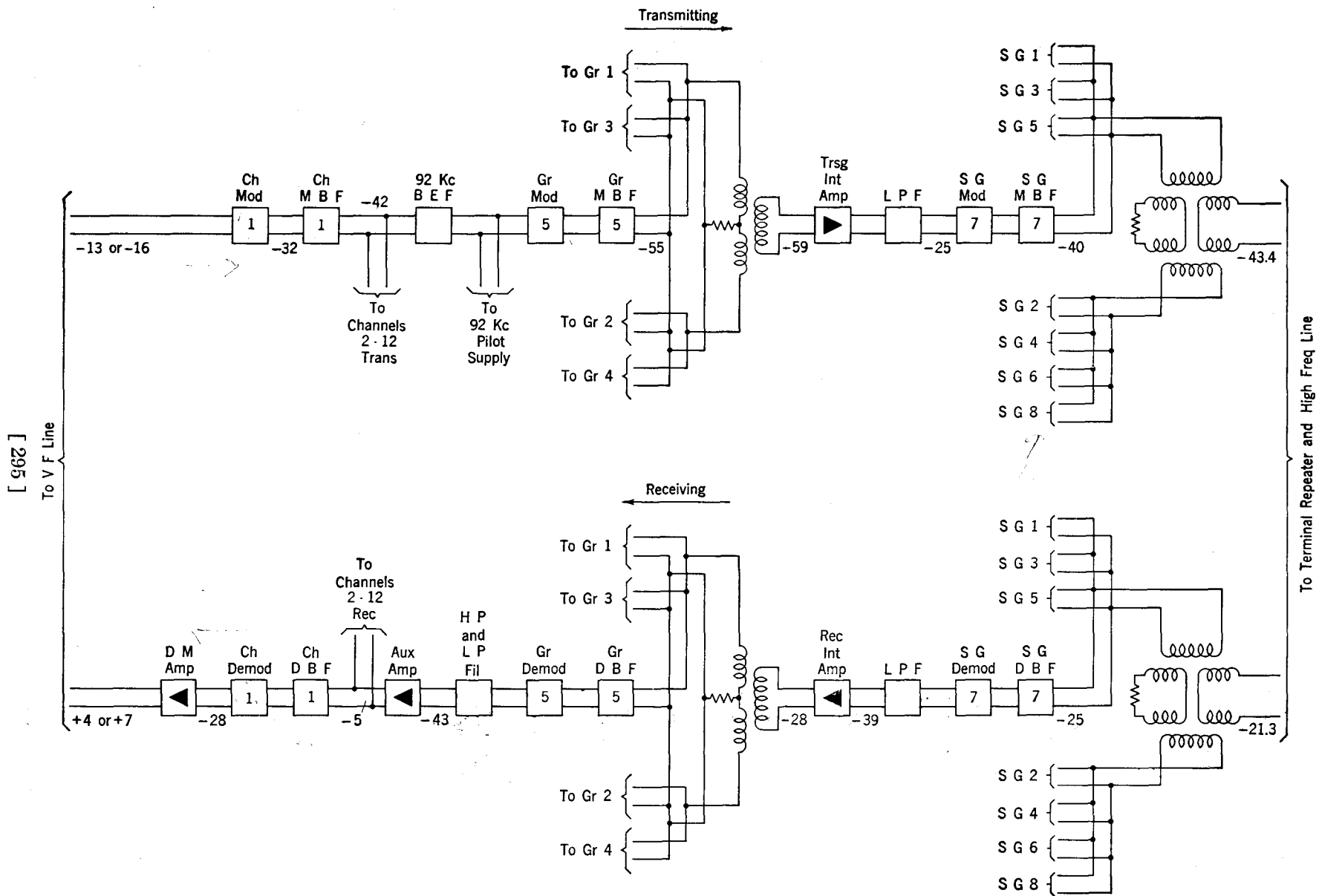


FIG. 32-2 TYPE-L1 CARRIER TERMINAL



Transmission over the high frequency line of the wide band of frequencies to which the 600 voice channels of the L1 system are translated requires the use of many amplifiers and associated regulating and equalizing equipment. The general arrangement for transmission in one direction is shown in block schematic by Figure 32-3. The line section illustrated is designated as a "switching section". Circuit dropping or branching cannot occur within a switching section. These sections, therefore, may extend from one terminal to a distant terminal; or to an intermediate point which may or may not be arranged for terminating some part of the carrier channels, or for connecting some or all of them to other transmission paths. In the case of .375 coaxial, amplifiers must be located along the line at intervals of approximately 8 miles. Most of these are "auxiliary repeaters" which consist only of amplifiers, basic equalizers, and associated regulating equipment.

At intervals usually in the order of about 100 miles but in no case greater than 165 miles, "main repeaters" which include adjustable equalizers as well as amplifying, regulating and power supply equipment, are provided. These main repeaters may be of the switching or non-switching type

depending on maintenance and overall circuit layout requirements.

Sixty-cycle a-c power for the operation of auxiliary repeaters is fed from terminal and main repeater points over a series loop made up of the two center conductors of the pair of coaxials used for the two directions of transmission. The 60-cycle currents are separated from the high frequency transmission currents by means of power separation filters. Since as many as eleven auxiliary repeaters and a terminal or half a main repeater must sometimes be so supplied in series, the voltage applied at the source may be quite high (up to 2000 or more volts) in order to overcome the resistance drops along the line, and the drops in the heater elements of the vacuum tubes of the repeaters served.

As indicated in Figure 32-3, the output of the transmitting terminal is supplied to the terminal repeater at a level of -50 db. It is transmitted to both the working line and a spare line by means of a hybrid coil. The spare line circuit is not shown in the diagram but is identical with that of the working line. At the transmitting end of a repeater section, a transmitting amplifier of the type described in Article 27.2 raises the level to

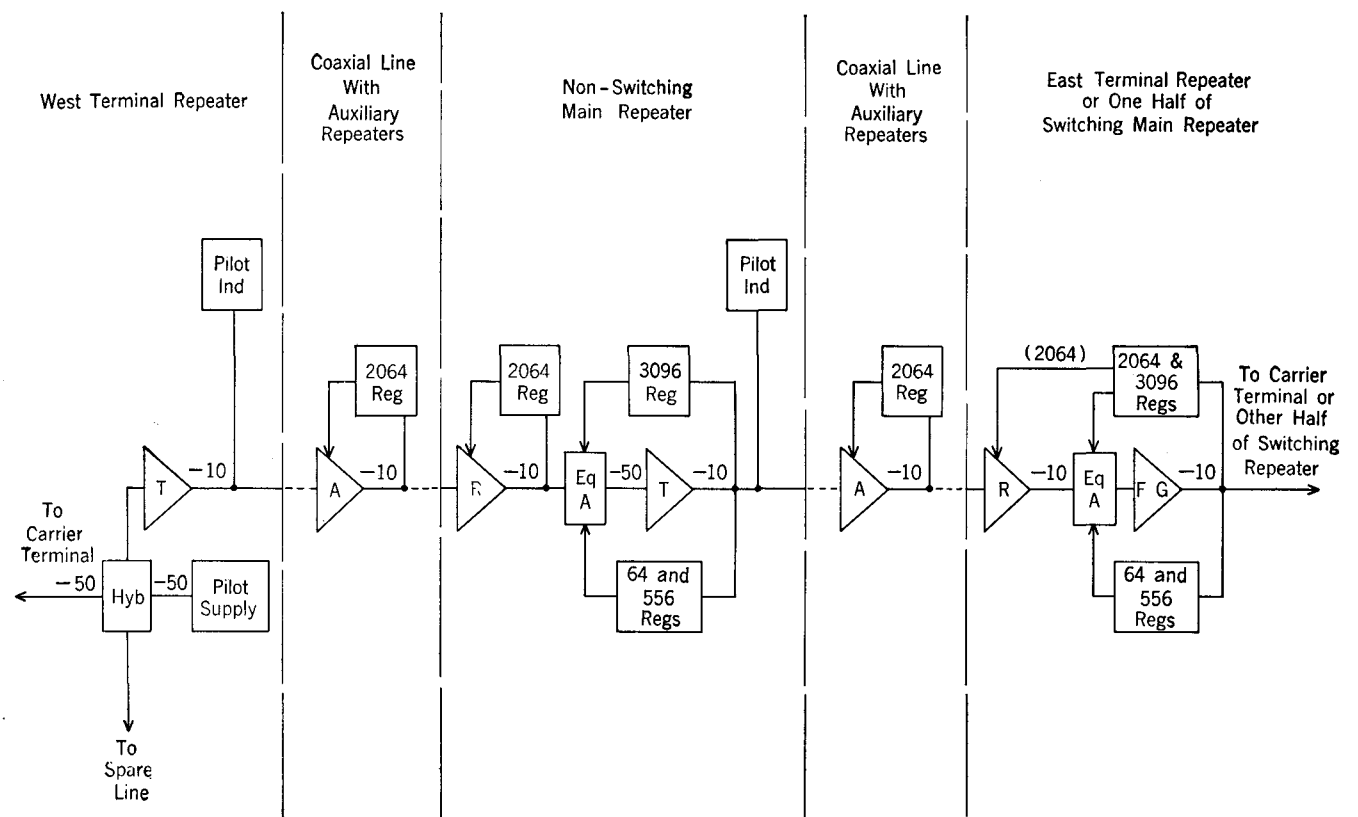
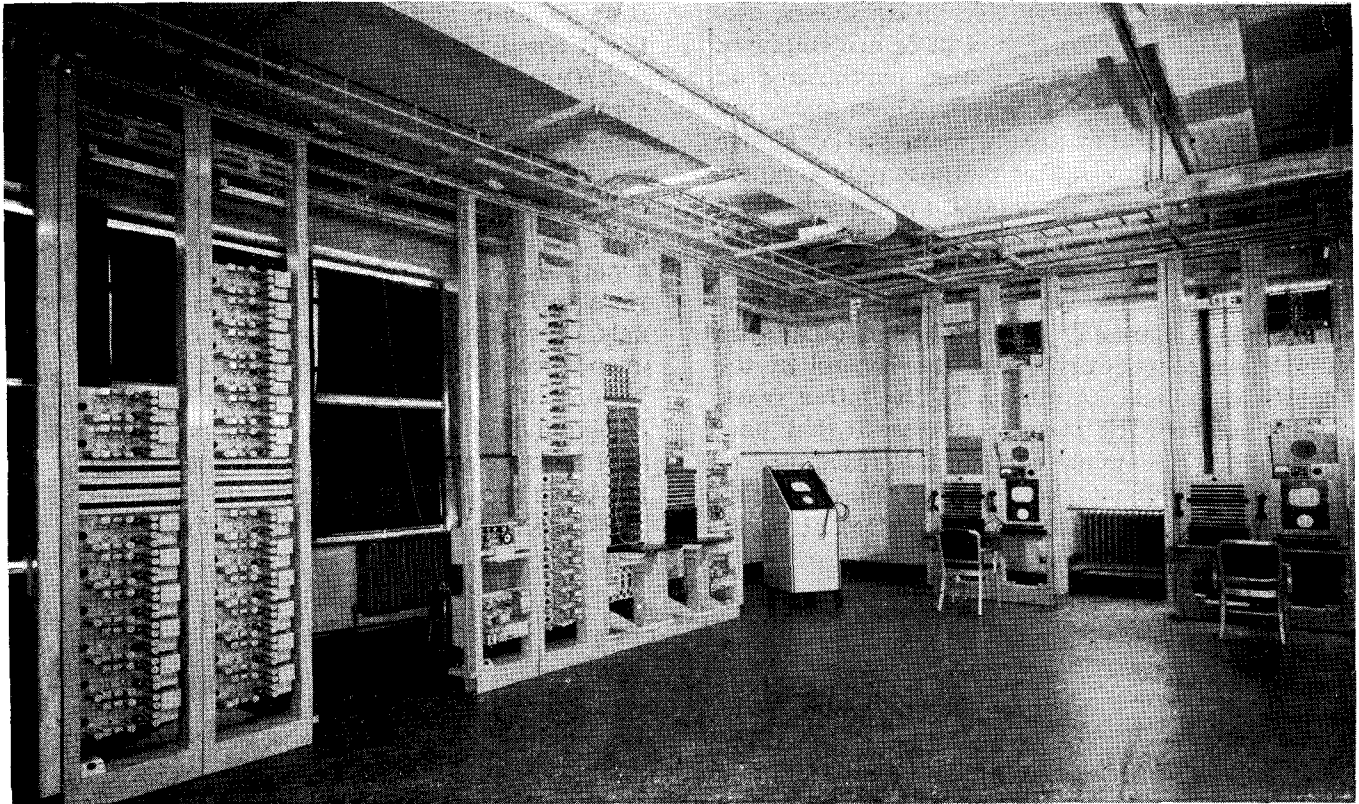


FIG. 32-3 BLOCK SCHEMATIC OF L1 CARRIER SWITCHING SECTION



TV PROGRAM NETWORK CONTROL AND SWITCHING CENTER

—10 db and applies it to the coaxial line. Auxiliary amplifiers along the line counteract the losses in each preceding 8-mile section of line, bringing the levels up again to —10 db at their outputs. At the receiving end of a section, switching main and terminal repeaters include a basic equalizer and receiving amplifier, which serves the same purpose as the auxiliary amplifiers, adjustable equalizers and one or more flat gain amplifiers. Non-switching main repeaters are similar except that flat gain amplifiers are not normally required.

At the transmitting hybrid coil, pilot voltages are supplied at the four frequencies of 64, 556, 2064 and 3096 kc. The 2064 kc pilot is used to vary the gains of the auxiliary and receiving amplifiers to compensate for line attenuation changes caused by temperature variations. The regulators are of the dynamic type, acting to hold the 2064 kc pilot power at the output of each regulated amplifier at a practically constant value. Their design is similar to that of the regulator shown in Figure 31-14. The 2064 kc pilot is picked off at the output of the amplifier by a selective filter, amplified and rectified (in this case by means of a vacuum tube). The rectifier output is applied to the grid of a 2000-cycle vacuum tube oscillator whose output in turn drives the heater element of

an indirectly heated thermistor inserted in the feedback circuit of the amplifier. As in the case of regulating circuits previously discussed, the design is such that a small change in the pilot input causes a relatively large change in the thermistor heating current and, consequently, in the amplifier gain.

Where coaxials are in underground or buried cable, it is the general practice to use automatic regulation only at every other auxiliary repeater. Alternate repeaters are then arranged for manual regulation. Such regulation is also effected by means of the thermistor in the amplifier feedback, which is indirectly heated by current supplied through a manually adjusted potentiometer. In this case, a disc-type thermistor is associated with the gain control thermistor to compensate for the effect of ambient temperature changes.

Because of the very wide band of frequencies employed, it is evident that equalization in Type-L systems presents a larger problem than in other carrier systems. Major equalization is effected, as in all other transmission systems, by means of fixed equalizers which are inserted at appropriate points, although not shown in Figure 32-3. To take care of variations in line and apparatus transmission characteristics that are functions of

temperature or frequency, special equalizers, designated equalizer *A* in Figure 32-3, are employed at main and terminal repeater points, and at "equalizing auxiliary stations". These latter are similar to the non-switching main station shown in Figure 32-3 except that power supply arrangements are omitted. The equalizers are actually a series of networks, each designed to modify the shape of the transmission characteristic over a partial range of the total frequency band. Some of them are manually adjustable and others are adjusted automatically under the control of the 64, 556 and 3096 kc pilots. In the latter cases, dynamic regulating circuits similar to that used by the 2064 kc pilot, control the resistance of thermistors which form a part of the equalizing networks. Pilot regulators are omitted at equal-

izing auxiliary stations and manual adjustment of the *A* equalizer at these points is necessary.

A Type-L carrier system, coded L3, has been designed to operate over a broader frequency band than the L1 system. This design provides for a maximum of as many as 1860 two-way telephone channels in the frequency range between 312 and 8284 kc. As shown in Figure 32-4, ten 60 channel supergroups are modulated with appropriate carriers to form a **master group** of 600 voice channels. The first master group is placed in the frequency range between 564 and 3084 kc, the second between 3164 and 5684 kc, and the third between 5764 and 8284 kc. In addition, a single supergroup may be transmitted below master group No. 1 in the basic supergroup range of 312 to 552 kc.

Where this system is applied to .375 inch co-

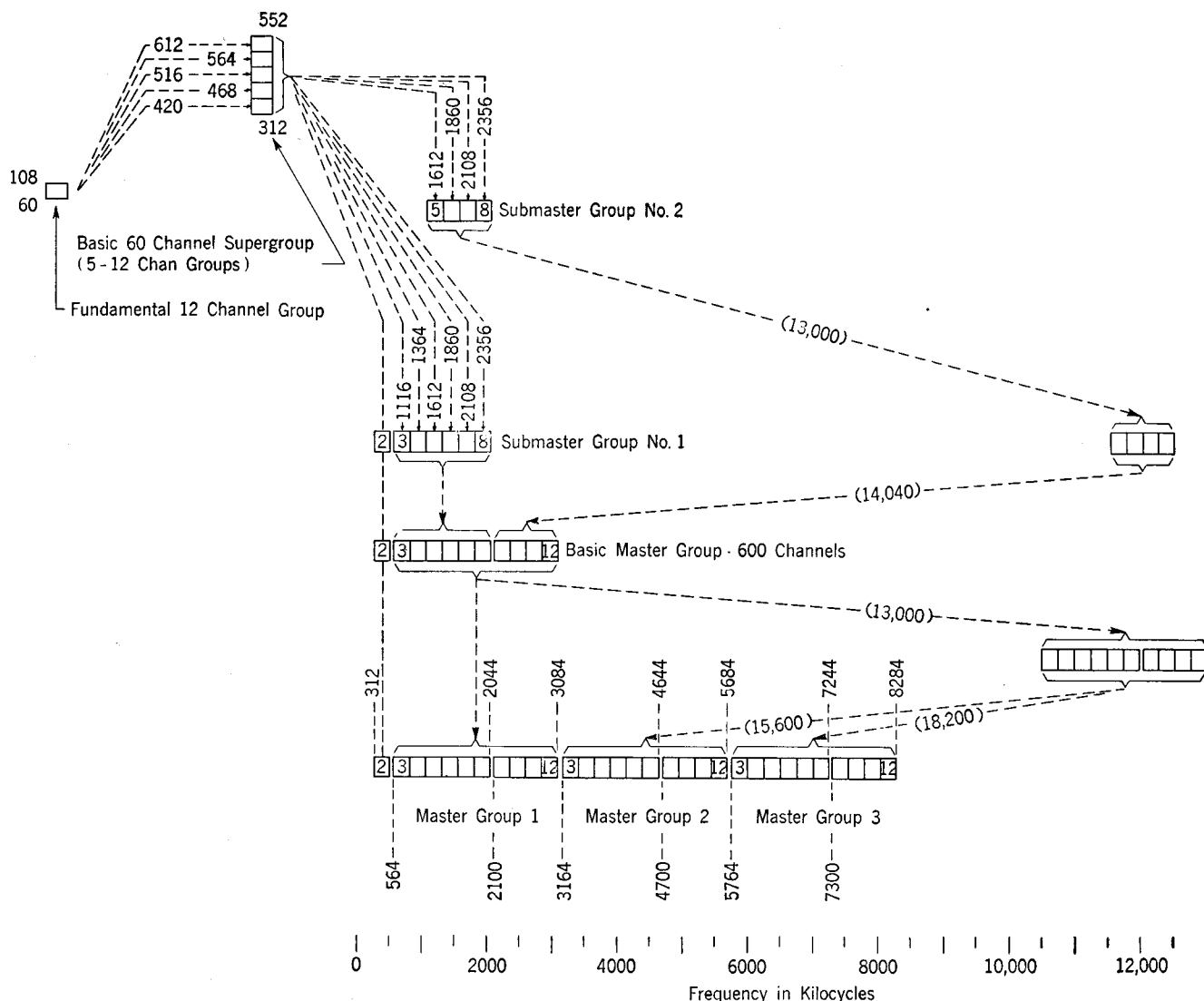


FIG. 32-4 FREQUENCY TRANSLATIONS OF L3 SYSTEM .

axials, it will require amplifier spacing at 4-mile intervals instead of the 8-mile intervals of the L1 system in order to maintain the high-frequency energy along the line at workable levels. The extremely broad frequency band also requires a different amplifier design, employing a new series of very high transconductance vacuum tubes developed especially for this purpose. The amplifier uses three tetrodes and two triodes of this special design. Unlike the amplifiers of the L1 system, the new tubes are socket-mounted and are not paralleled. Power for the operation of the auxiliary repeaters is supplied from main repeater stations in the same way as in the L1 system. Since the number of intermediate repeaters is doubled, however, the maximum voltage between the two coaxial center conductors will be considerably higher than in the L1 system.

Regulation in L3 systems employs six pilot frequencies at 308, 556, 2064, 3096, 7266 and 8320 kc. The 7266 kc pilot controls amplifier gains compensating for line attenuation changes due to temperature variations. The other five pilots control adjustable equalizers at main and terminal repeater points. Dynamic regulation under the control of the 7266 kc pilot is used, in general, at every other auxiliary repeater. The alternate repeaters employ a new method known as **thermometer control** regulation, in which a thermistor, mounted in a waterproof container, is buried near the underground coaxial cable to sense variations in ground temperature. This thermistor in turn controls a thermistor inserted in the amplifier feedback circuit.

### 32.2 Television Transmission on Type-L Carrier Systems

Type-L carrier systems are used for the transmission of television signals as well as for multiple channel voice transmission. Transmission of a television signal necessarily requires the employment of a very wide band of frequencies. This results from the fact that, as is well known, television depends upon the repetitive detailed scanning of a scene at extremely rapid intervals. Standard practice in the United States for black and white television calls for 525 horizontal lines for each complete scanning of the scene and for 30 complete scans per second, with the reproduced image having a width to height ratio of 4 to 3. In practice, a single complete scan or "frame" is accomplished in two steps. In the first step, the scene is scanned over the odd-numbered  $262\frac{1}{2}$

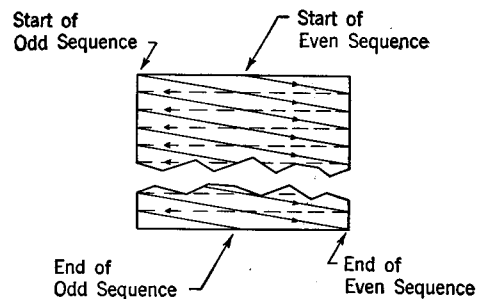


FIG. 32-5 TV SCANNING SEQUENCE

lines to form one "field"; and in the second step, it is again scanned over the even  $262\frac{1}{2}$  lines. This procedure, known as interlaced scanning, effects the eye of the viewer of the image as if the total scene were being reproduced 60 times per second instead of 30, and thus minimizes "flicker".

The scanning sequence is shown in Figure 32-5. For each line, the electron beam in the television camera and in the cathode-ray receiving tube moves horizontally across the image. At the same time it moves vertically downward a distance corresponding to two lines, under the control of the sweep circuit voltages applied to the deflecting plates or coils of the tubes. The scanning beam is blanked out at the completion of each horizontal line and returned quickly to the starting point of the next line, as indicated by the dotted lines in the Figure. The process is repeated until the bottom of the image is reached. The beam is then blanked out for a longer interval while it is returned to the top of the image for the start of the next scanning sequence. The duration of each scanning line is 54 microseconds and 9.5 microseconds are allowed for the horizontal retrace of the beam. The image is scanned at the rate of 15,750 lines per second.

To maintain the exact synchronization between the camera and the receiver that is obviously necessary, synchronizing pulses generated at the image pickup point are applied to the camera tube and transmitted to the receiver along with the image signals. The synchronizing pulses are superimposed on the signal blanking pulses in such a way that they can be "clipped" from the image signal and applied to the saw-tooth generators which control the deflections of the scanning beam. As previously noted, the horizontal synchronizing pulses must recur at the rate of 15,750 per second and the vertical pulses, which return the beam from the bottom to the top of the image, must recur at the rate of 60 times per second.

Figure 32-6 (A) indicates graphically the form

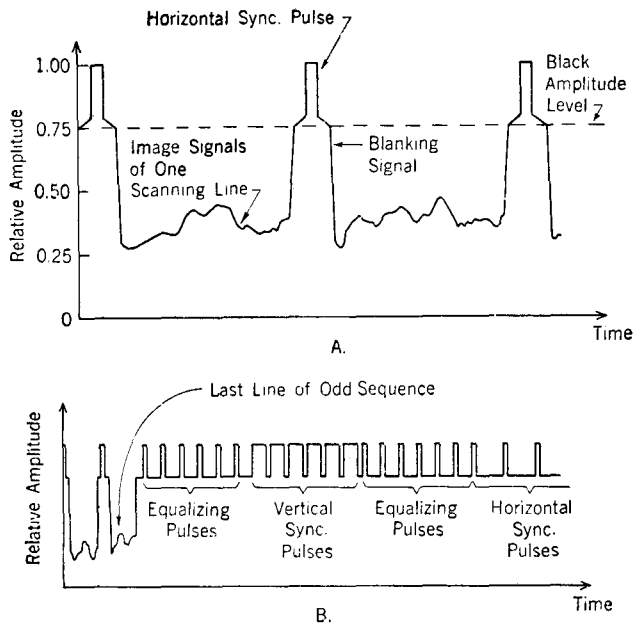


FIG. 32-6 TV SIGNAL FORM

of the TV signal at the receiver for two scanning lines covering a total time of 127.0 microseconds. The image signal, which is applied to the control electrode (grid) of the picture tube, may vary between zero amplitude for "white" and an amplitude which effectively blocks the electron beam to produce "black" in the image. The synchronizing signals, it may be noted, rise above the black level to a region sometimes called "blacker than black". Figure 32-6(B) illustrates the form of the longer vertical synchronizing pulse, which extends over a period of 190.5 microseconds. Ver-

tical and horizontal synchronizing pulses are separated for application to their proper respective deflecting coils by means of a simple *RC* timing circuit which recognizes the large difference in their time duration. The vertical pulse is "serrated" as shown so that the horizontal pulses will continue during the vertical deflection period to avoid the possibility of their falling out of step. A series of "equalizing" pulses is included before and after the vertical synchronizing pulse to take care of the time factors introduced by the fact that the first scanning field is completed in the middle of a line, and the second at the end of a line.

