TM 11-690

DEPARTMENT OF THE ARMY TECHNICAL MANUAL

BASIC THEORY AND APPLICATION OF TRANSISTORS

U. S. Dept. of the Army.

HEADQUARTERS, DEPARTMENT OF THE ARMY

MARCH 1959

TM 11-690

TECHNICAL MANUAL No. 11-690

HEADQUARTERS,
DEPARTMENT OF THE ARMY
WASHINGTON 25, D.C., *17 March 1959*

BASIC THEORY AND APPLICATION OF TRANSISTORS

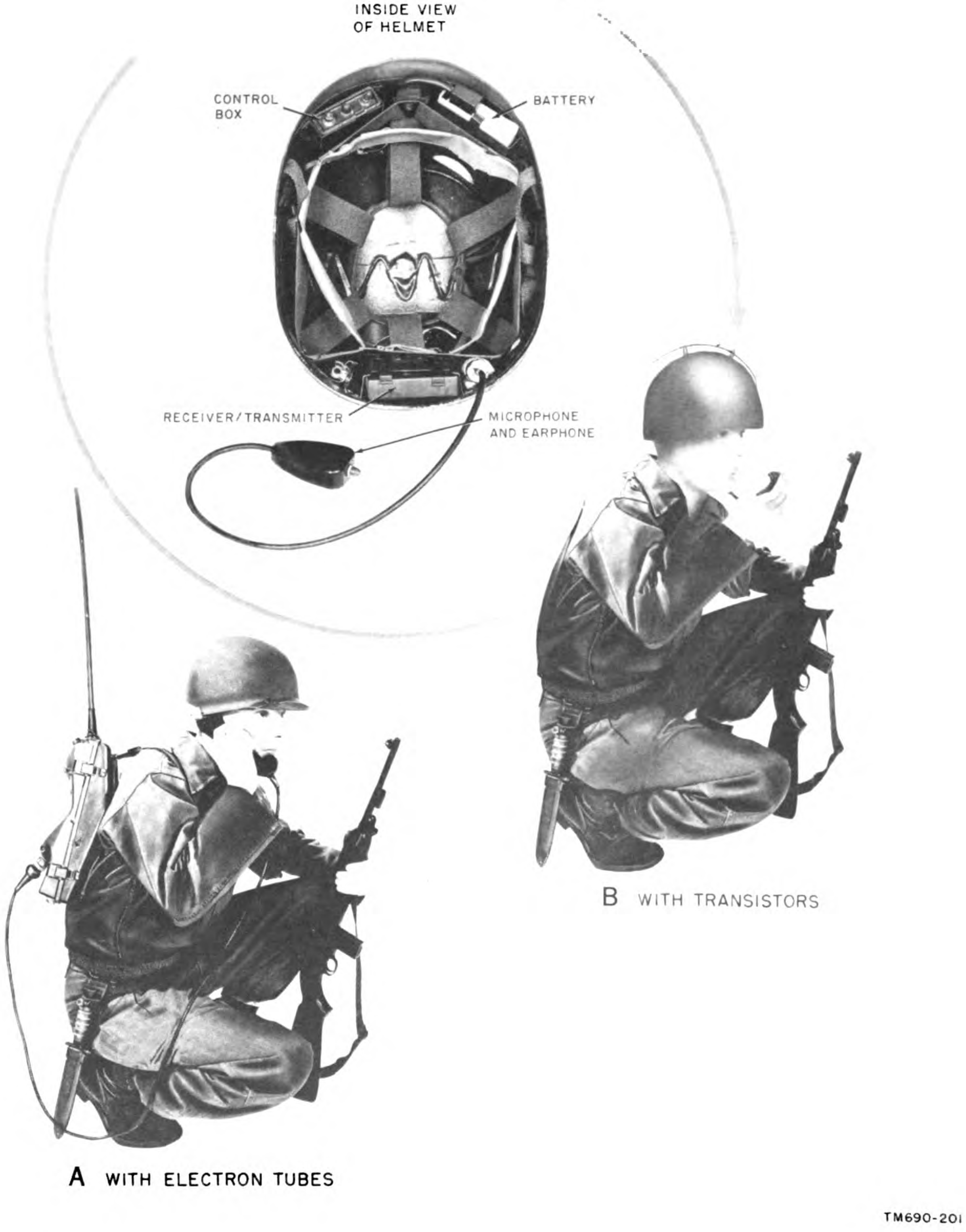


Figure 1. Portable radio transceivers.

CHAPTER 1

INTRODUCTION

Note. This manual is for the electronic technician who has a detailed knowledge of the theory and application of electron tubes.

1. General

Fundamentally, the transistor is a valve which controls the flow of current carriers (electrical charges in motion) through the semiconductor crystal material of which it is made. As the current carriers pass through the transistor, they are controlled as easily as if the same current carriers were passing through an electron tube. The transistor's ability to *control* current carriers and their associated voltages makes it potentially the most useful single element in modern signal communication equipment. In increasing numbers, transistors are being applied in military radio, sound, radar, facsimile, telephone, teletypewriter, and computer equipments.

2. History of Transistors and Semiconductors

a. Crystal Rectifier (fig. 2). The first use of a crystal semiconductor as a rectifier (detector) was in the early days of radio. A crystal was clamped in a small cup or receptacle and a flexible wire (cat-whisker) made light contact with the crystal. Tuning of the receiver was accomplished by operating the adjusting arm until the cat-whisker was positioned on a spot of the crystal that resulted in a sound in the headset. Tuning the variable capacitor provided maximum signal in the headset; frequent adjustment of the contact point was required.

b. Point Contact Diode (fig. 3). Point contact diodes (germanium rectifiers) were used during World War II for radar and other high-frequency applications replacing electron tube diodes. The point contact diode has a very low shunt capacitance and does not require heater power; these properties provide a definite advantage over the electron tube diode. The point contact diode is identical with the crystal rectifier (*a* above). The point contact diode consists of a semiconductor, a metal base, and a metallic

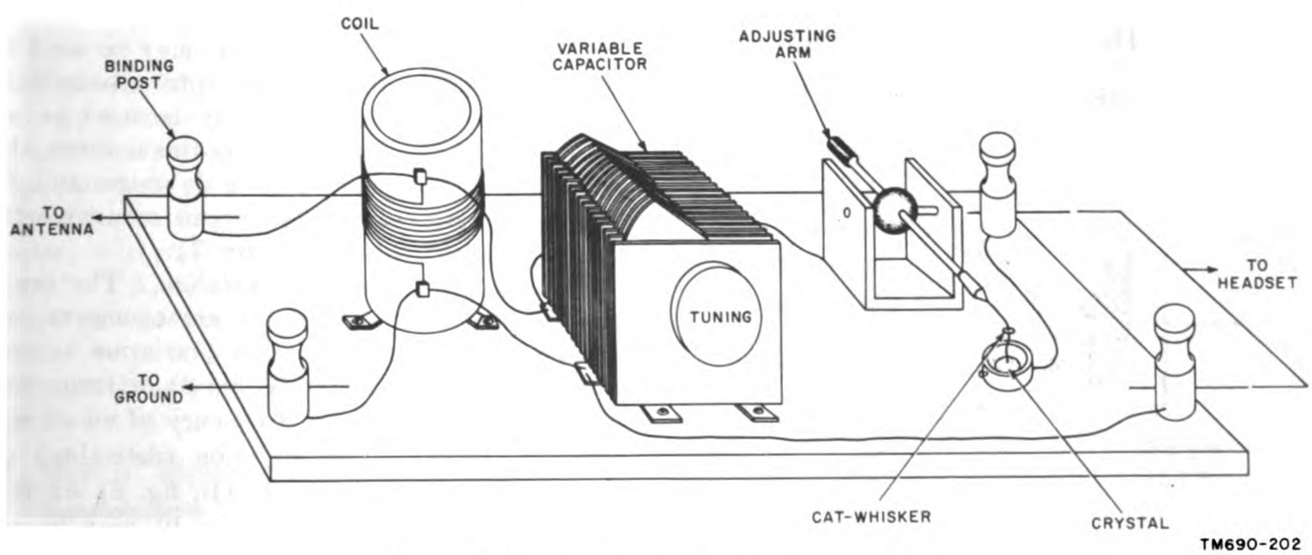


Figure 2. Radio receiver, using crystal rectifier.

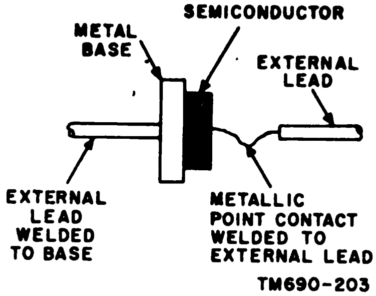


Figure 3. Physical construction of point contact diode.

point contact. The connections to the point contact diode are an external lead welded to the metallic point contact, and an external lead welded to the metal base.

c. Point Contact Transistor (fig. 4). The development of the point contact transistor was announced in 1948. The physical construction of the point contact transistor is similar to that of the point contact diode except that a third lead with a metallic point contact is placed near the other metallic point contact on the semiconductor. One lead is called an *emitter* lead; the other, a *collector* lead (ch. 2). When the two metallic points are property biased with respect to the metal base, the point contact transistor is capable of producing a power gain (ch. 13).

d. Junction Diode (fig. 5). The development of the junction diode was announced in 1949. The junction diode consists of a junction between two dissimilar sections of semiconductor material. One section, because of its characteristics, is called a *P-type semiconductor;* the other, an *N-type* (pars. 19 and 20). The connections to the junction diode consist of a lead to the P-type semiconductor and a lead to the N-type semiconductor. The junction diode handles larger

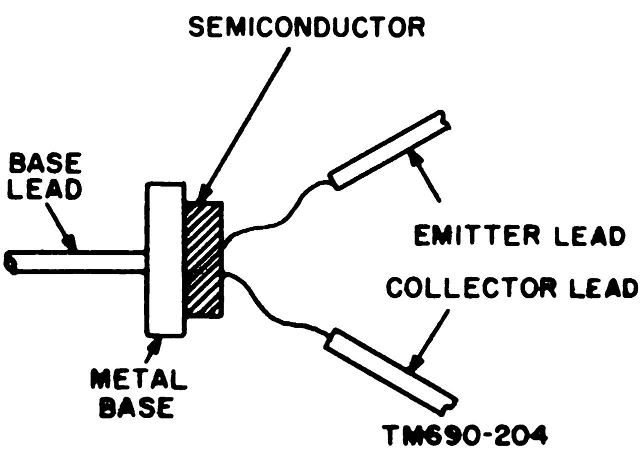


Figure 4. Physical construction of point contact transistor.

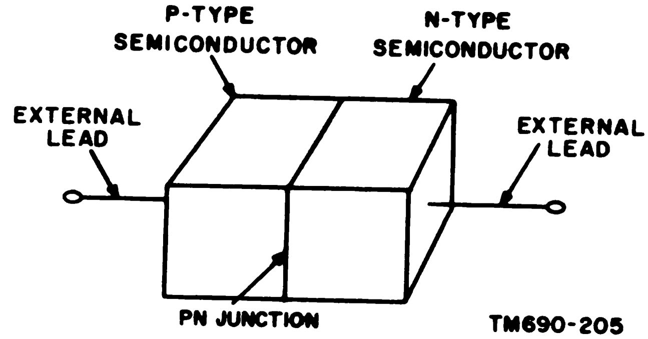


Figure 5. Physical construction of junction diode.

power than the point contact diode but the junction diode has a larger shunt capacitance.

e. Junction Transistor. The development of the junction transistor was announced concurrently with the development of the junction diode. The junction transistor (ch. 2) consists of two PN junctions (A and B, fig. 6). Operation of the junction transistor is similar to that of the point contact transistor. The junction transistor permits more accurate prediction of circuit performance, has a lower signal to noise ratio, and is capable of handling more power than the point contact transistor.

3. Transistor Functions

a. Amplification. The transistor may be used as a current, voltage, or power amplifier. For instance, a stronger signal current may be obtained from a transistor (A, fig. 7) than is fed into it. A signal of 1 milliampere fed into the input circuit of the transistor may appear as 20 milliamperes at its output. Various circuit arrangements provide for various amounts of signal amplification.

b. Oscillation. The transistor may be used to convert direct-current energy into alternating-current energy; that is, it may be used as an oscillator. When functioning in this manner, the transistor draws energy from a dc source and, in conjunction with a suitable circuit arrangement, generates an ac voltage (B, fig. 7).

c. Modulation and Demodulation. The transistor used in various circuit arrangements can provide amplitude modulation (variation in amplitude of an rf signal) (A, fig. 8) or frequency modulation (variation in frequency of an rf signal) (A, fig. 9). Demodulation (detection) of amplitude-modulated signals (B, fig. 8) or frequency-modulated signal (B, fig. 9) may be accomplished with transistors. These circuits are

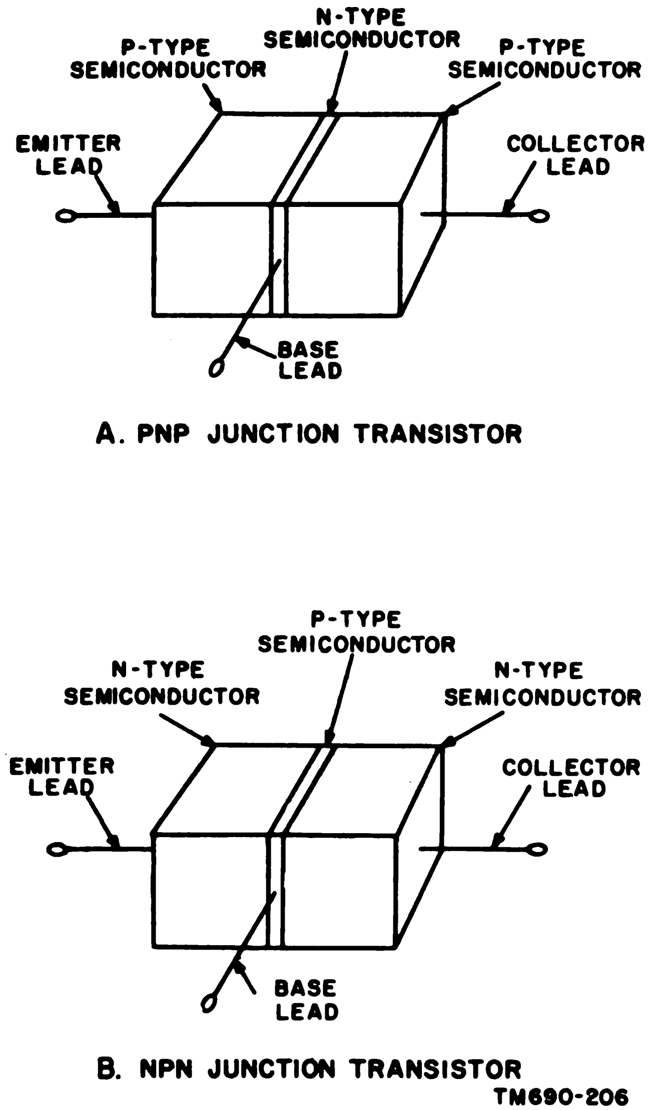


Figure 6. Physical construction of PNP and NPN junction transistors.

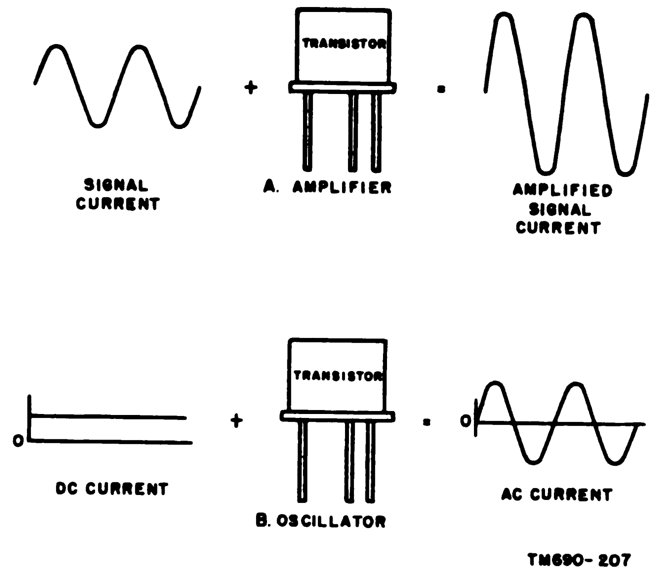


Figure 7. Transistor used as amplifier or oscillator.

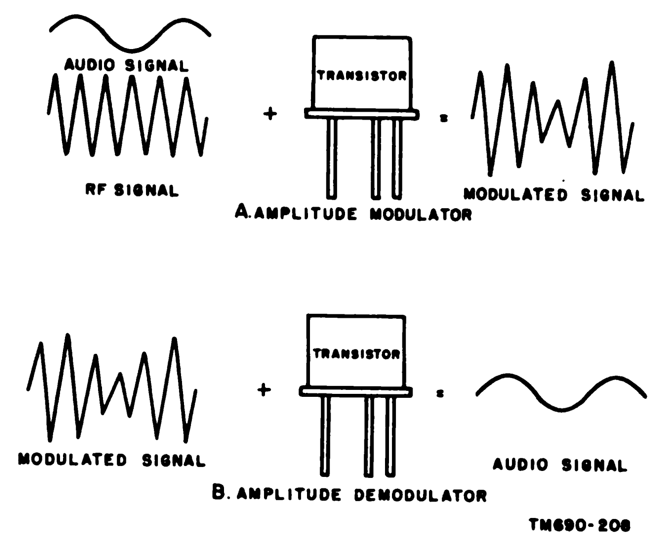


Figure 8. Transistor used as am modulator or am modulator.

well-suited for miniature transmitters intended for short range applications.

d. Miscellaneous. The transistor may also be used to modify the shape of signal wave forms. Wave form shaping is vital in various types of radar, teletypewriter, computer, and television circuits. A, figure 10, indicates the use of the transistor in transforming a sine wave into a square wave. B, figure 10, indicates the use of the transistor in clipping the negative alternations of a series of positive and negative pulses.

4. Use in Military and Commercial Equipments

a. Portable Equipment. The use of transistors opened an entirely new field in the development of portable equipment. Figure 1 illustrates the comparison of a transistorized *walkie-talkie* (transceiver) to a walkie-talkie employing electron tubes. The compactness and ruggedness of transistorized equipment permits its application in a variety of portable equipments. Transistors are now being used in mobile equipment, test equipment, tape recorders, meteorological equipment, and photographic equipment. Various portable equipments in commercial use, such as tape recorders, radios, television sets, hearing aids, automobile radios, and marine radios now employ transistors.