Considering the transmission of the total television signal, it is evident that the indispensable synchronizing pulses alone make the signal rather complex. The part of the signal carrying the image must be much more complex if satisfactory image detail (resolution) is to be obtained. Thus if a scene is to be analyzed as the horizontal beam crosses it in the same detail as is provided by the 525 line dissection of the image vertically, the signal might take  $4/3 \times 525$  or 700 different values for each horizontal trace. This would correspond to a variation at the rate of 350 cycles per line which would mean  $350 \times 525 \times 30$  or approximately  $5\frac{1}{2}$  million cycles per second. Furthermore, if the scene being televised was one in which there were many transitions between black and white, such as a black and white checkerboard pattern, the image signal would tend to take the form of a square wave. Accurate transmission in such a case would theoretically involve frequencies

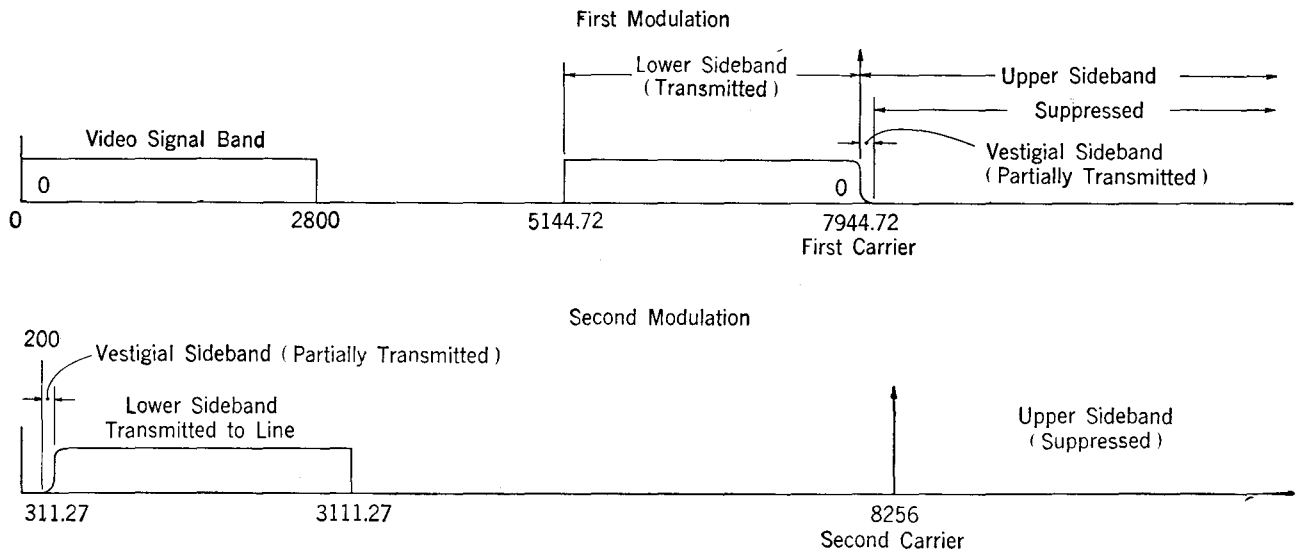


FIG. 32-7 VIDEO SIGNAL FREQUENCY TRANSLATIONS

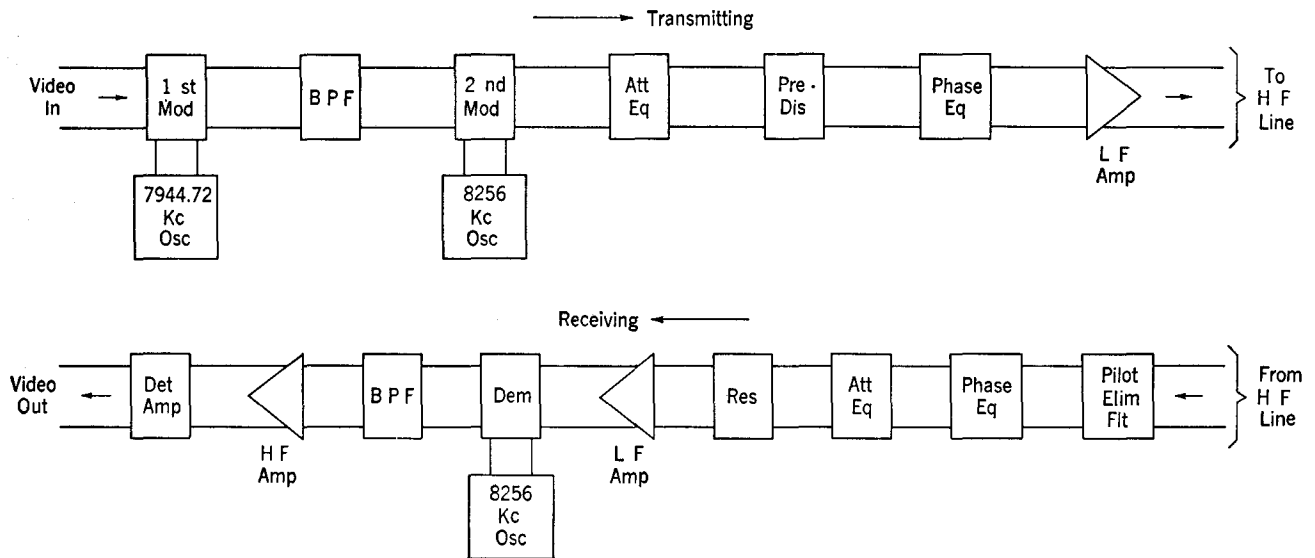
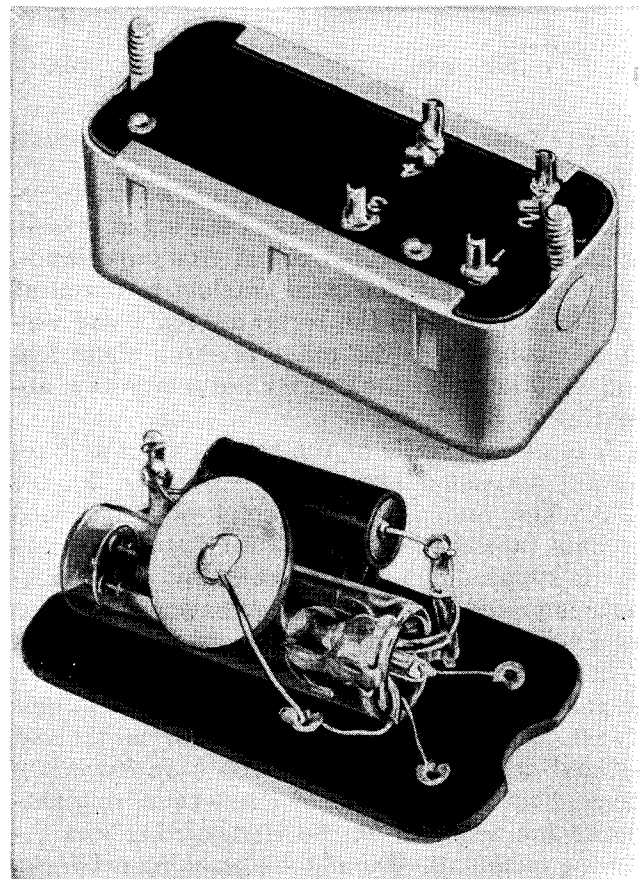


FIG. 32-8 BLOCK SCHEMATIC OF TELEVISION TERMINAL

extending toward infinity. Actually, practical experience indicates that entirely satisfactory resolution for black and white images is obtained from a video signal including frequencies up to a maximum of about 3 million cycles, although the standard broadcast TV signal is normally considered as 4 million cycles in width. In any event, it is to be noted that the lower frequencies are indispensable. Included here are the vital synchronizing pulses as well as the major values in the image structure. The higher frequencies become increasingly less important as they approach values which tend to enhance the detailed accuracy of the picture beyond the practical limit of perception of the normal eye. As might be expected also, the major energy content of the signal tends to be concentrated in the lower frequencies.

For transmission over the Type-L1 carrier system, the frequency range on the line between about 200 and 3100 kc is employed. The lower frequency is limited by equalization difficulties and the upper by the characteristics of the line repeaters. Since the standard video signal begins at about 30 cycles and may be considered as extending upward, in this case, to about 2800 kc, it is necessary to translate it by modulation procedures to place it in the proper position for transmission over the line. This is accomplished by two stages of modulation, as indicated in Figure 32-7. The carrier frequency of the first modulation stage is 7944.72 kc. A band-pass filter permits the passage of the lower side-band, extending from about 5100 kc up to the carrier frequency, plus a small part of the upper side-band, extending from the

carrier frequency up to about 8100 kc. This latter is known as a "vestigial sideband" and is included in the passed band to insure first, that there is no



THERMISTOR ASSEMBLY USED FOR GAIN REGULATION OF REPEATER AND GROUP AMPLIFIER

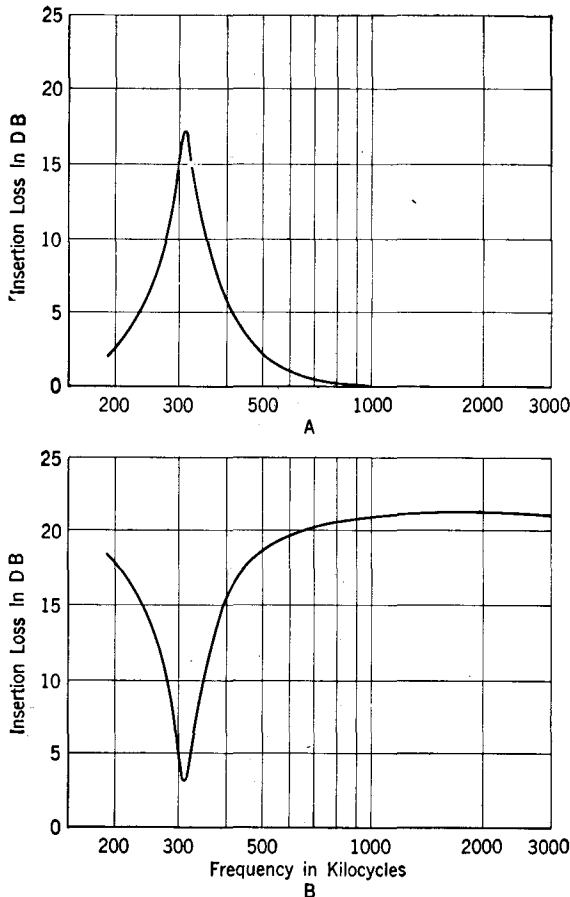


FIG. 32-9 CHARACTERISTICS OF PREDISTORTER AND RESTORER

clipping of the lower side-band; and second, to reinforce the lower frequencies of the signal which, as has already been noted, are of vital importance. The second modulation stage employs a carrier of 8256 kc to translate the foregoing main and vestigial sidebands to the range between about 200 and 3100 kc, with the carrier frequency now appearing at 311.27 kc.

A block schematic of the transmitting and receiving television terminals is given in Figure 32-8. The first modulator employs a balanced vacuum tube circuit to which the 7944.72 kc carrier is supplied by a crystal oscillator. The modulator is preceded by a two-stage balanced vacuum tube video amplifier, the second stage of which is a cathode-follower. The second modulator employs balanced vacuum tubes with input and output transformers. To maintain transmission flat and without delay distortion over the wide frequency range, both attenuation and phase equalizing networks are required at the transmitting and receiving terminals. The phase equalizing networks are needed principally to compensate for delay contributed by the band-pass filters.

The transmitting terminal also includes a special network known as a **pre-distorter**. As shown in Figure 32-9(A) this inserts a loss in the immediate neighborhood of the carrier frequency (311.27 kc) having a maximum value of about 18 db. The purpose is to reduce the relatively large energy values of the low frequencies and thus to

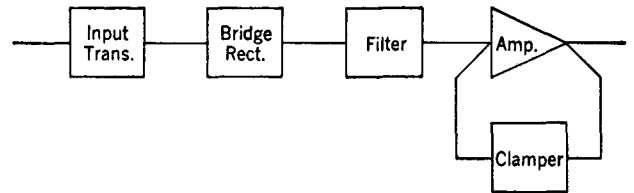


FIG. 32-10 BLOCK SCHEMATIC OF DETECTOR-AMPLIFIER

distribute the energy of the television signal applied to the line more evenly over the entire frequency range. At the receiving terminal, the signal is restored to its original form by a **restorer** network whose complementary characteristic is shown in Figure 32-9 (B). A three-stage line fre-

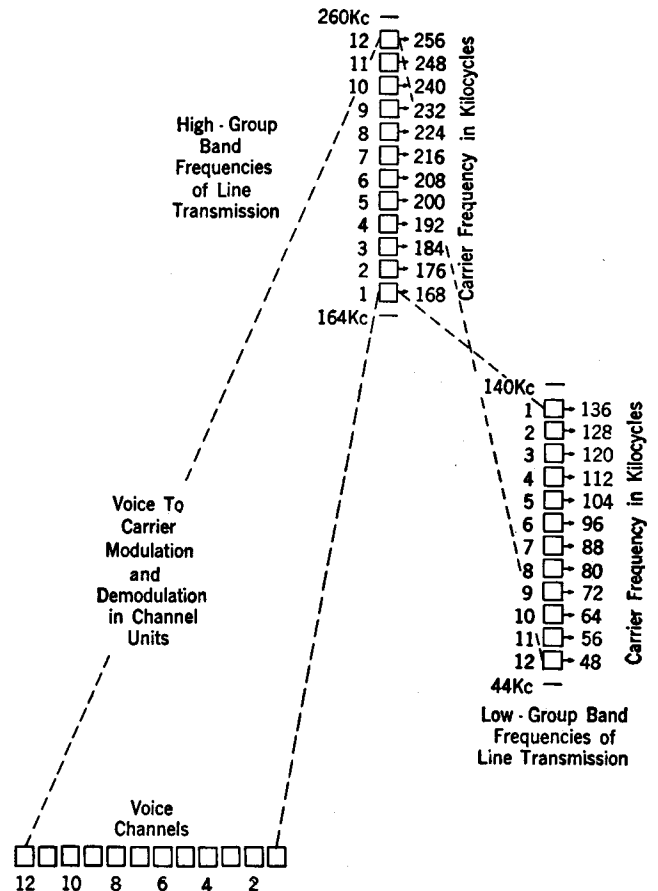


FIG. 32-11 TYPE-N CARRIER FREQUENCY TRANSLATIONS

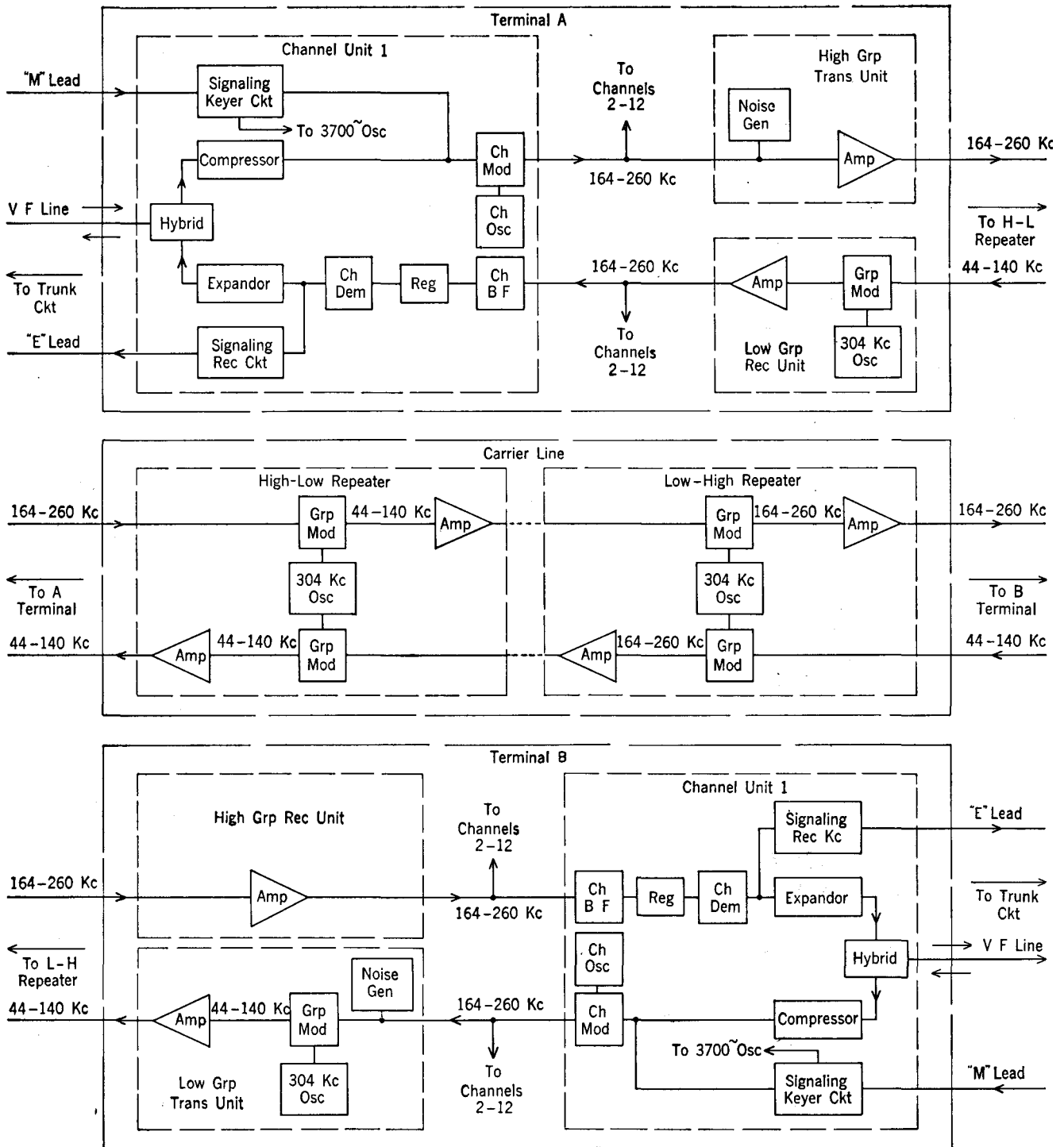


FIG. 32-12 BLOCK SCHEMATIC OF TYPE-N CARRIER SYSTEM

quency amplifier raises the signal level to the proper value for application to the high-frequency line.

The first apparatus unit in the receiving terminal is a pilot elimination filter which removes the 556, 2064 and 3096 kc pilot frequencies used

for regulation in the Type-L1 carrier system. These would naturally cause distortion of the received image if permitted to pass through to the demodulator. A demodulator, employing a carrier frequency of 8256 kc translates the incoming signal to the frequency range between about 5100



and 8100 kc. Its output, after passing through a band-pass filter, is amplified by a three-stage high-frequency amplifier having a nominal gain of about 31 db and transmitted to a detector-amplifier circuit which reproduces the original video signal. The employment of linear detection instead of a second stage of demodulation eliminates the difficult problem of supplying a local carrier frequency that would have the same frequency and the same phase as the carrier at the transmitting terminal.

As indicated by the block schematic of Figure 32-10, the detector-amplifier includes an input transformer, a germanium varistor bridge detector, a filter to eliminate the second harmonic of the transmitted carrier, and a clamper-amplifier. The amplifier proper is of three stages, with the third stage including four triodes connected for parallel operation. The clamper circuit, which is bridged around the amplifier, is a "d-c restorer". The TV signal at the output of the camera consists of a varying direct current which may be considered as an alternating current superimposed on a single-valued direct current. In passing through amplifiers and over the line, the direct-current component is naturally eliminated. At the TV receiver, however, the original varying direct current is needed. The clamper circuit restores the d-c component by effectively grounding the grid of the tube of the first stage of the amplifier during the horizontal synchronizing pulses of the signal. This automatically refers the entire signal to a reference base corresponding to the maximum synchronizing pulse amplitude.

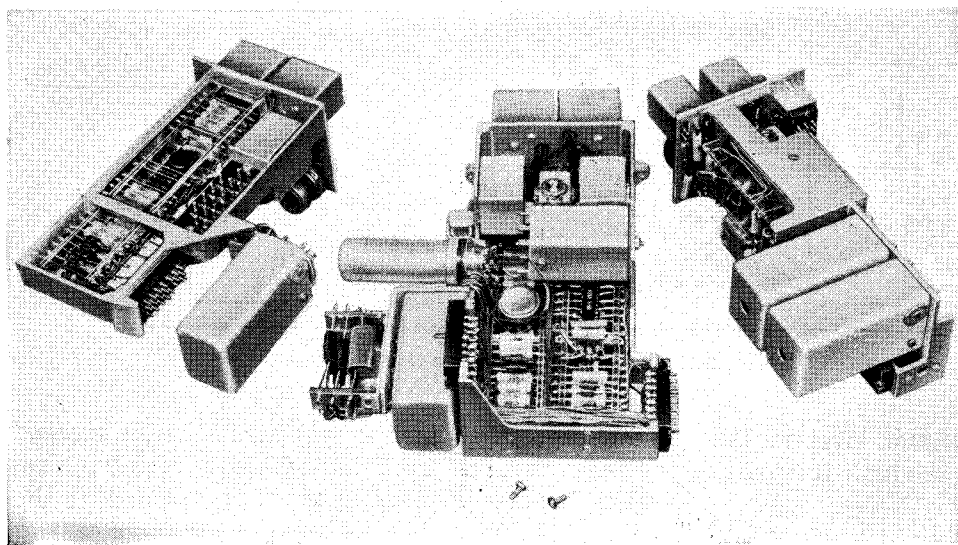
In other words, the output signal is held entirely above this base line instead of varying above and below a central reference. The clamping circuit employs two triodes connected as diodes, which become conducting during the synchronizing pulses that are picked off at the output of the amplifier.

### 32.3 Type-N Carrier System

The Type-N carrier system is designed to provide 12 two-way telephone channels on two non-loaded pairs of toll or exchange cables over distances from as little as 15 or 20 miles up to a nominal maximum of 200 miles. It operates on a 4-wire basis, and uses different frequency ranges on the line for transmission in the two directions, thus making possible the use of two pairs in the same cable. In order that carrier transmission over such short distances might be economically feasible, a number of interesting innovations have been introduced to keep the cost of terminal equipment at a minimum.

The system employs throughout miniaturized equipment components mounted compactly in lightweight plug-in unit assemblies providing for maximum flexibility and ease of maintenance. Devices known as "compandors" are employed in the channel units to compress the range of speech volume on the line and thus obtain a better signal to noise and crosstalk ratio. This eliminates the need for special crosstalk balancing and noise treatment of the cable pairs. The system uses double side-band transmission about carriers

spaced at 8 kc intervals to give channel bandwidths of about 250 to 3100 cycles. Double side-band transmission permits the use of relatively simple receiving channel band filters, and eliminates the need for transmitting channel band filters altogether. The channel carriers are transmitted on the line and serve to control the regulation of repeaters, thus obviating the need of a pilot for regulating purposes. A technique known as "frequency



TYPE-N CARRIER UNIT SUBASSEMBLIES. COMPRESSOR AT LEFT; EXPANDOR AND SIGNALING AT CENTER; CARRIER AT RIGHT

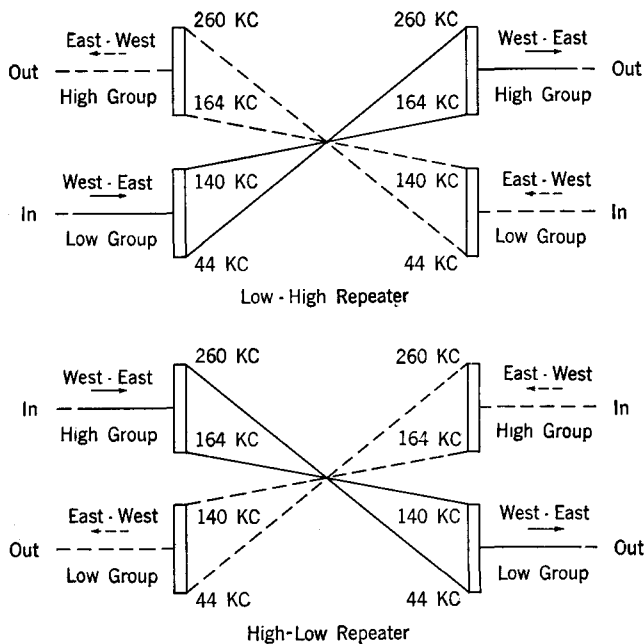


FIG. 32-13 FREQUENCY TRANSLATIONS AT REPEATERS

frogging” is employed whereby the frequency groups in each direction of transmission are transposed and reversed at each repeater so that all repeater outputs are always in one frequency band, and all repeater inputs in the other. This minimizes the possibility of “interaction crosstalk” around the repeaters through paralleling voice-frequency cable pairs. The reversal of the channel

groups at each repeater provides automatic self-equalization. The N-system also includes a unique arrangement for the transmission of dial pulses and supervisory signals, in which the signaling equipment is built-in as a component part of the terminal channel unit.

The channel frequency allocations used are indicated in Figure 32-11. The channel modulators and demodulators at the terminals use only the high group of carriers between 168 and 256 kc. But either this high group or the low group may be placed on the line, depending on the number of repeater sections and other factors. It is accordingly necessary to arrange each group terminal so that it can transmit or receive either high-group or low-group frequency bands. The four necessary group units are designated HGT, LGT, HGR and LGR. A high group transmitting unit always works in conjunction with a low group receiving unit and vice versa.

The general layout of the overall system is shown in block schematic in Figure 32-12. Here at terminal A, the high group is transmitted, and the HGT unit consists only of an amplifier and, when required, a controlled source of noise which may be introduced to over-ride crosstalk that might otherwise be heard on unusually quiet circuits. The incoming transmission at this terminal will be in the low group band. The LGR unit must therefore include a group modulator employing a

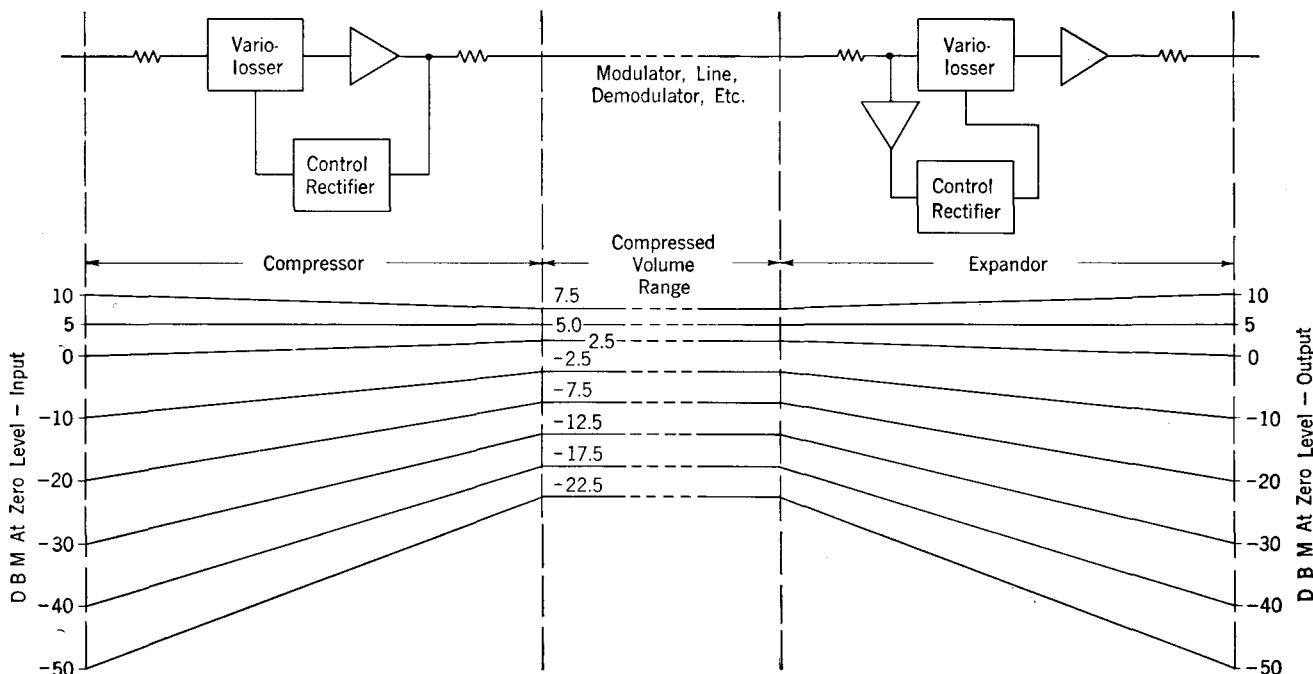
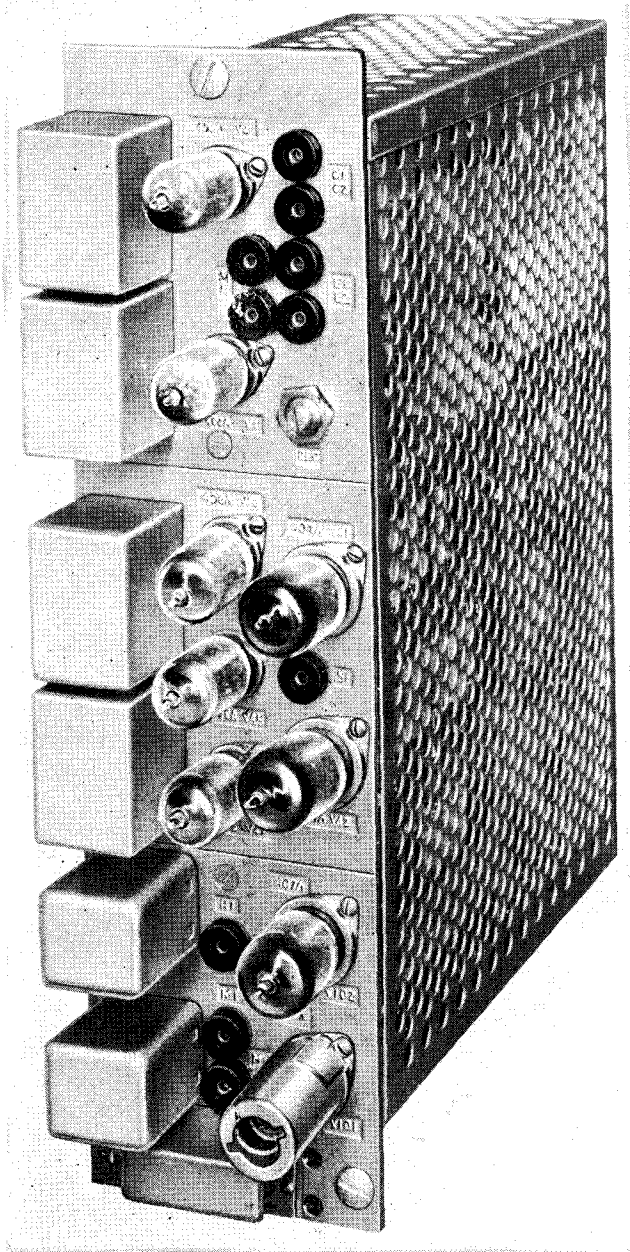


FIG. 32-14 COMPANDOR ACTION ON SIGNAL LEVELS



TYPE-N CHANNEL UNITS—FRONT VIEW

304 kc carrier to translate the incoming transmission to the 164-260 kc band before it is amplified and transmitted to the 12 channel units. At terminal *B*, on the other hand, the incoming reception in this case is in the high-frequency band and the receiving unit (HGR) consists only of an amplifier. The transmitting unit (LGT) here, however, includes a group modulator to translate the high-frequency band to the 44-140 kc band.