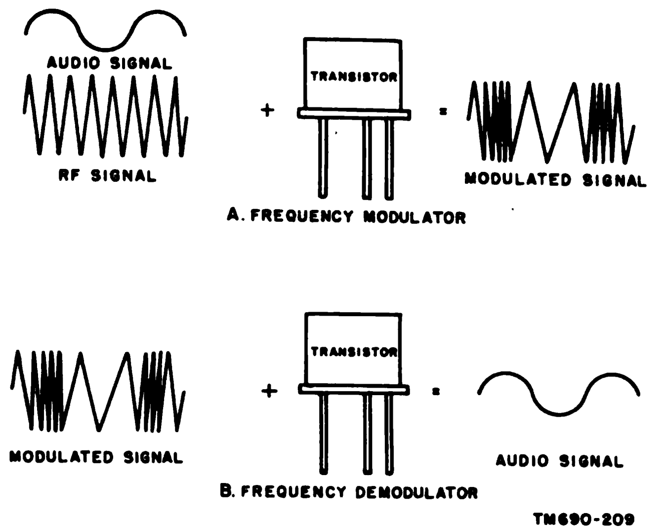


Figure 9. Transistor used as fm modulator or fm demodulator.

b. Fixed Equipments. Fixed equipments that are transistorized take up less space, and use less power. Military and commercial equipment such as wire systems (telephone or teletypewriter), facsimile, radar, computers, public address systems, intercommunications systems, television and radio receivers and transmitters employ transistorized circuitry.

5. Comparison With Electron Tubes

a. Efficiency and Voltage Requirements. The transistor power efficiency is greater than that of the electron tube because the transistor does not require heater power. In addition, it does not require warm-up time, and it does not require a large dc voltage to operate. Other advantages of the transistor are its useful life (*b* below), its noise level (*c* below), and its size and construction (*d* below).

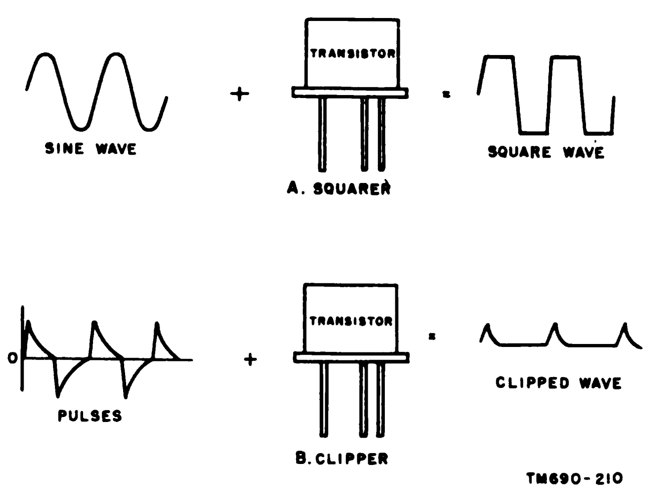


Figure 10. Transistor used to modify wave forms.

b. Useful Life. Life expectancy is a very important consideration in the application of any electronic device. A transistor that is hermetically sealed in glass or metal will withstand a variety of situations that an electron tube cannot withstand. For example, a transistor, even though it is immersed in water, will operate for long periods of time with very little noticeable effect on its operating frequency. It also will withstand centrifugal force, gravity, and impact tests that would completely shatter an electron tube. Although transistors are a comparatively new development, and complete data on their life expectancy are not yet available, it has been estimated that they can operate *continuously* for approximately 8 years, a time much greater than the life of the average electron tube.

c. Noise Level. The noise level of a transistor is approximately 20 db with a frequency input of 1,000 cycles per second. In comparison, the average electron tube has a lower noise level for the same frequency input. Although the noise level of a transistor is higher than that of an electron tube at this particular frequency, the noise level of the transistor is inversely proportional to the audio-frequency input. When a transistor is used with a higher frequency input, the noise level becomes considerably lower.

d. Size and Construction. A power amplifier electron tube is shown in A, figure 11, and a power amplifier transistor is shown in B, figure 11. The construction of the electron tube permits efficient dissipation of heat. Although the transistor must also dissipate heat, the size is noticeably smaller. The flange type construction of the transistor cover provides heat dissipation. In some cases a special metallic heat dissipator must be used. A medium-power electron tube and a medium-power transistor are shown respectively, in C and D, figure 11. Note that the construction of the electron tube is much larger than that of the transistor. A miniature electron tube and a miniature transistor are shown in E and F, figure 11. The construction of the electron tube is again much larger than the transistor. Notice that the power transistor (B, fig. 11) is smaller than the miniature electron tube (E, fig. 11).

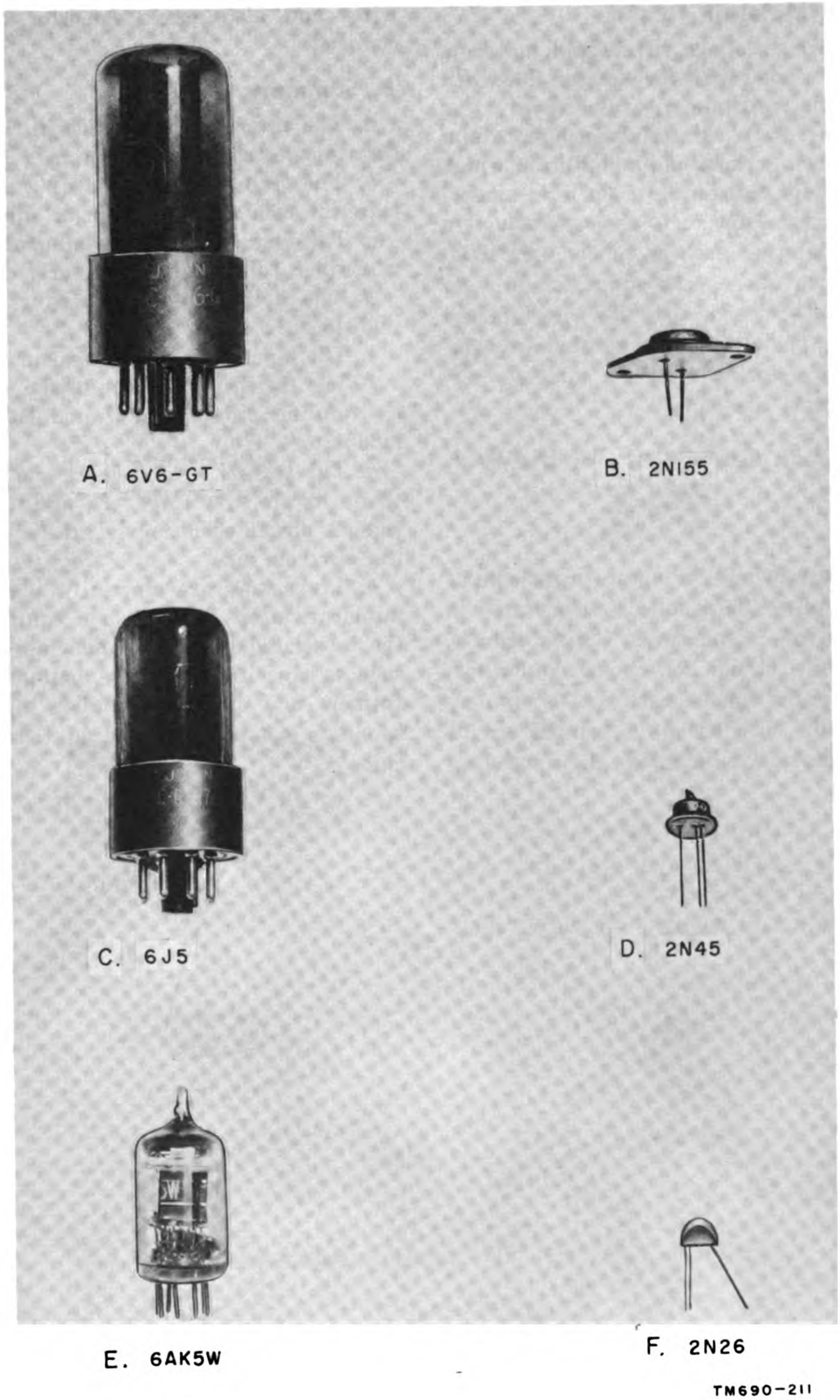


Figure 11. Comparison of transistors and electron tubes.

6. Transistor Material

Materials, such as copper, silver, gold, and iron, which provide a good path for electron flow with little opposition (resistance), are referred to as *conductors* (par. 15*a*). Materials such as carbon in diamond form, germanium, and silicon, which provide a path for electron flow but offer moderate opposition, are referred to as *semiconductors* (par. 15*c*). Materials such as rubber, porcelain, and glass, which offer great opposition and do not provide a path for electron flow, are referred to as *insulators* (par. 15*b*). Transistors are composed of semiconductor material such as germanium or silicon (par. 15*c*).

7. Special Circuit Components for Use With Transistors

(fig. 12)

The lower current and voltage requirements of transistors simplify the problems encountered in miniaturization of components. Capacitors (B, fig. 12), resistors (C, fig. 12), chokes, and transformers (A, fig. 12) for use with transistors may be miniaturized. The use of a printed circuit board (not shown) eliminates all connecting wires and also make transistor circuits more compact.

8. Summary

a. The transistor is a *valve* in the same sense that the electron tube is a valve.

b. The transistor is made of a semiconductor material such as germanium or silicon.

c. The first crystal semiconductor was used as a rectifier (detector) in the early days of radio.

d. The point contact transistor was invented in 1948.

e. The junction transistor was invented in 1949.

f. The transistor may be used in circuits such as amplifiers, oscillators, modulators, and demodulators.

g. The transistor is smaller and more rugged than the electron tube. In addition, the transistor's power efficiency is greater than that of the electron tube.

h. Transistors are especially adaptable to the miniaturization of electronic equipment. Even the associated circuit components of transistors such as capacitors, resistors, batteries, and transformers may be made smaller than the corresponding components used in electron-tube circuits.

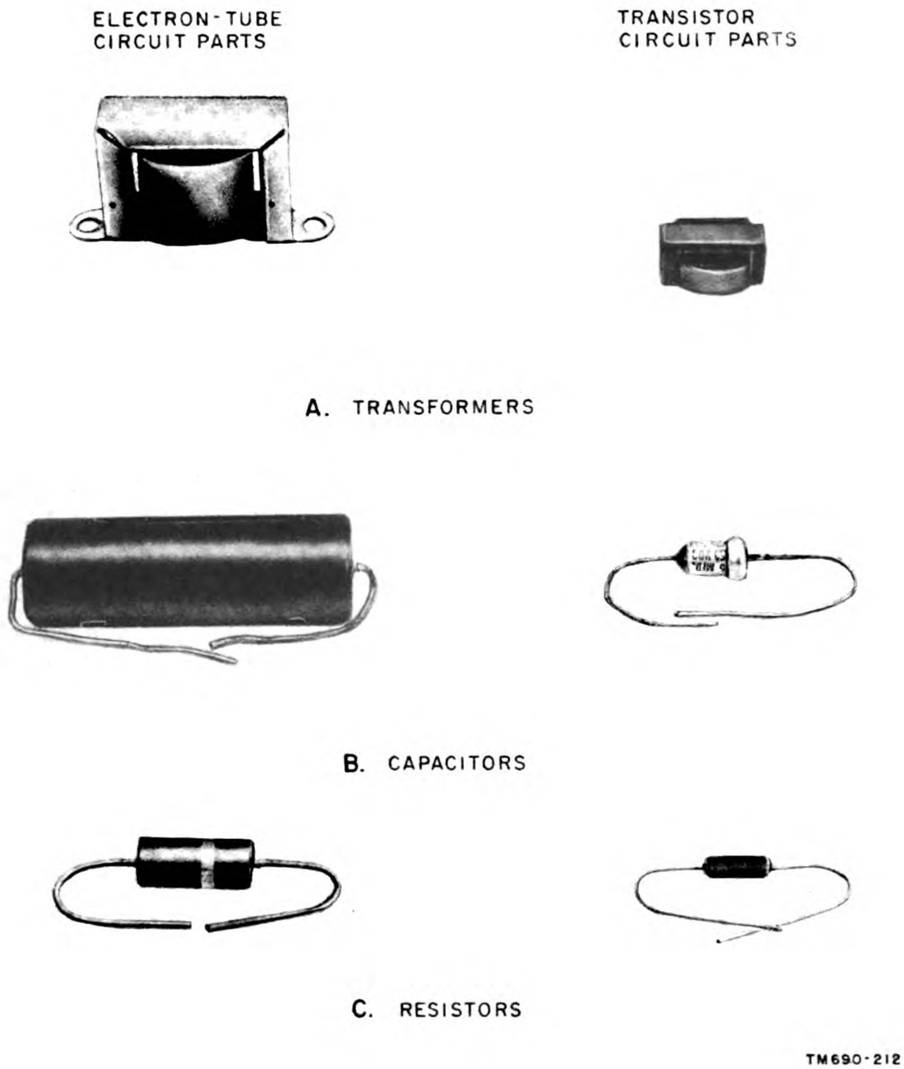


Figure 12. Circuit components used with transistors and electron tubes.

CHAPTER 2

FUNDAMENTAL THEORY OF TRANSISTORS

Section I. STRUCTURE OF MATTER

9. General

a. A knowledge of the theory of the structure of matter is required (pars. 10 through 14) to understand the theory of the internal conduction mechanism that occurs in the transistor.

b. Transistors are constructed from semiconductor materials. The properties and characteristics of conductors and insulators are covered in TM 11–661. A comparison of the properties of conductors, semiconductors, and insulators is given in paragraph 15. Detailed properties and characteristics of semiconductor materials used in transistors are covered in paragraphs 16 through 20.

10. Matter, General

Matter may be defined as any substance that has weight (mass) and occupies space. Examples of matter are—air, water, clothing, and one's body. These examples demonstrate that matter may be found in any one of three states: gaseous, liquid, or solid. Matter is found in nature as elements (*a* below), or compounds (*b* below). The elements and compounds are made up of molecules (*c* below), atoms (*d* below), and subatomic particles (*e* below).

a. Element. Matter consists of one or more basic materials which are called elements. Scientists have definite proof that approximately 102 elements exist and believe that there are several additional elements. In chemistry, an element is defined as a substance that can be neither decomposed (broken up into a number of substances) by ordinary chemical changes nor made by chemical union of a number of substances. Copper, iron, aluminum, and gold are examples of metallic elements; oxygen, hydrogen, and sulphur are nonmetallic elements.

b. Compound. A substance containing more than one constituent element and having properties different from those of its elemental constituents is called a *compound.* For example, water is made up of two parts hydrogen and one part oxygen. Therefore, water is a compound.

c. Molecule. A *molecule* is defined as the smallest particle of matter which can exist by itself and still retain all the properties of the original substance. If a drop of water, a compound, is divided until the smallest possible particle is obtained and is still water, that particle is known as a molecule. An idea of the size of a molecule may be obtained by imagining that a stone is first broken into two pieces, that the two pieces are then broken into four pieces, and that this process is carried on. The smallest particle of stone which could be obtained by this process would be a molecule. Actually, it is impossible to crush a stone into its molecules; we can only crush it into dust. One small particle of dust is composed of thousands of molecules.

d. Atom. An atom is defined as the smallest part of an element that can take part in ordinary chemical changes. The atoms of a particular element are of the same average mass, but their average mass differs from that of the atoms of all other elements. For simplicity, the atom may be considered to be the smallest particle that

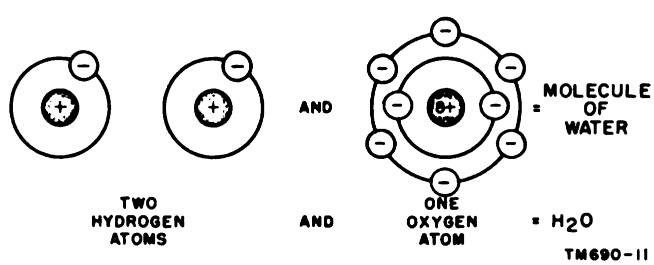


Figure 13. Molecule of water.

retains its identity as part of the element from which it is divided. Figure 13 shows that the molecule of water is made up of two atoms of hydrogen and one of oxygen. Since there are approximately 102 known elements, there must be 102 different atoms or a *different* atom for each element. All substances are made of one or more of these atoms. Just as thousands of words can be made by combining the proper letters of the alphabet, so thousands of different materials can be made by chemically combining the proper atoms.

e. Subatomic Particles. Although it was formerly believed that the atom was the smallest particle of matter, it is known now that the atom itself can be subdivided into still smaller, or subatomic particles. Paragraph 11 explains the nature of these subatomic particles.