Repeaters along the carrier line alternate between "high-low" and "low-high". Each repeater

includes two group modulators for making the necessary translations from high group to low group and vice versa. Each such translation also reverses the location of the channels in the group bands, as indicated in Figure 32-13. These reversals automatically balance out the slope characteristic of two adjacent line sections to give a very nearly constant net loss across the 12-channel group. When there are an even number of line repeater sections, equalization is therefore practically complete. When the total number of sections is odd, equalization for the odd section may be taken care of by an appropriate equalizer at one terminal. The frequency reversals also reduce the maximum repeater gain to that required at the mean frequency instead of the highest frequency, and permit the use of two-stage rather than three-stage amplifiers.

With frequency frogging, the repeater spacings required for opposite directions of transmission are identical, which permits wider spacing than would be possible if transmission were in the high frequency group over the entire line. Normal repeater spacing for N-systems on 19-gage conductors is about 8 miles but must of course be less than that when smaller conductors are used. Power for the operation of repeaters can be supplied from a repeater office to an adjacent repeater on either side by a simplex connection over the line. Such repeaters may be arranged for pole mounting in cabinets. All amplifiers used in the system employ two stages with stabilized feedback. A thermistor is inserted in the feedback of the repeater amplifiers and the group receiving amplifiers to provide automatic regulation against line temperature variations. The resistance of these thermistors is controlled by the total value of the 12 channel carriers, whose power is at a much higher level than that of the voice currents. All thermistors are guarded against changes in the local ambient temperature by associated disc-type thermistors. The group modulators are of the double-balanced type employing copper-oxide disc varistors.

In addition to the compandor and signaling circuits, which are discussed later, the channel units include channel modulators and their associated oscillators on the transmitting side, and a channel band filter, regulator, and demodulator on the receiving side. The channel modulator is a bridge type arrangement of point-contact germanium varistors with its carrier supplied from a crystal controlled electron-coupled oscillator. In the receiving side of the circuit, the regulator is a two-

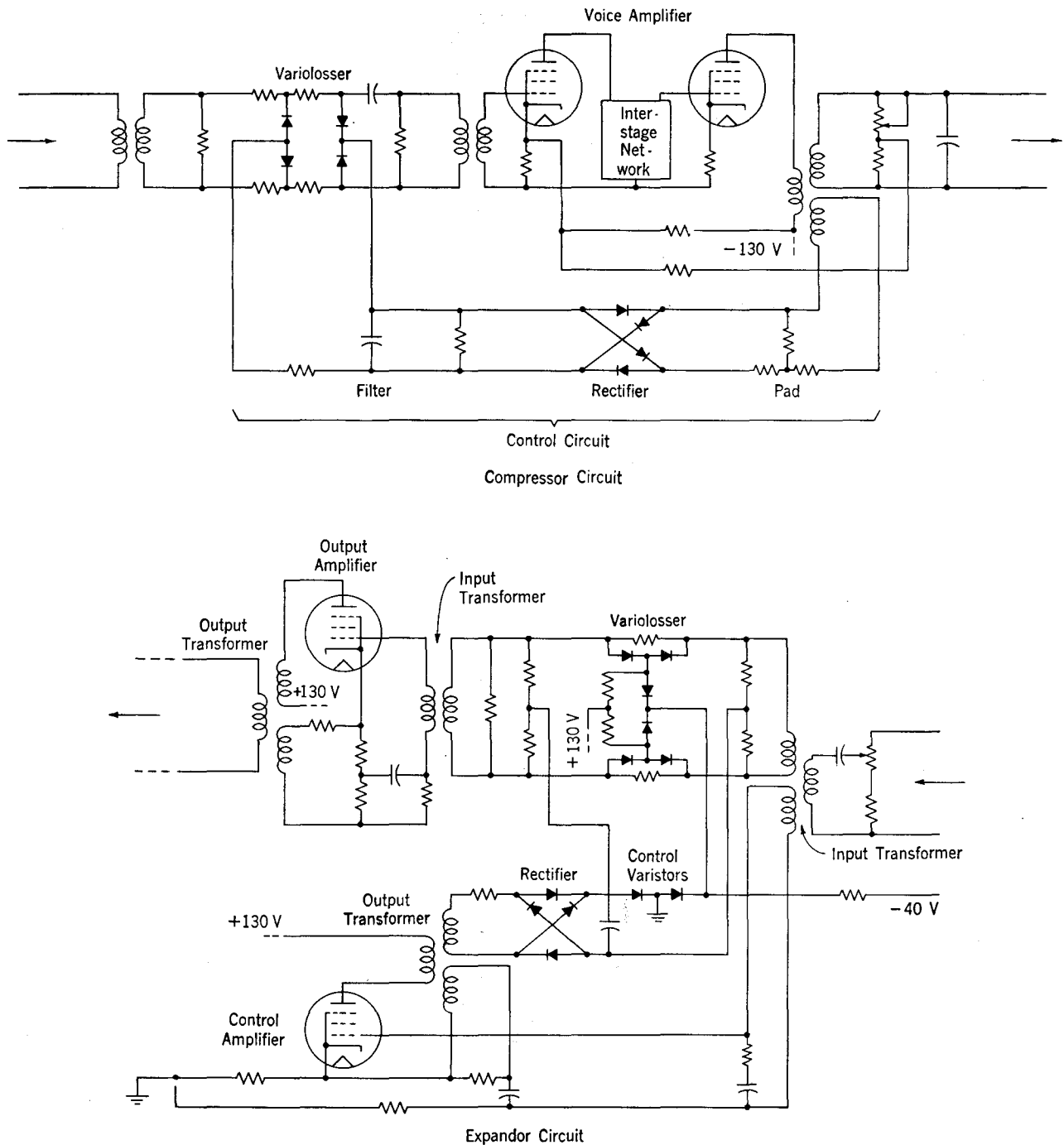


FIG. 32-15 COMPANDOR SCHEMATICS

stage amplifier arranged to provide delayed automatic volume control by the application to the grid of the input tube of rectified carrier picked off at the output of the demodulator. The demodulator is a lattice arrangement of germanium varistors, which operates as a linear detector.

As Figure 32-12 indicates, the **compandor** consists of two separate devices—a **compressor** and

an **expander**. Devices of this type have sometimes been used for noise reduction in long distance radio circuits and in Type-J carrier circuits, but are incorporated as an integral part of a system for the first time in Type-N carrier. The compressor provides a variable gain, the value of which depends upon the strength of the incoming voice signal, and the expander provides a variable

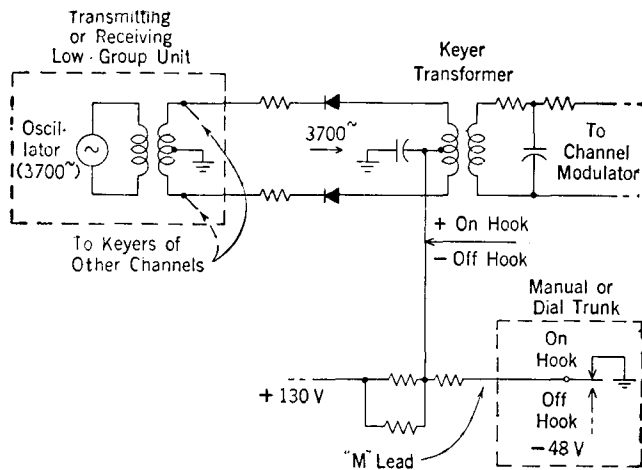


FIG. 32-16 SIGNALING KEYER CIRCUIT

loss which depends upon the same factor. The overall action of the two devices is illustrated in Figure 32-14 in terms of the signal levels. From this, it may be noted that a signal of  $+5$  dbm is applied to the line without modification by the compressor. A signal input of  $-50$  dbm, however, receives 27.5 db gain in the compressor and is applied to the line at  $-22.5$  dbm. The expander in turn introduces a loss of 27.5 db to return the signal to its original level. At intermediate values of input, the compression and expansion is proportionately smaller. For each signal input

weaker by 2 db than  $+5$  dbm, the compressor introduces 1 db more of gain and the expander 1 db more of loss. In other words, the compression ratio is 2 to 1.

Schematics of the compressor and expander circuits are shown in Figure 32-15. The overall gain of the compressor circuit is determined by the **varioloesser** which precedes the two-stage amplifier. This is effectively a resistance pad having two shunt arms whose resistance values vary in accordance with the value of direct current flowing through the germanium varistors of which they are composed. The direct current is obtained from the control circuit, which rectifies a portion of the amplifier output voltage. Thus, if the amplifier output rises, the direct current in the varioloesser increases, which increases its attenuation so that the input to the amplifier decreases. The design is such that a 2 db change in the varioloesser input results in only 1 db change in the varioloesser output. The control circuit contains a filter, whose time constants are such that the direct current varies in accordance with the syllabic envelope of the speech in the amplifier output. The expander circuit operates on the same general principles as the compressor. Its control circuit, however, is operated by the input speech signal rather than by the output signal.

The built-in signaling circuits of the N-system

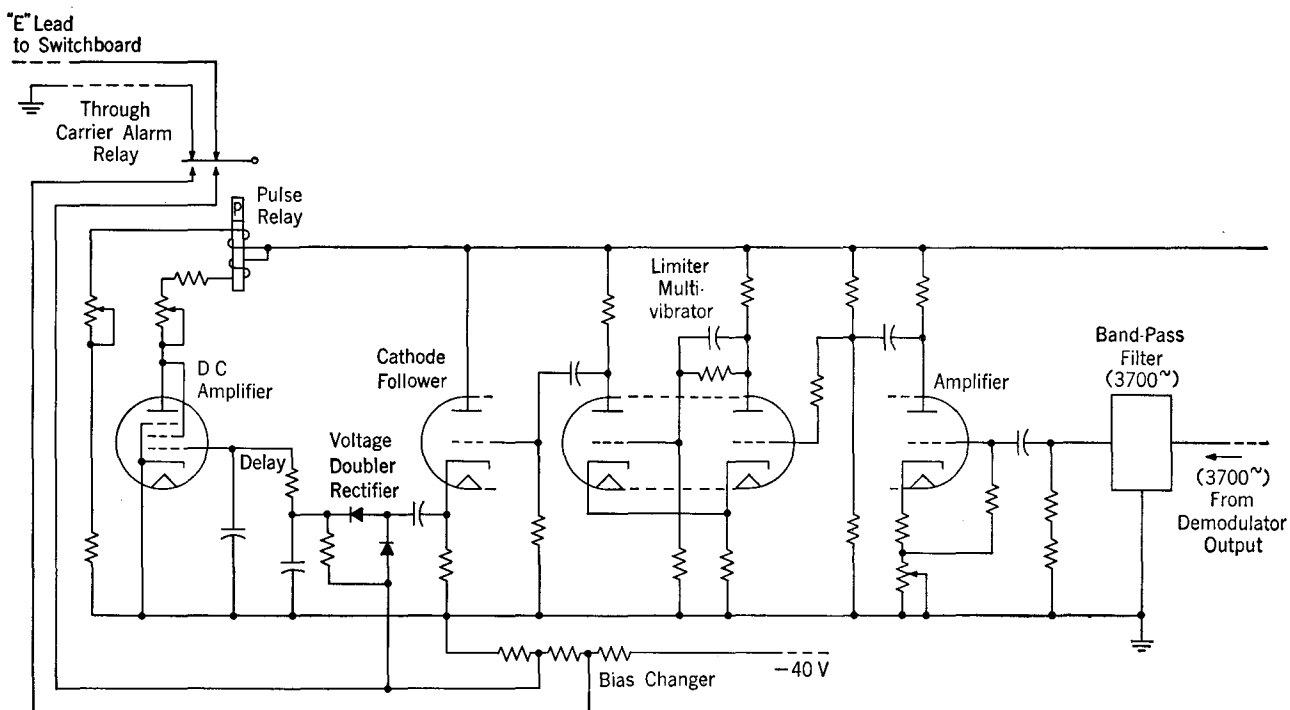


FIG. 32-17 SIGNALING RECEIVER CIRCUIT

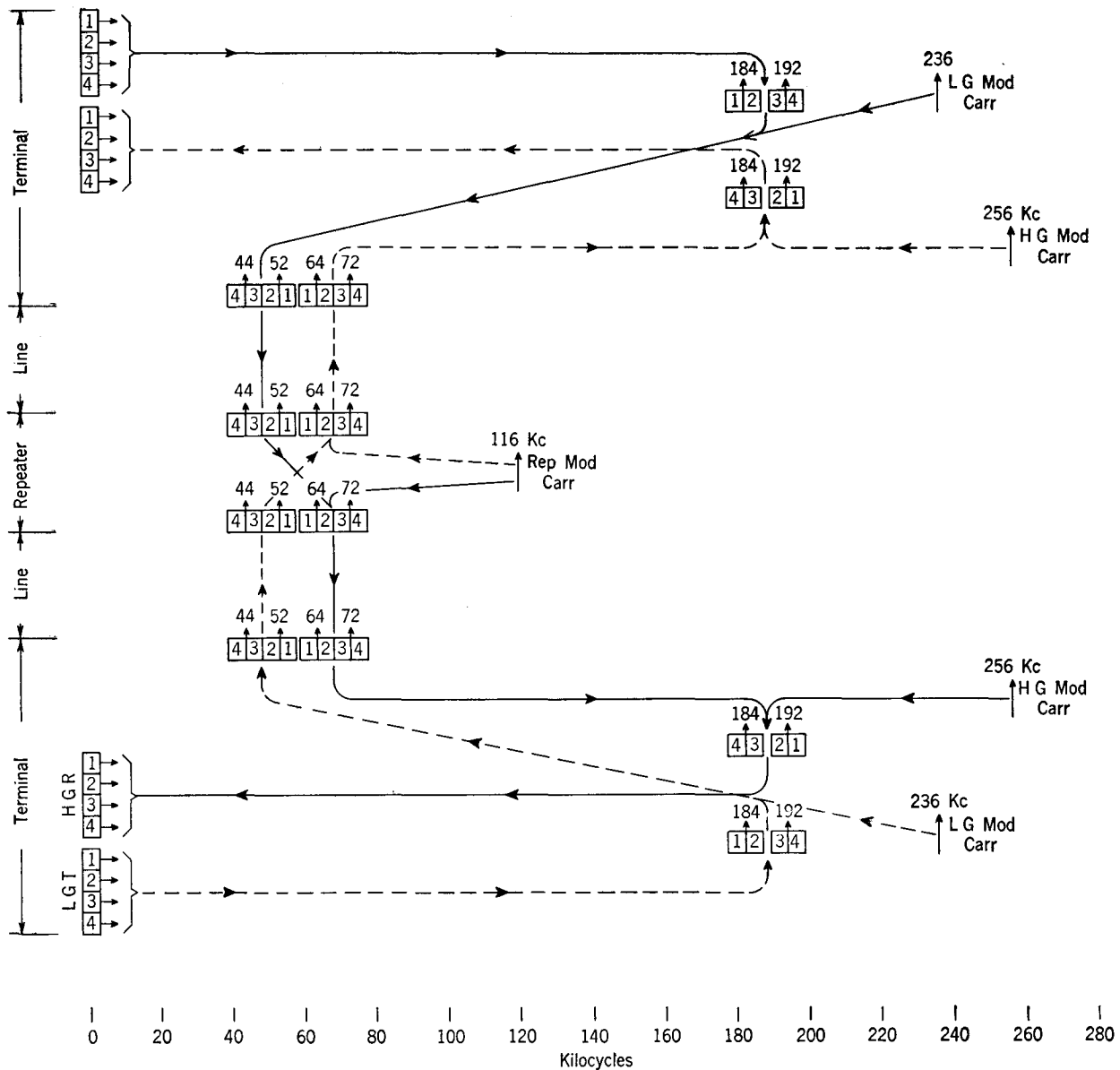


FIG. 32-18 TYPE-OB1 CARRIER FREQUENCY TRANSLATIONS

are designed primarily to handle dial pulse and supervisory signals. Ringdown signals at 1000 cycles and multi-frequency key pulsing signals are passed directly over the voice channel. Dial pulses and supervisory signals are transmitted by turning on and off a 3700 cycle signaling tone which is injected at the transmitting terminal following the compressor, and removed at the receiving end before reaching the expander. The 3700 cycle tones after modulation appear on the line as side frequencies 3700 cycles above and below the channel carrier frequencies. They are kept separated from the voice channels, which extend only up to 3100 cycles, by appropriate low-

pass and high-pass filters (not shown in Figure 32-12).

Figure 32-16 is a schematic of the signaling keyer circuit, which is included in each transmitting branch of each channel unit. It includes two germanium varistors which act to transmit, or to block, the 3700-cycle oscillator output. Thus for supervisory signaling, when ground is connected to the "M" lead from the trunk, positive potential causes current to flow in the varistors in such a direction as to make their resistance low and permit the transmission of signal tone to the channel oscillator. In the off-hook condition, on the other hand, the negative potential from the —48

volt battery is controlling. This causes the varistors to have high resistance, blocking the transmission of the signal tone.

On the receiving side, the 3700-cycle tone is picked off after the channel demodulator and applied to the signaling receiver circuit shown schematically in Figure 32-17. After selection by the band-pass filter, the 3700-cycle signal is amplified and applied to a limiter-multivibrator that transforms the sine wave signal, which may vary somewhat in amplitude, into a square wave signal of constant amplitude. This reshaping of the signal wave prevents the time-delay circuit, which follows, from introducing pulse width distortion that might cause erroneous registration of dial pulses. The output of the limiter-multivibrator is fed through an impedance-matching cathode-follower to a voltage-doubling rectifier whose output is connected through a delay network to the grid of a "d-c amplifier". The time constant of the delay network is such as to make the overall response time of the circuit 15 to 20 milliseconds, which prevents short noise bursts or other tran-

sients from causing false operation. The plate circuit of the d-c amplifier is connected to the operating winding of a mercury-contact polarized relay, which in operating disconnects ground from the "E" lead to the switchboard trunk. When no 3700-cycle tone is being received, the d-c amplifier tube is biased by -40 volts, no plate current flows, and the pulse-relay is held in its non-operated position by the current in its biasing winding. When 3700-cycle tone is present, the rectifier output voltage overcomes the tube bias permitting the flow of a large plate current. This operates the pulse-relay and removes ground from the "E" lead.

### 32.4 Type-O Carrier System

The Type-O carrier system is designed to provide relatively short-haul carrier channels over open wire conductors on an economic basis. It makes use of miniaturized equipment and many of the other features of the Type-N system including compandors, frequency-frogging and built-in

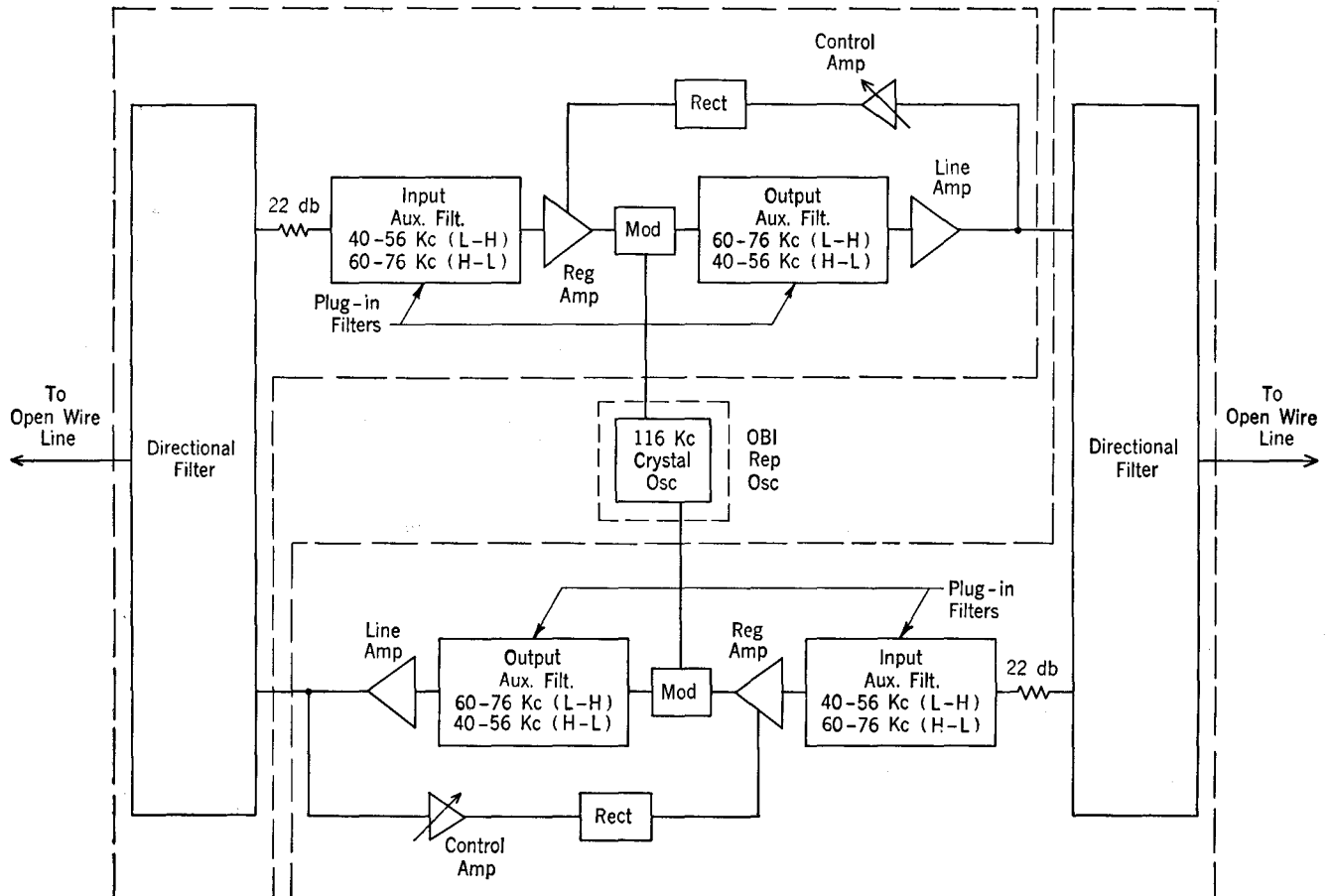


FIG. 32-19 BLOCK SCHEMATIC OF TYPE-OB1 REPEATER

3700-cycle signaling. The system will ultimately furnish 16 two-way voice channels on one pair of wires. The 16 channels are divided into four 4-channel groups designated OA, OB, OC and OD. The only group equipment in manufacture and use at this time is that coded OB which uses the frequency band from 40 to 56 kc for transmission in one direction, and the band from 60 to 76 kc for transmission in the opposite direction. As in the N-system, terminals are arranged to transmit either the low or high group of frequencies, and to receive the corresponding opposites. Repeaters, which are spaced at intervals of 40 to 50 db, are arranged alternately for low-high or high-low transmission.

Unlike the N-system, however, single side-band transmission is used, with the upper and lower side-bands of a single carrier providing two channels transmitting in the same direction. Thus

only two carriers, spaced 8 kc apart, are required to obtain the 4 voice channels. Figure 32-18 indicates the frequency translations employed in the terminal channel and group modulators, and at the repeaters. The two carriers are transmitted over the line, and their combined power is used for regulation of the amplifiers at repeater and group receiving terminals to correct for line attenuation variations. Regulation is accomplished by means of a type of automatic volume control in which a part of the output of a line amplifier following the modulator is picked off and amplified in a "control amplifier", rectified and fed back to the input of a regulating amplifier that precedes the modulator. A block schematic of the arrangement of a repeater is given in Figure 32-19. Other units employed in the O-system are practically identical with their counterparts in the N-system, which have already been discussed.



## CHAPTER 33 RADIO SYSTEMS

### 33.1 Types of Radio Systems

The essential elements of any radio system are (1) a transmitter for modulating a high-frequency carrier wave with the signal, (2) a transmitting antenna that will radiate a maximum amount of the energy of the modulated carrier wave, (3) a receiving antenna that will intercept a maximum amount of the radiated energy after its transmission through space, and (4) a receiver

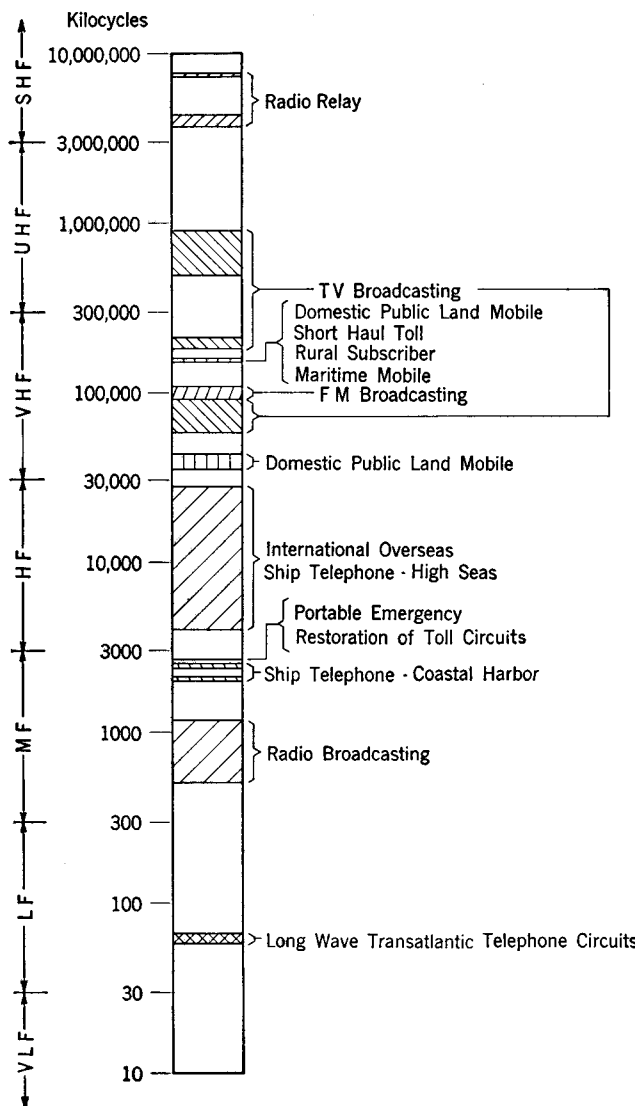


FIG. 33-1 RADIO FREQUENCY SPECTRUM

to select the carrier wave and detect or separate the signal from the carrier. Although the basic principles are the same in all cases, there are many different designs of radio systems. These differences depend upon the types of signal to be transmitted, the distances involved, and various other factors, including particularly the part of the frequency spectrum in which transmission is to be effected.

Figure 33-1 is a chart of the radio spectrum indicating at the left the commonly accepted classification of radio frequency ranges; and showing at the right the more important frequency ranges of special interest in current telephone practice. It will be noted that telephone practice makes use of some part of nearly all of the major frequency ranges. It must accordingly employ a corresponding variety of types of radio facility. It is not practicable or desirable to attempt to describe all of these in this book, and what follows will therefore be limited to a brief general discussion of principles applicable to all radio systems, with a few examples of specific facilities.

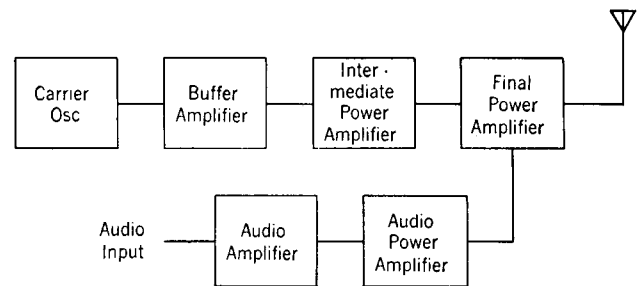


FIG. 33-2 AM RADIO TRANSMITTER-HIGH-LEVEL MODULATION

### 33.2 Radio Transmitters

The principal components of a typical amplitude-modulated radio transmitter, such as might be used in radio broadcasting or for relatively low-powered point-to-point transmission, are indicated in the block diagram of Figure 33-2. Here the amplified input signal plate-modulates the carrier in the output circuit of a Class C power amplifier which represents the final stage in a chain of amplifiers that increases the power of the

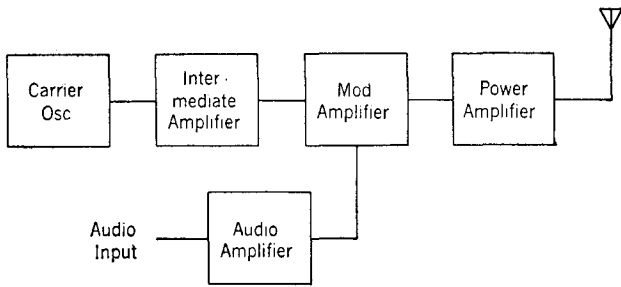


FIG. 33-3 AM RADIO TRANSMITTER—LOW-LEVEL MODULATION

carrier to an appropriate value for application to the antenna.