11. Structure of Atom

Figure 14 illustrates the relationship of the atom to matter, and the atom to its subatomic particles. The material illustrated in all parts of the figure, except A and B, is hypothetical. The portions of matter which they represent are so small that they are invisible to the eye, even with the aid of the most powerful microscope. By experiment and study scientists have been able to learn about subatomic particles and, from developed facts, have been able to produce an understandable explanation of the make-up of the atom. Thus, diagrams C, D, E, and F, figure 14 can be called representations of what scientists believe the element aluminum might look like if tiny bits of aluminum could actually be seen. If an *imaginary microscope* were available which would allow the examination of tiny particles of matter, observations made through this microscope would probably indicate that the representations of figure 14 are almost exact if the matter under scrutiny is a piece of aluminum segment.

a. In A, figure 14, the magnification is unity, that is, one is simply looking at a piece of aluminum.

b. In B, figure 14, the magnification has been increased to 100 diameters. It becomes apparent that aluminum does not have a perfectly smooth surface but rather is a crystalline substance, with the size of the crystals being dependent on the previous heat and mechanical treatment of the metal. Note that the crystals are extremely small, nonuniform in shape, and irregularly arranged. The crystals present a specific granular structure, known as a *polycrystalline structure.* Conductors such as copper and aluminum reveal a polycrystalline structure when viewed under a microscope. The properties and characteristics of polycrystalline materials are quite different from the properties and characteristics of *single* crystal materials. Germanium and silicon when processed for use in transistors are single crystal materials (par. 17*b*).

c. When the imaginary microscope is adjusted to a magnification of 100,000 diameters (C, fig. 14), evidence of the presence of individual atoms or subatomic particles can be detected.

d. In D, figure 14, magnification is 10 million diameters. At this magnification, individual dots, or little bodies, of approximately spherical shape and a rather fuzzy outline would appear. These dots are aluminum atoms and no detectable difference is apparent between them. It is possible to measure the diameter of any aluminum atom but because the edges of the atoms are fuzzy, it is difficult to say just where one atom ends and the next atom begins.

e. A magnification of 100 million diameters is required to eliminate this fuzziness. A single aluminum atom now fills the entire area (E, fig. 14). This single atom of aluminum resembles the solar system to some degree, for there is a central body, called a *nucleus*, about which a number of smaller particles (electrons) move in approximately elliptical orbits. To determine the number of electrons in an atom is exceedingly difficult because of the way electrons move in their elliptical orbits. However, it has been found that there are 13 electrons in the aluminum atom. Careful observation disclosed that each of the electrons in this atom has a charge of electricity that is identical with the charge on any of the other electrons. The charge associated with an electron is the smallest electrical charge which has yet been discovered. It is called the *elemental charge.* (The charge on the electron was first measured by R. A. Millikan, an American physicist and it was arbitrarily called a *negative charge.*) Thus, the aluminum atom is said to contain 13 elemental charges of negative electricity. The nucleus also has a positive electrical charge, and it has been determined that the quantity of this positive charge is 13 times as great as the quantity of the negative charge on one electron. Therefore, since there are 13 elemental

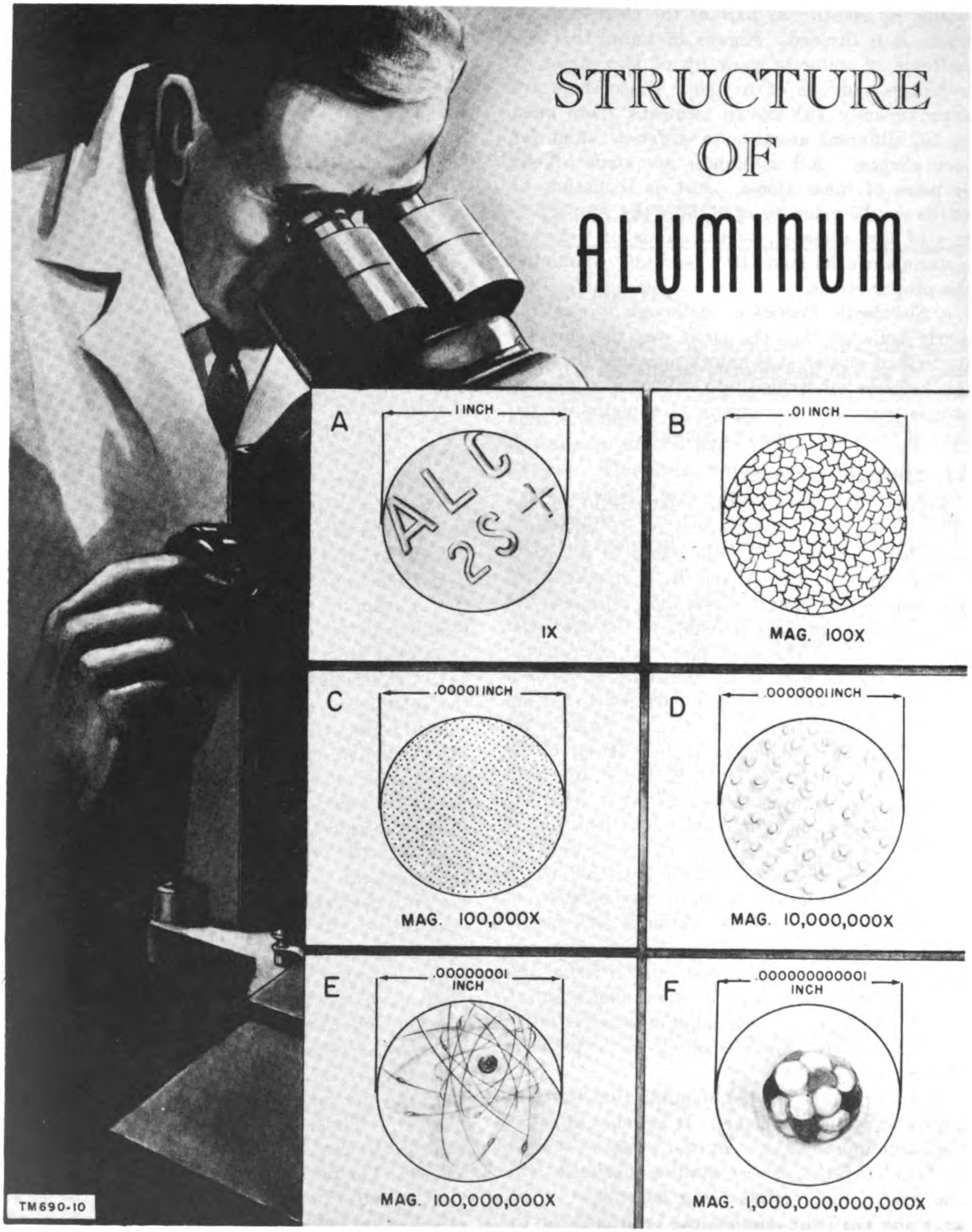


Figure 14. Structure of aluminum.

positive charges on the nucleus, the entire atom is found to be electrically neutral. That is, *the aluminum atom contains an equal number of positive and negative charges—an equal amount of opposite kinds of electricity.*

f. By increasing the magnification until only the nucleus fills the viewing area it is observed that the enlarged nucleus looks like a bunch of grapes (F, fig. 14) consisting of 27 particles, 13 of which carry an elemental positive charge of electricity. These elemental positive charges are called *protons*. The other 14 particles in the nucleus are uncharged and are called *neutrons*.

g. Figure 15 illustrates the situation when the the aluminum is magnified so that only a few atoms of aluminum cover the entire viewing area. The electrons which revolve in the outer, or third shell (near the edges of the atoms), do not always remain confined to the same atom and some, moving in a random manner, may travel from atom to atom. (The shells are also referred to as *orbits*, or *rings*.) Outer orbit electrons that move at random through the material are called *free electrons* (par. 14).

12. Electrons, Protons, and Neutrons

In paragraph 11, the aluminum atom was shown to consist of a positively charged nucleus with negatively charged electrons that revolved at a very great speed around the nucleus. The electron theory states that the atoms of *all* elements (copper, gold, oxygen, etc.) are similarly constructed of a central nucleus and revolving electrons.

a. Examples of Atomic Structure.

(1) A, figure 16, represents the atomic structure of the simplest of all atoms, the hydrogen atom. It contains one electron revolving around one proton which acts as a nucleus. Because the negative charge on the electron is exactly equal to the positive charge on the proton, the atom is electrically balanced or neutral.

(2) B, figure 16, represents the aluminum atom. The nucleus of the aluminum atom contains 14 neutrons and 13 protons. The positive charges of the 13 protons are just balanced by the negative charges of the 13 revolving electrons and the electrical charge on the entire atom is again neutral. Note that in this arrangement, the outermost shell has three electrons. The importance of this arrangement is explained in paragraph 20.

(3) The phosphorous atom (C, fig. 16) is a more complex structure. Note that the 15 orbital electrons revolve in three separate rings, or *shells*, and in this atom the outermost ring has five electrons. The importance of this arrangement is explained in paragraph 19.

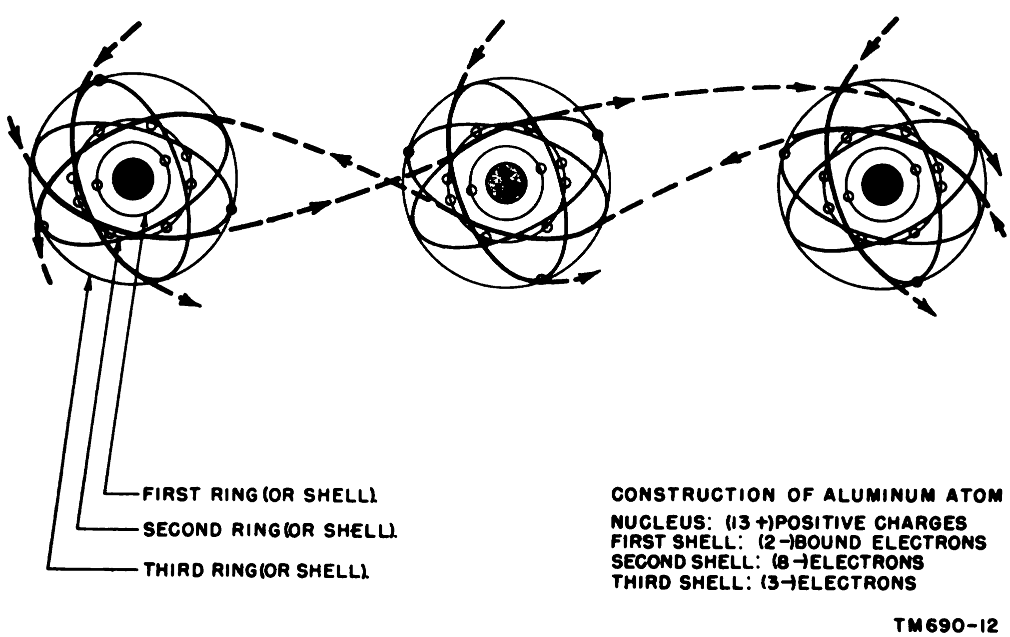


Figure 15. Three atoms of aluminum.

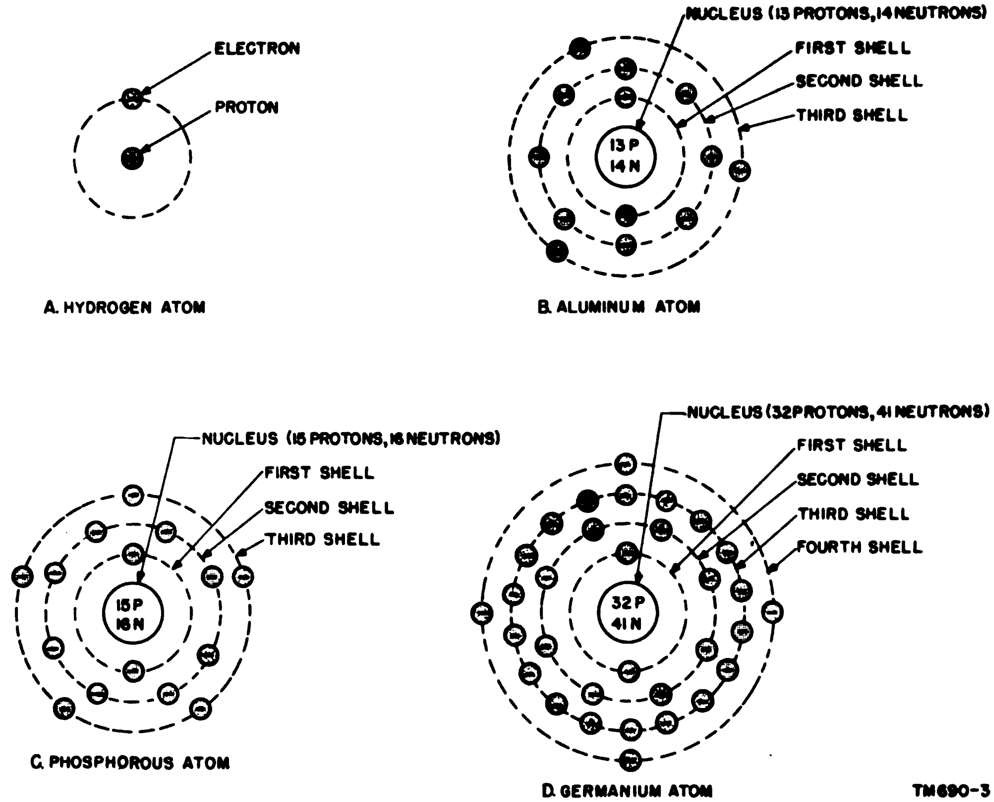


Figure 16. Structure of atoms.

(4) In D, figure 16, an even more complex atom is shown, the germanium atom. The nucleus is composed of 32 protons and 41 neutrons. The 32 orbital electrons revolve in four separate shells, with four electrons revolving in the outer shell. Paragraph 17 explains the importance of this arrangement.

b. Building Blocks. The electron theory states that the difference among the various elements is in the *number* and *arrangement* of the electrons, protons, and neutrons of which each atom is composed. There is no difference between an electron in an atom of copper and an electron in an atom of aluminum, or any other element. There is no difference between a proton in one atom and a proton in another atom of a different element. Likewise, the neutrons in the atoms of various elements are thought to be identical. Since all matter is composed of atoms and all atoms are composed of positively charged particles called protons, negatively charged particles called electrons, and uncharged particles called neutrons, it follows that the proton, the electron, and the neutron are the fundamental building blocks of the universe.

c. Characteristics of Subatomic Particles.

(1) *Electrical.* The electrical charge of the proton is exactly equal and opposite to that of the electron; that is, the proton and the electron are exactly equal amounts of opposite kinds of electricity. Because it is believed that no smaller amount of electricity exists, the charge on the electron and the proton is considered the elemental unit of electrical charge. However, the elemental unit is too small a quantity of electricity for practical purposes and a larger unit of charge called the coulomb is commonly used. One coulomb of electricity contains over 6 million, million, million (6.28×10^{18}) electrons.

(2) *Physical.* Electrons and protons are approximately spherical particles of matter. The diameter of an electron, approximately 0.00000000000022 inch (22×10^{-14}), is about three times the diameter of a proton. Despite its smaller diameter, a proton's mass is 1,850 times greater than the mass of the electron. The diameter and mass of a proton and a neutron are

approximately the same. Relatively speaking, there are great distances between the electrons and the protons of an atom even in solid matter. It has been estimated that if a copper one-cent piece could be enlarged to the size of the earth's path around the sun (approximately $186{,}000{,}000 \times 3.14$ miles), the electrons would be the size of baseballs and would be about 3 miles apart.

13. Electrical Balance and Unbalance

a. Electrical Balance. Because the protons and electrons of the atom carry positive and negative charges of electricity, respectively, they are particles of energy. These changes form the electric field of force within the atom, and since the number of positive charges (protons) is always equal to the number of negative charges (electrons) an atom is electrically balanced.

b. Electrical Unbalance. It is possible to transfer some of the electrons from one substance to another. When this happens, the normally equal distribution of positive and negative charges in each substance no longer exists, and because each contains more of one kind of electricity than of the other they are said to be electrically charged. For example, when a glass rod is rubbed with silk, some of the electrons which are loosely held to the atoms in the glass are transferred to the silk. Then when the glass and silk are separated, the glass rod (B, fig. 26) will have more positive charges (protons) than negative charges (electrons)and will be *positively* charged. The silk, on the other hand, will have more electrons than protons and will be *negatively* charged. When a hard rubber rod is rubbed with cat's fur, the cat's fur loses electrons to the rubber rod (B, fig. 25). In this case, the cat's fur becomes positively charged and the rubber rod becomes negatively charged.

14. Free Electrons

a. General. The electron theory states that all matter is composed of atoms which, in turn, are composed of subatomic particles called protons, electrons, and neutrons. In conductors, all the electrons are tightly bound to the nucleus except those that revolve in the outer orbit. Figure 15 shows how the electrons in the outer orbits of the aluminum atoms may move from one atom to another in a haphazard manner. Electrons that are able to move in this fashion are known as *free electrons.* All matter is composed of positive and negative charges of electricity, and the atomic structure of a material will determine whether or not the material will have many or few free electrons.

b. Electron Flow or Current. If the haphazard movement of free electrons in a material is controlled so that the electrons move generally in the same direction an electron flow (or drift) results. This electron flow is called an electric current.

15. Conductors, Semiconductors, and Insulators

(fig. 17)

In general, all materials may be placed in one of three major categories—conductors, semiconductors, or insulators. These categories were evolved from a consideration of their ability to allow an electric current to flow. This, in turn, depends on their atomic structure.

a. Conductors. A good conductor is a material that has a *large number of free electrons.* All metals are conductors of electricity to some extent, but some are much better conductors than others. Silver, copper, and aluminum, are conductors which have resistances of less than 3 millionths of an ohm per centimeter cube (cm^3). That is a cube each edge of which measures one centimeter. Silver is a better conductor than copper, but copper is more widely used because it is less expensive. Aluminum is used as a conductor where weight is a major consideration, for example, on high-tension lines with long spans between supports. Usually these lines are stranded cables and have a small steel wire core to provide the necessary tensile strength. The ability of a material to conduct electricity also depends on its dimensions. Conductors may be in the form of bars, tubes, or sheets; but the most common conductors are in the form of wire. Many sizes of wire are used, from the fine hair-like wire used in the coils of sensitive measuring instruments, to the large bus-bar sizes used for carrying high current in electric power-generating plants. In order to make wire easier to handle and also less subject to changes in weather and other external conditions, it is often covered with some other material such as rubber, cotton, plastic, or enamel. These coverings provide protection against short circuits and leakage and are known as insulators.

495909 O—59——2

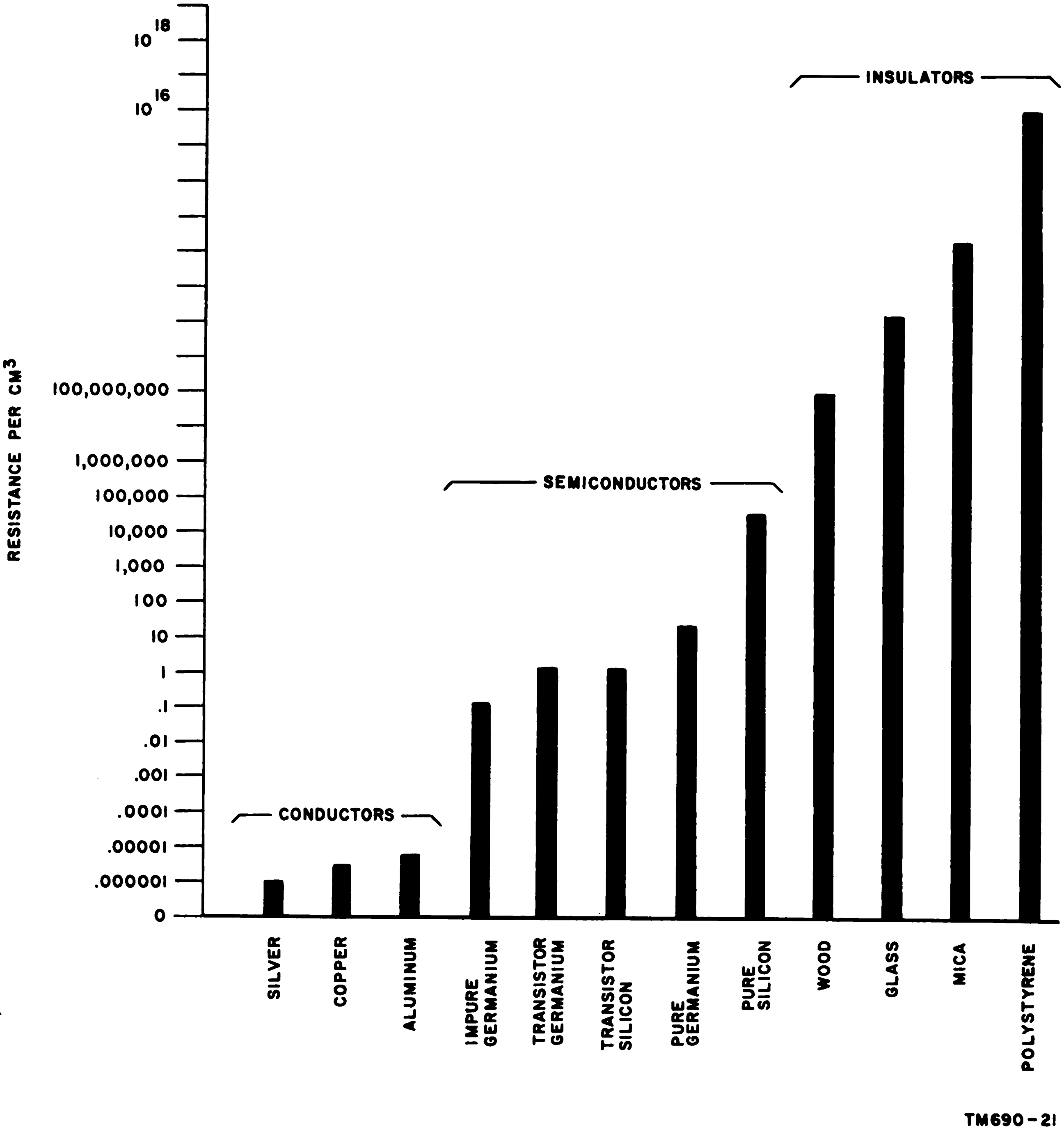


Figure 17. Chart of resistance per cm³ of conductors, semiconductors, and insulators.

b. Insulators. An insulator (also known as a dielectric) is a material, or combination of materials, the atomic structure of which is such as to limit almost to nothing the movement of electrons from atom to atom. In other words, *an insulator is a material that has few loosely held electrons.* No material known is a perfect insulator but there are materials which are such poor conductors that for all practical purposes they are classed as insulators. Glass, dry wood, rubber, mica, and certain plastics such as polystyrene are insulating materials which have a resistance of several millions of ohms per centimeter cube.

c. Semiconductors. Between the extremes of good conductors and good insulators are a number of materials which are neither good conductors nor good insulators. Germanium and silicon fall into this category and are called semiconductors. Pure germanium is not a very good conductor (about 60 ohms per centimeter cube at room temperature). Pure silicon is a somewhat worse conductor (about 60,000 ohms per centimeter cube at room temperature). Both impure germanium and impure silicon have a resistance of 0.2 through 0.4 ohm per centimeter cube, depending on the type and amount of impurities present. Germanium and silicon, for use in transistors, have carefully controlled amounts of impurities added (par. 18), and each has a resistance of 2 ohms per centimeter cube at room temperature. This resistance decreases rapidly as the temperature rises. Note that while only germanium and silicon as semiconductors are discussed there are many other semiconductors.

Section II. CRYSTALS, DONORS, ACCEPTORS, AND HOLES

16. Crystals, General

a. General. Most solids (*c* below), except those exhibiting a biological structure of cells such as leaves, branches, and bone, reveal a crystal structure when studied under a microscope. Many substances such as rocks and metals which are not usually considered crystalline, reveal a specific crystal pattern when studied under a microscope.

b. External Characteristics. The most commonly known characteristics of crystals are their angles and their planes. Snow crystals, for instance, although formed in an infinite number of geometric patterns, contain only 60° angles. Some materials, such as common salt (sodium chloride), form cubes; other materials form long needles, rhomboids, or variations of hexagonal or rectangular structures. Each material has a characteristic form.

c. Internal Structure. X-rays have been used to investigate the internal structure of crystals. The wavelength of X-rays approximate the distance between the atoms or molecules of crystals. When X-rays are beamed through a crystal, the rays are deflected and distributed in accordance with the specific arrangement of the atoms or molecules of the crystal. When the resultant ray is photographed, the photograph invariably shows a specific pattern depending upon the substance of the crystal. With the pattern indicated on the photograph, and through a complex mathematical analysis, scientists have been able to construct models of the internal structure of a given crystal. These analyses have indicated that the atoms of crystals are arranged in specific patterns; that one atom is not closely related with another atom only, but rather is related equally to a number of adjacent, equidistant atoms. The specific arrangement of atoms depends on the size and number of atoms present and on the electrical forces between them. The physical, electrical, optical, and mechanical characteristics of the crystal depend on the forces between the atoms. Crystals used in transistors are discussed in paragraph 17.

17. Pure Germanium Crystal

a. General. Figure 18 shows a pure germanium crystal. Each sphere represents a germanium atom less the four electrons (valence or outer orbit electrons) that are in the outer (fourth) shell of the atom (fig. 16). The sphere contains the nucleus of the atom, and the 28 tightly bound electrons that orbit around the nucleus. The nucleus contains 32 protons; that leaves a net positive charge of four (4) on the sphere. Throughout this text, the sphere will be referred to as the *germanium core.*

b. Single Crystal Structure. The dashed lines in the illustration form two cubes. Note that the four germanium cores between the two cubes are shared equally by the cubes. If the illustration were to be extended in all directions, the sharing of the corner germanium cores would be extended to all adjacent cubes. This repeated uniform, *cubical structure* constitutes a *single germanium crystal.* The properties and characteristics of *single crystal* materials such as germanium and silicon (as prepared for use in transistors) are quite different from the properties and characteristics of *polycrystalline materials* such as copper and aluminum (par. 11*b*). The term crystal used throughout this text will refer to *single crystal material* only.

c. Lattice Structure.

(1) It has been established that electrons rotate constantly in relatively fixed orbits about the nucleus. In a crystal, the rotation of one valence electron of a given atom is coordinated with the

rotation of one valence electron of an adjacent atom. The coordinated rotation of two valence electrons (one from each of two adjacent atoms) results in the formation of an *electron-pair bond*. The electron-pair bonds, shown diagrammatically in figure 18, are also referred to as *valence bonds*. The electron-pair bonds cause the cores to be attracted toward each other. The positive charges on the cores cause the cores to repel each other. When a balance of the forces of attraction and repulsion is obtained, the crystal is said to be in a state of equilibrium.

(2) Each germanium core is equidistant from four adjacent germanium cores. Note that each germanium core is interconnected with adjacent germanium cores by four electron-pair bonds. This condition exists since each germanium atom contains four valence electrons in its outer shell. This arrangement of germanium cores and electron-pair bonds is referred to as a *lattice*. To avoid congestion, figure 18 shows only a few of the electron-pair bonds.

d. Conductivity. The valence electrons of good conductors such as copper or aluminum, are loosely bound to the nucleus of the atom, and move quite readily through the conductor under the influence of a potential field. Valence electrons which form part of an electron-pair bond, however, are bound in the electron-pair bond and are not free to take part in conduction. Crystalline materials such as carbon, germanium, and silicon, the valence electrons of which are bound, are poor conductors (par. 15*c*) under normal conditions. Only if the material is subjected to high temperatures or strong radiation will the electron-pair bonds separate and partial electrical conduction occur.

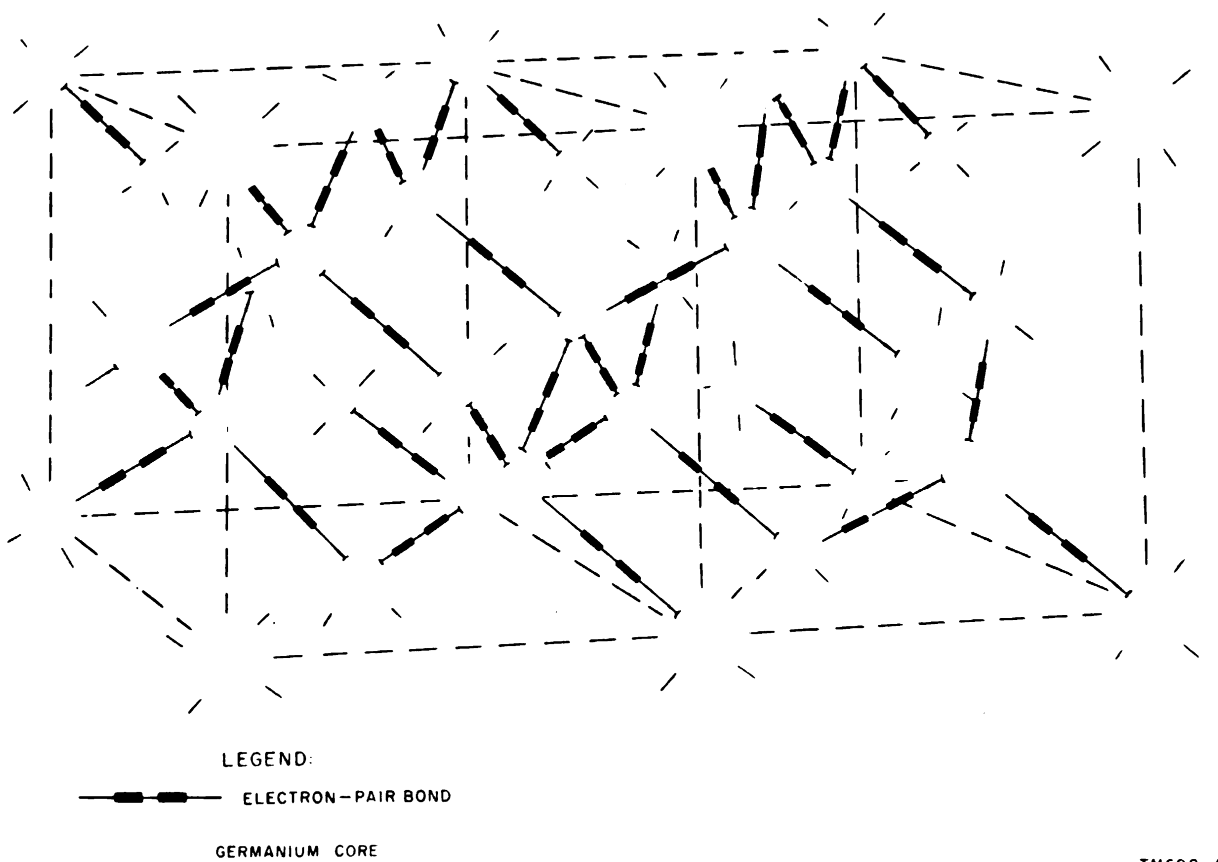


Figure 18. Pure germanium crystal, lattice structure.

18. Impurities

a. General. It is possible for the atoms of substances other than germanium to join the crystal lattice structure of germanium. These substances whether found in germanium in its natural state, or added intentionally during the processing of germanium for use in transistors, are referred to as *impurities*.

b. Donor and Acceptor Impurities. Two groups of substances exhibit the important characteristic of joining the lattice structure of germanium. The substances in one group are called *donors;* in the second group, they are called *acceptors.*

(1) The atoms of presently used substances classified as donors have five valence electrons in the outer incomplete shell. Some of the substances that have been used as donors are arsenic, phosphorous (C, fig. 16), antimony, and boron. The characteristics and properties of germanium containing donor atoms are discussed in paragraph 19.

(2) The atoms of presently used substances classified as acceptors have three valence electrons in the outer incomplete shell. Some of the substances that have been used as acceptors are aluminum (B, fig. 16), gallium, and indium. The characteristics and properties of germanium containing acceptor atoms are covered in paragraph 20.

19. N-Type Germanium, Donors

a. Figure 19 shows a germanium crystal in which one of the germanium atoms has been replaced by a donor impurity (par. 18). The dark sphere in the illustration represents the nucleus of the donor atom and all the tightly bound electrons that orbit around the nucleus. The valence electrons are not included in the sphere. The donor impurity contains five valence electrons. Note that four of the valence electrons of the donor form electron-pair bonds with electrons of four neighboring atoms of germanium.

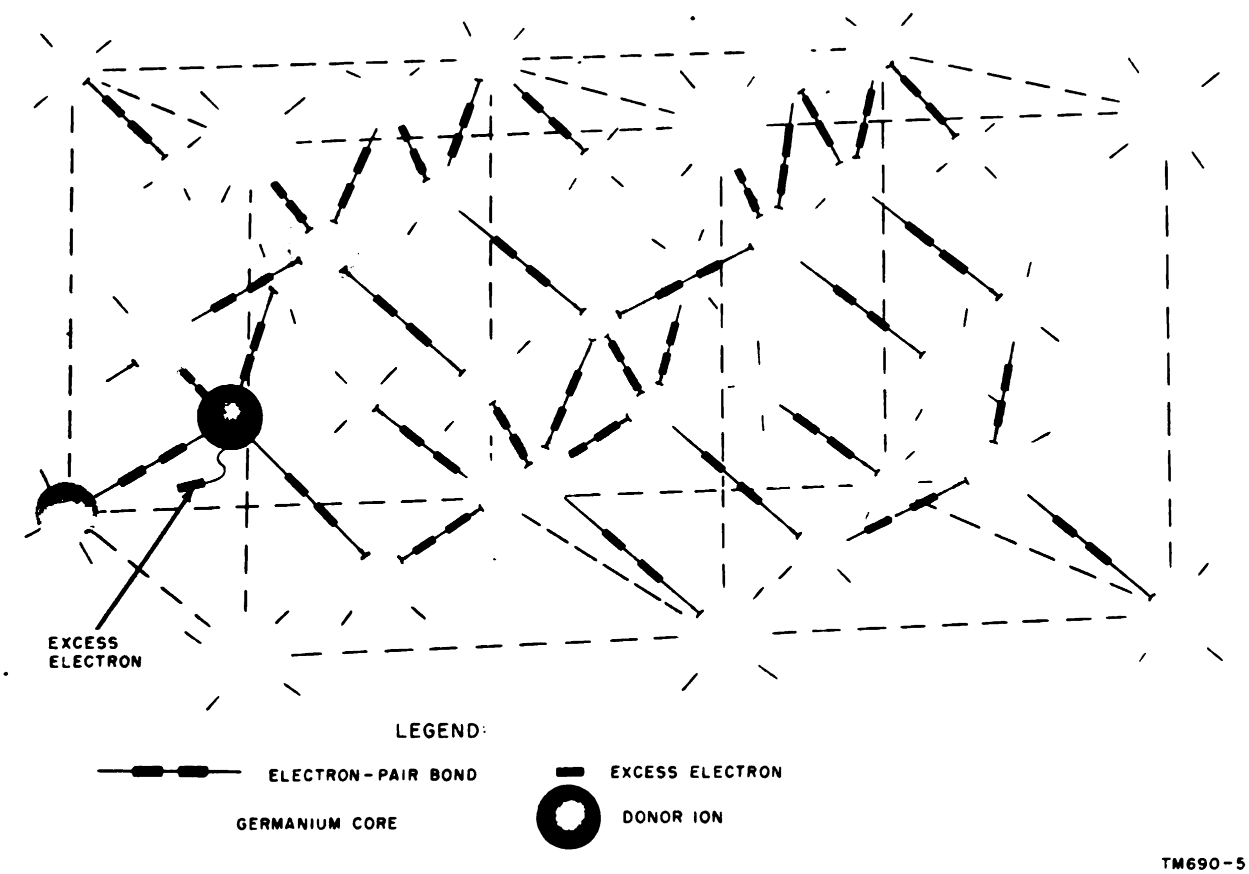


Figure 19. Germanium crystal with donor atom.

The electrons of both the germanium and the donor atoms that enter into electron-pair bonds form a very stable structure, and are not removed readily from the bonds.

b. The fifth valence electron of the donor cannot form an electron-pair bond since there are no adjacent electrons available. This electron becomes an excess electron. The donor nucleus has a very weak influence over the excess electron. Actually, one one-seventieth of the energy required to remove an electron from an electron-pair bond is required to remove the excess electron from the donor. At normal room temperature (70° F.), enough thermal (heat) energy is present to cause the excess electron to break away from the donor and wander through the space between the crystal lattices (fig. 20).

c. When the excess electron leaves (ionizes from, or is *donated* by), the donor atom, the donor atom then possesses a positive charge equivalent to the negative charge of one electron. An atom that loses or gains an electron is called an *ion.* For that reason, the spheres that represent the donor (fig. 19) are called donor ions.

d. Note that the germanium crystal that contains a donor ion (positive) also contains an excess electron (negative). The germanium crystal taken as a whole therefore is electrically neutral; that is, the crystal posseses a net charge of zero.

e. Germanium containing donor impurities is referred to as *N-type germanium.* The letter *N* refers to the *negative* charge of the excess electron.

20. P-Type Germanium, Acceptors

a. Figure 21 shows a germanium crystal in which one of the germanium atoms has been replaced by an acceptor impurity (par. 18). The cross-hatched sphere contains the nucleus of the acceptor atom and all the tightly bound electrons that orbit around the nucleus; the valence electrons are not included in the sphere. The acceptor impurity contains three valance electrons. Note that the three valence electrons of the acceptor form electron-pair bonds with electrons of the neighboring atoms of germanium.

b. One valence electron of the fourth neighboring germanium atom cannot form an electron-pair bond since the acceptor has only three

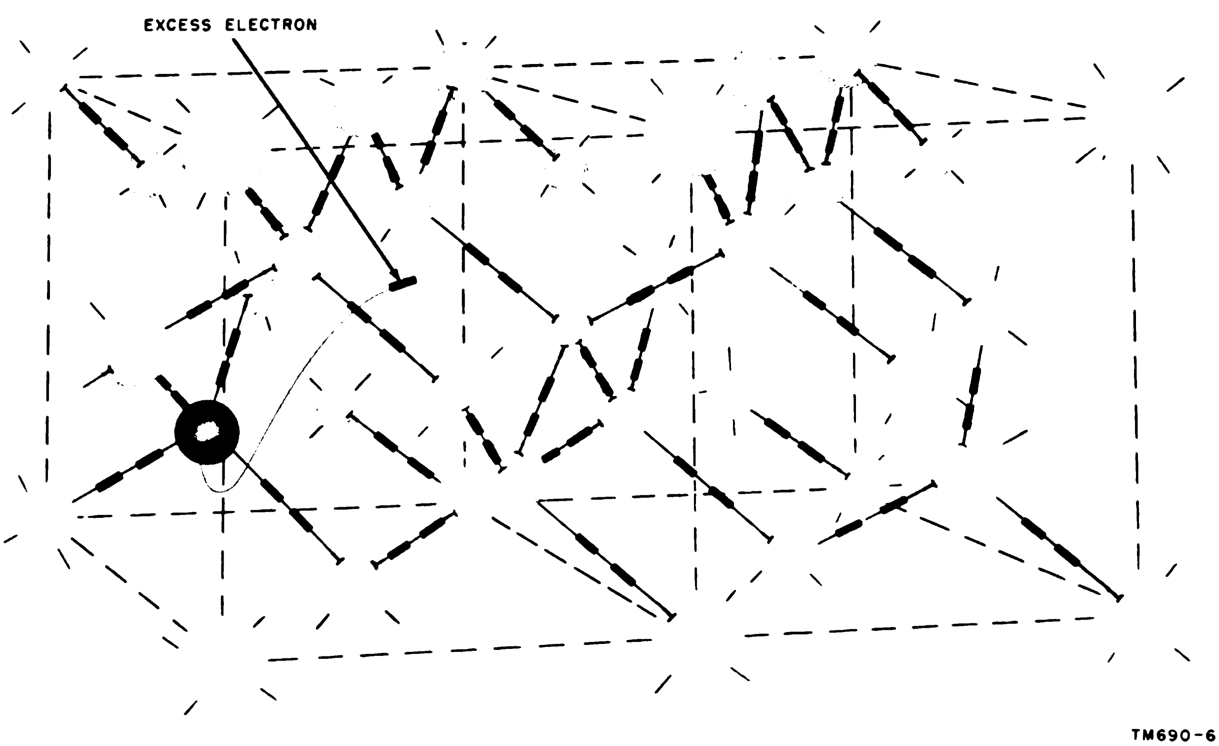


Figure 20. Germanium crystal with donor atom, showing movement of excess electron.