Another common design of transmitter for relatively low power work applies the signal to the plate circuit of an intermediate modulating amplifier, as indicated in Figure 33-3. As compared with the arrangement of Figure 33-2, this has the advantage of requiring less amplification of the audio signal to secure full modulation. On the other hand, it necessitates the use of a linear amplifier in the final power stage instead of the more efficient Class C amplifier that may be used when the modulating signal is applied to the final power stage as in Figure 33-2.

For very long distance point-to-point radio telephone circuits, such as those used in transoceanic service, transmitter design is naturally somewhat more elaborate. Figure 33-4 is a block schematic showing the major components of a transmitter (coded LD-T2) used in the Bell System for multi-channel single-sideband transmission in the 4 to 28 megacycle range. As may be seen, this transmitter employs a triple modulation system fol-

lowed by a six-stage linear amplifier. It is arranged to accept two independent voice-frequency bands, each extending from 100 to 6000 cycles. These bands are broad enough to permit superimposing two separate voice channels on each by appropriate carrier methods. The two band groups are transmitted as the upper and lower side-bands of a single carrier frequency, thus providing a total of four telephone circuits on one radio channel. The carrier itself is also transmitted at reduced volume for automatic frequency control and automatic volume control at the receiver.

As Figure 33-4 indicates, the two voice-frequency bands are applied separately to two low-frequency modulators, together with a 100 kc carrier. Band-pass filters following the LF modulators select two side-bands, the upper of which extends from 100.1 to 106 kc, and the lower from 99.9 to 94 kc. The carrier is at this point eliminated. The two side-bands are combined in a hybrid and applied to a two-stage amplifier at the output of which the 100 kc carrier is re-supplied at reduced volume. A second, medium-frequency modulator stage translates the total signal to an intermediate frequency range extending from 2794 to 2806 kc, which is the upper side-band of a 2700 kc conversion frequency. The original 100 kc carrier is now at 2800 kc. Precision of signal range is secured by a series of four 2794 to 2806 kc band-pass filters working in tandem with three medium-frequency amplifiers.

The final frequency conversion takes place in the high-frequency modulator where the signal centered about 2800 kc modulates a conversion

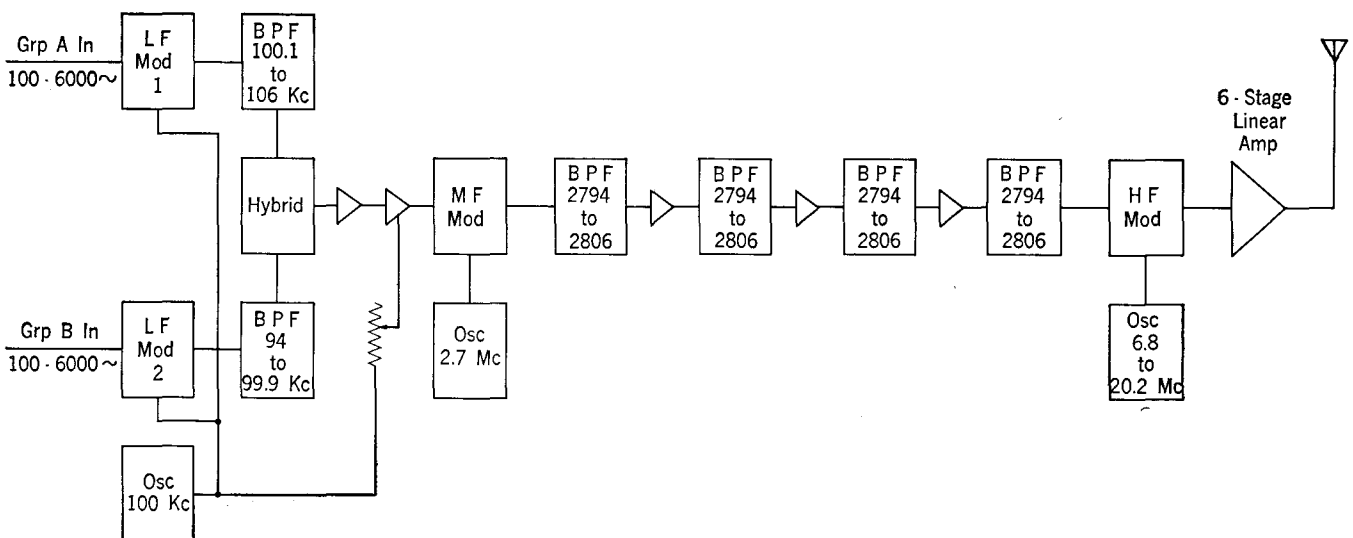
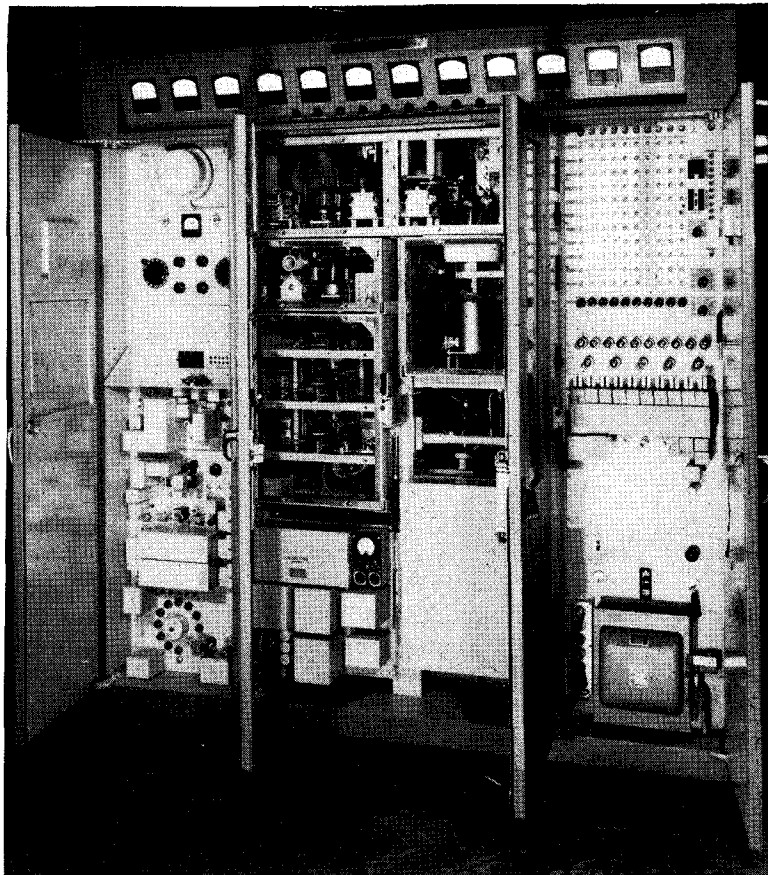


FIG. 33-4 LD-T2 RADIO TRANSMITTER  
[ 313 ]



LD-T2 RADIO TRANSMITTER (FRONT VIEW)

frequency, which may have any one of a number of values ranging between 6.8 and 20.2 megacycles. These frequencies are supplied by an oscillator circuit whose output frequency is determined by the controlling crystal that is connected into the circuit. The transmitter is equipped with a set of crystals which provide ten different frequencies. After final modulation, the signal wave is applied to a six-stage linear amplifier which is capable of delivering a peak power

of 4 kw. The first four stages of the amplifier employ pentodes, the fifth stage tetrodes, and the final stage a 2.5 kw air-cooled triode operating with grounded grid.

As was pointed out earlier, there are a number of different types of circuits in use for frequency modulating a radio carrier wave. Figure 33-5 indicates in block schematic the arrangement of an FM transmitter in which modulation is effected by means of a reactance tube circuit. Because FM radio (voice) transmission is generally in the very high-frequency range, it is necessary to use frequency multiplying circuits to bring the basic frequency generated by the master oscillator up to the desired value. The oscillator usually operates in the neighborhood of 5 mc, and its output frequency must be multiplied by factors in the order of 10 to 20 to reach the frequencies prescribed for FM transmission. Two or three frequency-multiplication stages are usually employed for this purpose.

In order to maintain the carrier frequency at a fixed value, the reactance tube type of transmitter requires the use of an automatic frequency control arrangement as shown in the lower part of Figure 33-5. This includes a crystal-controlled oscillator of highly stable frequency. A portion of the modulated carrier is picked off and compared in a mixer (modulator) circuit with the output of the crystal oscillator. The difference between the two frequencies, if any, is fed to a discriminator, the output of which, after rectification, is applied to the grid of the reactance tube. The polarity of this rectified output will be such as to

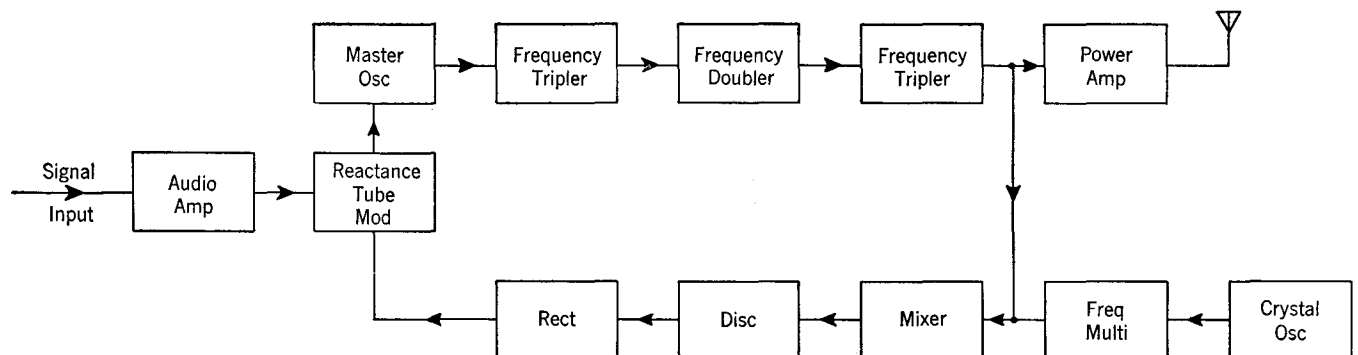


FIG. 33-5 FM TRANSMITTER USING REACTANCE TUBE

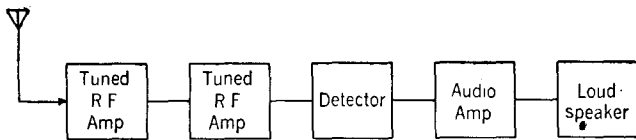


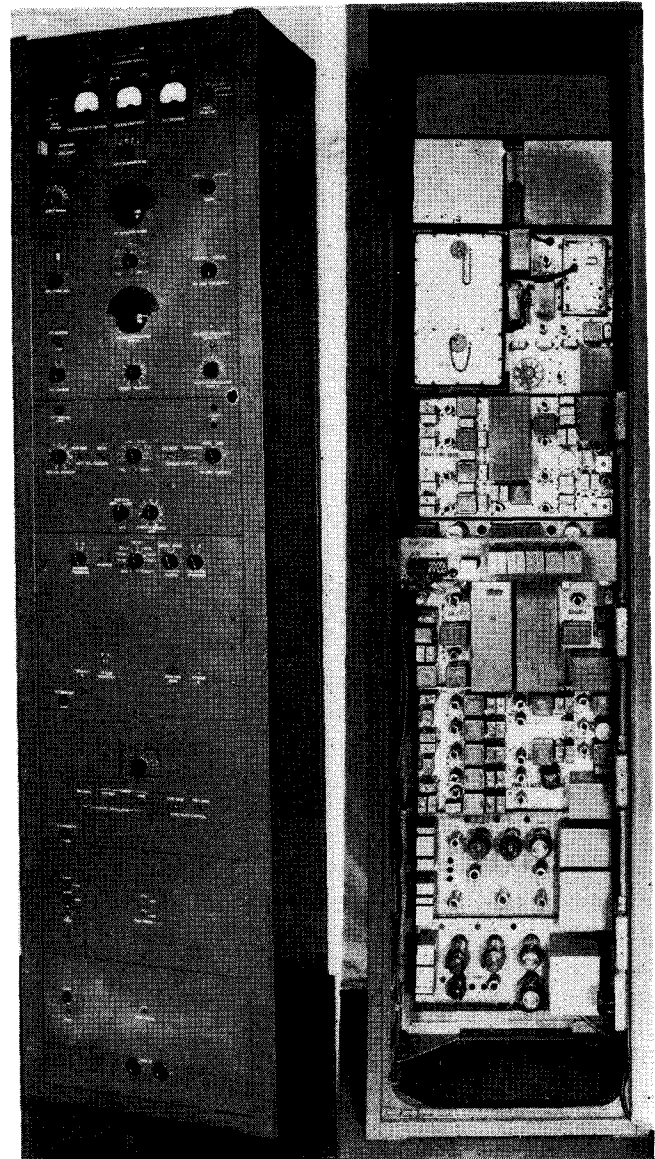
FIG. 33-6 TRF RADIO RECEIVER

hold the mean frequency of the master oscillator effectively constant under the control of the crystal oscillator.

### 33.3 Radio Receivers

The radio signals that are picked up by the antenna of a radio receiver are usually very weak so that the receiver circuit must ordinarily include one or more amplifiers. Figure 33-6 is a block schematic of a tuned radio-frequency (TRF) receiver for amplitude-modulated signals. With this arrangement, the r-f signal taken from the antenna passes through from one to three or more stages of amplification before reaching the detector. Associated with the input of each amplifier is a tank circuit which can be tuned by a variable capacitor to the frequency of the desired incoming signal. After detection, one or more stages of audio amplification are employed to bring the detected signal up to the strength required to operate the loud-speaker or other sound-reproducing device. TRF receivers, while once in general use, are used at the present time only on a limited basis, chiefly because of the difficulty of securing sufficiently sharp selectivity with respect to adjacent channels.

The vast majority of modern receivers are of the "superheterodyne" type illustrated in block schematic in Figure 33-7. Here, before detection, the r-f signal is converted by a modulation process to a fixed intermediate frequency value in which most of the required amplification takes place. The incoming signal is selected by a variable tuned circuit, which in some cases may include one amplification stage. A local oscillator, which supplies the mixer or demodulator (also sometimes called a converter or first detector), is tuned simultane-



(FRONT) (REAR)  
LD-R1 RADIO RECEIVER

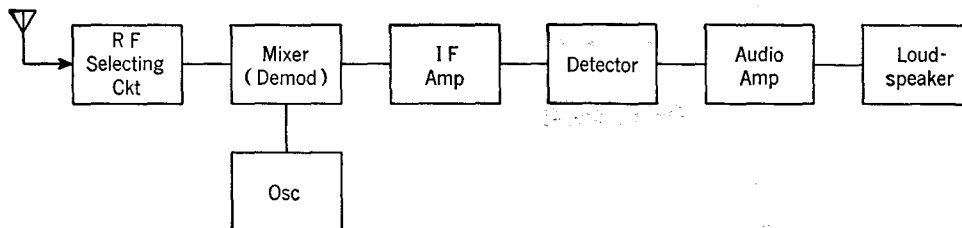


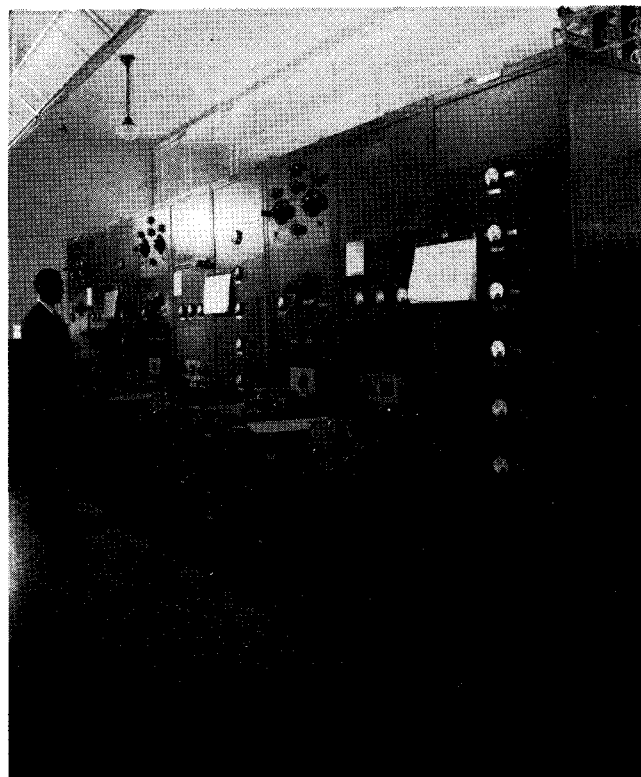
FIG. 33-7 SUPERHETERODYNE RADIO RECEIVER

ously with the signal selecting tuner so that the frequency of the mixer output, which is the difference between the frequencies of the oscillator and the incoming carrier, is always the same. The intermediate frequency amplifier circuits, accordingly, require no adjustment and may employ coupled circuits double-tuned to a single constant frequency. Radio broadcast receivers are usually designed for an intermediate frequency value of about

450 kc. Figure 33-7 indicates a local oscillator separate from the mixer circuit, but in most broadcast receivers a single multi-grid tube known as a "pentagrid converter" performs both the oscillator and mixer functions.

The superheterodyne receiver, although its advantages are sufficient to warrant its general use, has some inherent capacity to produce spurious responses under certain conditions. One of the major sources of such undesired responses is the possible presence at the antenna of a signal whose value is the "image frequency" of the tuned-in signal. The frequency of the image signal is greater than the frequency to which the receiver is tuned by twice the value of the intermediate frequency if, as is normally the case, the local oscillator is operating at a higher frequency than that of the desired signal. Such an image signal will mix with the oscillator frequency to produce a difference frequency that is exactly equal to the intermediate frequency. Thus, both signals would be amplified in the IF section and appear simultaneously in the receiver output.

Receiver response to image signals can only be avoided by blocking the image in the selecting circuits that precede the mixer. This is facilitated by the use of an intermediate frequency which will cause the image of the desired signal frequency to lie at a considerable distance in the frequency band from the desired frequency. For example, if the intermediate frequency is 450 kc, and the receiver is tuned to a signal carrier at 800 kc, the oscillator frequency must be 1250 kc. With this oscillator frequency, the image signal that would produce a 450 kc intermediate fre-



SINGLE SIDEBAND OVERSEAS RADIO RECEIVERS

quency would be 1250 kc plus 450 kc or 1700 kc. This is sufficiently removed from the 800 kc signal so that little, if any, is likely to pass through the tuned selecting circuit to reach the mixer.

For reception of long distance point-to-point signals, as in transoceanic service, a more elaborate receiver employing two intermediate frequencies is often used. This arrangement is sometimes

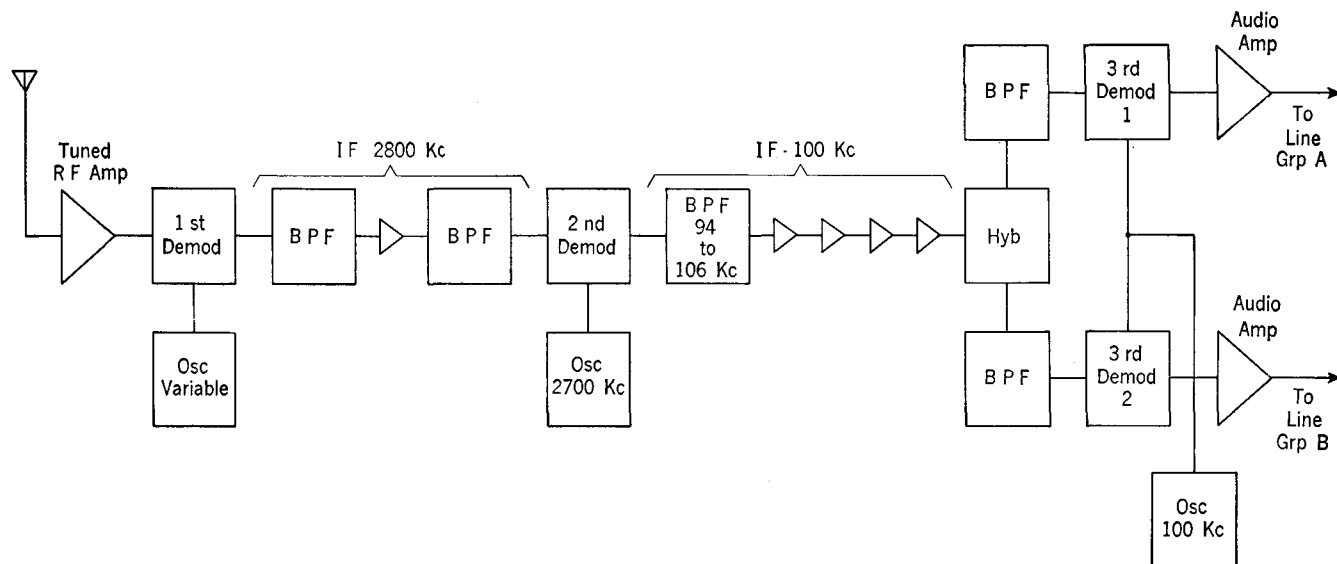


FIG. 33-8 LD-R1 SINGLE SIDEBAND RADIO RECEIVER

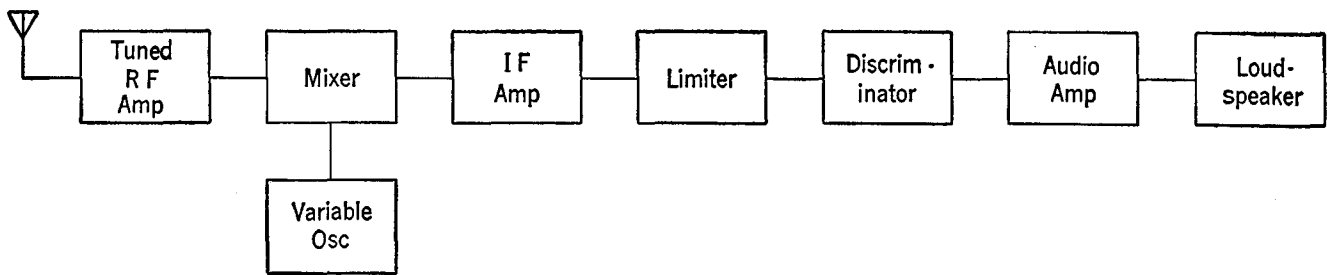


FIG. 33-9 FREQUENCY-MODULATION RECEIVER

known as a triple-detection receiver. The first intermediate frequency has a relatively high value to permit maximum image suppression, and the second intermediate frequency is comparatively low to provide high adjacent channel selectivity.

The Bell System single-sideband receiver coded LD-R1, which is shown in block schematic in Figure 33-8, is of this general type. This receiver operates in the frequency range from 4 to 28 mc. After the first demodulation, the signal is centered around an IF of 2800 kc, a frequency high enough to give good suppression of the image 5600 kc away from the desired signal. The band-pass filters on each side of the amplifier have high rejection characteristics for frequencies 200 kc below 2800 kc in order to avoid the possible transmission of a second image frequency into the 100 kc section of the circuit.

The second demodulator, which is supplied from

a 2700 kc oscillator, translates the signal into the 100 kc range, a band-pass filter selecting frequencies between 94 and 106 kc. The main four-stage amplifier delivers the signal through a hybrid to two band-pass filters which select the upper and lower side-bands of the 100 kc carrier. Each side-band group is then converted to voice frequencies by a balanced copper-oxide demodulator supplied with 100 kc from a local oscillator. Each group is sufficiently wide to provide for two telephone channels, as in the case of the comparable transmitter previously discussed.

The pre-detection stages of receivers for FM signals are generally the same as those of AM receivers. As shown in Figure 33-9, however, detection in this case is effected by means of limiting and discriminating circuits. For best results, IF amplification should be great enough to raise all peaks of the signal above the cut-off point of

the limiter. This will automatically eliminate any amplitude variations that may be present so that the signal at the output of the limiter will have a uniform fixed amplitude and will vary only in frequency. By eliminating amplitude variations, noise and unwanted energy which are the chief causes of such variations are minimized. The discriminator converts the constant amplitude frequency-variations into an audio-frequency signal, as described in Article 28.3, which is then amplified in the usual manner.



OVERSEAS HF RADIO TRANSMITTERS

### 33.4 Antennas

The effectiveness of a transmitting antenna is measured by its ability to convert a maximum amount of the power developed by a radio transmitter into radiant energy in the form of electromagnetic waves, which will be transmitted in such a direction as to produce maximum field strength at the receiver. In point-to-point transmission, and to some extent in broadcast transmission, the factor of major importance is usually the degree of directivity or antenna gain (Article 20.4) that can be secured. Transmitting and receiving antennas are sometimes alike but there are many situations where economic considerations and other factors require quite different designs for the two conditions.

There are many different antenna designs and arrangements in practical use in radio work. The various designs may be grouped into a limited number of major types, however, in which the controlling design factors are the frequency range in which the antenna is to operate and the degree of directivity desired. In general, antenna effectiveness can be more economically increased as the frequency of the radio wave is increased.

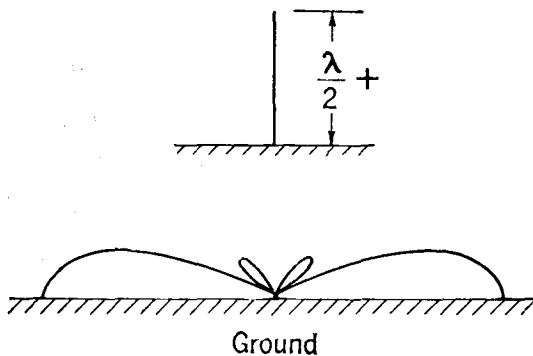


FIG. 33-10 RADIATION PATTERN OF HALF-WAVE VERTICAL ANTENNA

Transmitting antennas for medium radio frequencies quite commonly employ a simple vertical radiator which may consist of a single wire or a slender steel tower. Such radiators are tuned to resonance for the carrier frequency to be transmitted by making their total height equal to an appropriate fraction of the carrier wavelength; or, if the wavelength is too long to make this practicable, by adding lumped reactance in series with the radiator at its top (capacitive) or base (inductive). In either case, the antenna when energized at its resonant frequency, behaves like

a resonant transmission line with a standing wave of current extending along the conductor.

The typical radio broadcast antenna is a steel tower slightly more than one-half wavelength in height and effectively grounded at its base where the energy is applied. The vertical radiation pattern of such an antenna is shown in Figure 33-10, where the lobes indicate the relative field strengths radiated in various directions. The Figure of course represents the pattern of a single vertical plane through the antenna and the total pattern would be represented by rotating this plane through  $360^\circ$  about the antenna. This antenna,

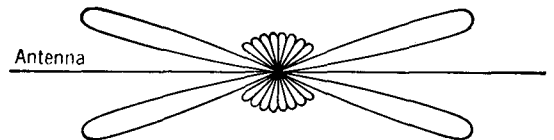


FIG. 33-11 RADIATION PATTERN OF LONG-WIRE ANTENNA OF LENGTH  $5\lambda$

accordingly, radiates with equal strength in all horizontal directions but concentrates the field along the horizontal, with relatively low radiation upward at high angles. Both of these features are generally desirable for ordinary broadcasting purposes. Where, as may be the case even in broadcasting work, some directivity in azimuth (horizontal) is desired, one or more additional radiators may be employed. The geometrical arrangement and spacing of the additional radiators and the phase relationships of the currents with which they are driven will determine the direction of maximum radiation.

In the high-frequency range (3 to 30 mc) extensive use is made of "long-wire" antennas for point-to-point transmission. A single horizontal conductor mounted a half-wavelength or more above the ground is directional along the axis of the wire when the wire is several wavelengths long. Thus, Figure 33-11 shows the radiation pattern of an isolated resonant wire of length  $5\lambda$ . While the pattern includes numerous minor "spurious" lobes, the axes of the major lobes are at relatively small angles to the antenna, and these

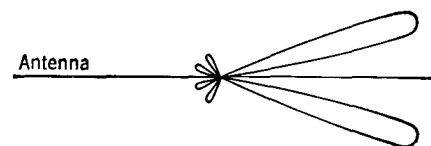


FIG. 33-12 RADIATION PATTERN OF NON-RESONANT LONG-WIRE ANTENNA

angles decrease with increasing antenna length. If the antenna is open at both ends and resonant, the net directional pattern extends equally in both directions. Where maximum radiation in only one direction is desired, the long-wire antenna may be terminated at one end in its characteristic impedance so as to be non-resonant. The radiation will then be largely concentrated in one direction, as indicated in Figure 33-12, where the driven end of the antenna is at the left and the terminated end at the right.

Combinations of long wires in appropriate geometrical arrangements are commonly used in long distance point-to-point practice for both transmitting and receiving antennas. One of the most effective such arrangements is the **rhombic antenna** illustrated in Figure 33-13. This, as shown, consists of four wires arranged in a diamond or rhomboid pattern terminated in a resistance  $R$  which makes the system non-resonant. Each of the four wires is from two to four wavelengths long, and each sets up an individual radiation

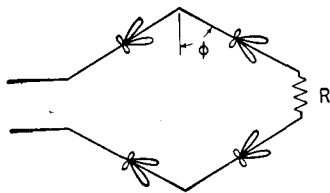


FIG. 33-13 RHOMBIC ANTENNA SHOWING RADIATION PATTERNS

pattern, as indicated in the drawing. When the "tilt-angle"  $\phi$  is properly chosen, four of the major radiation lobes point in the direction of the main axis of the rhombus while the other four major lobes tend in part to cancel each other out. The rhombic antenna operates satisfactorily over a wide frequency range because its impedance has a constant resistive value. This is of great importance in short-wave radio transmission over long distances where it is frequently necessary to change the frequency to obtain satisfactory transmission under changing radio propagation conditions. Rhombic antennas of identical design are used for both transmitting and receiving. A number of separate rhombics may be used, however, for "diversity reception" in which the energy received from several appropriately located antennas is combined to drive a single receiver.

For radio transmission in the very high and ultrahigh-frequency ranges, the basic form of antenna is a half-wavelength resonant wire com-

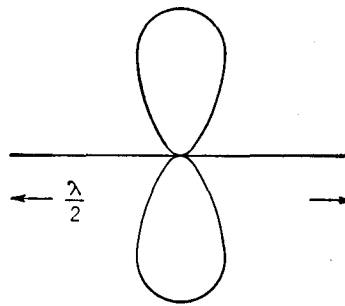


FIG. 33-14 RADIATION PATTERN OF ISOLATED HALF-WAVE ANTENNA

monly known as a **half-wave dipole**. The radiation pattern for this simple antenna, when isolated in space, is indicated in Figure 33-14. If the antenna is placed vertically, radiation strength is maximum along the horizontal and is equal in all directions. If the radiator is parallel to the ground, on the other hand, the field varies with azimuth and is zero in the direction of the wire axis. The vertical pattern is largely determined by the height of the antenna above the ground.

Increased directivity can be obtained by the use of arrays of half-wave antennas properly spaced and phased. One relatively simple arrangement employs one or more "parasitic antennas" in conjunction with a driven half-wave radiator. If maximum radiation in a particular direction is desired, a wire of such length as to be resonant at a frequency somewhat lower than the transmitted frequency, placed parallel to and at a distance about  $1/10$  wavelength away from the driven antenna on the side opposite to the preferred direction of radiation, will act as a reflector. This parasitic reflector is not directly driven but the currents induced in it by the driven antenna set up a radiated field which combines with that of the driven antenna to produce a net field in the desired forward direction. Directivity may be further enhanced by using one or more additional parasitic wires in front of the driven antenna and spaced in tandem in the direction of transmission at separations of about  $.1\lambda$ . These wires are made of such length as to be resonant at a frequency higher than the transmitted frequency and act as "directors" to further concentrate the field in the desired direction.

When no directivity in the horizontal direction is desired, as in most broadcasting and mobile radio applications, a vertical half-wave radiator may be employed to develop vertically polarized waves. This is the so called "flag-pole" type of antenna. Horizontally polarized omnidirectional



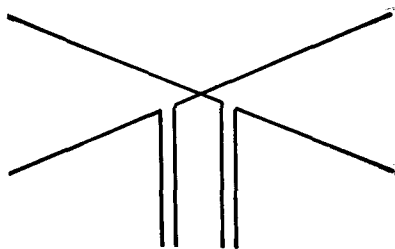


FIG. 33-15 TURNSTILE ANTENNA

waves are usually produced by antennas employing the "turnstile" principle. As indicated in Figure 33-15, the basic turnstile antenna consists simply of two half-wave resonant wire radiators placed crosswise at right angles to each other in the same plane. When the two radiators are excited by equal currents 90° out of phase with each other, the directional radiation pattern in the horizontal plane is roughly circular. Antennas of this type are advantageous in that they can be stacked vertically at half-wavelength intervals to increase horizontal radiation. They are used extensively with TV broadcast transmitters. The TV signal, however, is a very broad one (6 mc) and it is necessary that the antenna be able to operate over this broad frequency range. The broad-banding effect required is obtained by making the antenna itself thick so as to increase the effective diameter of the radiator. This may be accomplished by employing for the arms of the dipoles such structures as hollow cylinders, cones or cigar-shaped spheroids.

Radiation of electromagnetic energy in the super-high frequency range involves principles and methods that are analogous to, if not identical with, those of light transmission. Antennas used at these frequencies may be grouped under the general designation of **aperture radiators**. Their basic function is to transform the spherical wavefront, which is normally developed from a point source of radiation, to a plane wavefront. To the extent that this is accomplished, the radiated energy may be projected in a very narrow beam in the desired direction. In microwave practice, an-

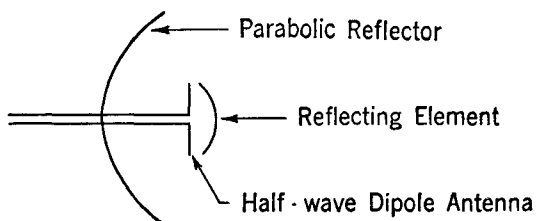


FIGURE 33-16

tenna gains in the order of 30 to 40 db are commonly obtained.

One method of obtaining a directive plane wavefront employs the optical technique of a parabolic reflector comparable to that of the ordinary searchlight. The geometrical characteristics of the parabola are such that waves emanating from a point source at the focus will be reflected in parallel straight lines that will all reach the plane of the mouth of the parabola at the same time. If, as illustrated in Figure 33-16, an auxiliary reflecting surface is placed in front of the energy source to prevent any direct radiation, all of the energy will be reflected from the paraboloid and will appear as a plane wave across its mouth. Since, as has already been pointed out, electromagnetic energy can escape directly from the open end of a waveguide, the same effect can be produced by means of a waveguide leading to the focal point of the parabola with its open end turned back to direct escaping energy toward the reflecting surface, as illustrated in Figure 33-17.

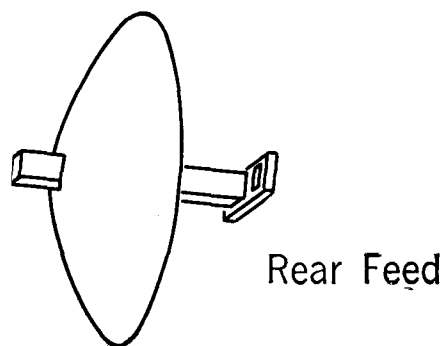


FIGURE 33-17

A plane wavefront across an aperture may also be attained by simply flaring out the end of a waveguide to form a long horn in which the fields in the guide can expand gradually to produce a uniform field across the mouth of the horn. To accomplish this result, however, the flare angle of the horn must be small and an aperture comparable in size to that of the parabolic reflector could only be reached with a horn that would be too long to be practicable. In the major Bell System radio relay systems (TD-2), this difficulty is overcome by using a comparatively short horn having a high flare-angle but with a "lens" across its mouth that acts like an optical lens to produce a plane wavefront. This arrangement, known as a **delay lens antenna**, is illustrated in Figure 33-18. As indicated there, the lens cross-section is of

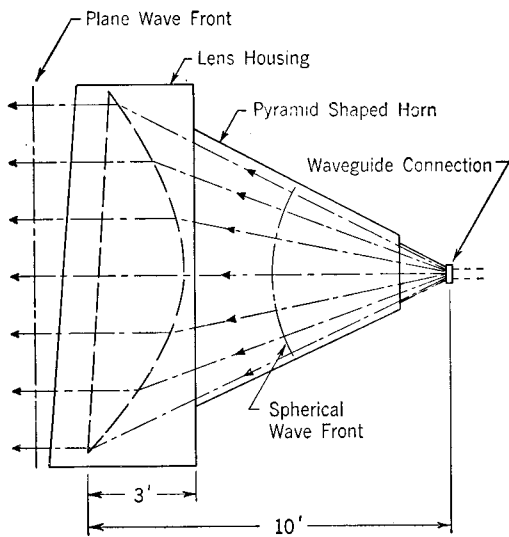


FIG. 33-18 DELAY LENS ANTENNA

plano-convex form and is placed at the mouth of a pyramid-shaped horn in a housing 10 feet square and 3 feet deep. The effect of the lens is much the same as that of a plano-convex lens on the transmission of light. Superhigh-frequency radio waves coming from the waveguide connected to the rear of the horn diverge, as indicated by the arrowed broken lines, to form a spherical wavefront. When they reach the lens, however, the velocity of the rays near the center is decreased enough by the thicker lens structure at that point to cause delay equal to the time required for the outer rays to traverse their geometrically longer path. All of the rays accordingly reach the front of the lens at the same time and proceed in the parallel paths of a plane wavefront. The entire lens is tilted back slightly, as shown in the Figure. This is done to prevent any energy that might be reflected from the face of the lens from being focused back to the waveguide outlet. This lens, although its action is very similar to

that of a glass lens, is actually made up of a very large number of narrow aluminum strips held in place by slabs of foamed polystyrene. The effect of these metal strips is essentially the same as that of the molecules of a glass lens in retarding light waves.

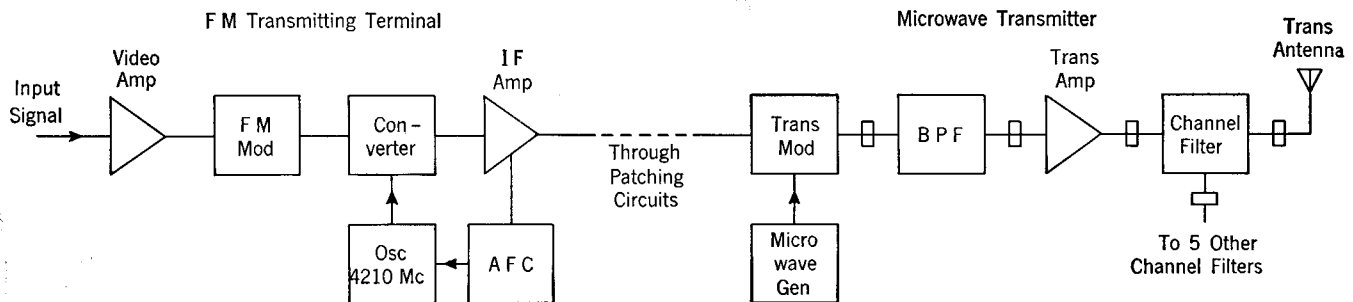
### 33.5 Microwave Radio Relay Systems

A number of types of radio relay systems employing frequencies in the SHF range are in use in long distance telephone work. As the basic principles involved are essentially the same in the various types, discussion here will be confined to the system coded TD-2, which is the type currently in use in the Bell System for major long distance routes.

The TD-2 system operates in the frequency range between 3700 and 4200 mc. This 500 mc band is divided into 12 channels, each with a bandwidth of 20 mc and a separation of 20 mc between adjacent channels. Each 20-mc channel can handle one one-way television signal or one direction of transmission of the many voice channels of a Type-L telephone carrier system. All transmitting channels are served by one antenna, and all receiving channels are served by a second antenna of identical design. The system therefore requires only two antennas at terminals and four at relay points. In practice, two 20-mc channels are used for each of six two-way channels numbered as indicated below:

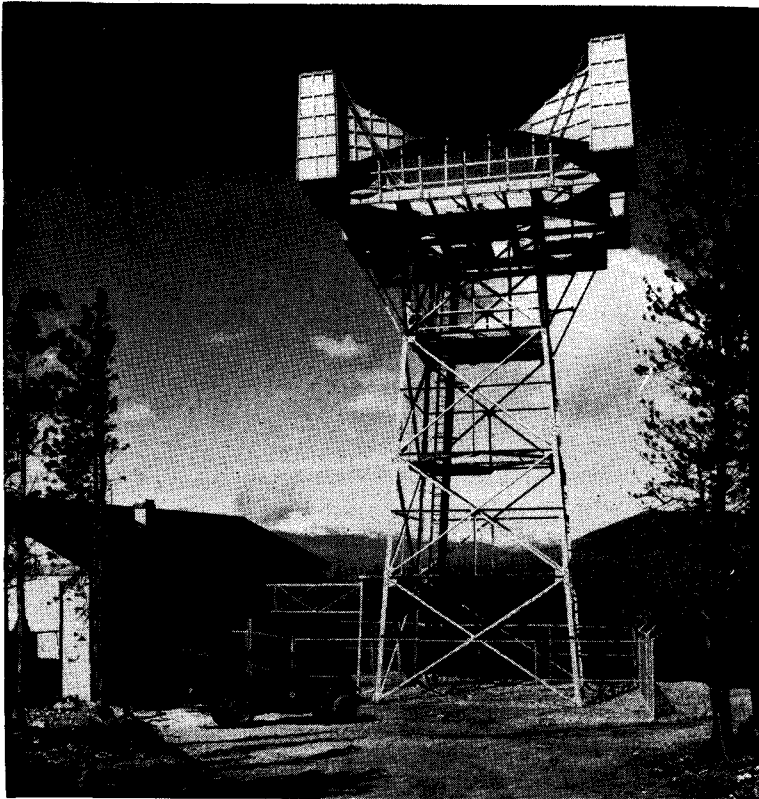
Channel No.	Frequencies
1	3720-3740 mc and 3760-3780 mc
2	3800-3820 mc and 3840-3860 mc
3	3880-3900 mc and 3920-3940 mc
4	3960-3980 mc and 4000-4020 mc
5	4040-4060 mc and 4080-4100 mc
6	4120-4140 mc and 4160-4180 mc

At each relay or repeater station, the channel



—□— Indicates Waveguide

FIG. 33-19 TD-2 TRANSMITTING TERMINAL



RADIO RELAY STATION IN ROCKY MTS.

frequencies are interchanged so that the outgoing signal is always 40 mc removed from the incoming signal. This shifting of the channel frequencies avoids possible difficulties that might be caused by local feedback between transmitting and receiving antennas.

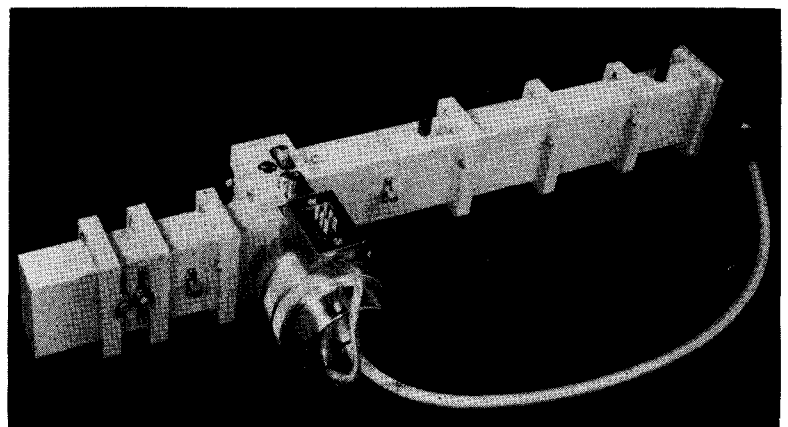
Figure 33-19 is a simplified block schematic of the TD-2 transmitting terminal. As there indicated, the input signal, which may cover all or part of a band between 30 cycles and about 4 megacycles, is first amplified by a video amplifier, which increases the amplitude of the signal waves to a maximum of 8 volts peak-to-peak. When multiplex telephone signals are being transmitted, the output of the video amplifier is applied directly to the FM modulator. For television, a clamping circuit is included between the amplifier and modulator. This adds a d-c component to the signal to clamp it to the amplitude value of the tips of the horizontal synchronizing pulses as a base line.

Frequency modulation is obtained in a comparatively simple manner by applying the signal directly to the repeller

electrode of a reflex-klystron oscillator, which is tuned to a nominal frequency of 4280 mc. A signal level of 8 volts peak-to-peak at the repeller produces  $\pm 4$  mc frequency modulation. The frequency-modulated carrier is then applied to a converter, whose active component is a point-contact varistor, where it is mixed with the output of a 4210-mc beating oscillator. The resultant converter output is an intermediate frequency normally centered about 70 mc and ranging between 74 and 66 mc. This of course represents the difference between the two frequencies applied to the converter. The four-stage IF amplifier raises the level of the modulated IF signal to +13 dbm. A small portion of the amplifier output is picked off and applied to a slow acting automatic frequency control circuit, which, in the case of telephone transmission, measures the average frequency at the amplifier output and adjusts the beating oscillator so that this average frequency remains constant at 70 mc. For video transmission, the AFC circuit measures the am-

plifier output frequency only during the horizontal synchronizing pulses and adjusts the beating oscillator frequency to hold this frequency constant at 74 mc.

All the components discussed above comprise what is generally known as the FM transmitting terminal. The modulated signal output of this terminal, after passing through appropriate switching or patching circuits, is applied to the microwave transmitter itself. The transmitter modulator translates the IF signal to the desired



MICROWAVE TRANSMITTER-MODULATOR

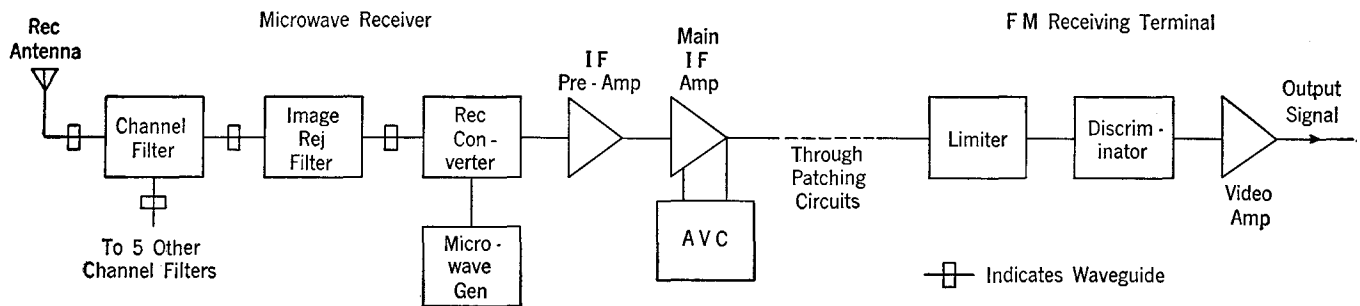


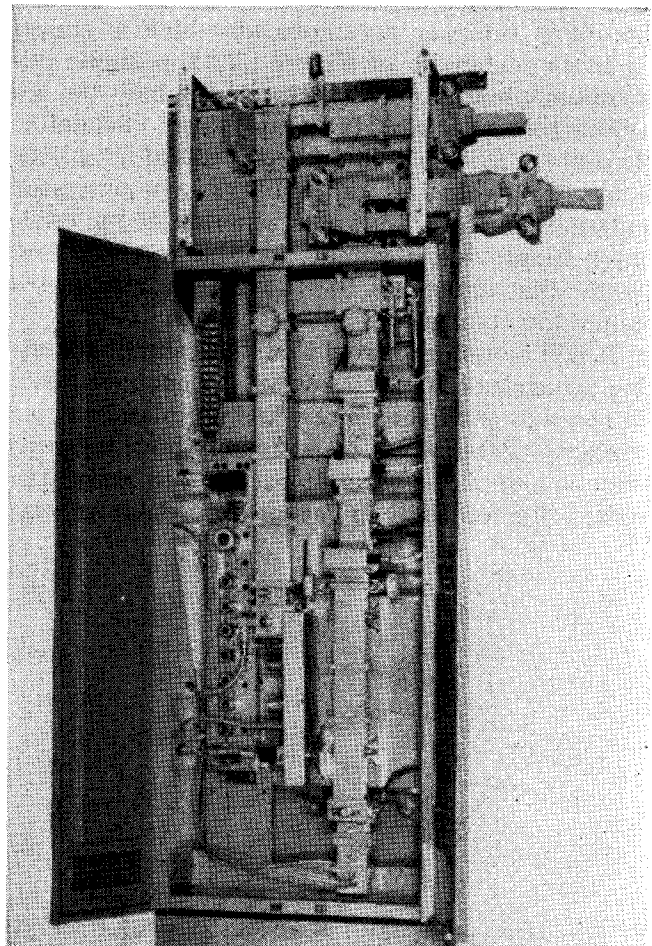
FIG. 33-20 TD-2 RECEIVING TERMINAL

microwave frequency. The modulator employs a 416 type tube, and is supplied with the proper beat-frequency from a microwave generator whose output frequency is 70 mc removed from the desired microwave channel frequency. The basic unit of the microwave generator is a very stable crystal-controlled oscillator operating in the frequency range from 17.5 to 19.0 mc, depending on the frequency of the crystal employed. The basic oscillator is followed by a series of frequency-multiplying stages providing a total multiplication factor of 216. The modulated output is led through a band-pass waveguide filter, which selects the upper side-band. The microwave transmitter amplifier, which was described in Article 27.3, is capable of producing an output of +27 dbm (slightly more than  $\frac{1}{2}$  watt). This output is fed through a channel filter where it is joined by the outputs of five other transmitting channels and applied collectively to the transmitting antenna.

At the receiving terminal of a microwave channel, the incoming channels are separated by channel filters, as indicated in Figure 33-20. The incoming energy for each channel then passes through an image rejection filter. This is a band-pass filter designed to have particularly high suppression characteristics in the neighborhood of the image frequency 140 mc away from the signal frequency. The receiver converter is a demodulator employing point-contact varistors, in which the SHF signal is mixed with the output of a microwave generator to again produce the 70 mc IF. This is passed through an IF pre-amplifier having a gain of about 12 db to an 8-stage main IF amplifier having a maximum overall gain of about 60 db. Associated with this amplifier is an automatic volume control circuit which compensates for differing input levels due to fading and holds the output power constant at approximately +9 dbm. The FM receiving terminal includes limiting and discriminating circuits which convert the

frequency-modulated 70 mc signal back to its original amplitude-varying form in the frequency range between 30 cycles and 4 megacycles. This signal is applied to a video amplifier whose push-pull output voltage is about 2.0 peak-to-peak.

Repeater stations, which are located at intervals of about 25 miles along the radio relay routes, are of two types—main and auxiliary. Main stations include switching and branching circuits,



PORTION OF TD-2 TRANSMITTER-RECEIVER BAY SHOWING WAVEGUIDE ELEMENTS AND CONNECTIONS

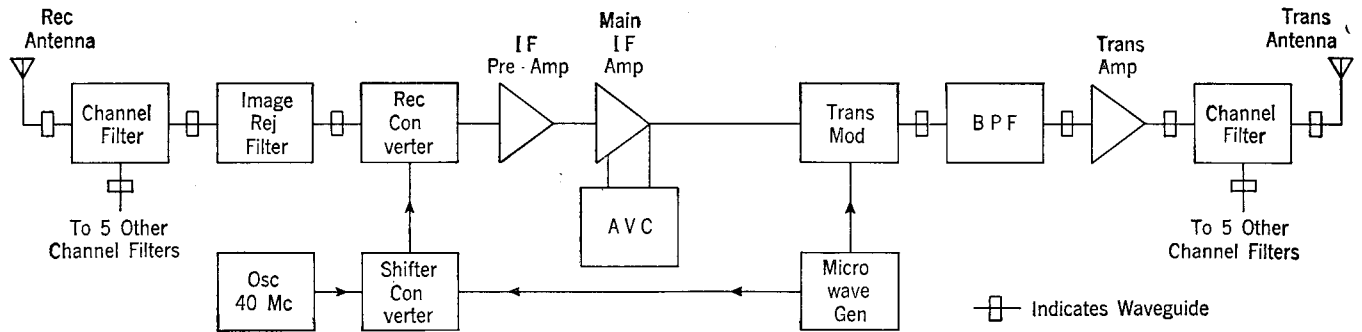
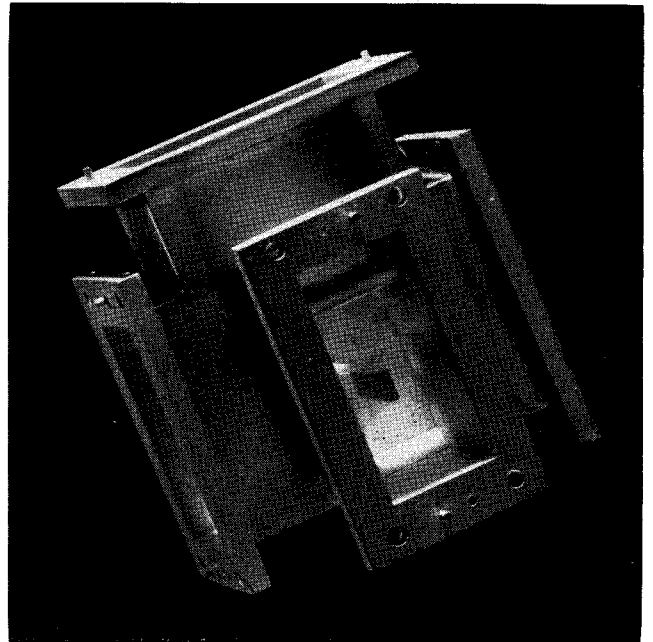


FIG. 33-21 TD-2 AUXILIARY REPEATER

while auxiliary stations are arranged only to receive, amplify and re-transmit the radio signal. The layout of the main repeater station may be represented by connecting together, through patching circuits, a microwave receiver as shown schematically at the left of Figure 33-20 and a microwave transmitter as shown at the right of Figure 33-19. The 40-mc frequency shift that is made at all repeaters is obtained by using conversion frequencies that differ by 40 mc. In the auxiliary repeater arrangement which is shown in block schematic in Figure 33-21, a single microwave generator is used to supply both the receiver converter and the transmitter modulator. The 40-mc shift is secured by the use of a "shifter converter" which mixes the microwave generator output with the output of a separate 40-mc oscillator to provide a supply for the receiver converter that differs by 40 mc from the microwave generator frequency.

It will have been noted that the general operating principles of the microwave system have much in common with the more usual radio and carrier systems. Because of the extremely high frequencies employed, however, most of the apparatus units differ radically in design from those of rela-

tively low-frequency systems. One of the most interesting examples of such design difference is found in the channel filters of the TD-2 system. The key component of these filters is a device known as a **waveguide hybrid**, one form of which



ONE FORM OF WAVEGUIDE HYBRID

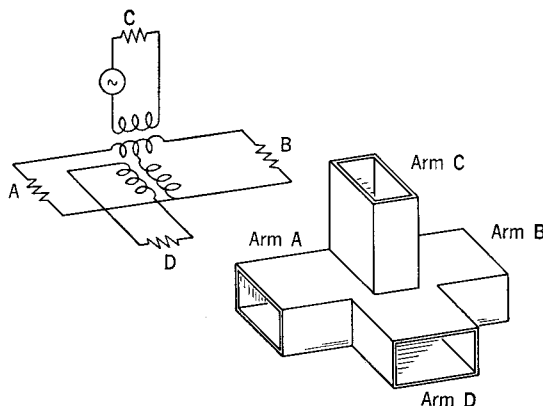
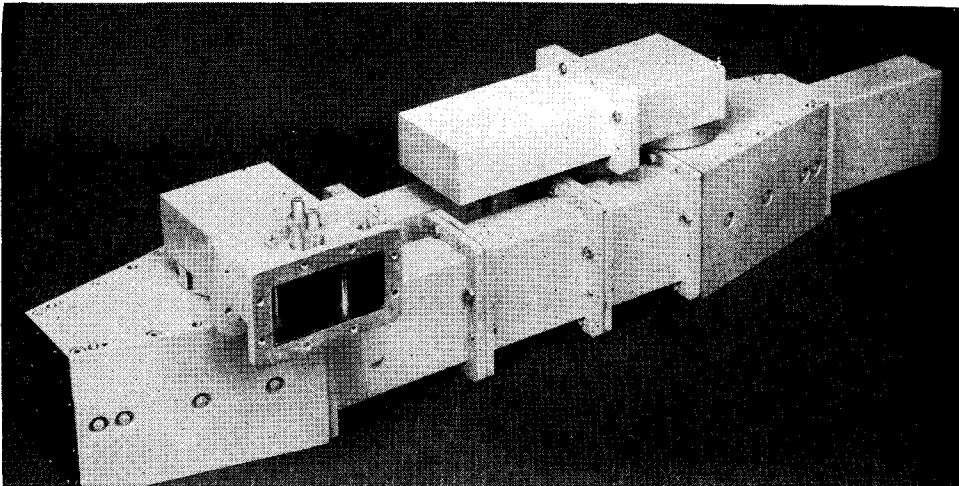


FIG. 33-22 WAVEGUIDE HYBRID AND CIRCUIT ANALOG

is illustrated in Figure 33-22 together with its circuit analog. When the impedances of the four waveguide arms are properly matched, energy entering arm C will divide equally between arms A and B and none will reach arm D. Similarly, energy applied at D will divide equally between arms A and B with no output to C. However, when the input is to arm C, the outputs of arms A and B are in phase opposition, while with the input at D, the outputs at A and B are in phase. This may be understood by referring to the circuit analog. It follows that when equal and in-phase voltages



CHANNEL SEPARATION NETWORK

are applied across arms *A* and *B* there will be no output to arm *D* and full output to arm *C*. On the other hand, if the equal inputs of arms *A* and *B* are  $180^\circ$  out of phase, there will be no output at *C* and full output at *D*.