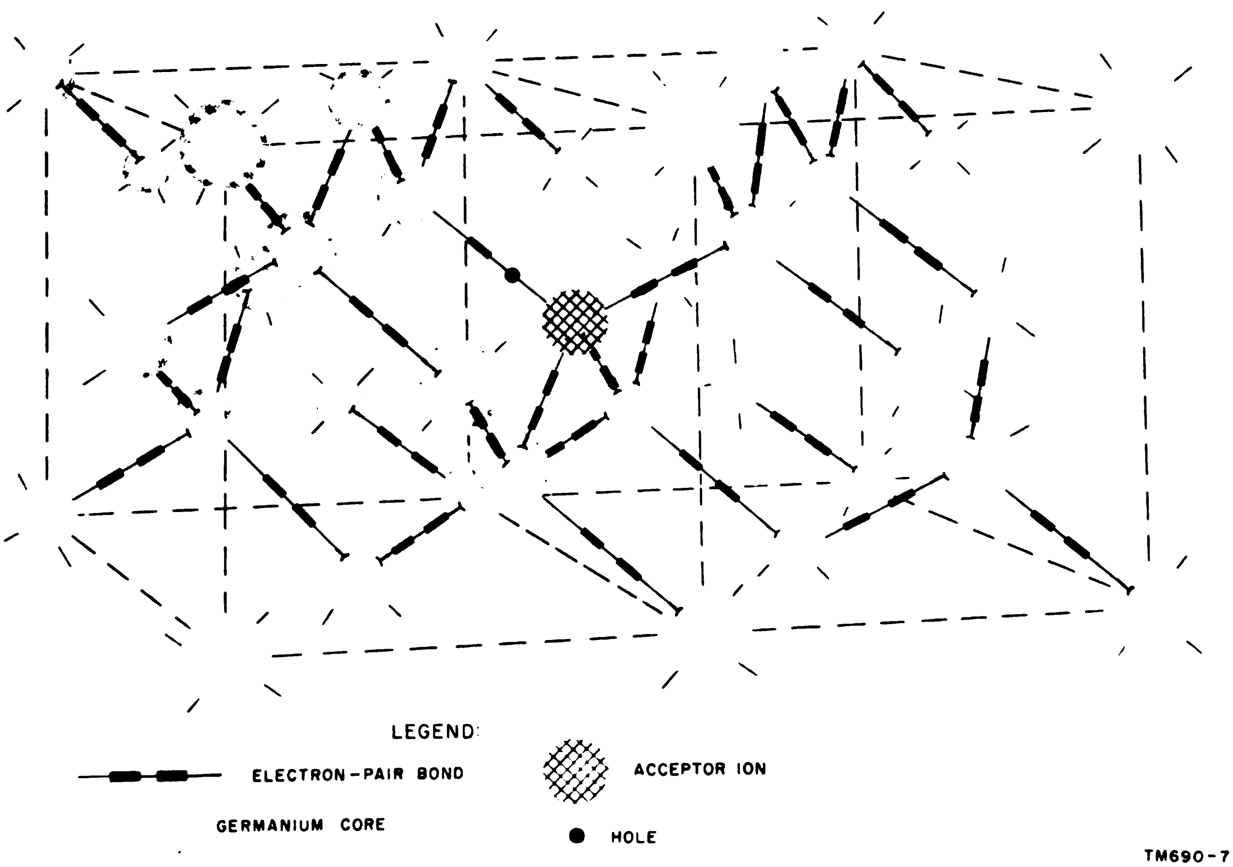


Figure 21. Germanium crystal with acceptor atom.

valence electrons. In this condition, an electron-hole arrangement exists. The position that would normally be filled with an electron is designated a *hole*.

c. It is possible for an electron from an adjacent electron-pair bond to absorb enough energy to break its bond (fig. 22) and fill in the hole in the original electron-hole arrangement (fig. 23). Note that the hole has moved to a new position. When the hole moves to the new position, two important changes take place.

(1) The first change is that the acceptor atom has been ionized; that is, the acceptor has acquired (or *accepted*) an electron and is now an ion. A negative charge exists in the immediate vicinity of the acceptor.

(2) The second change is that the germanium atom, which requires four valence electrons, is left with only three valence electrons. The germanium atom, lacking an electron, has a net positive charge equivalent to the negative charge of the electron. Because of the existence of the electron-pair bond structure, the positive charge of the germanium atom is not diffused or scattered, but is concentrated in the hole in the electron-hole arrangement. Furthermore, laboratory experiments have shown that the positive hole moves within the crystal in the same manner that a free electron moves within the crystal. The concept of holes is very important in understanding the operation of transistors; the properties and characteristics of holes are presented more thoroughly in paragraph 22.

d. Note that the germanium crystal that contains an acceptor ion (negative) also contains a hole (positive). The germanium crystal taken as a whole therefore possesses a net charge of zero.

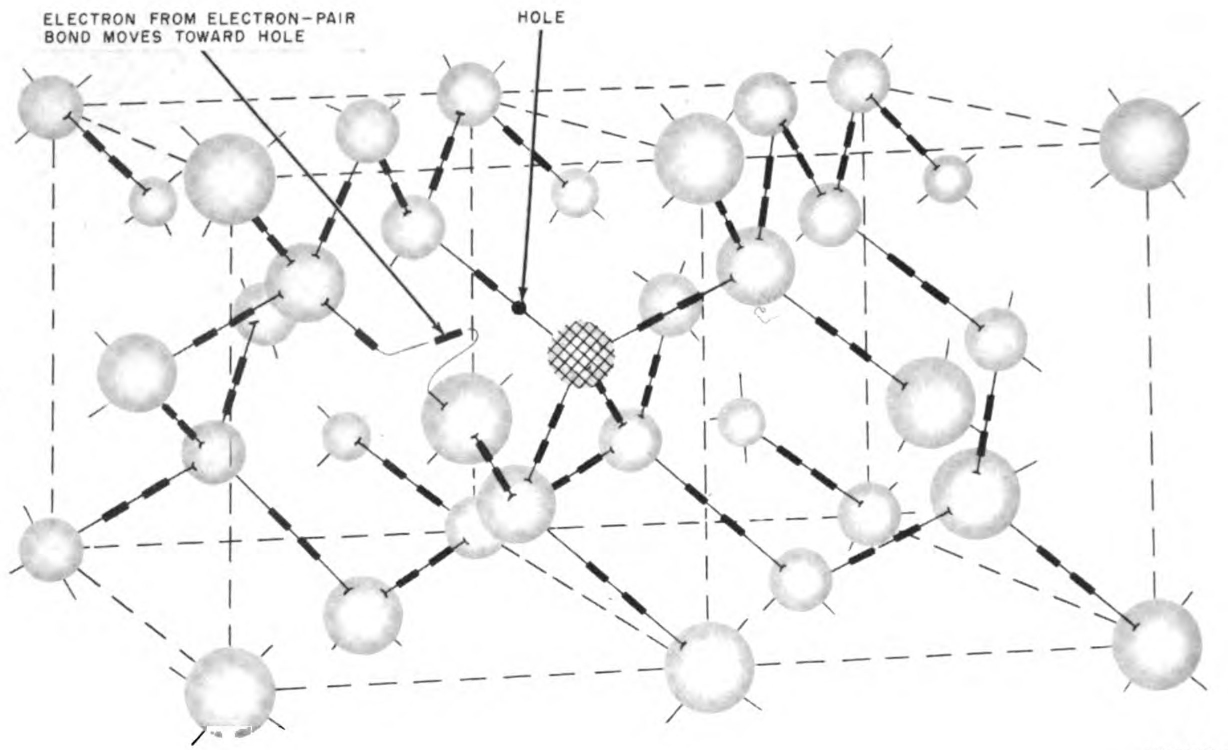


Figure 22. Germanium crystal with acceptor atom, showing electron from electron-pair bond moving toward hole.

e. Germanium containing acceptor impurities is referred to as *P-type germanium.* The letter *P* refers to the positive charge of the hole.

21. Movement of Hole

Figure 24 is a two-dimensional representation of the mechanism involved in the movement of a hole through a crystal. In A, figure 24, the hole is in the upper left hand corner. An electron from an adjacent electron-pair bond moves to the position of the hole. The hole (B, fig. 24) is now midway between top and bottom of the crystal and slightly to the right of its original position. This process is repeated in C and D, figure 24, until the hole is at the right hand side of the crystal (E, fig. 24). The complete path of the hole through the crystal is shown in F, figure 24.

22. Holes, Properties and Characteristics

For the understanding of transistors and transistor theory, it is convenient for the reader to think of the hole as a specific particle. Holes, in motion, constitute an electrical current to the same extent that electrons in motion constitute an electrical current (par. 14). There are differences, however, which must be kept in mind.

a. The hole can exist only in a semiconductor material such as germanium or silicon. This is because the hole depends for its existence on a specific arrangement of electrons (electron-pair bonds) and atoms as are found in crystal substances (par. 20). Holes *do not* exist in conductors such as copper and aluminum.

b. The hole is deflected by electric and magnetic fields in the same manner that electrons are deflected. Because the hole possesses a charge equal and opposite to that of the electron, the direction of deflection of the hole is opposite to that of the electron. In an electric field, for instance, the electron moves toward the positive pole; the hole moves toward the negative pole.

c. In the field of electronics, the electron is considered indestructible. When a hole is filled by an electron from an adjacent electron-pair bond, the hole is considered as having moved from one position to another (fig. 24). When a hole is filled by a free or excess electron, the hole no longer exists. This statement is supported by

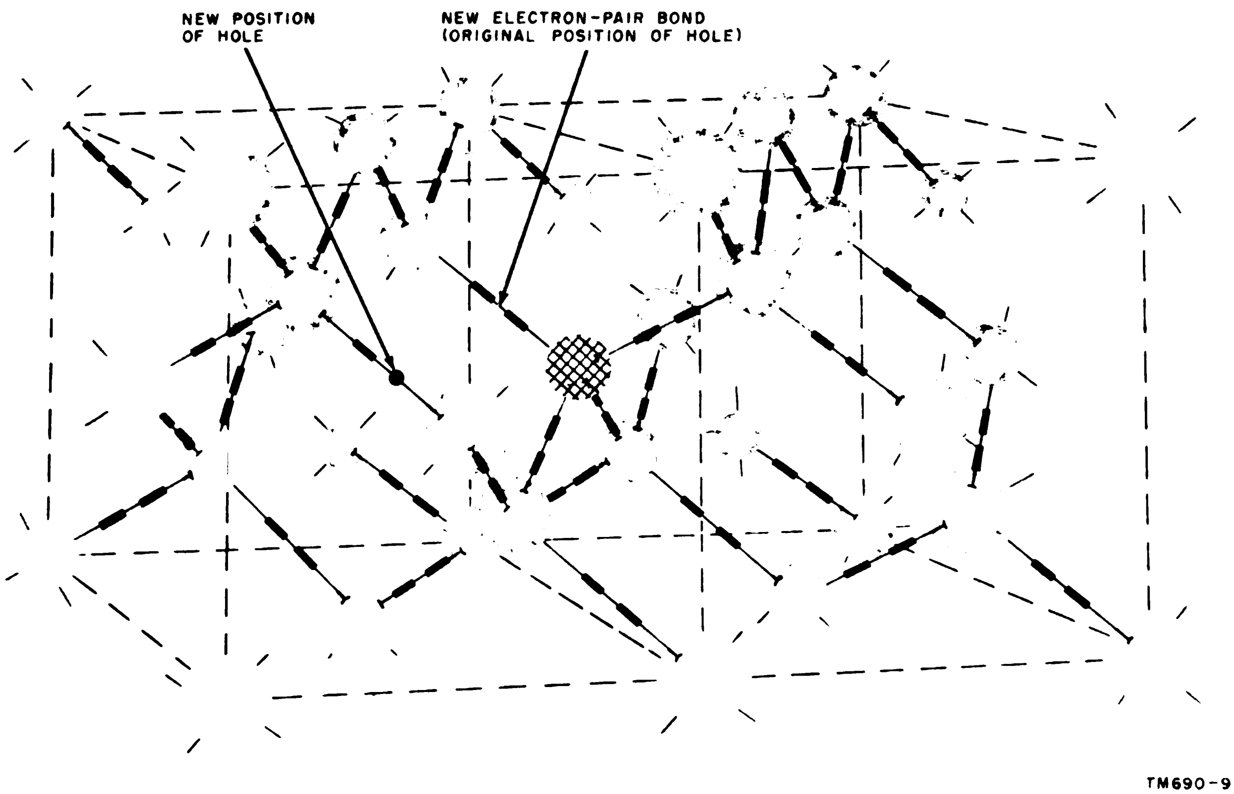


Figure 23. Germanium crystal with acceptor atom showing new position of hole.

the fact that germanium containing an equal number of donor and acceptor atoms has none of the properties of P-type or N-type germanium.

23. Movement of Positive Charges

a. The fact that a positive charge in motion causes an electric current to the same extent that a negative charge (electrons) in motion causes an electric current has been known for many years. In 1899, a scientist by the name of H. A. Rowland established this fact by a simple experiment. Rowland's experimental setup was *similar* to that shown in figures 25 and 26.

(1) In A, figure 25, a doughnut-shaped neutral body (a conducting material) is suspended by a string. A negatively charged rubber rod (B, fig. 25) is brought into contact with the neutral body. The electrons on the rubber rod flow to the neutral body and the rubber rod is removed. The electrons adhere to the surface of the doughnut-shaped body. While the negatively charged body is at rest (C, fig. 25), a compass needle suspended by a string is brought near the negatively charged body. The compass needle does not deflect. When the negatively charged body is rotated rapidly (D, fig. 25), a magnetic field is generated with the direction of the magnetic lines of force as shown. The generation of the magnetic lines of force in this case is similar to the generation of magnetic lines of force when electrons flow through a stationary conductor. The compass needle deflects. The direction of rotation of the negatively charged body is *clockwise*.

(2) The experiment is repeated, starting again with a neutral body (A, fig. 26). However, the neutral body is now touched with a *positively* charged glass rod (B, fig. 26). Electrons flow from the doughnut-shaped body to the glass rod, and the glass rod is removed. This action leaves a net *positive* charge on the doughnut-shaped body. While the positively charged body (C, fig. 26) is

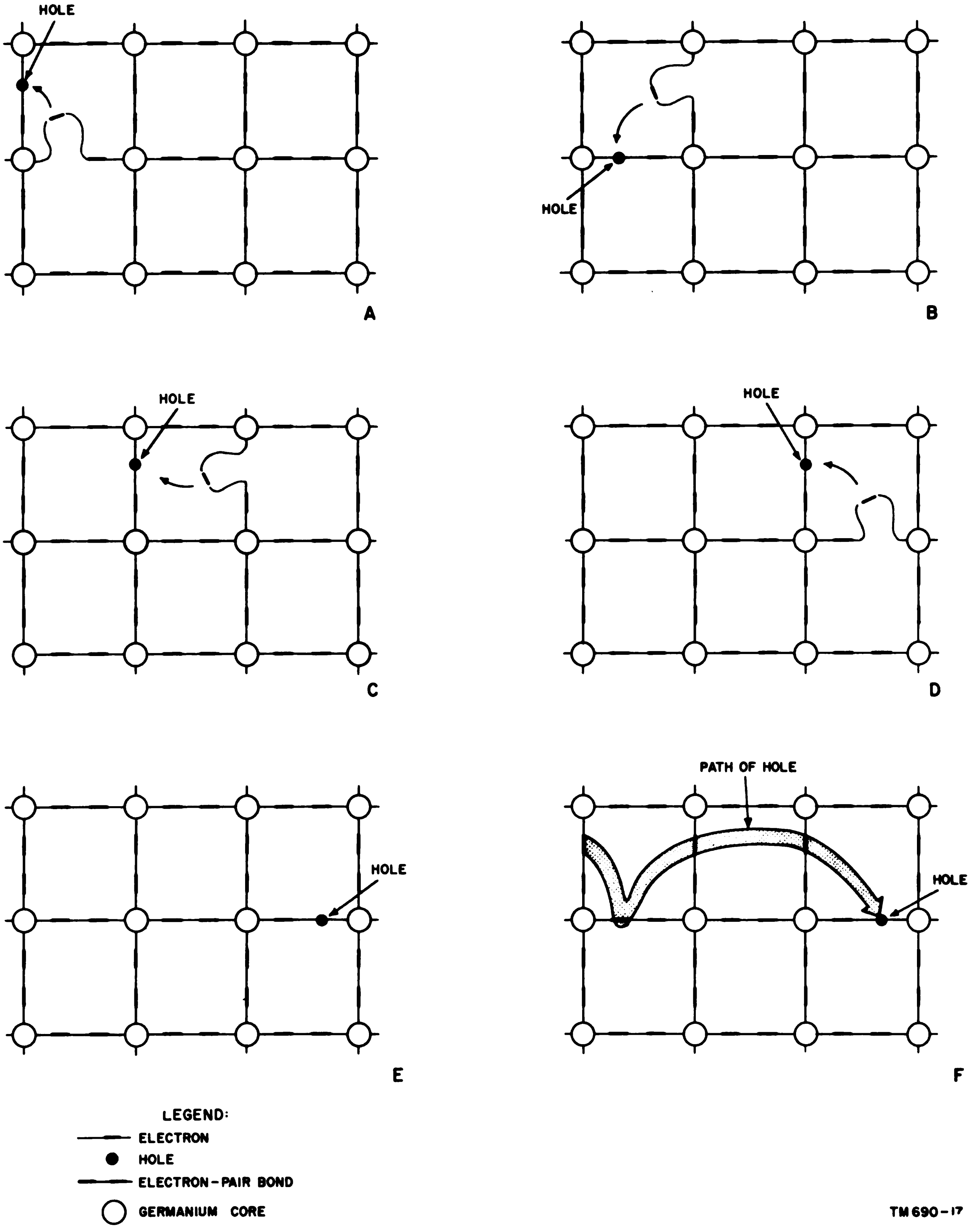


Figure 24. Movement of hole through crystal.

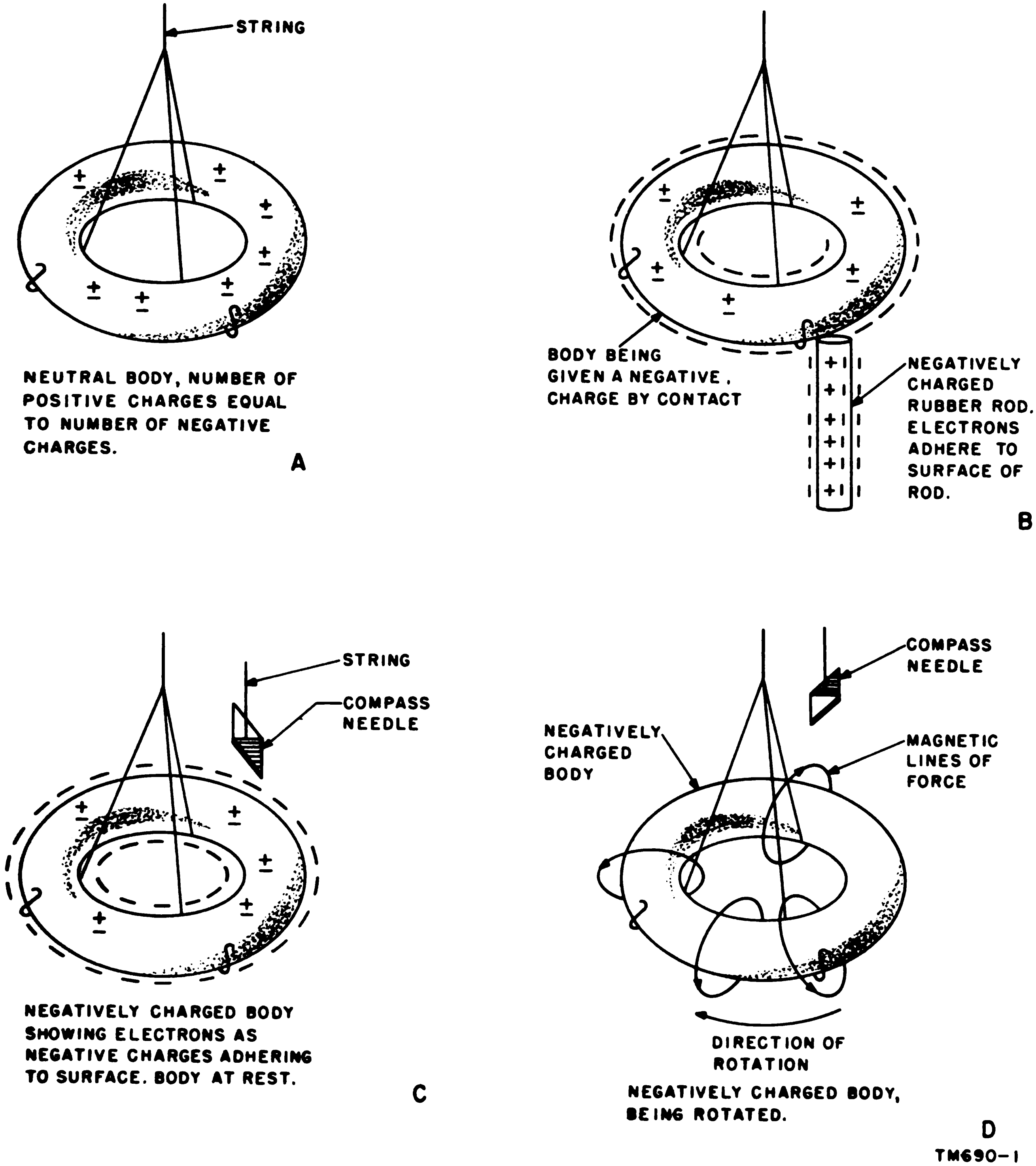


Figure 25. Neutral body being negatively charged and rotated.

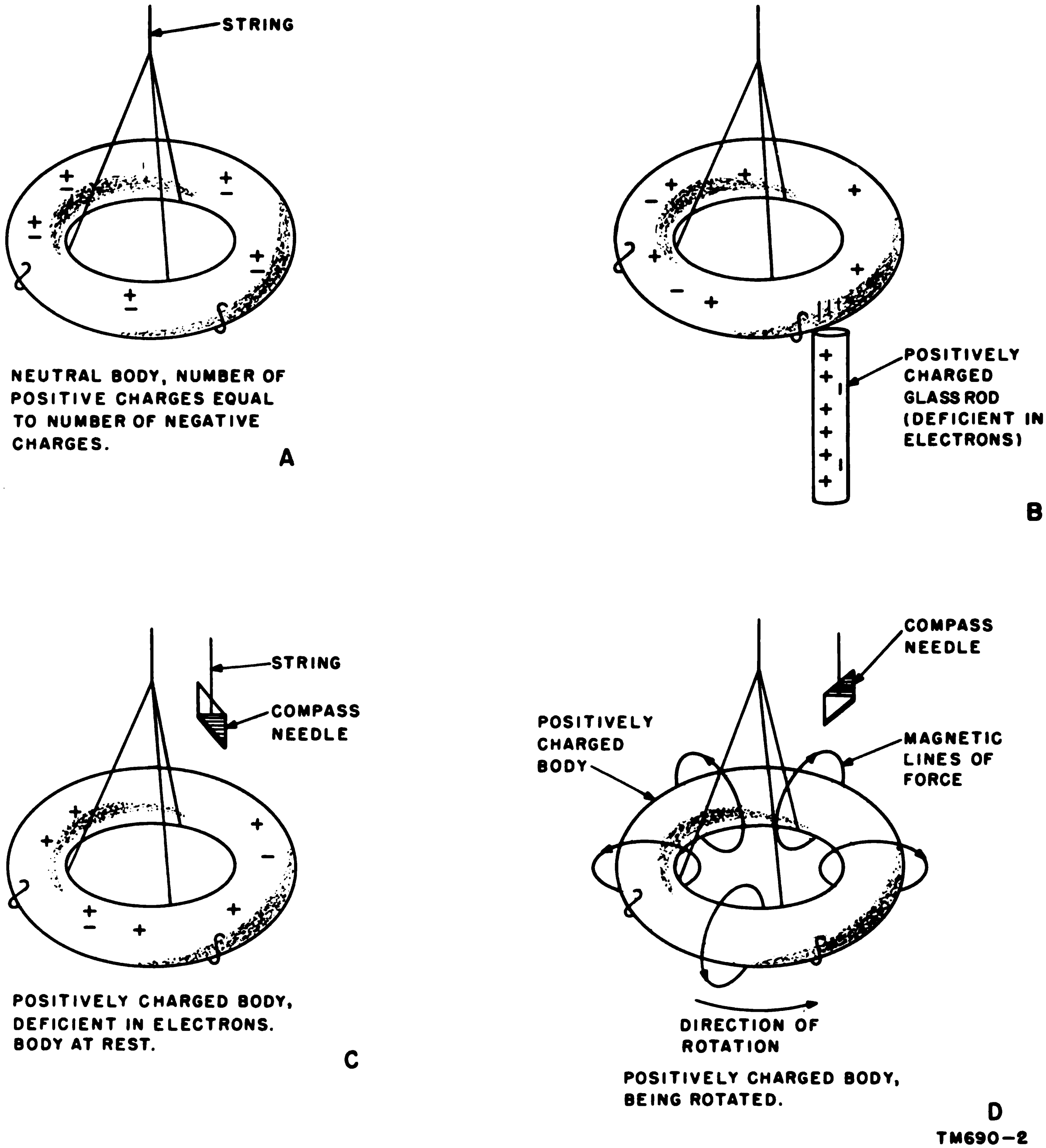


Figure 26. Neutral body being positively charged and rotated.

at rest, a compass needle is brought near the positively charged body. The compass needle does not deflect. When the positively charged body is rotated rapidly (D, fig. 26), a magnetic field is generated with the direction of the magnetic lines of force as shown. The compass needle deflects. The direction of the magnetic lines of force is the same as that for the negatively charged body (D, fig. 25), *but the direction of rotation of the positively charged body is opposite to that of the negatively charged body.*

b. This experiment proves that the movement of positive charges constitutes an electric current. Although the positive charges involved in this experiment are not holes in the semiconductor sense of the word, the experiment proves that positive charges are *current carriers.*

24. Holes as Current Carriers

Figure 27 is a *simplified* version of a laboratory experiment performed to confirm the theory of holes and demonstrate that they act as current carriers.

a. In A, figure 27 a battery is placed across a slab of P-type germanium.

(1) Because a hole possesses a positive charge, it is attracted toward terminal 2 (negative), and repelled by terminal 1 (positive). When a hole reaches termi-

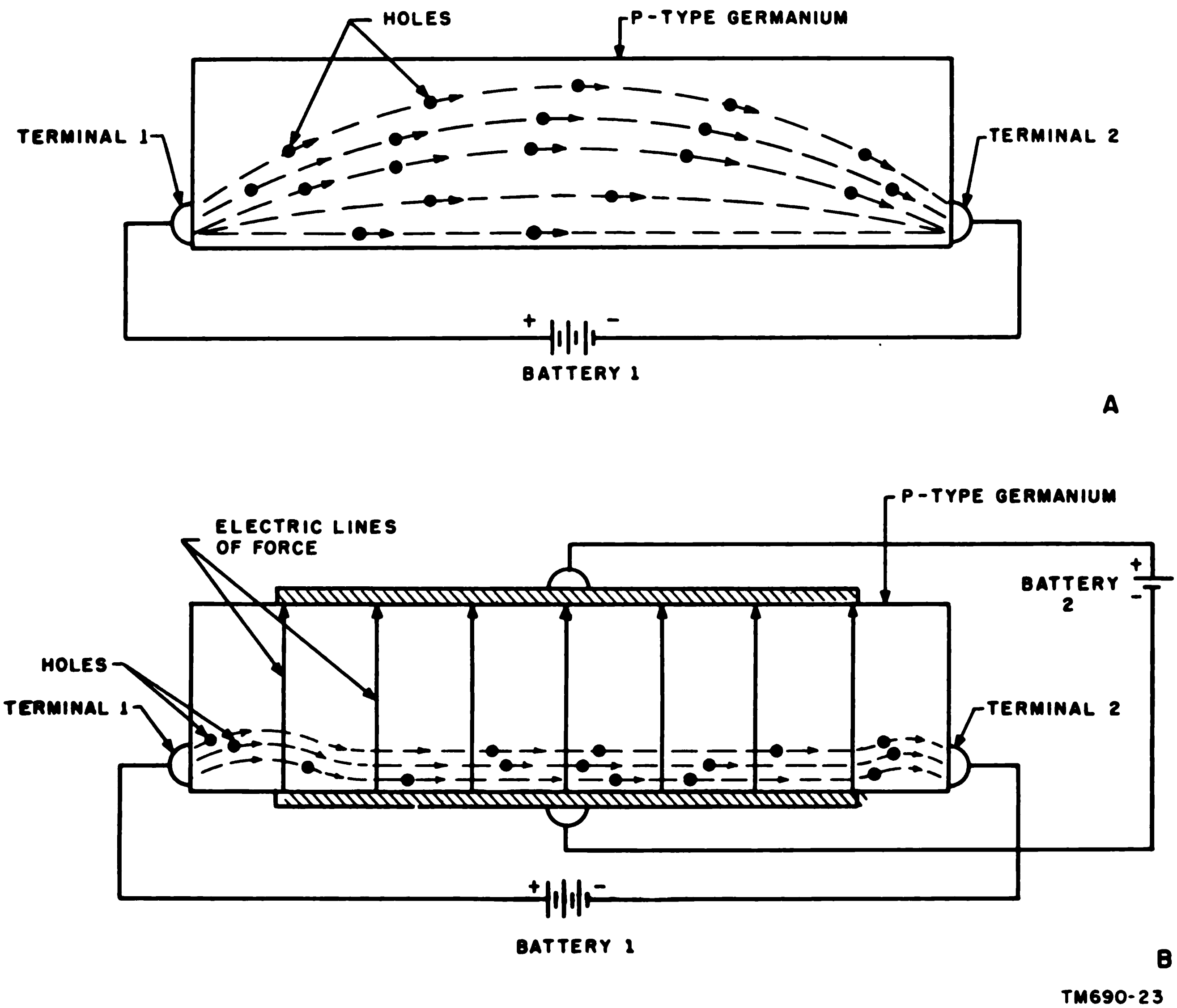


Figure 27. Deflection of hole current in P-type germanium.

nal 2, an electron from the battery enters the germanium and fills the hole. When this happens, an electron from an electron-pair bond in the crystal and near terminal 1 breaks its bond and enters terminal 1. The breaking of the bond creates another hole; the hole flows toward terminal 2. This action provides a continuous flow of electrons in the external circuit and holes in the germanium.

(2) Observe that the holes flow in an arc from terminal 1 to terminal 2. Under precise laboratory conditions, scientists measured the time required for the hole to flow from terminal 1 to terminal 2.

b. In B, figure 27, the circuit arrangement shown in A is repeated, but in addition, a field of electric lines of force is introduced into the body of germanium. The field is established by connecting a second battery (battery 2) across the germanium slab as shown.

(1) The electric lines of force are in the direction that an electron (negative) would be deflected if placed in the field. A hole placed in the field of battery 2 would be deflected in a direction opposite to that of the electron. The resultant force on the hole, by the fields of battery 1 and battery 2, causes the hole to move from terminal 1 to terminal 2 in essentially a straight line.

(2) Under precise laboratory conditions, the time required for the hole to flow from terminal 1 to terminal 2 was measured and it was found that the time was less than that in *a*(2) above.

c. The explanation for the reduced time in *b*(2) above, compared to that in *a*(2) above, is that the path of the holes has been changed from an arc to an approximately straight line. If electrons had been the current carrier, the electrons would have been deflected upward by the electric lines of force and the path would have been increased. The flow time would also have been increased. Similar experiments yielding the same results have been performed by scientists using magnetic lines of force to deflect the hole current.

Section III. PN JUNCTIONS

25. General

When N-type germanium (par. 19) and P-type germanium (par. 20) are joined in the same crystal, an unusual but very important phenomenon occurs at the surface where contact is made between the two types of germanium. The contact surface is referred to as a *PN junction.* The phenomenon that occurs at PN junctions permits the use of semiconductors, such as germanium and silicon, in circuits normally employing electron vacuum tubes. The detailed theory of PN junctions is covered in paragraphs 26 through 29.

26. PN Junction, General

a. Figure 28 shows a section of P-type germanium and a section of N-type germanium. Note that the sections are separated.

(1) For simplicity, the electron-pair bonds are not shown. Only the holes, the excess electrons, the germanium cores, and the donor and acceptor ions are represented.

(2) For discussion purposes, the figure shows a large number of acceptor ions in the P-type germanium and a large number of donor ions in the N-type germanium. In practice, however, transistor germanium contains approximately one impurity atom per 10 million germanium atoms.

(3) If one could actually look inside the bulk germanium material, one would see the germanium cores and the impurity ions vibrating within their lattice positions because of thermal energy. However, the cores and the ions *do not* leave their lattice positions and therefore do not constitute a current. The cores and the ions may be considered to be stationary. The holes and the excess electrons would be seen to move haphazardly within the bulk germanium material. The movement of the holes and the electrons is also due to thermal energy; this movement of charges in the absence of an applied field is called *diffusion.* Even though the holes are in motion they are evenly distributed throughout the P-type

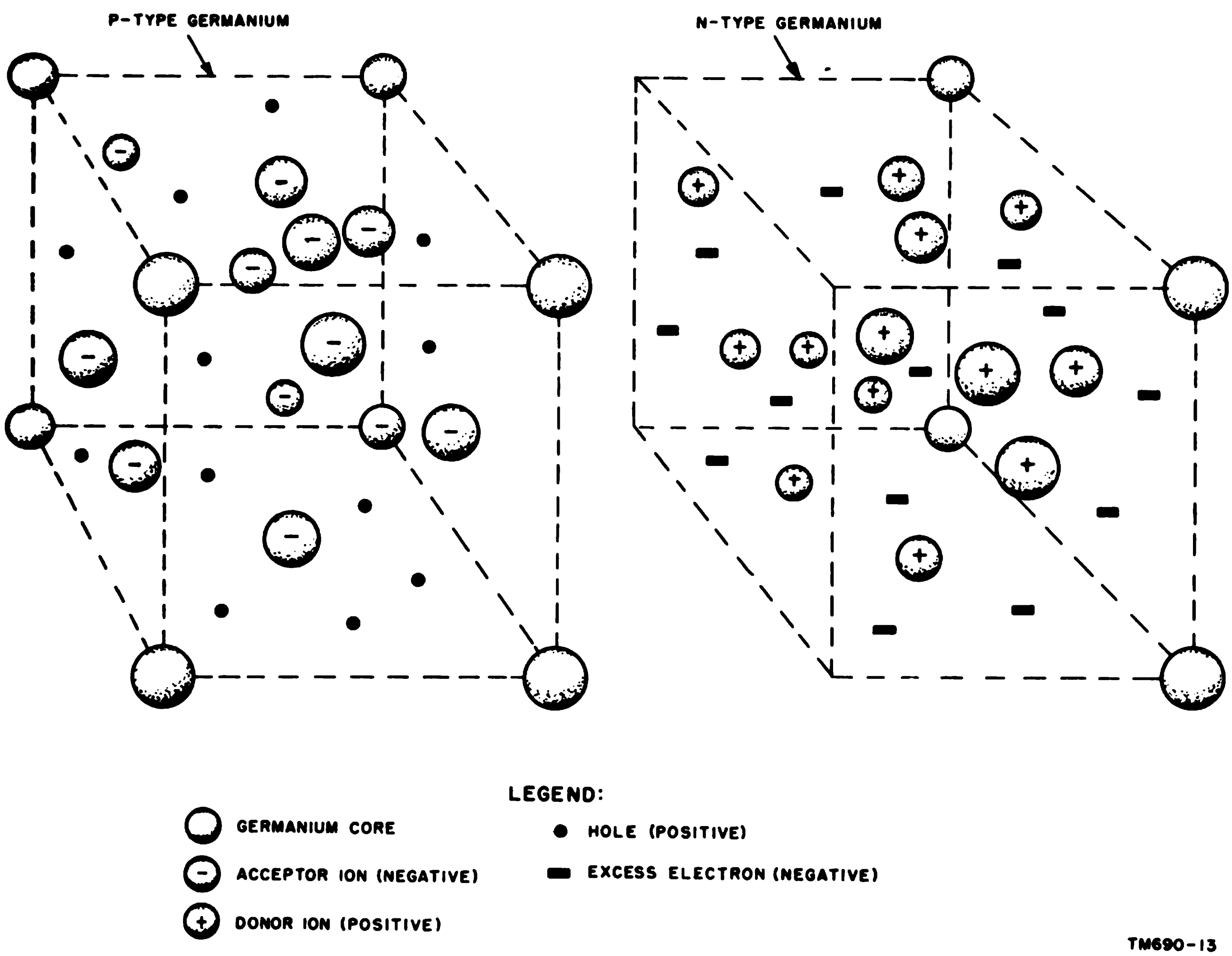


Figure 28. Separated sections of P-type and N-type germanium.

germanium; the excess electrons are evenly distributed throughout the N-type germanium.

b. Figure 29 shows the same two sections of germanium (*a* above), joined to form a PN junction.

(1) Note that no external circuits or voltages have been connected to the germanium; nor is the germanium exposed to external electric or magnetic fields.

(2) One would normally expect the holes in the P-type germanium and the electrons in the N-type germanium to flow towards each other, combine and eliminate all holes and excess electrons. When the two types of germanium are joined, however, after a few combinations of holes and electrons result, a restraining force is set up automatically to preclude total combination. This restraining force is called a *barrier*. The cause and nature of the barrier are discussed in paragraph 27.