The arrangement for employing these waveguide hybrids to obtain filter action is illustrated schematically in Figure 33-23. The total microwave energy coming from the antenna enters the upper hybrid at arm *C* and divides equally between *A* and *B* with no transmission at *D*. Inserted in series with both arms *A* and *B* are identical band-reflection filters (Fig. 23-22) tuned to reflect the frequency band of one particular microwave channel but to pass all other frequencies. The frequencies of the reflected channel band travel back to the hybrid and are applied to arms *A* and *B*. Because one of the band-reflection filters is located one-quarter wavelength farther away from the hybrid than the other, the energy reflected by one has to travel a half-wavelength farther than that traveled by the other in going

from the hybrid to the reflection filter and back. The reflected waves are therefore  $180^\circ$  out of phase when they reach the hybrid and the total reflected energy is therefore transmitted to arm *D* and thence to the channel receiver.

The energy that was not reflected by the filters passes on to arms *A* and *B* of the second hybrid in phase and is accordingly transmitted to arm *C*. This is connected to another arrangement identical to that of Figure 33-23 except that its filters are tuned to reflect a different channel band; and so on until all channels have been dropped off to their respective channel receiving circuits.

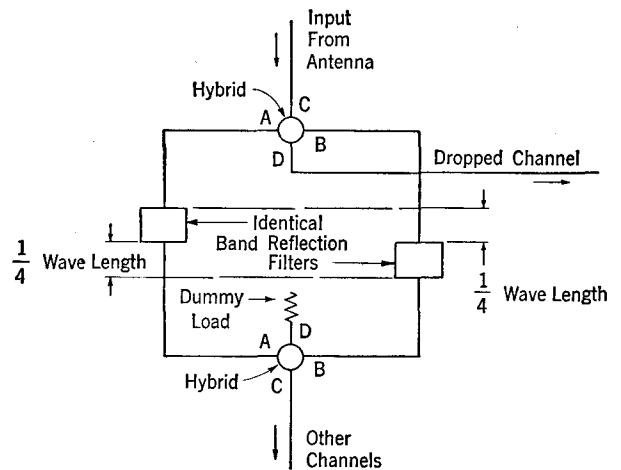


FIG. 33-23 CHANNEL BRANCHING FILTER PRINCIPLE

## CHAPTER 34

### NOISE AND CROSSTALK

#### 34.1 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 electric 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 the desired signal 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 electric circuits such as power lines. It is necessary, therefore, that telephone circuits not only be efficient in transmitting electric energy without distortion and without too great a loss, but also that they be protected against induced voltages caused by adjacent telephone circuits or other electric circuits.

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.

#### 34.2 Causes of Crosstalk

As we know, 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 34-1. 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

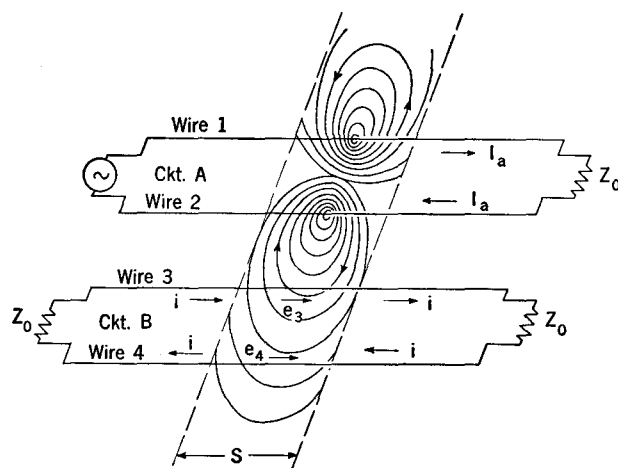


FIGURE 34-1

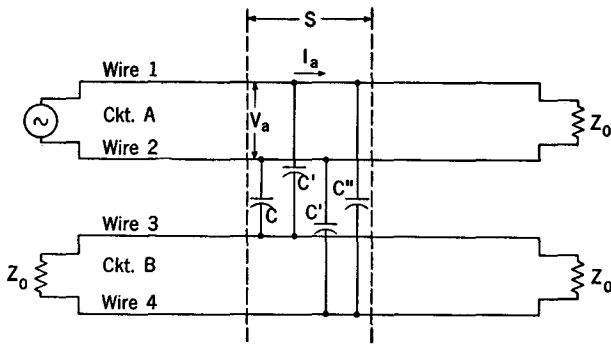


FIGURE 34-2

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 unbalance 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 unbalance voltage induced in a short section of the circuit may be written as—

$$i = \frac{e_3 - e_4}{2Z_0} \quad (34:1)$$

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 energy 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 in the disturbed circuit appears at both ends of the circuit. The cross-talk 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 34-1 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*.

The crosstalk effect of electric induction may also be analyzed by consideration of the capacitive relationships between the wires of the disturbing and disturbed circuits. Thus, referring to Figure 34-2, we know that in any unit length of the two circuits there is a definite capacitance 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, and there is therefore the same capacitance between wires 2-3 as between the wires of either pair. This is represented in the Figure by the small capacitor designated *C*. Similarly, the capacitances between wire 1 and wire 3 and between wire 2 and wire 4 are designated by capacitors, *C'*, less in value than *C* because the separation between these wires is greater. There remains also the still smaller capacitance between wire 1 and wire 4, which is indicated by *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 several capacitances to the wires of circuit *B*. The net effect can best be analyzed by redrawing the diagram of Figure 34-2 in the form of a Wheatstone bridge network as shown in Figure 34-3. A study of the capacitance 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 34-4, 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

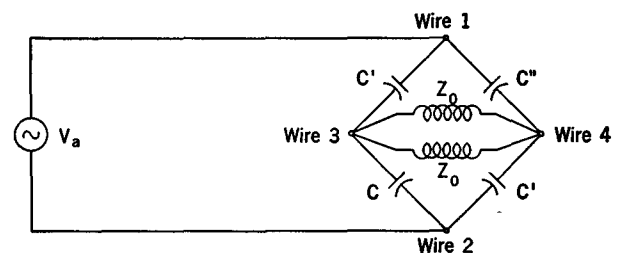


FIGURE 34-3



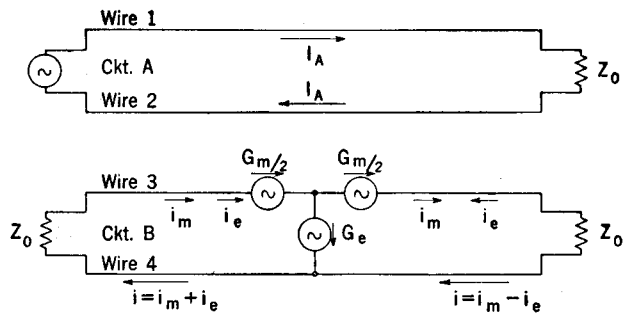


FIGURE 34-4

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 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.

### 34.3 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 34-5. For a number of reasons, however, such wire configurations are not usually practicable.

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 34-1 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

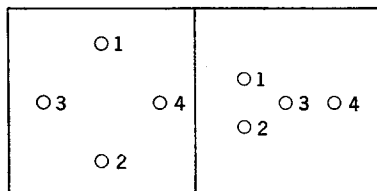


FIGURE 34-5

each other. Similarly, if the two wires of the disturbed pair are close together, the effect of any field set up by the disturbing pair will be practically the same on both wires of the disturbed pair, so that there will be no resultant unbalance 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 34-6, which shows the same four wires as were indicated in Figure 34-1. 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

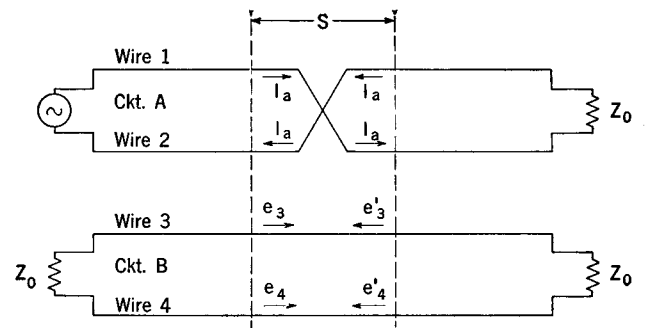


FIGURE 34-6

the two wires of circuit A, the voltages induced in wires 3 and 4 will be in opposite directions on the two sides of the point where circuit A is transposed. Thus, as indicated in the Figure, while  $e_3$  is larger than  $e_4$  and  $e'_3$  is larger than  $e'_4$ ,  $e_3$  is exactly equal and opposite to  $e'_3$  and  $e_4$  is exactly equal and opposite to  $e'_4$ . There is therefore no net voltage induced in either wire 3 or wire 4 and, consequently, no crosstalk from circuit A.

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 34-7. 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,  $e_4$

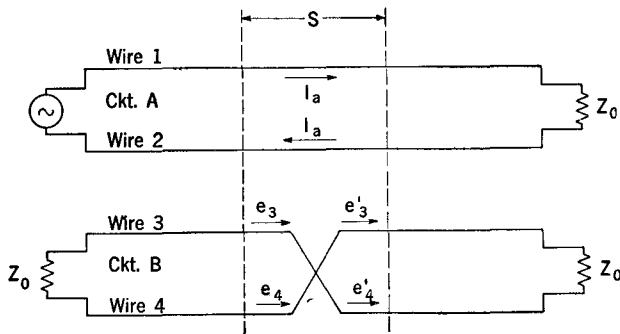


FIGURE 34-7

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—

$$i = \frac{(e_3 + e'_4) - (e_4 + e'_3)}{2Z_0} \quad (34:2)$$

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 (34:2) is therefore equal to zero and there is no resultant crosstalk.

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 wavelengths 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 wavelength of the propagated current. This is illustrated by Figure 34-8. 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  $90^\circ$ . The transposition would, therefore, not decrease the induced crosstalk. However, a number of transpositions within a single wave-

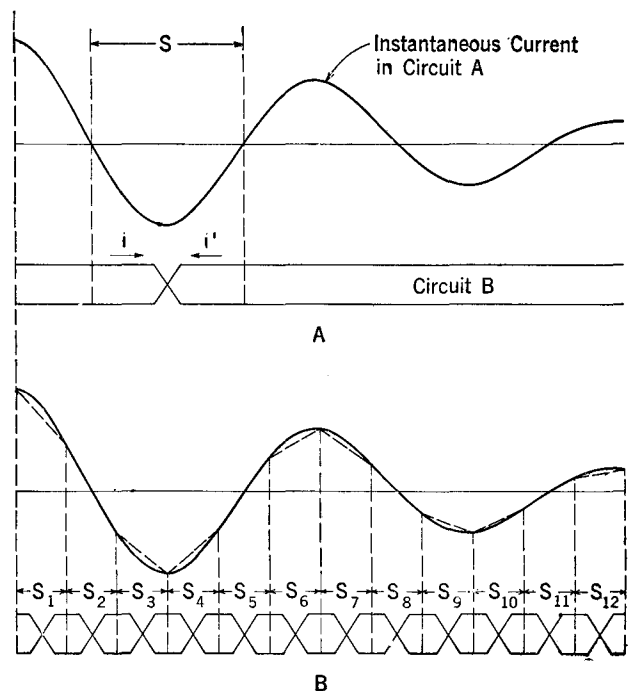


FIGURE 34-8

length, as illustrated at B, will reduce the crosstalk to practical limits, although the crosstalk could be 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 wavelengths 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 wavelengths are so short as to require 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 34-6 and 34-7 that the currents flowing in the two wires of the disturbing pair, or the induced voltages 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 34-6 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 voltages 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 unbalance in the 3-4 pair would cause a similar result. In this case, any resistance unbalance would cause one of the induced voltages 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.

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 34-9. 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

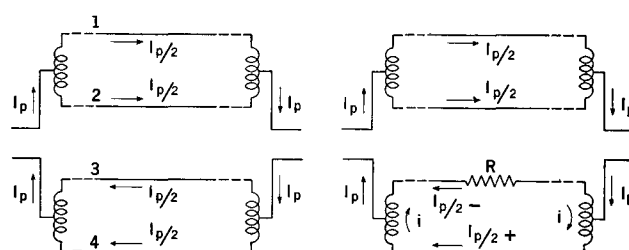


FIGURE 34-9

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.

### 34.4 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 case of Type-K carrier circuits, in different cables. A comparable 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.

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 21-4, where it may be noted that the separation of the wires of each pair is 6 inches and the horizontal separation on the crossarm between any two wires of different pairs is at least 28 inches.

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 eight miles. The purpose of the

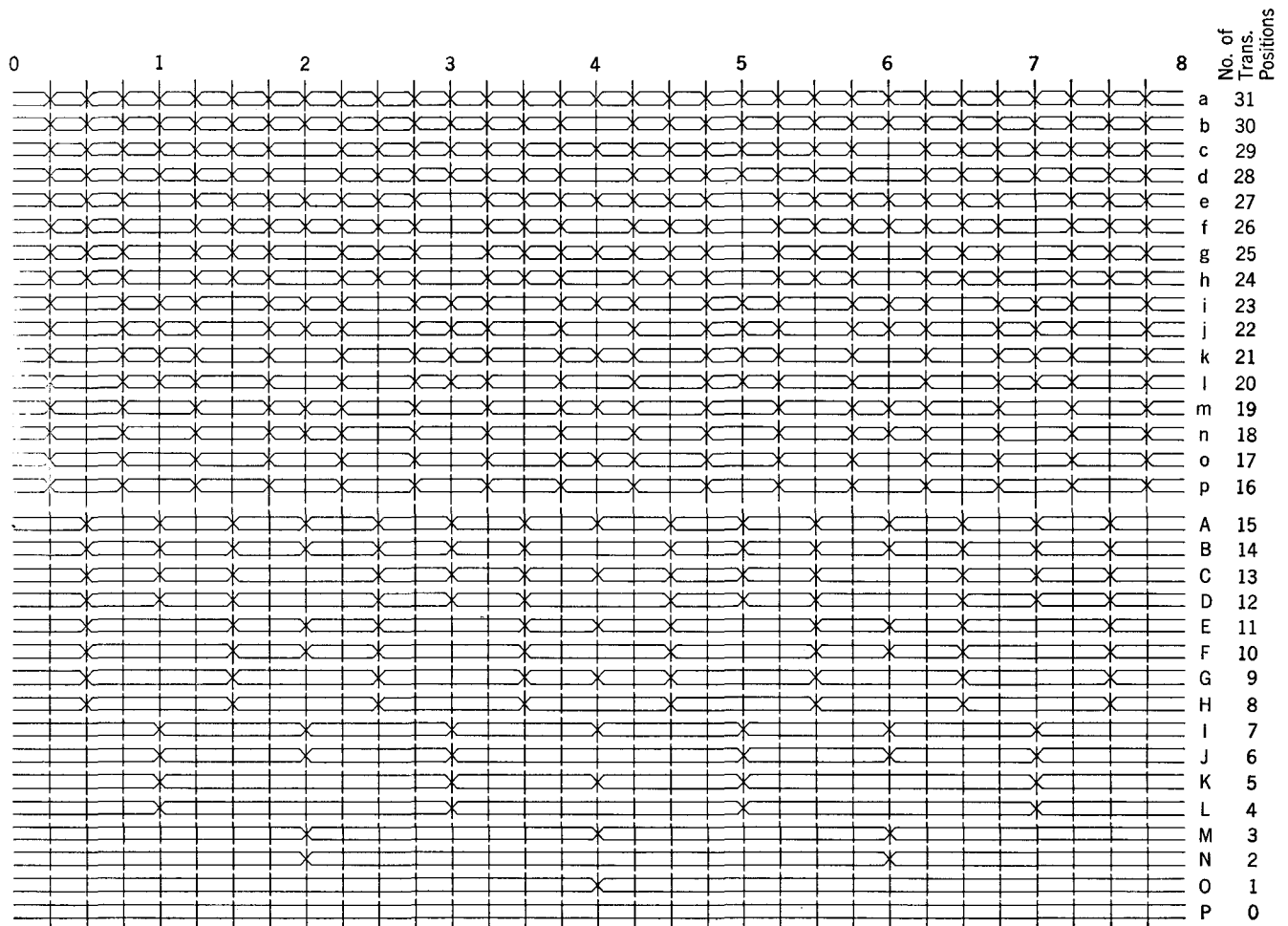


FIG. 34-10. STANDARD TRANSPOSITION CODE

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 the result 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 34-10 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 34-11 and 34-12. 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 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.

At voice frequencies, magnetic induction (inductive coupling) between circuits in a cable is normally so small as to be of relatively little importance in causing crosstalk. The same cannot be said of electric induction (capacitive coupling). Despite the most careful manufacturing methods, the capacitance unbalances between cable conductors usually remain large enough to cause objectionable crosstalk in long circuits. This 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 capacitance unbalances of each quad at the

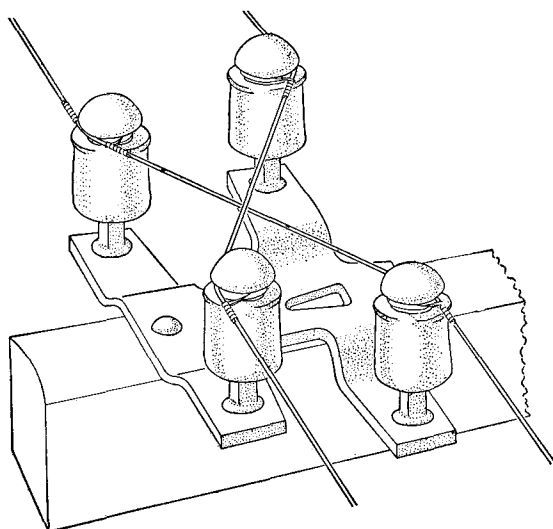


FIG. 34-11. POINT-TYPE TRANSPOSITION

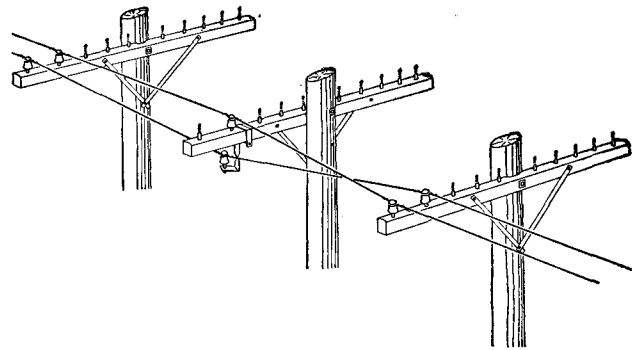


FIG. 34-12. DROP-BRACKET TRANSPOSITION

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. One depends 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 is counteracted by

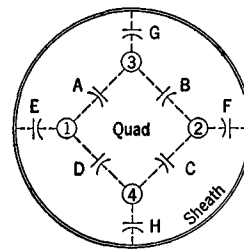
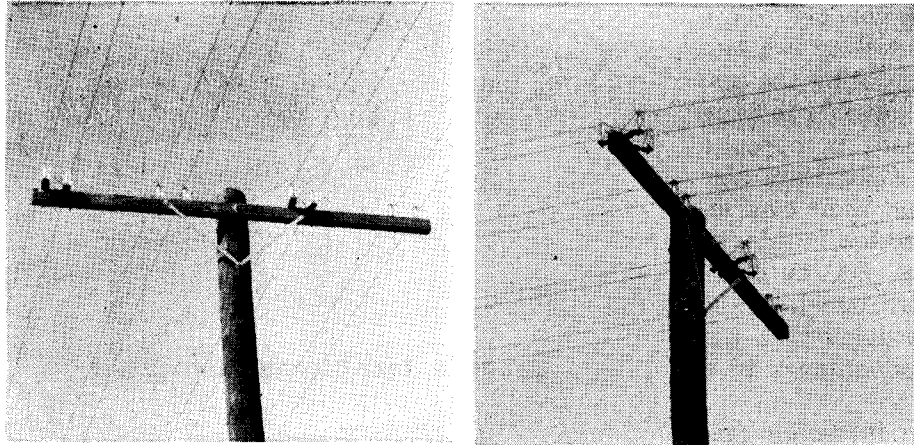


FIGURE 34-13

an equal and opposite unbalance in the adjacent section. This will perhaps be made clearer by referring to Figure 34-13 where the four wires of a quad are shown in cross-section, with the capacitances between the wires indicated by small capacitors. The wires marked 1 and 2 form one pair of the quad, and the wires marked 3 and 4 the other. (The capacitances 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 capacitances  $A$ ,  $B$ ,  $C$ , and  $D$  shall be equal, and that capacitance  $E$  shall equal capacitance  $F$  and capacitance  $G$  shall equal capacitance  $H$ . In this case there is no unbalance within the quad and no crosstalk. However, if it is found, for example, that capacitance  $A$  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 capacitance  $A$  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.

In the second method, the unbalances are counteracted in part by connecting small balancing capacitors into the circuits at one or two



TRANSPPOSITIONS IN 8-INCH SPACED OPEN WIRE LINE USING POINT-TYPE FIXTURES

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 capacitance unbalance tests that have to be made when a cable is installed. These balancing capacitors consist of short lengths of two parallel insulated fine-gage wires wound helically around a non-conducting core. Two terminals of this tiny capacitor are connected across the two line conductors whose capacitance it is desired to increase, and the other ends of the wires can be cut off at whatever point is necessary to give the capacitor the precise value required. A large number of these capacitors can be included within the 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 capacitances for a whole repeater section by adding capacitors of this type at one end.

The capacitance 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 more severe. In this case, while capacitive 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 capacitive 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 such 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. 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 capacitive and inductive couplings. In addition, special precautions are taken to prevent interaction 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 34-14, 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 carrier frequency currents over the voice-frequency conductors.

The methods of balancing out capacitive coupling between the carrier pairs themselves are essentially the same as were discussed above in connection with voice-frequency transmission. In balancing out crosstalk due to inductive coupling, different methods must be used. 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 in effect 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

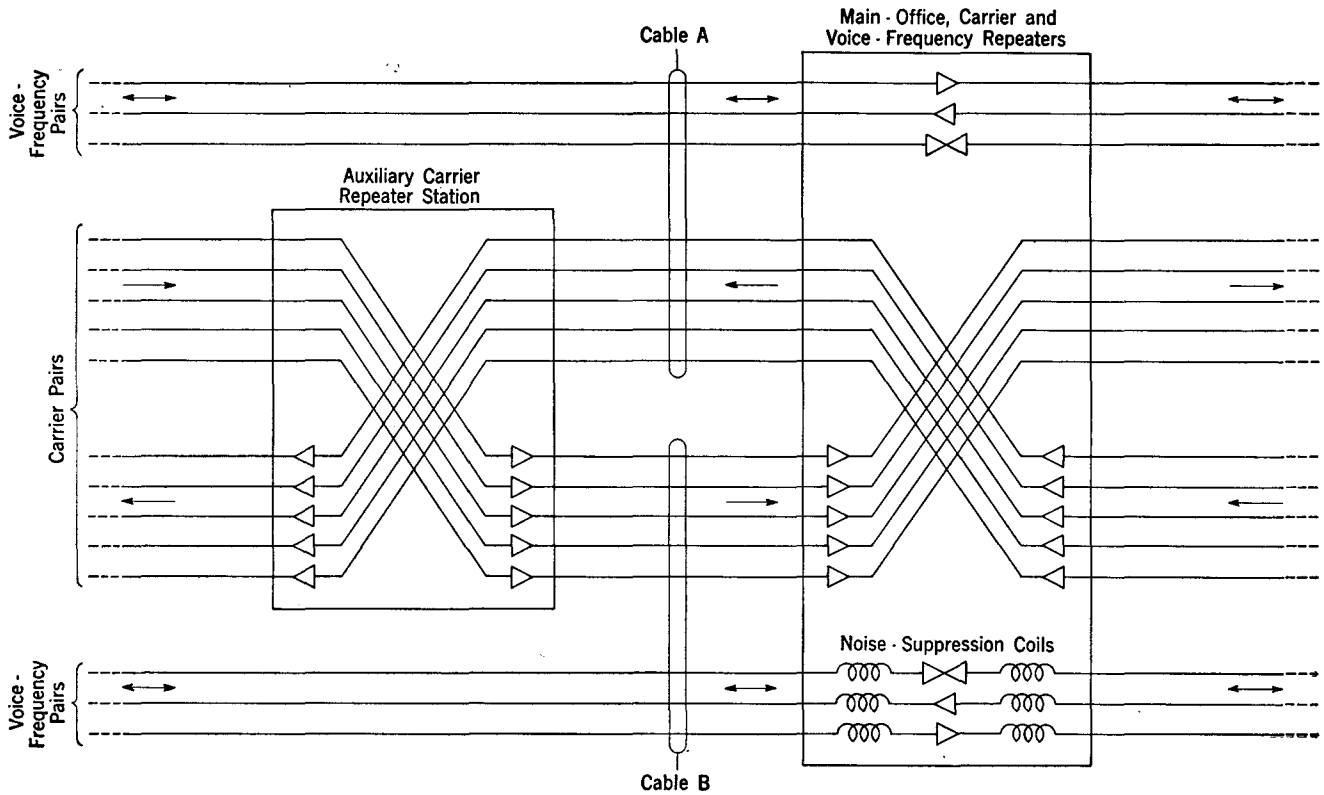


FIG. 34-14. ARRANGEMENTS FOR REDUCING NOISE AND CROSSTALK IN TYPE-K CARRIER SYSTEMS

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 34-15. 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 cir-

cuit. 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.

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 cylindrical containers arranged for rack mounting. The position of the coil cores is controlled by a screw by means of which the core can be moved through its maximum travel of  $\frac{1}{2}$  inch in about 16 complete turns.

In using these coils to balance out crosstalk, measurements of the inductive coupling between

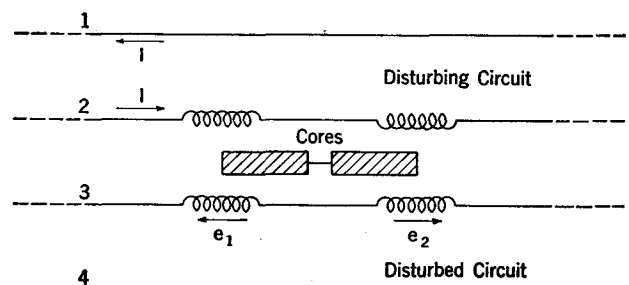


FIG. 34-15. PRINCIPLE OF CROSSTALK BALANCING COIL



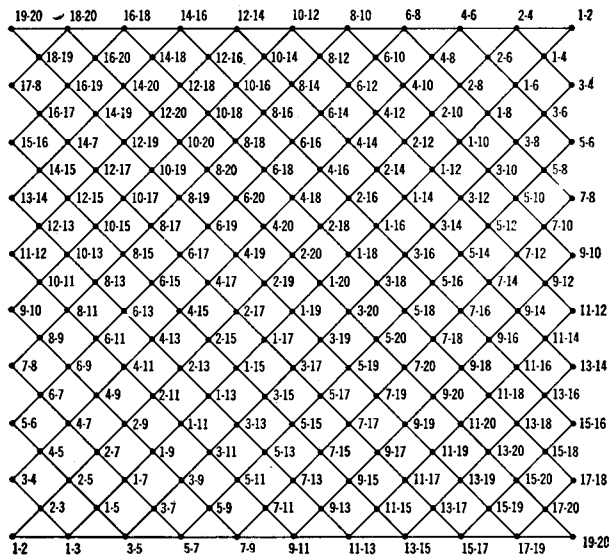


FIG. 34-16. METHOD OF CONNECTING CROSSTALK BALANCING COILS

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 34-16, 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.

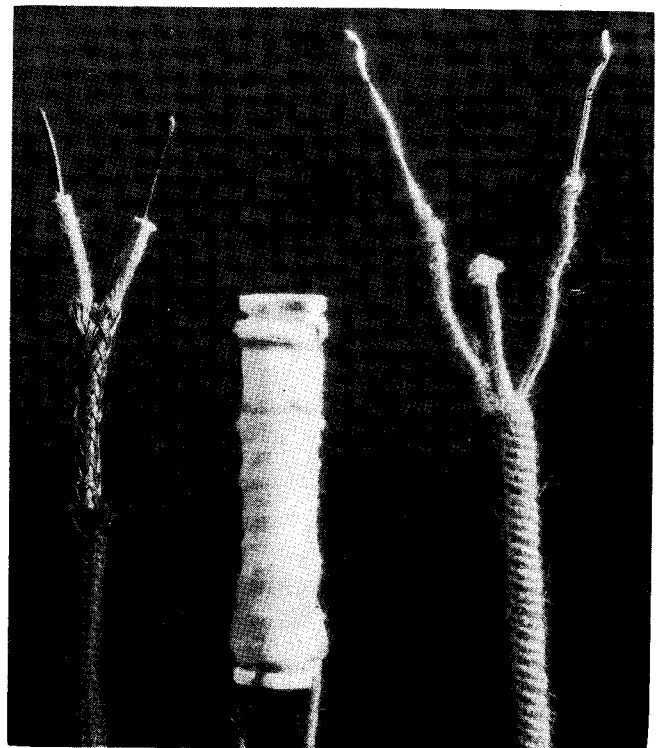
### 34.5 Noise

Voltages within the voice-frequency range, induced in a telephone circuit by electric power circuits, are manifested to a listener on the 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 dis-

turbed 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.

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 34-17 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



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

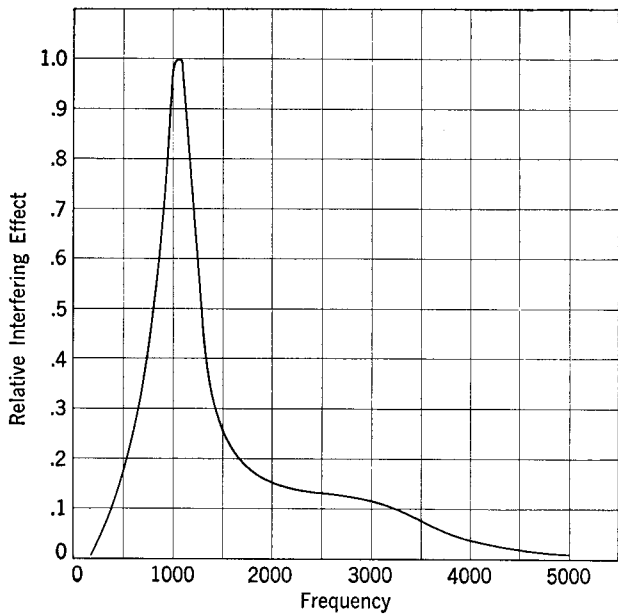


FIG. 34-17. RELATIVE INTERFERING EFFECT OF NOISE AT DIFFERENT FREQUENCIES

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.

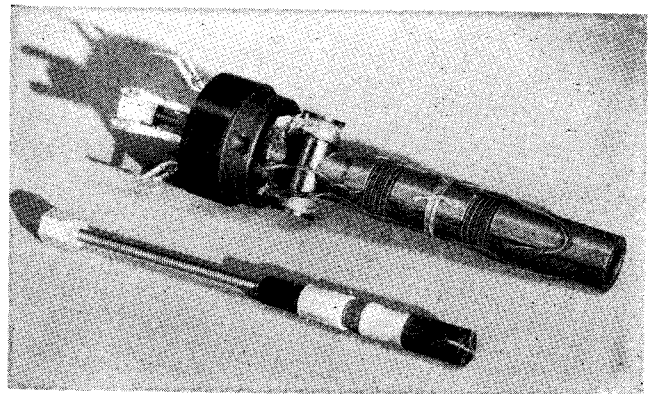
The ideal way of eliminating such interference is to avoid any parallels of small separation and appreciable length between power and telephone lines. This is done whenever practicable. In many cases, however, parallels with fairly close separation, such as lines on the opposite sides of a highway, cannot 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.

### 34.6 Crosstalk and Noise Measuring Units

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 desirable 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.

The measure of either crosstalk or noise that would be of major significance as a practical matter is the extent of the interference or annoyance to which a listener on a disturbed circuit is subjected. Since such a measure is obviously affected by numerous subjective factors, it is clear that completely objective quantitative measurements of crosstalk and noise effects are practically impossible. It is possible, however, to make precise quantitative measurements of the **crosstalk coupling** between a given sending point on a disturbing circuit and a given receiving point on a disturbed circuit. Essentially this is simply the measure-



CONSTRUCTION OF CROSSTALK BALANCING COIL

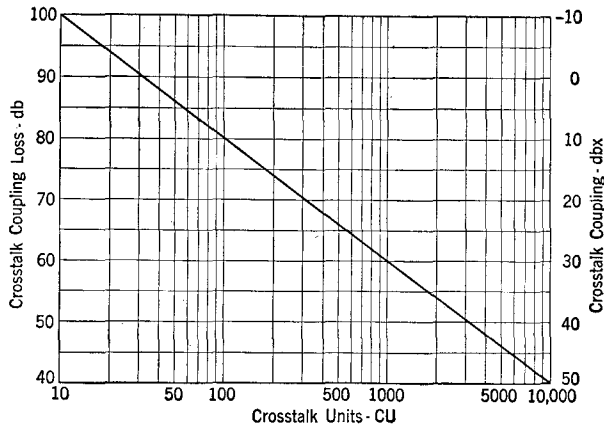


FIG. 34-18. RELATIONS BETWEEN CROSSTALK MEASURING UNITS

ment of the transmission loss between the two points, and like any other transmission measurement it may be made at one or more frequencies as desired. Such a measurement gives a value of what is known as "crosstalk coupling loss" in db. A more commonly used measure of crosstalk coupling employs a unit designated dbx, which expresses the coupling in db above "reference coupling". Reference coupling is equivalent, broadly speaking, to a crosstalk coupling loss of 90 db and is formally defined as "the coupling which would be required to give a reading of zero dba on a 2-type noise measuring set connected to the disturbed circuit when a test tone of 90 dba (using the same weighting as that used on the disturbed circuit) is impressed on the disturbing circuit". Another unit is sometimes used for measuring crosstalk coupling. This is the "crosstalk unit" abbreviated CU. The number of crosstalk units representing any given coupling is  $10^6$  times the ratio of the current or voltage in the disturbed

circuit to the current or voltage in the disturbing circuit at the two points under consideration; or, if the circuit impedances are not the same,  $10^6$  times the square root of the power ratio. The relationships between the three measures of crosstalk coupling are shown graphically in Figure 34-18.

For measuring noise, a basic reference point has been selected, which is equal to  $10^{-12}$  watts of 1000-cycle power. This corresponds to 90 db below 1 milliwatt ( $-90$  dbm). Noise may then be measured in terms of number of decibels above this reference point. However, the interfering effect of noise on a listener varies with both the power level and the frequency; and the relative importance of the components of noise at the different frequencies must be taken into consideration in determining the total amount of interference. The interfering effect also varies according to the sensitivity of the receiving device that converts the noise currents into audible sound. For these reasons, in measuring noise, it is desirable to employ "weighting networks" which act to integrate the noise power over the voice-frequency range by giving each small band of frequencies a weighting proportional to its contribution to the total interfering effect. Different weighting networks may be used with different receiving devices. Even so, equal values of db reading will not necessarily indicate equal interfering effects without some adjustment of the calibration constants. In practice, an adjusted unit designated dba is employed, which measures the acoustic interfering effect of the frequency-weighted noise energy. Equal values of dba measured across any receiving device, with proper weighting used, should indicate approximately equal interfering effects.

## CHAPTER 35

### ALTERNATING-CURRENT MEASUREMENTS

#### 35.1 Alternating-Current Meters

Meters of the galvanometer type, in which the current to be measured flows through the windings of a movable coil suspended in the field of a permanent magnet, are commonly used for the measurement of direct currents. Such meters are not suitable for measuring alternating currents because they would tend to indicate the instantaneous values of the rapidly changing current, both positive and negative.

For the measurement of alternating currents of relatively low frequency (up to a few hundred cycles), direct-reading meters employing the **dynamometer** principle are frequently used. These are similar in external appearance to the usual d-c meters, but the permanent magnet of the galvanometer type meter is replaced by a stationary coil through whose windings the current to be measured flows in series with the windings of the movable coil. Current reversals thus occur simultaneously in both coils and the interaction between their magnetic fields is such as to cause the movable coil to rotate in only one direction regardless of the direction of the current in the windings. The deflection of the movable coil is proportional to the **effective** value of the current through its winding. Within the frequency limits indicated, this type of instrument may be used to measure alternating current, voltage or power provided that the values are of sufficient magnitude to register. For power measurement, however, one of the coils must be connected in series so that the current in it varies as the line current, while the other coil is connected across the circuit so that the current in it is proportional to the voltage. This automatically takes care of any phase difference between current and voltage, permitting the meter to register in watts.

The direct-reading a-c meter is ordinarily not adequate for measuring current, voltage and power at the higher frequencies generally employed in communications work because of the extremely small values of the quantities involved and other difficulties inherent in the design of a-c meters at high frequencies. Since it is possible to build d-c meters with a much higher degree of

sensitivity, the great majority of a-c measurements in communications work are made indirectly by rectifying the alternating current and then measuring the rectified current with a d-c meter. If the a-c quantities are still too small to register on a meter, an appropriate amplifier may be inserted in front of the rectifying device. Where it is not necessary to use an amplifier, the conversion of small alternating currents to proportional direct currents is usually effected by a thermocouple. The arrangement is illustrated schematically in Figure 35-1 where the alternating current to be measured, flowing through the wire *ab*, heats the junction *H* of two dissimilar metal conductors. Due to thermo-electric effect, this produces a difference of potential between the two dissimilar conductors, which causes a direct current to flow through the d-c meter. When the meter is properly calibrated by means of the potentiometer *P*, this current is exactly equal to the effective value of the applied alternating current. Where the a-c quantities are so small as to require the use of an amplifier for their measurement, rectification is usually effected by means of either disc or point-contact rectifiers.

For most purposes, a-c measurements in communications work are not made in terms of volts or amperes but in logarithmic units such as the db or dbm. Many of the meters used are accordingly designed with scales reading in these units. The ordinary d-c ammeter is designed to have equal spacing on its scale for equal increments of current. Since the movable coil rotates in an air-gap of uniform flux density, the torque on the coil is directly proportional to the current. If such a meter is calibrated to read in decibels, the points on the scale for successive units above or below zero will be spaced increasingly close together, as shown in Figure 35-2 (A), because the decibel is a

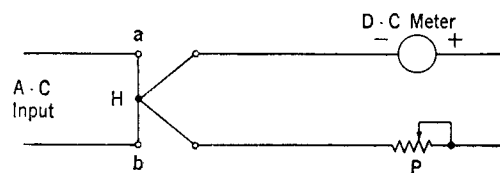


FIG. 35-1 THERMOCOUPLE A-C METER

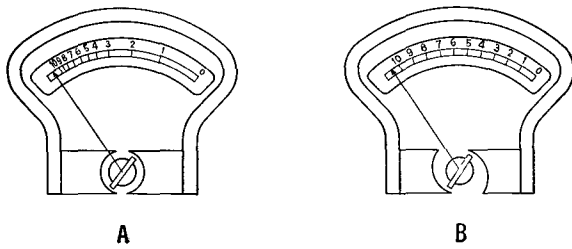


FIG. 35-2 DECIBEL METER

logarithmic unit. To obtain the even spacing on the db or dbm scale that is generally desirable, an ammeter may be made as shown in Figure 35-2 (B), with the pole-pieces of the permanent magnet cut in an eccentric pattern so that the flux across the air-gap increases logarithmically as the needle moves away from the zero position. The advantages of such a design are somewhat offset by the fact that this meter requires a higher operating current because of the lower flux at the zero position.

It should be noted that the zero of a db or dbm scale does not have the same significance as that of an ordinary linear scale because the decibel by definition is the logarithm of a ratio, and zero is the logarithm of 1. When an ordinary ammeter reads zero, it means that no current is flowing in its winding. When a db meter reads zero, it means only that the ratio between the power being measured and some **reference power** is 1. This reference power, as we know, is 1 milliwatt for measurements in dbm. Under other conditions, the relationship between meter reading and reference power may be complicated.

Although not explicitly stated, it has been assumed in the foregoing that the alternating currents and voltages dealt with were in the form of sine waves. When this is not the case, measurements in terms of the units discussed above may have little meaning. Speech waves, for example, vary with time in a complex manner and it is not possible to measure their precise values in terms of watts or decibels except on an instantaneous basis. It is frequently desirable, however, to know the overall average strength of transmitted speech or music waves. A device known as a **volume indicator** may be used for this purpose. This is essentially a high impedance voltmeter made up of an attenuator, a copper-oxide rectifier, and a d-c milliammeter having specified dynamic characteristics, which may be bridged across a line without appreciable effect on normal transmission. The meter is calibrated in terms of VU. This is a

logarithmic unit that measures strength or volume above or below a specified reference level. In general, this reference level, designated 0 VU, indicates no precise electrical quantity, but the volume indicator is calibrated to read 0 VU on 1 milliwatt of 1000-cycle power dissipated in a 600-ohm resistance. The VU represents the same power ratio as the db and the volume indicator may therefore be used to measure transmission losses or gains when the current being measured is a 1000-cycle sine wave, although its primary purpose is to measure the volume of complex waves.

Another example of measurements where the ordinary electrical units cannot be practicably applied is presented by the video signal of television program transmission. This signal is not only non-sinusoidal, but its positive and negative peaks are usually unequally displaced from the zero axis. Under these conditions, a measurement of the signal amplitude in ordinary power terms is meaningless. As has been indicated earlier in this book, video signal amplitude is commonly defined in terms of peak-to-peak voltage—that is, the voltage difference between the maximum swing of the signal in one direction and that in the other direction. For general measurements of video signal strength, it is convenient to use a unit known as the dbv. This expresses the ratio in db of the peak-to-peak voltage at any point to a reference voltage of 1 volt peak-to-peak. The 1 volt peak-to-peak reference voltage is zero dbv and for a sine wave its value is equivalent to an effective value of 0.354 volts.

### 35.2 The Cathode-Ray Oscilloscope

The cathode-ray oscilloscope is fundamentally a voltage measuring device that is particularly useful for observing and measuring a-c voltages of non-sinusoidal waveshape. Its basic element is an electrostatic cathode-ray tube, which, because of the almost negligible inertia of its electron beam,

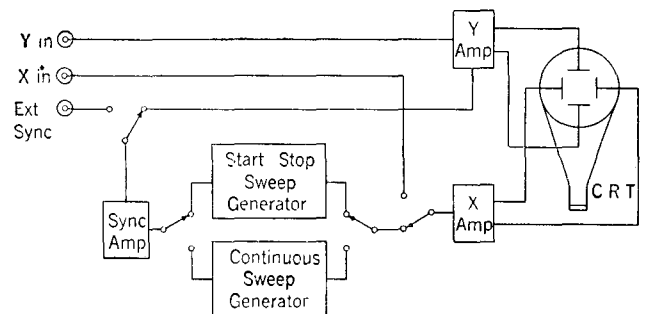


FIG. 35-3 BLOCK DIAGRAM OF CATHODE-RAY OSCILLOSCOPE

makes possible the visual observation of the characteristics of extremely rapid voltage changes, provided that they recur with sufficient frequency to produce a persistent light pattern on the screen of the tube.

Figure 35-3 shows in block diagram the essential circuit elements that are ordinarily associated with the cathode-ray tube to make up an oscilloscope that is adaptable for the observation and measurement of signals of many types. For most observations, the voltage under test is applied to the vertical or *Y* deflecting plates of the tube while a linear sweep voltage of the same frequency as the voltage under observation is applied to the horizontal or *X* deflecting plates. The amplifiers designated *Y* and *X* are used to bring the deflecting plate voltages up to the values necessary to cause the desired swings of the electron beam. There are two sweep generators, both usually of the multivibrator type. One is free-running to give a continuous saw-tooth wave output; the other is of the start-stop type, which stops at the end of each saw-tooth cycle until started by a "triggering" signal. The latter makes it possible to obtain a steady trace of a signal whose recurrence is not uniform with time. Switches are provided as shown in the diagram for connecting either generator into the circuit. Both sweep generators are arranged with appropriate controls for adjusting their frequencies over a wide range.

In order to hold the sweep circuits in the exact synchronism that is necessary to produce a stationary pattern on the tube screen, synchronizing pulses are applied to the multivibrators through a SYNC amplifier, as indicated. These pulses are ordinarily obtained internally from the output of the *Y* amplifier, but provision is made for picking them up from an external source by operating the switch to the EXT SYNC position. When a linear time scale is applied to the *X* plates of the tube, as discussed above, the screen gives a direct picture of the shape of the wave under investigation. By placing an appropriately calibrated grid scale on the face of the tube, reasonably accurate quantitative measurements may also be read directly from the picture.

For most of the purposes for which it is used in communications work, the oscilloscope is operated with a time scale applied to the *X* axis. It is possible, however, to apply an independent external voltage to the *X* plates as well as to the *Y* plates. Such a procedure is employed when the oscilloscope is used for measuring frequencies, as discussed briefly in the following Article.

### 35.3 Frequency Measurements

Methods for measuring the frequency of sine wave alternating currents depend upon the frequency range under investigation and the degree of accuracy required. At frequencies in the "power" range, up to about 200 cycles, direct-reading meters employing the principle of mechanical resonance are commonly used. Such meters have a comb-like arrangement of metal reeds of differing lengths, each of which has a different natural period of mechanical resonance. The comb is mounted in the gap of an electromagnet to the windings of which the voltage to be measured is connected. A scale above the ends of the reeds is calibrated in cycles to show the frequency of the reed which vibrates with maximum intensity. At higher frequencies, the principle of electrical resonance is quite generally employed in frequency measurements. Any resonant circuit with adjustable capacitance or inductance values can be used for determining frequency. One illustration of such an application is the wave meter shown in Figure 35-4. In circuit *A*, if an inductance *L* and a capacitance *C* are so adjusted as to give resonance, the meter *M* will give a maximum reading. Similarly, circuit *B* will be resonant when maximum sound is heard in the telephone receivers which are connected in series with a rectifying device, *D*. To know the frequency at the resonant point, it is of course necessary that the variable capacitor be previously calibrated in such terms.

Wave meters of such comparatively simple design are widely used in connection with radio circuits where frequency determination of precise accuracy is rarely necessary. There are many situations in the voice and carrier frequency ranges where frequency measurements accurate to a few cycles, or even to one cycle, are desirable. One device employed for measuring frequencies in these ranges is the bridge circuit shown schematically in Figure 35-5. This also makes use of the principle of electrical resonance but employs a null method of measurement. The bridge arms

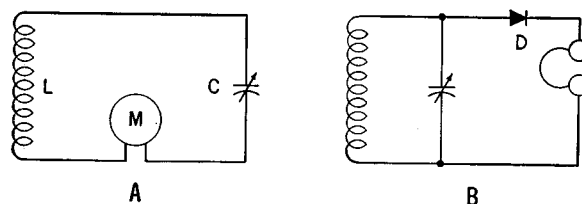


FIG. 35-4 WAVE METERS

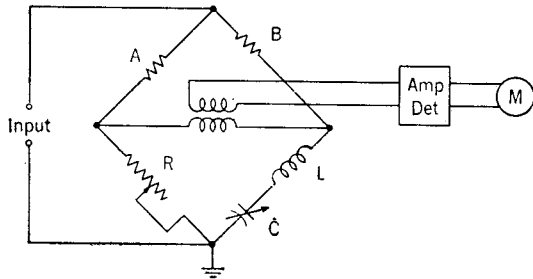


FIG. 35-5 BRIDGE-TYPE FREQUENCY METER

*A* and *B* are of equal resistance and when the variable capacitor *C* and the variable resistor *R* are adjusted so that the meter registers no current, the *LC* arm of the bridge is in resonance with the input frequency. The scales of *R* and *C* are factory calibrated and a table is supplied with the bridge from which the frequency values corresponding to their settings may be read. The accuracy of this device of course depends upon the original accuracy, and the permanency, of this calibration.

Where greater precision is necessary, a meter employing a cathode-ray oscilloscope may be used. The essential elements of this device are shown in the block diagram of Figure 35-6. Its principle depends upon comparing the voltage under test with the output of a precisely calibrated variable oscillator by applying the two voltages respectively to the vertical and horizontal plates of a cathode-ray tube. When the two voltages have precisely the same frequency, a **stationary** pattern will appear on the screen of the tube. As was pointed out in Article 25.3, the form of the stationary pattern will depend upon the phase rela-

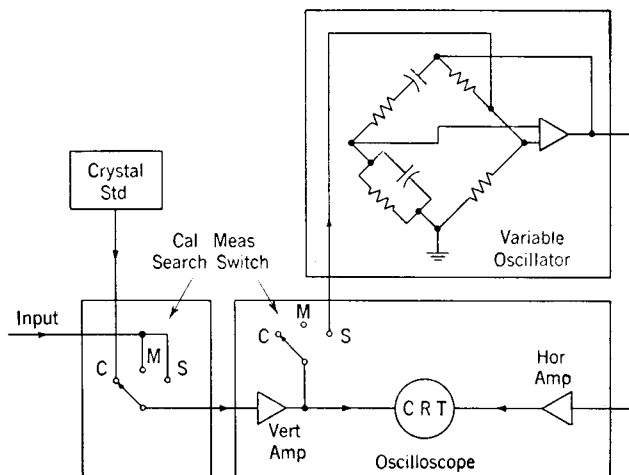


FIG. 35-6 OSCILLOSCOPE CIRCUIT FOR FREQUENCY MEASUREMENTS

tionship of the two voltages, being a straight diagonal line when the voltages are in phase or  $180^\circ$  out of phase; a circle when the voltages are  $90^\circ$  out of phase; and elliptical at other phase angles. When such a stationary 1-to-1 frequency pattern is obtained on the oscilloscope screen, the frequency adjusting dials of the variable oscillator will read the frequency directly to 1 cycle.

The stability of the bridge-type variable oscillator is very high. Nevertheless, it is necessary to calibrate it each time the meter is placed in operation and at intervals of about an hour during operation, if maximum precision is required. For this purpose, the meter includes a crystal standard whose known single frequency may be compared in the oscilloscope with several key frequencies in the variable oscillator's range, that are related to the fixed standard frequency by integral units. In these calibrating comparisons, the oscillator circuit constants are adjusted by "trimmer"

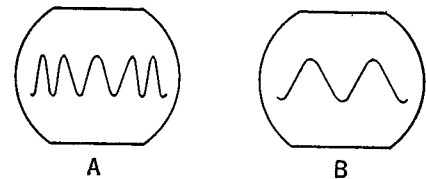


FIG. 35-7 OSCILLOSCOPE FREQUENCY PATTERNS

resistors, if necessary, until a stationary pattern is obtained on the screen. Since there is only one crystal standard frequency, these stationary patterns will not represent 1-to-1 ratios but other integral relationships. Figure 35-7, for example, shows the patterns for a 10-to-1 and a 4-to-1 relationship, which correspond to oscillator frequencies of 400 and 1000 cycles respectively when the crystal standard is 4000 cycles.

When the frequency of the voltage being measured is not even approximately known, the meter circuit can be arranged to search for the unknown frequency by operating the switch to the contacts marked *S*. Operation of the search switch converts the oscillator to a tuned amplifier. The amplifier output will then be negligible until and unless its tuning circuit is adjusted to a frequency that is close in value to the unknown input frequency. As long as the frequencies are far apart, only a straight vertical line will appear on the oscilloscope screen. When the tuning of the amplifier approaches the unknown frequency, this vertical line will take the form of an ellipse that will bend over and approach closer and closer to a straight diagonal line, as shown in Figure 35-8.

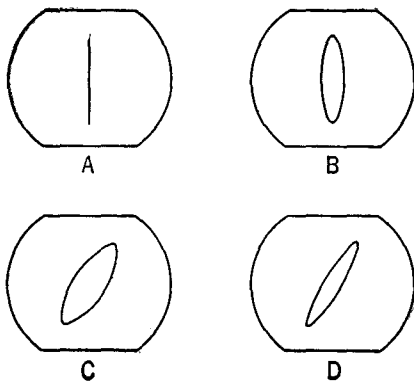


FIG 35-8 SEARCH OSCILLOSCOPE PATTERNS

At this point, the switch may be thrown back to *M* and the exact frequency determined.

### 35.4 Impedance Measurements

Many times throughout this text, attention has been called to the necessity for matching impedances at junction points of communication circuits in order to eliminate unnecessary transmission losses or other undesirable effects. This makes it necessary, for practical maintenance purposes, to have available a device by means of which impedances can be accurately measured.

Figure 35-9 indicates the principle of a simple bridge circuit widely used in the telephone plant for measuring impedances in the voice-frequency range between 100 and 3000 cycles. As shown, the unknown impedance is connected in one arm of the bridge and the balancing arm consists of a variable resistor and a variable inductor (inductometer) in series. Arms  $R_a$  and  $R_b$  are resistors of equal value. Measuring current is supplied from a variable oscillator capable of delivering satisfactory waveshape and output through the range of voice frequencies for which the bridge is de-

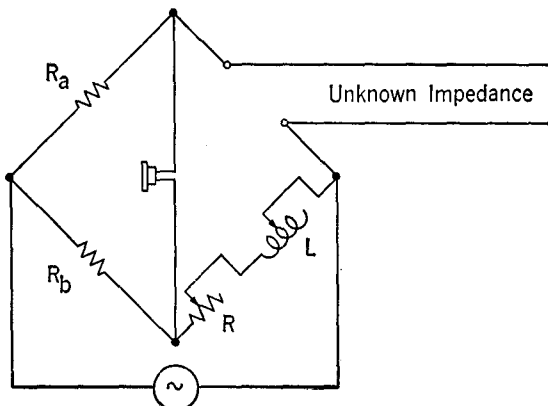


FIG. 35-9 SIMPLE IMPEDANCE BRIDGE

signed. The values of  $R$  and  $L$ , when they are adjusted so that there is no current in the telephone receiver, will be equal to the corresponding values of the unknown impedance. The circuit as shown in the diagram could measure only an inductive impedance. The practical circuit, however, is arranged so that the variable inductor may be switched into the other arm of the bridge in series with the unknown impedance. When the bridge is balanced in this condition, the inductometer in effect gives a measure of negative inductance, which is equivalent to capacitance. The variable units are actually calibrated to read resistance in ohms and inductance in millihenries, but the readings may readily be converted into

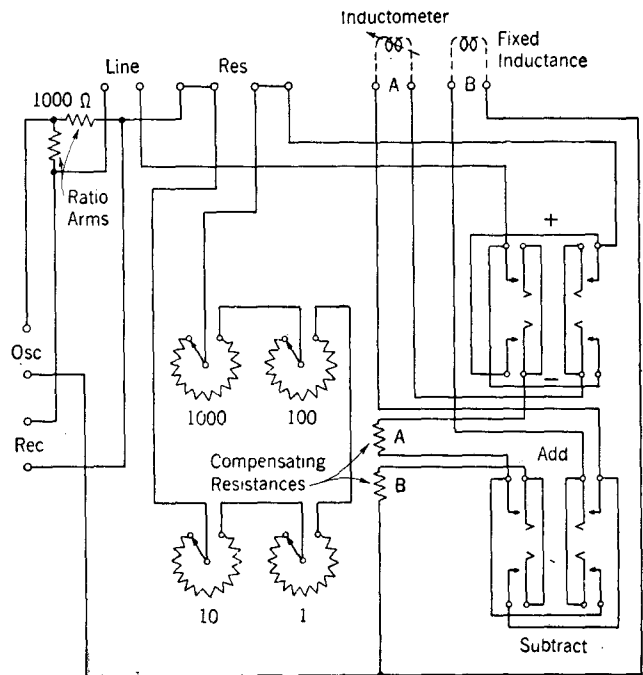


FIG. 35-10 CIRCUIT OF A LINE IMPEDANCE BRIDGE

reactance and impedance values by the application of basic a-c equations.

A detailed wiring diagram of the bridge is given in Figure 35-10, which shows a number of features not indicated in Figure 35-9 but necessary to the practical operation of the bridge. Thus, the fact that it is not possible to build an inductometer that may be adjusted to zero inductance requires the use of a second fixed inductor which also may be switched from one arm of the bridge to the other. Zero inductance may then be secured by switching the fixed inductance to one arm of the bridge and the inductometer to the opposite arm so that the fixed inductor neutralizes its value



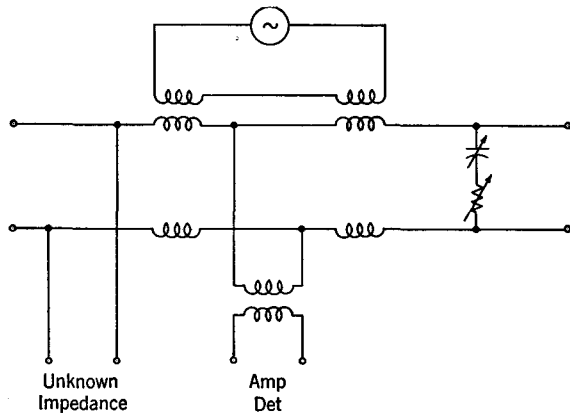


FIG. 35-11 HYBRID-TYPE IMPEDANCE BRIDGE

on the scale of the inductometer. Since either inductor may be switched to either arm, the total reactance values that can be measured range from zero to  $\pm 2\pi f (L_a + L_b)$ , where  $L_a$  and  $L_b$  are the values of the two inductances. The diagram also shows two compensating resistors, whose values are equal to the resistances of the inductors. These are connected to the switch contacts in such a way as to be always in the opposite arm from the corresponding inductor, thus neutralizing the resistances of the inductors.