27. Junction Barrier

a. When the P-type germanium and the N-type germanium are joined (fig. 29), some of the holes in the P-region and some of the excess electrons in the N-region diffuse toward each other and combine. Each combination eliminates a hole and an *excess* electron; the excess electron is now part of an electron-pair bond. This action occurs for a short time in the immediate vicinity of the junction. Negative acceptor ions in the P-region and positive donor ions in the N-region and near the junction are left uncompensated. Additional holes that would diffuse into the N-region are repelled by the uncompensated positive charge of the donor ions. Electrons that would diffuse into the P-region are repelled by the uncompensated

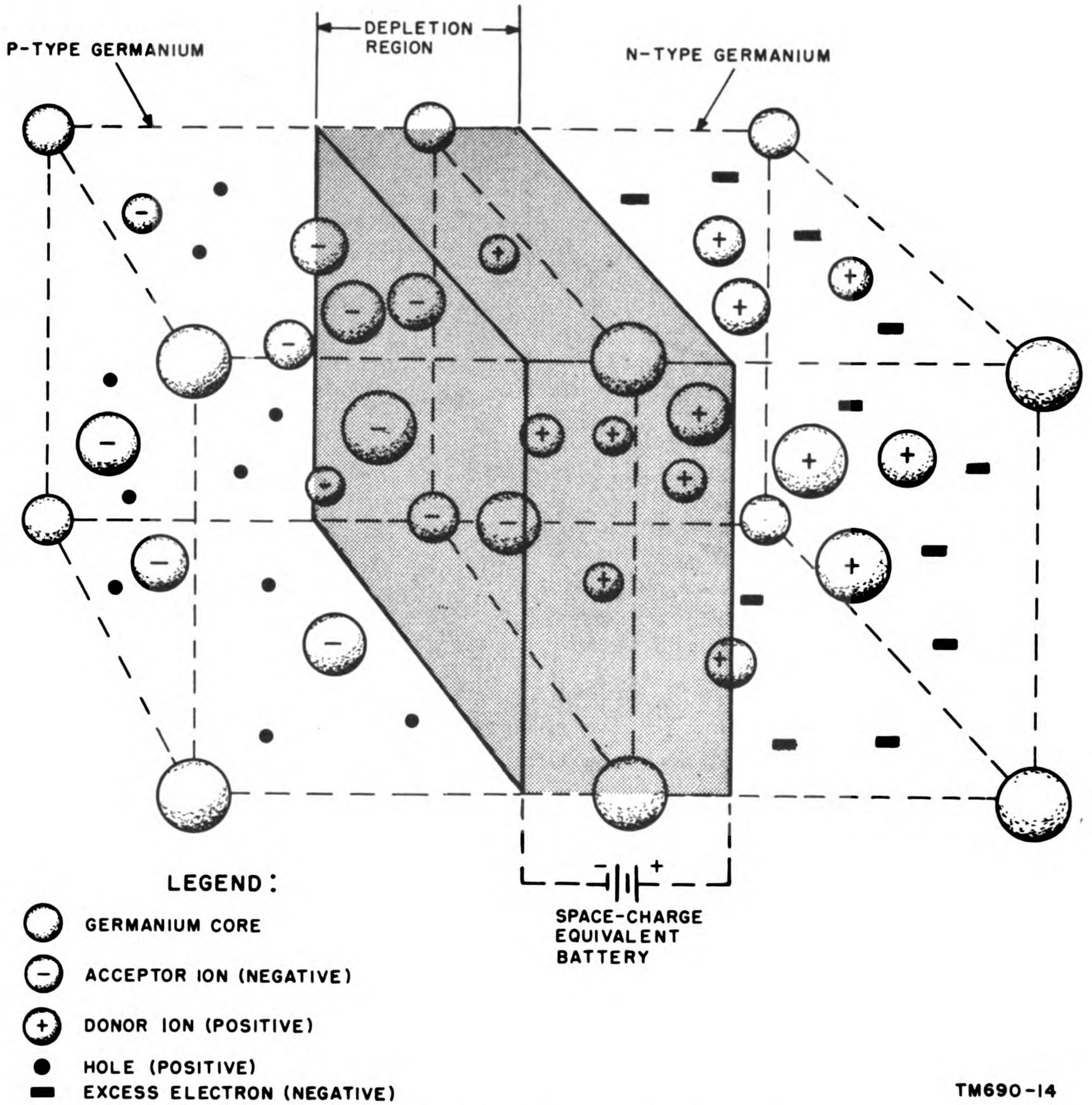


Figure 29. Joined sections of P-type and N-type germanium.

negative charge on the acceptor ions. As a result, *total* recombination of holes and electrons cannot occur.

b. The region containing the uncompensated acceptor and donor ions is referred to as the *depletion region*. That is, there is a depletion of holes and a depletion of excess electrons in this region. Since the acceptor and the donor ions are immobile (fixed) and are charged electrically, the depletion region is also referred to as the *space charge region*. The electric field between the acceptor and the donor ions is called a *barrier*. The effect of the barrier is represented by the imaginary space-charge equivalent battery. The physical distance from one side of the barrier to the other is referred to as the *width* of the barrier. The width of the barrier depends on the density of holes and excess electrons in the germanium crystal. The difference of potential from one side of the barrier to the other is referred to as the *height* of the barrier. The *height* of the barrier is the intensity of the electric field (voltage of space-charge equivalent battery) and is measured in volts. With no *external* batteries connected, the barrier height is on the order of tenths of a volt.

c. It is stated in *a* above, that total recombination of electrons and holes cannot occur. Inspection of the polarity of the space-charge equivalent battery confirms this statement. Note that the electrons in the N-type germanium are already at the *highest* positive potential (positive terminal of space-charge equivalent battery) within the crystal. The holes in the P-type germanium also are at the highest negative potential (negative terminal of space-charge equivalent battery) within the crystal. This condition precludes the movement of holes or electrons across the PN junction.

28. PN Junction, Reverse Bias

a. Figure 30 shows what happens when an external battery with the indicated polarity is connected to a PN junction. Note that the negative terminal of the battery is connected to the P-type germanium and the positive terminal of the battery is connected to the N-type germanium. The holes are attracted toward the nega-

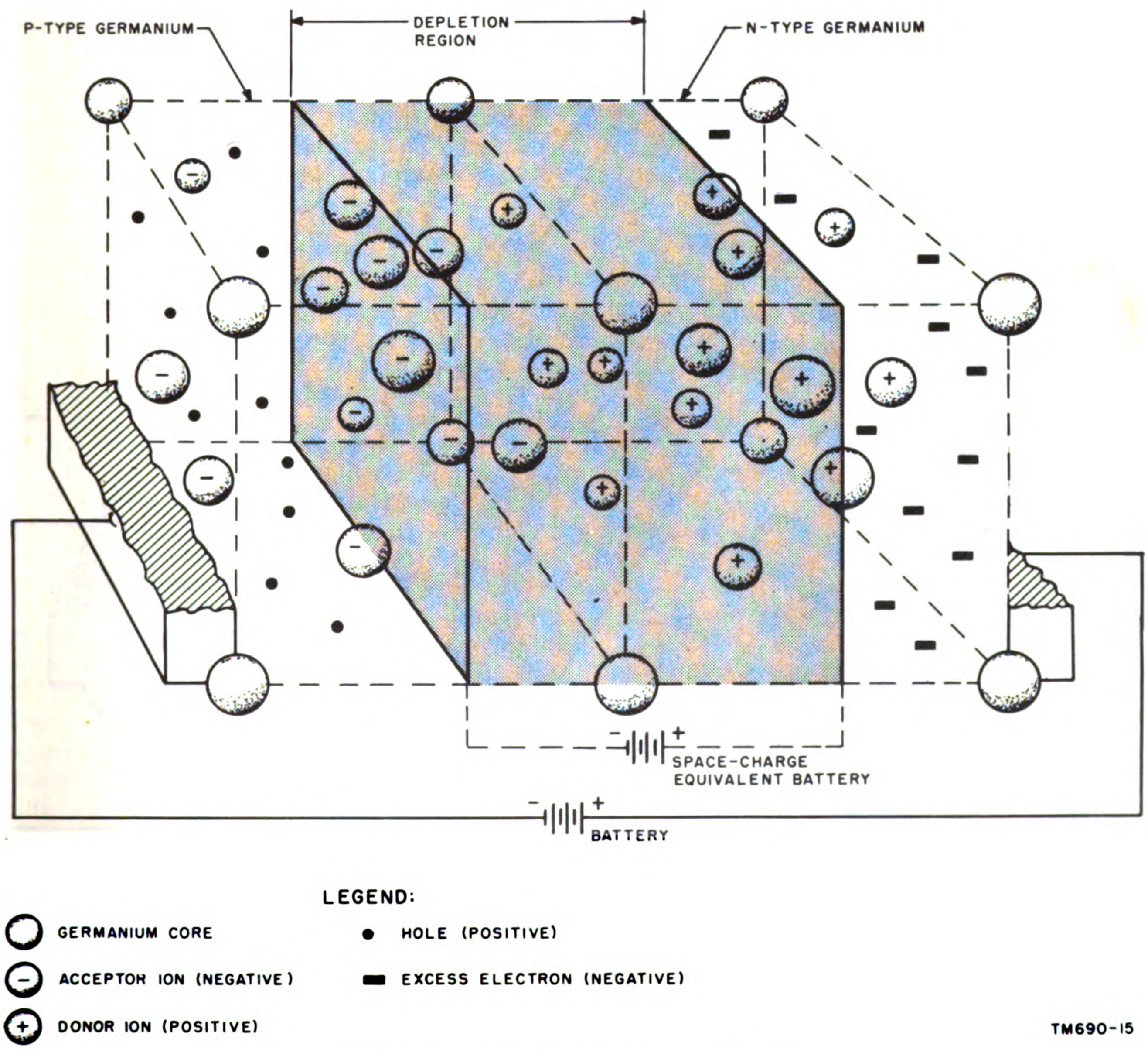


Figure 30. PN junction showing reverse bias.

tive terminal and away from the junction. The electrons are attracted toward the positive terminal and away from the junction. This action widens the depletion region and increases the barrier height (potential). Compare the width of the depletion region of figure 29 and that of figure 30.

b. Since the depletion region widens until the barrier height (potential of space-charge equivalent battery) equals the potential of the external battery, no current flow of holes or electrons occurs because the battery voltages are in opposition. In this condition, the PN junction is biased in the *reverse* direction; or simply, a *reverse bias* is placed across the PN junction.

c. It is possible to apply a reverse bias greater than the largest possible barrier height. However, if this is done, the crystal structure will break down. In normal applications, this condition is avoided. The crystal structure will return to normal when the excess reverse bias is removed, provided that overheating does not permanently damage the crystal.

29. PN Junction, Forward Bias

a. Figure 31 shows what happens when an external battery with the indicated polarity is

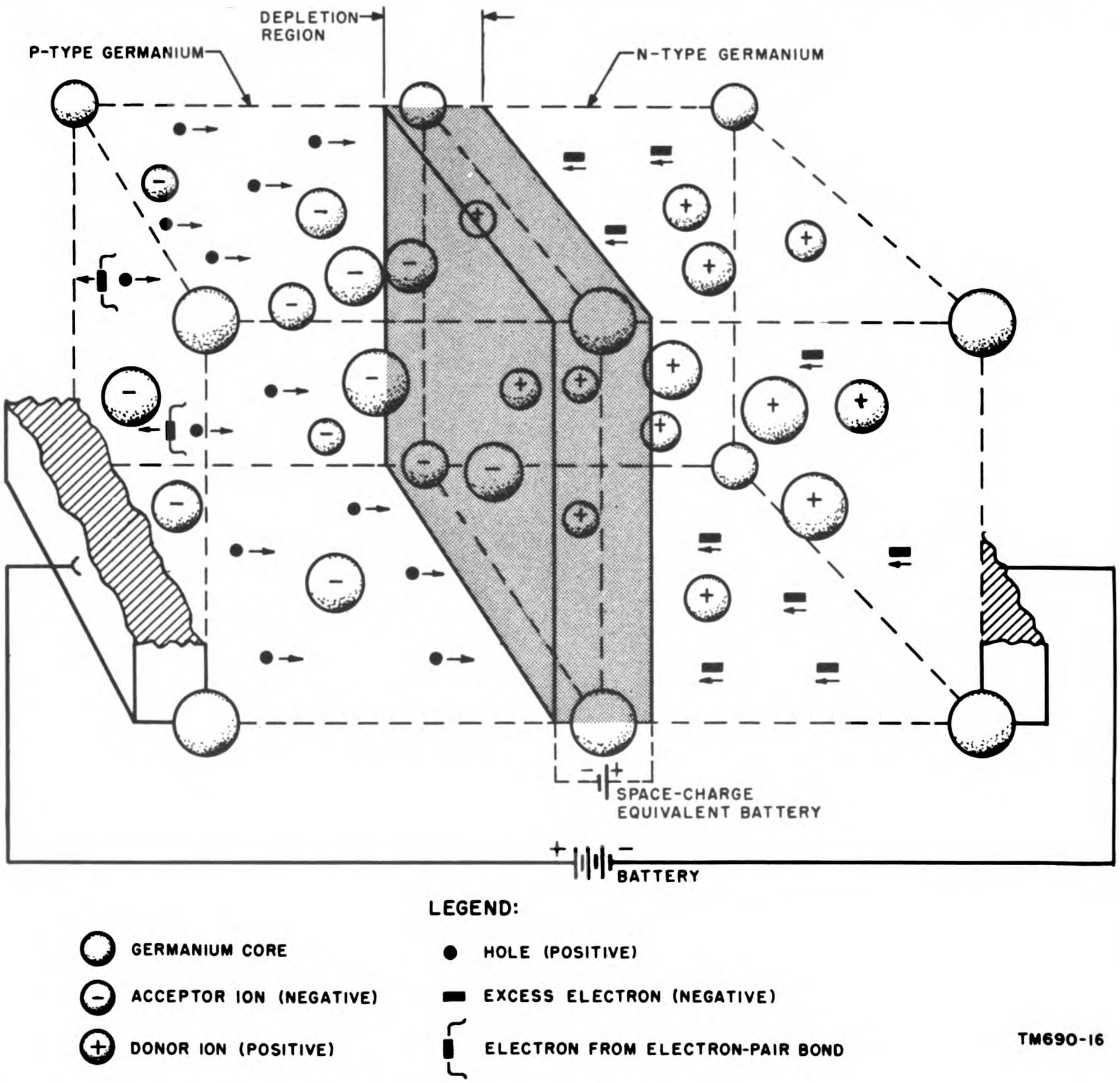


Figure 31. PN junction showing forward bias.

connected to a PN junction. Note that the positive terminal of the battery is connected to the P-type germanium and the negative terminal of the battery is connected to the N-type germanium. The holes are repelled from the positive terminal of the battery and drift toward the junction. The electrons are repelled from the negative terminal of the battery and drift toward the junction. Because of their acquired energy, some of the holes and the excess electrons penetrate the depletion (space charge) region and combine (par. 27*a*).

b. For each combination of an excess electron and a hole that occurs, an electron from the negative terminal of the external battery enters the N-type germanium and drifts toward the junction. Similarly, an electron from an electron-pair bond in the crystal, and near the positive terminal of the external battery, breaks its bond and enters the positive terminal of the external battery. For each electron that breaks its bond, a hole is created which drifts toward the junction. Recombination in and about the space-charge region continues as long as the external battery is applied.

c. Note that there is a continuous electron current in the external circuit. The current in the P-type germanium consists of holes, the current in the N-type germanium consists of electrons. In this condition, the PN junction is said to be biased in the forward direction. If the forward bias is increased, the current increases.

d. In paragraph 27*b* it was stated that the barrier potential with no external battery connected is on the order of tenths of a volt: It would appear, therefore, that an external battery of very low voltage (about 1 volt) would eliminate the barrier completely. However, the larger the voltage of the external battery, the greater the current flow through the crystal. Since the crystal has a relatively high resistivity (several hundred thousand times that of copper (par. 15)), increased current causes increased voltage drop on both sides of the barrier. The remaining voltage of the external battery does not overcome the barrier completely. Normally, 1 to 1½ volts is used to bias the PN junction in the forward direction. If excessive forward bias is used, excessive current will cause excessive thermal agitation and breakdown of the crystal structure.

30. Diode Action

a. Paragraphs 26 through 30 cover the mechanism of rectification through a PN germanium diode. Figure 32 is a plot of current flow versus voltage applied to a practical PN junction. Note that current flow in the forward bias direction is quite high (measured in ma). However, current flow in the reverse bias direction, although low (measured in μa), is not zero as might have been expected (par. 28). The reverse-bias current flow occurs because some acceptor ions and their associated holes occur in the N-type germanium (par. 19), and some donor ions and their associated excess electrons occur in the P-type germanium (par. 20). The holes found in N-type germanium and the excess electrons found in the P-type germanium are called *minority carriers* because they are so few in number compared to the *majority carriers*, holes in P-type, and excess electrons in N-type germanium.

b. When the PN junction is biased in the reverse direction for the majority carriers (par. 28), the PN junction is biased in the *forward* direction (par. 29) for the minority carriers. The internal mechanism of conduction for the minority carriers when *forward* biased (majority carriers reverse biased) is identical with that for forward-biased majority carriers.

c. Note (fig. 32) that when a very high reverse bias is applied, a high reverse current flows. This high current is not due to the minority carriers. A breakdown of the single crystal structure occurs (par. 28*c*).

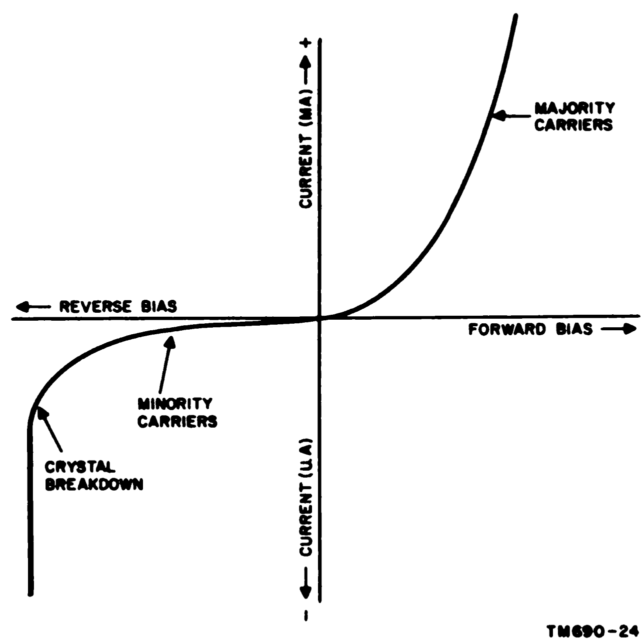


Figure 32. Chart of current through and voltage across a PN junction.