Other bridge designs, operating on a basically similar principle, are used for impedance measurements at higher frequencies. One of these, which is satisfactory for measurements between 1800 and 35,000 cycles, is shown schematically in Figure 35-11. The bridge here is the familiar hybrid coil. When the unknown impedance connected to the "line" side of the coil is matched by the adjustable impedance connected to the "net" side of the coil, voltage applied to the series winding from an oscillator will produce no current in the bridge connection to the amplifier-detector. It will be noted that the reactance adjustment in this circuit is made by means of a variable capacitor rather than an inductometer. If the reactance of the unknown impedance is inductive, the variable capacitor is transferred by an appropriate switch to the line side of the coil in series with the unknown impedance. Another bridge, designed for making measurements between 1 and 100 kc, is shown in Figure 35-12 in simplified schematic. This bridge differs from the usual circuit in that the ratio arms are four pairs of equal resistances, and the variable and unknown impedances are connected between mid-points of opposite pairs. The impedance is measured when the bridge is balanced in terms of resistance and capacitance

in parallel rather than in series, and switches are provided to transfer the variable elements to the opposite side of the bridge if this should be necessary to secure balance.

One of the major uses of the impedance bridge in practical communications work is the location of impedance irregularities in long wire circuits. The impedance of a long line that is free from irregularities and terminated in its characteristic impedance, when measured over a wide band of frequencies, will appear as a smooth curve over the measured frequency range. If, however, there is an impedance irregularity along the line, such as might be caused by a defective or improperly located loading coil, some part of the energy applied to the line at the sending end will be reflected back from the point of irregularity. The reflected wave will add to or subtract from the initial applied wave, depending on its phase relationship when it reaches the sending end. The sending end impedance will be affected accordingly. The phase of the reflected wave with respect to the initial wave of course depends on the time it takes to travel from the irregularity to the sending end or, since the velocity of propagation is a constant for a particular type of facility, on the distance from the irregularity to the sending end.

The measured resistance and reactance of the line under these conditions, when plotted over a range of frequencies, will produce curves of the type illustrated in Figure 35-13. At points *M*, *N*, and *O*, the measured resistance values are maximum, indicating that the reflected current, when it reaches the sending end, is exactly opposite in phase to the initial current; while at *P*, *Q*, and *R*, the two currents are exactly in phase. Adjacent humps, either maximum or minimum, occur at frequency intervals corresponding to a change of

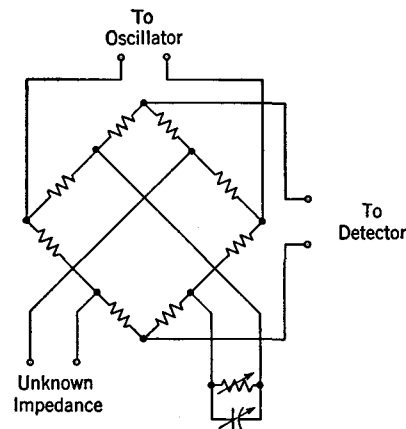


FIGURE 35-12

one-half wavelength in the distance from the sending end to the point of irregularity. The distance may, therefore, be determined by the use of the following equation:—

$$d = \frac{W}{2(f_2 - f_1)} \quad (35:1)$$

where  $d$  is the distance to the irregularity,  $W$  is the velocity of propagation, and  $f_2 - f_1$  is the average frequency interval between adjacent humps of the curve.

Where a large number of impedance-frequency measurements are to be made, particularly when the measurements need to be made quickly because the fault condition causing the irregularity is not constant, there are advantages in using a rectifying bridge of the type shown in Figure 35-14 that employs only one variable element. In this bridge, the unknown impedance is balanced by the variable resistance  $R$  and the bridge will therefore measure the magnitude but not the angle of the impedance. Current from the oscillator can flow in the meter circuit  $M$  only during the half-cycle when  $a$  is positive. In the negative half-cycle, current will flow through paths  $bca$  and  $bda$  but no current can flow in the meter circuit since any voltage applied across the rectifiers  $X$  in series with the meter will be of the wrong polarity for either one or the other of them. The bridge, therefore, acts as a half-wave rectifier. During the positive half of the cycle, current from  $E$  divides at  $a$  and flows through the paths  $acemb$  and  $adfmb$ . If the resistances of these two paths are equal, there is no potential difference between  $e$  and  $f$  and no current will flow through the meter. The bridge is calibrated by adjusting the potentiometer  $m$  so that the meter will read zero when

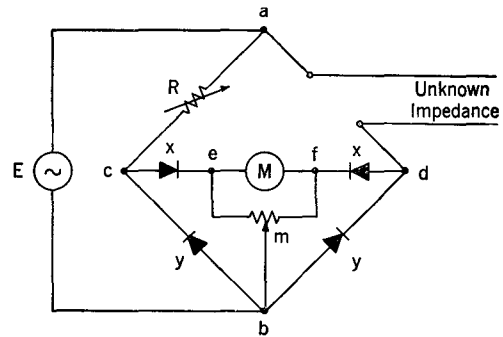


FIG. 35-14 RECTIFYING-TYPE IMPEDANCE BRIDGE

equal resistances are inserted in the arms  $ac$  and  $ad$ . This corrects for any differences that may exist between the resistances of the two  $X$  varistors. The calibrating resistances should be of a value comparable, as nearly as possible, to the magnitude of the unknown impedance to be measured.

### 35.5 Transmission Measurements

Most widely used of the many types of measurements required in communications work are those known as transmission measurements. These are measurements of the ratio of the power at the receiving end of a transmission line to the power applied to the transmitting end. As is well known, they indicate the loss or gain of a circuit in terms of decibels or comparable logarithmic units. Two basic methods of making transmission measurements are commonly employed. The first is a direct method in which a known amount of power (generally 1 milliwatt) is applied to the sending end of the circuit under test and the power at the receiving end is measured by a direct-reading meter in terms of db or dbm. This is obviously the simpler method and is used wherever practicable. In situations where it is not feasible to supply a known fixed power at the sending end of the circuit, a comparison method is used in which the loss or gain of the circuit under test is measured by comparing it with a known, calibrated loss or gain.

For routine checking of telephone circuits, transmission measurements are usually made at a single frequency of 1000 cycles and, in most cases, the direct method of measurement is employed. Fixed testing power of 1 milliwatt is supplied at the sending terminals from a 1000-cycle source of power, which consists of a small magneto-generator. At the receiving end, the power is amplified, rectified by copper-oxide varistors,

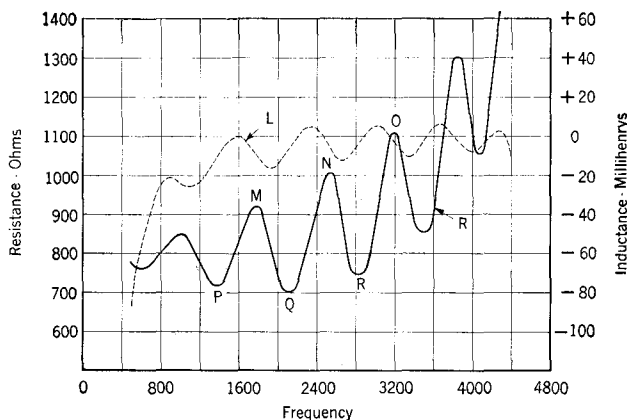


FIG. 35-13 IMPEDANCE CHARACTERISTIC OF CIRCUIT WITH SINGLE IRREGULARITY

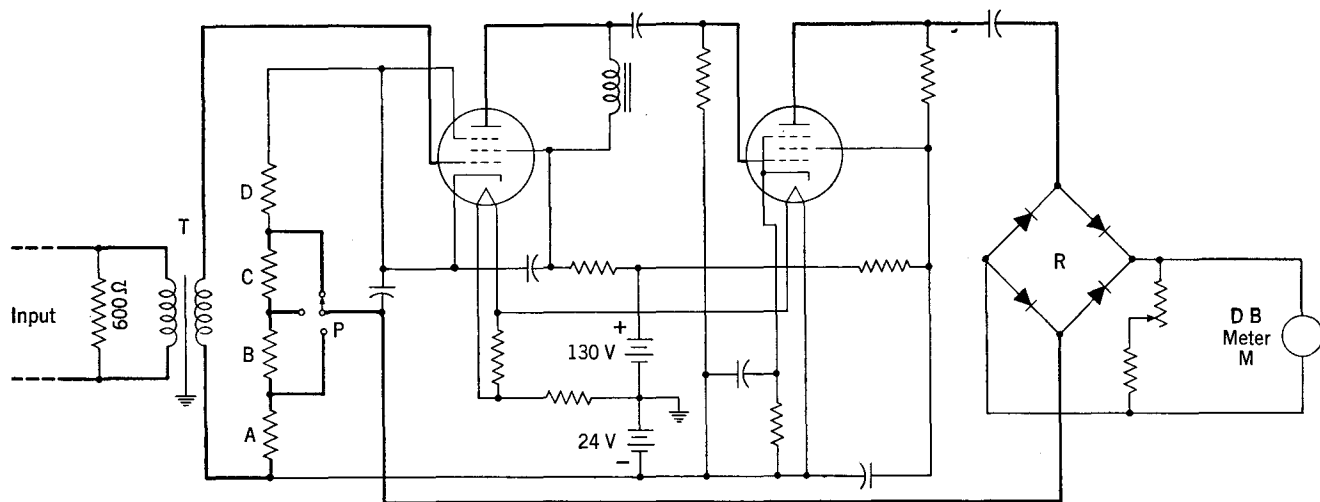


FIG. 35-15 DIRECT READING TRANSMISSION MEASURING SET WITH AMPLIFIER

and supplied to a d-c meter reading directly in db or dbm. The detailed circuit arrangement is shown in Figure 35-15. Where measurements at frequencies other than 1000 cycles are required, the same receiving circuit may be used but the sending power is furnished by an appropriate variable oscillator. To insure that the test power is at 1 milliwatt, the oscillator output must be calibrated against a fixed 1000-cycle generator output for each series of measurements at other frequencies.

In situations where a fixed known testing power source is not available, as would ordinarily be true for example in the case of portable transmission measuring sets, the comparison method mentioned

above may be employed. The general principle of this type is illustrated in Figure 35-16. 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 passed 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 cali-

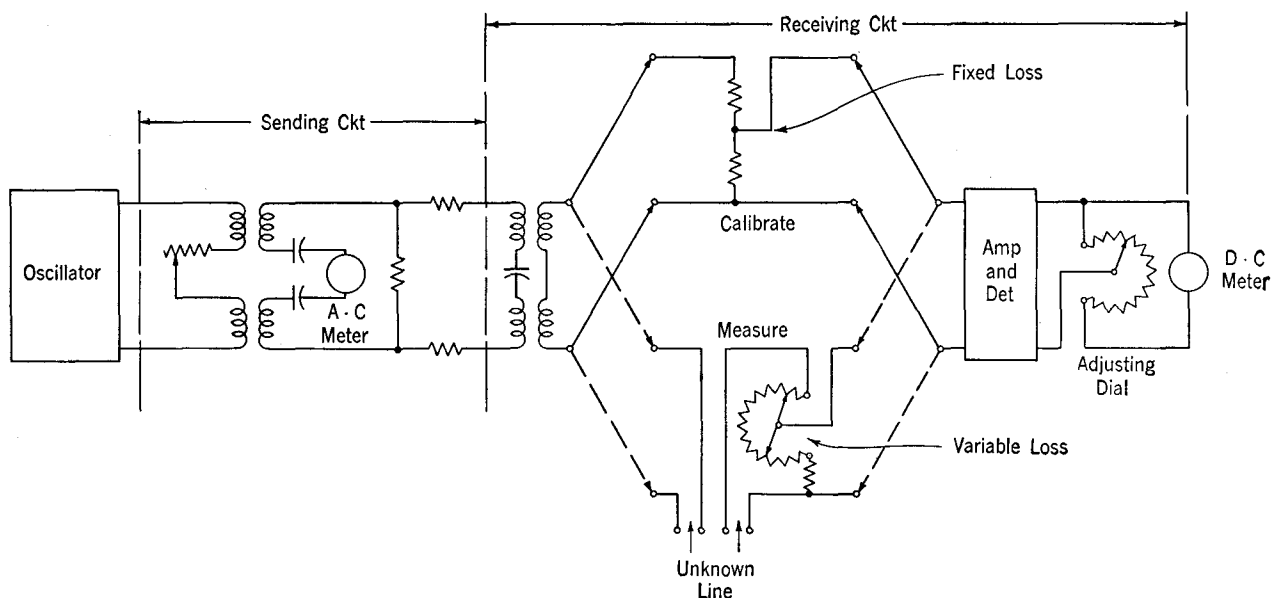


FIG. 35-16 PRINCIPLE OF TRANSMISSION MEASURING SET

brating 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.

For transmission measurements at higher frequencies up to 3500 kc, both comparison type and direct-reading sets are extensively used in the telephone plant. The principles involved are not essentially different from those already discussed for measurements at voice-frequencies, although the measuring sets themselves are necessarily somewhat more elaborate in design. The comparison type sets generally employ thermocouple detectors to drive a direct-reading meter. The receiving circuits of the direct-reading sets are essentially superheterodyne detectors, the outputs of which are fed to d-c milli- or microammeters reading directly in dbm. Appropriate types of variable oscillators must of course be employed with each measuring set.

In practical transmission maintenance work, it is often necessary to check transmission losses over a considerable range of frequencies to insure that a line is properly equalized to provide uniform transmission over the entire frequency range of the signals to be transmitted. This is particularly true in the case of television signals, where attenuation distortion must be carefully guarded against. Such "attenuation-frequency runs" can be made with most of the types of measuring sets discussed above, but the process would obviously be extremely laborious if separate measurements had to be made over a frequency range up to about 3500 kc and the frequency-attenuation curve plotted. To obviate this difficulty, a visual type of transmission measuring set may be used, which plots the attenuation over the entire frequency range under study almost instantaneously on the screen of a cathode-ray oscilloscope. The arrangement of such a transmission measuring set is shown in block diagram in Figure 35-17.

The sending circuit here is designed to apply to the line under test 1 milliwatt of power over a range of frequencies sweeping linearly with time between 70 kc and 4500 kc, at a repetition rate of 50 times per second. As the drawing indicates, the sweep is developed from a heterodyne oscillator in which the 90 megacycle output of a fixed oscillator is combined in a modulator circuit with the output of a variable oscillator, which sweeps linearly over the range 85 to 90 mc. The linear sweep of the variable oscillator is controlled by a 50-cycle saw-tooth oscillator whose output controls the effective capacitance of a reactance tube

in the resonant circuit of the variable oscillator. The modulator output includes the sum and difference products of the two oscillators. The sum is in a very high frequency range and is heavily attenuated, while the difference product between zero and 5 mc is transmitted to the amplifier, which has a flat gain over the range 70 to 4500 kc.

The amplifier output is supplied to a "gate tube" which is normally merely a wideband amplifier. A portion of the amplifier output is also supplied to an automatic volume control circuit which holds the net amplifier output constant over

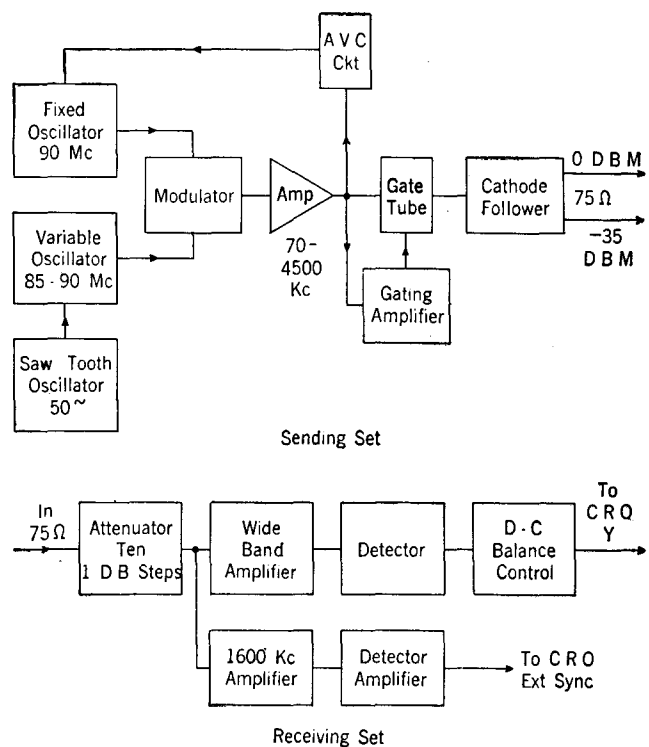


FIG. 35-17 VISUAL TRANSMISSION MEASURING SET

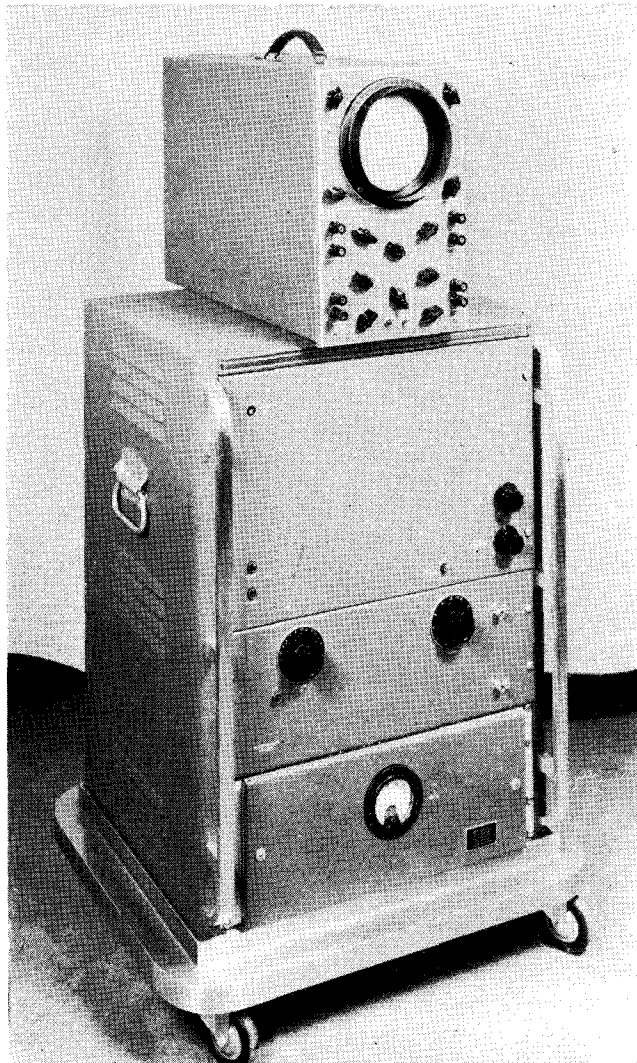
the frequency range. Another portion is supplied to the "gating amplifier". The purpose of this device is to prevent frequencies at, or in the immediate neighborhood of, the L-carrier pilot frequencies from reaching the line where they would naturally tend to upset the line-up of the whole system. The gating amplifier circuit contains sharply resonant circuits tuned to the pilot frequencies, by means of which it is able to recognize when the sweeping frequency of the sending set is approaching a pilot frequency. Upon such recognition, it delivers a "squenching pulse" to the gate tube, which effectively blocks transmission through that tube's circuit of a narrow band of frequencies extending about 50 cycles on each side

of the pilot frequency. The output of the gate tube is delivered to the line through an impedance-matching cathode-follower circuit. The output power is normally held at 1 milliwatt, although a 35 db pad may be switched into the circuit to provide a -35 dbm output when a lower level of output power is desirable.

The receiving circuit includes an attenuator, a wideband amplifier and a germanium crystal detector circuit whose output is connected through a d-c balance control circuit to the Y plates of a cathode-ray oscilloscope. The purpose of the d-c balance control circuit is merely to center the pattern on the oscilloscope screen. The receiving circuit also includes a synchronizing circuit for keeping the horizontal time sweep of the oscilloscope in phase with the oscillator sweep of the sending circuit. For this purpose, a frequency of

1600 kc (any other single frequency in the band would do equally well) is picked off in front of the wideband amplifier by an amplifier tuned sharply to 1600 kc. This is rectified in the detector-amplifier to apply a synchronizing pulse to the EXT SYNC connection of the oscilloscope each time the received sweeping input passes through 1600 kc.

As each frequency over the 70 to 4500 kc range is received, a d-c voltage proportional to its amplitude is produced by the receiving circuit and applied to the Y axis of the oscilloscope. When the X axis of the oscilloscope is swept at the same rate as that of the sending sweep oscillator, a point is plotted for each frequency and a continuous line is therefore drawn (except for the very narrow bands suppressed at the pilot frequencies). If attenuation over the whole frequency range is uniform, the oscilloscope pattern will be a straight horizontal line; irregularities in the straight line pattern will indicate corresponding deviations in the attenuation at frequencies which may be easily read from the oscilloscope scale. Such deviations may be rapidly corrected by adjusting the appropriate equalizer circuits while observing the pattern on the oscilloscope screen.



VISUAL TRANSMISSION MEASURING SET

### 35.6 Measurements at Superhigh Frequencies

Figure 35-18 is a simplified schematic of a **frequency-power meter** that may be used to measure frequency and power levels in the range between 3600 and 4600 megacycles. As may be noted, the meter circuit includes three principal elements—a waveguide system shown at the upper left, a temperature controlling oscillator circuit below this, and an indicating circuit at the right.

Microwave power is measured by the amount of heat the waves generate in a resistive load consisting of the thermistor bead  $T_1$  indicated at the right of the waveguide system. Frequency is measured by means of the adjustable cavity resonator FM which is coupled to the guide and calibrated to read directly in megacycles. As the movable plunger of the cavity is adjusted to resonance, the reading of the power meter AM will show a sharp dip as a result of most of the energy being absorbed in the cavity and not reaching the thermistor bead. The setting of the dial which mechanically controls the movable plunger then indicates the frequency directly. The input microwave power can be reduced up to 6 db, if desired, by means of an attenuator which consists of a vane of insulating material with a thin

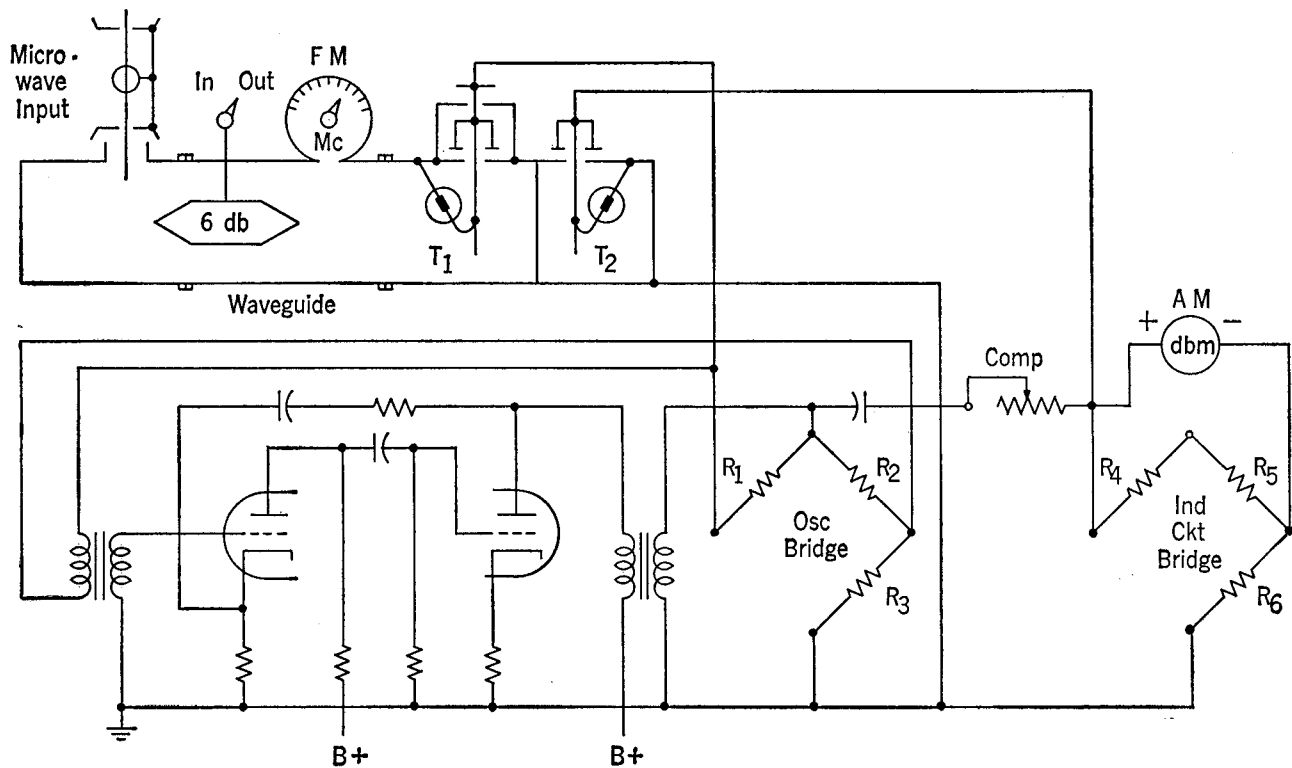


FIG. 35-18 MICROWAVE FREQUENCY-POWER METER

layer of carbon on one side that may be moved toward the center of the waveguide by a calibrated control knob.

The thermistor  $T_1$ , which is the power measuring element, serves also as a matched resistance load. Its resistance must therefore be held at a constant value in order to properly terminate the waveguide. Since the resistance of a thermistor changes sharply with temperature variations, this means that the temperature of  $T_1$  must be precisely held. The temperature controlling oscillator performs this function, and the microwave power absorbed by the thermistor is determined indirectly by measuring the change in oscillator output power that is required to hold the thermistor's temperature constant.

The oscillator, which operates at 85 kc, is of the bridge-stabilized type. It is comprised of a 2-stage amplifier stabilized by negative feedback, with its output connected to its input through a bridge as shown. Thermistor  $T_1$  forms one arm of the bridge and the other three arms are fixed resistors,  $R_2$  being equal to  $R_3$ . The resistance of  $R_1$  is approximately equal to that of  $T_1$  when the thermistor bead is at its normal control temperature. When the oscillator is first turned on, the thermistor is below its normal temperature and the bridge is unbalanced. This causes a large positive

feedback current to flow, which in turn increases the oscillator output voltage and hence the current through  $T_1$ . The temperature of the thermistor accordingly increases and its resistance decreases until the bridge approaches perfect balance and a stable condition is reached. The resistance and temperature of the thermistor bead then remains constant as long as the oscillator continues in operation. Any tendency of the temperature of the thermistor to change as a result of the absorption of microwave power is immediately counteracted by an offsetting change in the oscillator output.

The temperature controlling oscillator will also hold the temperature of  $T_1$  constant despite changes in ambient temperature to which the thermistor bead is naturally exposed. In doing so, however, the oscillator output is affected by ambient temperature changes, as well as by microwave power input. The oscillator output cannot, therefore, be used as a direct measure of the microwave power input. To overcome this difficulty, a second "compensating thermistor"  $T_2$  is employed. This thermistor is electrically and thermally similar to  $T_1$  and is so mounted as to be subjected to the same external temperature conditions.

The compensating thermistor is connected

across the oscillator output transformer through a capacitor and the COMP potentiometer and is therefore also heated by the oscillator output. Although it is not in the oscillator positive feedback circuit like  $T_1$ , proper adjustment of the COMP potentiometer will cause both thermistors to be held at the same fixed temperature as long as no microwave input power is supplied.

When microwave power is sent into the waveguide system, only thermistor  $T_1$  is affected since  $T_2$  is shielded from the incoming microwaves. To prevent the temperature of  $T_1$  from changing, the oscillator output decreases in proportion to the

incoming microwave power. This, however, decreases the current in  $T_2$  as well as in  $T_1$  and the temperature of  $T_2$  accordingly decreases proportionately. The resultant proportional increase in the resistance of  $T_2$  unbalances the indicating circuit bridge whose fixed resistance arm  $R_6$  is equal to the normal control resistance of  $T_2$  ( $R_4$  and  $R_5$  being equal fixed resistors). The unbalance causes a flow of current in the microammeter which is directly proportional to the change in resistance of  $T_2$  and hence to the microwave power reaching  $T_1$ . The meter is calibrated to indicate the microwave power directly in dbm.

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