31. General

Observation of figure 32 reveals that a PN junction biased in the forward direction is equivalent to a *low-resistance* element (*high* current for a given voltage). The PN junction biased in the reverse direction is equivalent to a *high-resistance* element (*low* current for a given voltage). For a given current, the power developed in a high-resistance element is greater than that developed in a low-resistance element. (Power is equal to the current squared multiplied by the resistance value, or simply: $P=I^2R$.) If a crystal containing *two* PN junctions were prepared, a signal could be introduced into one PN junction biased in the forward direction (low resistance) and extracted from the other PN junction biased in the reverse direction (high resistance). This would produce a *power gain* of the signal when developed in an external circuit. Such a device would *transfer* the signal current from a low *resistance* circuit to a high *resistance* circuit. Contracting the terms *transfer* and *resistor* results in the term *transistor*. The detailed theory covering the operation of the transistor is covered in paragraphs 32 through 34.

32. PNP Transistor

a. To form two PN junctions (par. 26), three sections of germanium are required. A, figure 33, shows the three sections separated. When the three sections are combined (B, fig. 33), a *transistor* is formed. This paragraph covers the combination of two sections of P-type germanium and one section of N-type germanium. This transistor is referred to as a *PNP transistor*. Paragraph 33 covers the combination of two sections of N-type germanium and one section of P-type germanium to form an *NPN* transistor.

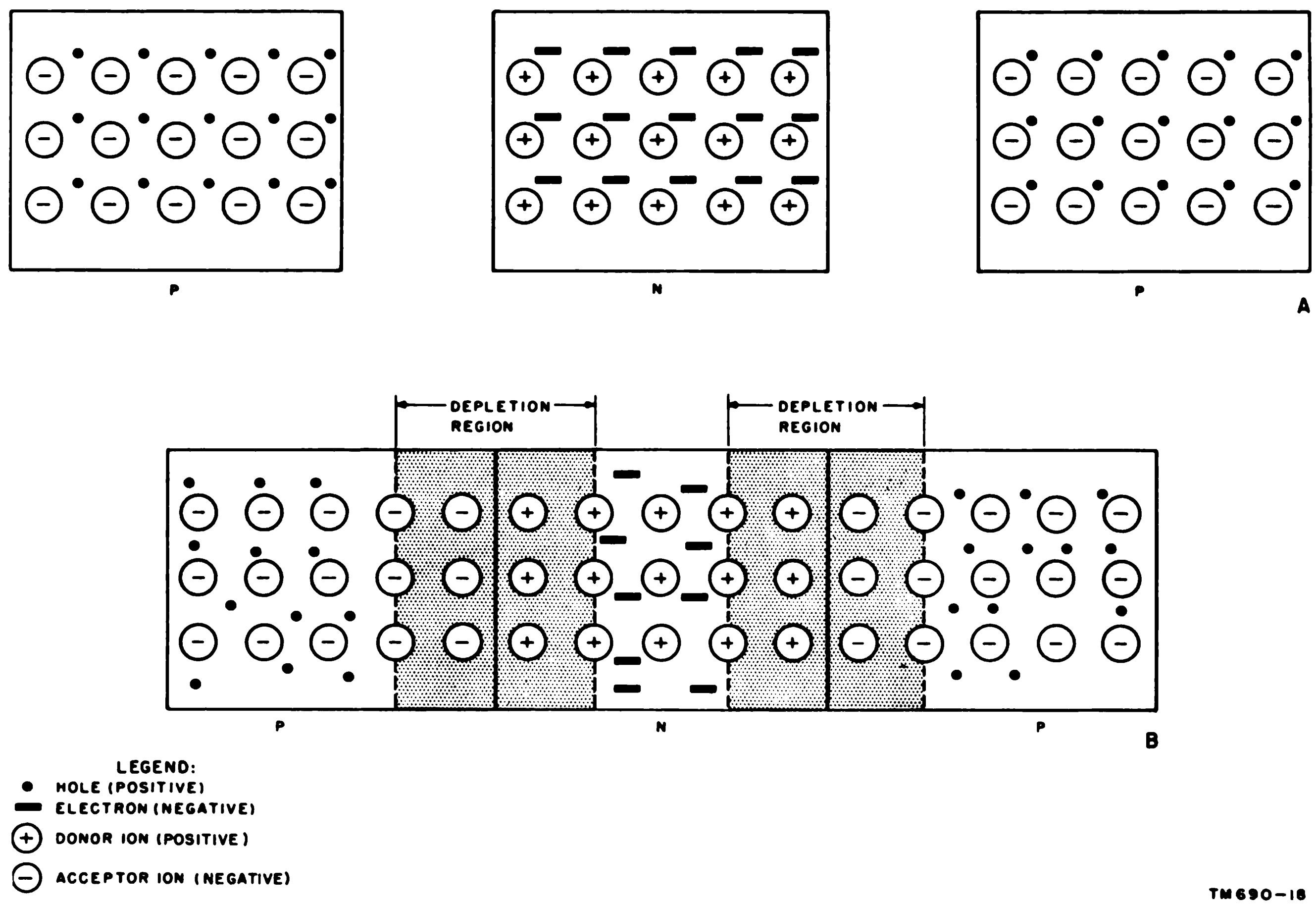


Figure 33. Two sections of P-type germanium and one section of N-type germanium, separated (A) and combined (B).

b. Note that when the three sections are combined, two depletion regions (barriers) occur at the junctions even though there is no application of external voltages, or fields. This phenomenon is the same as that which occurs when *two* sections of germanium are combined (par. 26).

c. Paragraph 31 states that transistor action requires that one junction be biased in the forward direction and the second junction be biased in the reverse direction.

(1) A, figure 34, shows the first junction biased in the forward direction. The second junction is not biased. Note that the depletion region (barrier) at the first junction is considerably reduced while the depletion region at the second junction is unchanged. The conduction mechanism is identical with that covered in paragraph 29.

(2) B, figure 34, shows the second junction biased in the reverse direction. The first junction is not biased. Note that the depletion region (barrier) at the second junction increases. Except for minority carriers (par. 30), not shown here, no current flows across the junction. This phenomenon is the same as that covered in paragraph 28.

d. A, figure 35, shows what happens when junctions are biased simultaneously. Because of

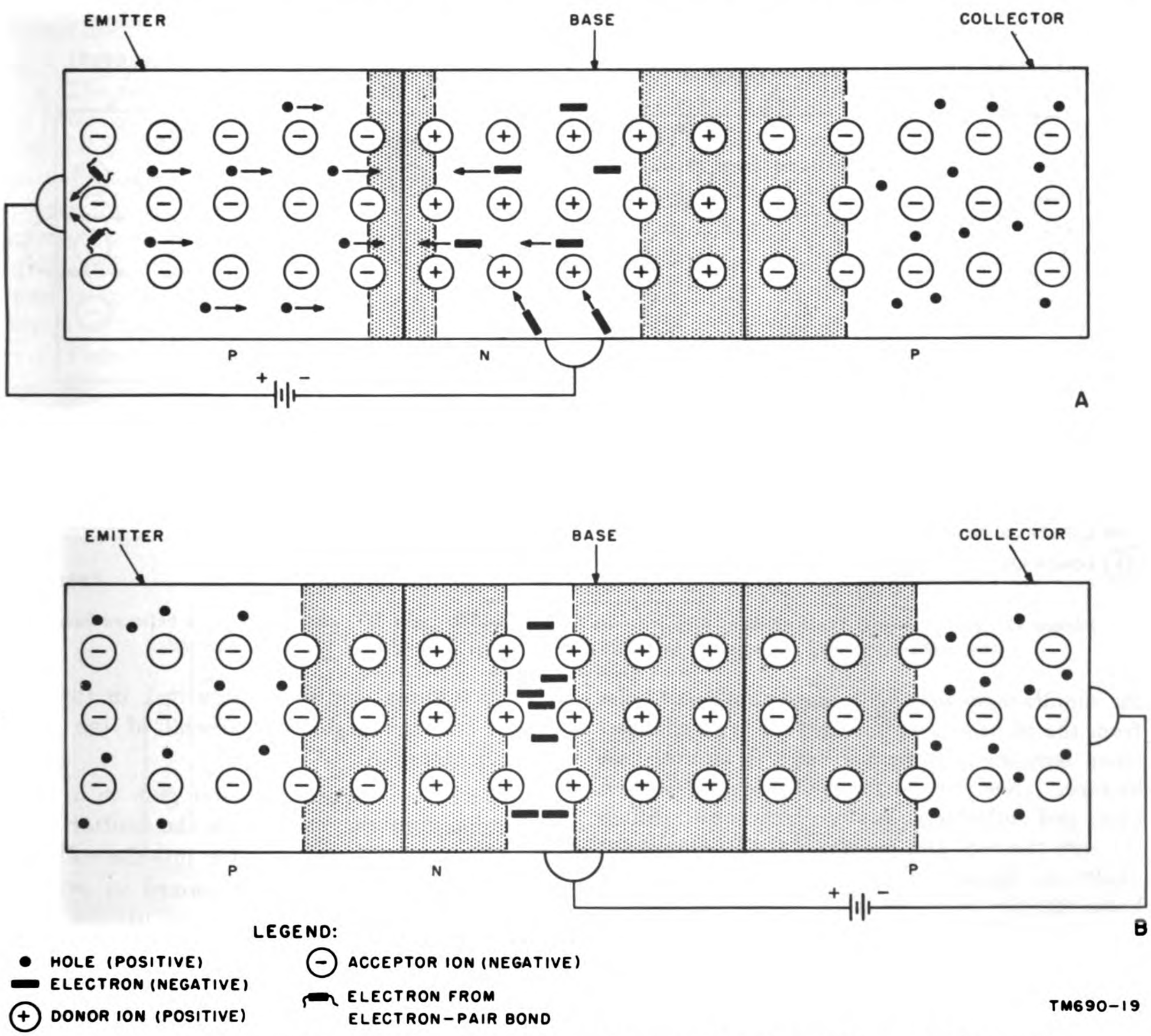


Figure 34. Forward bias between emitter and base (A); reverse bias between base and collector (B).

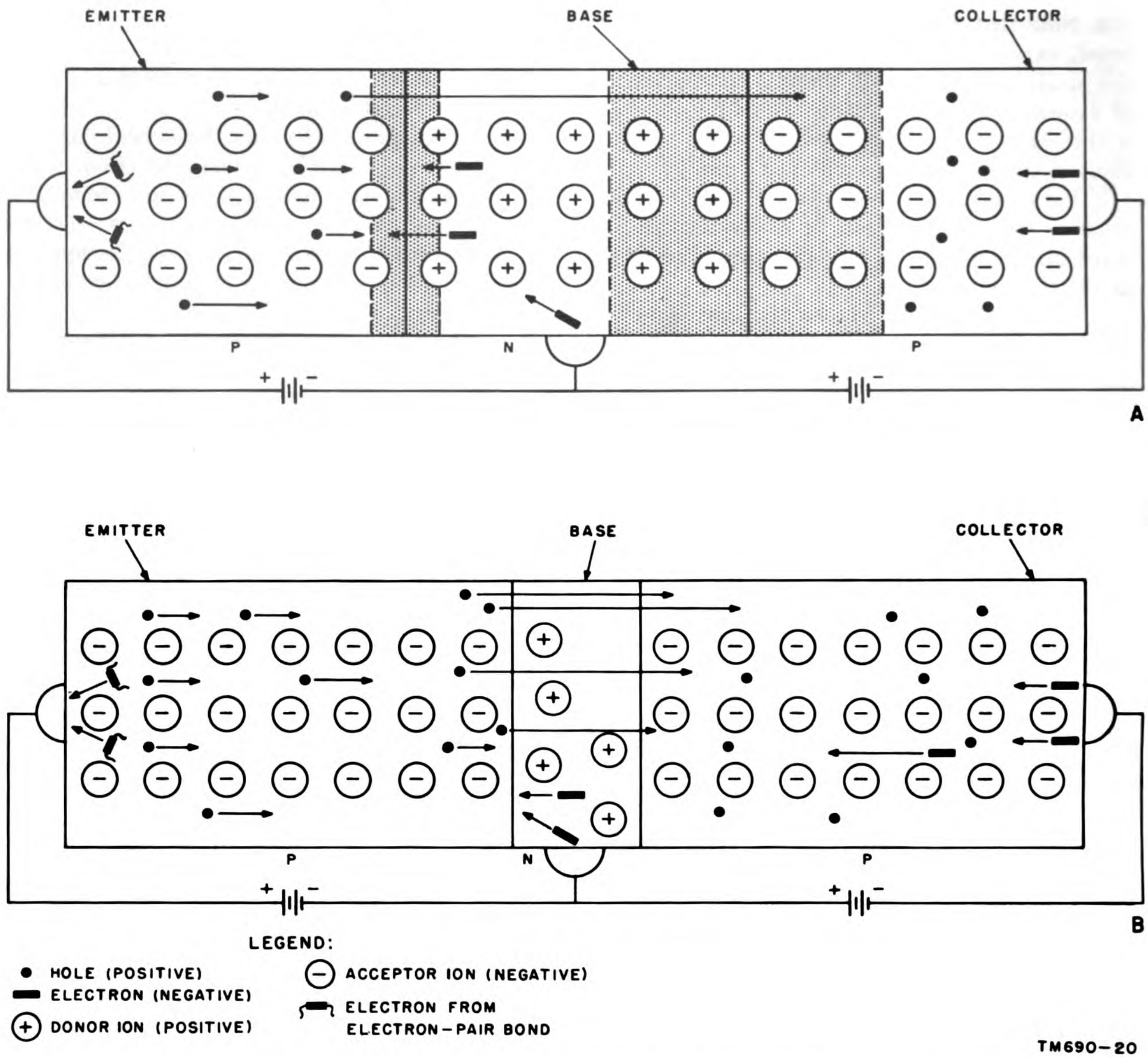


Figure 35. Simultaneous application of forward bias between emitter and base and reverse bias between base and collector of PNP transistor; theoretical transistor (A), and practical transistor (B).

the simultaneous biasing, a large number of holes from the emitter does not combine with the electrons entering the base from the emitter-base battery. (See par. 34 for definitions of emitter, base, and collector.) Many of the holes diffuse through the base and penetrate the base-collector depletion region. In the collector region the holes combine with electrons that enter the collector from the negative terminal of the base-collector battery. If the holes that enter the base from the emitter-base junction avoid combination with electrons entering the base from the battery, the holes are attracted to the collector by the acceptor ions (negative) in the collector and the negative potential of the base-collector battery.

e. To obtain maximum power gain in a transistor, most of the holes from the emitter must diffuse *through* the base region into the collector region. This condition is obtained in practice by making the base region very narrow compared to the emitter and the collector regions (B, fig. 35). In practical transistors, approximately 92 to 99 percent of the current from the emitter reaches the collector.

33. NPN Transistor

The theory of operation of the NPN transistor is similar to that of the PNP transistor (par. 32). However, inspection and comparison of figures 35 and 36 will reveal two important differences.

a. The emitter-to-collector carrier in the PNP transistor is the hole. The emitter-to-collector carrier in the NPN transistor is the electron.

b. The bias voltage polarities are reversed. This condition is necessitated by the different positional relationships of the two types of germanium as used in the two types of transistors.

34. Transistors and Electron Tubes

Some of the differences and similarities between electron tubes and transistors are discussed in *a* through *e* below.

a. The main current flow in an electron tube is from cathode to plate (fig. 37). In a junction transistor, the main current flow is from emitter to collector. The current in the electron tube passes through the grid. In the transistor, the current passes through the base. The cathode, grid, and plate of the electron tube are comparable to the emitter, base, and collector respectively of the transistor.

b. Plate current is determined mainly by grid-cathode voltage. Collector current is determined mainly by emitter-base voltage. In this respect, the collector current-collector voltage characteristic of a transistor with fixed emitter-base voltage (A, fig. 38) is similar to the plate current-plate voltage characteristic of the *pentode* tube with fixed grid-cathode voltage (B, fig. 38). In the transistor, the steady rise of current from O to X with a rise of collector voltage indicates that a sufficient supply of carriers is flowing from the emitter-base barrier through the base-collector barrier to satisfy the applied collector voltage. This section of the characteristic represents low collector resistance. From X to Y, the collector current remains relatively constant in spite of the rising collector voltage, because the supply of carriers through the emitter-base barrier does not increase. This section of the characteristic represents high collector resistance. In the pentode tube, the leveling off of plate current from X to Y is due to the shielding of the plate by the screen and the suppressor grids.

c. The electron tube requires heater current to *boil* electrons from the cathode. The transistor has no heater.

d. For current flow in an electron tube, the plate is always positive with respect to the cathode. For current flow in a transistor, the collector may be positive or negative with respect to the emitter, depending on whether electrons or holes respectively are the emitter-to-collector carriers.

e. For most electron tube applications, grid-cathode current does not flow. For most transistor applications, current flows between emitter and base. Thus, in these cases, the input resistance of an electron tube is much higher than

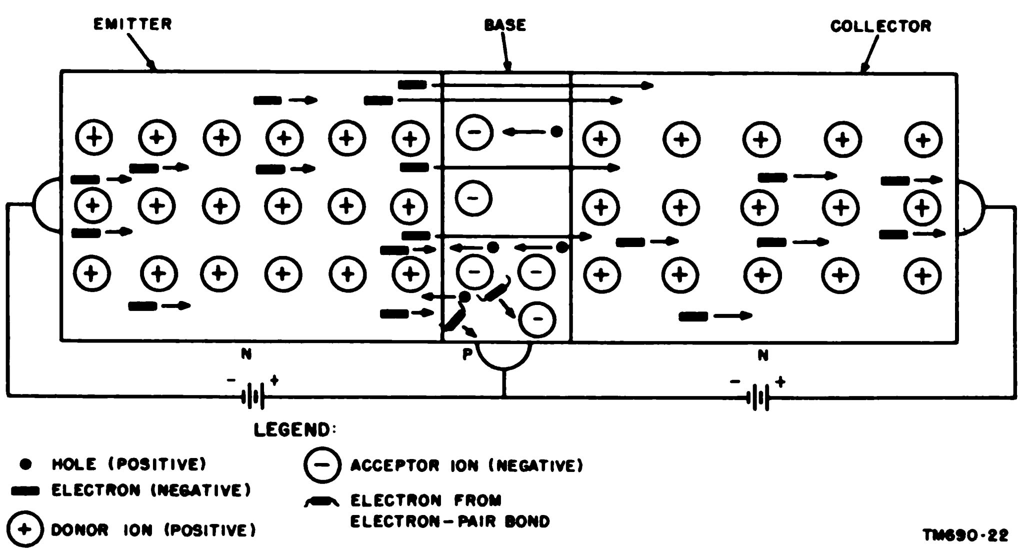
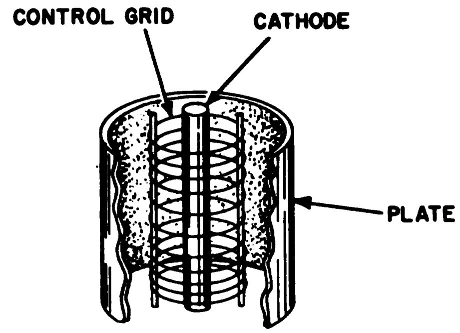
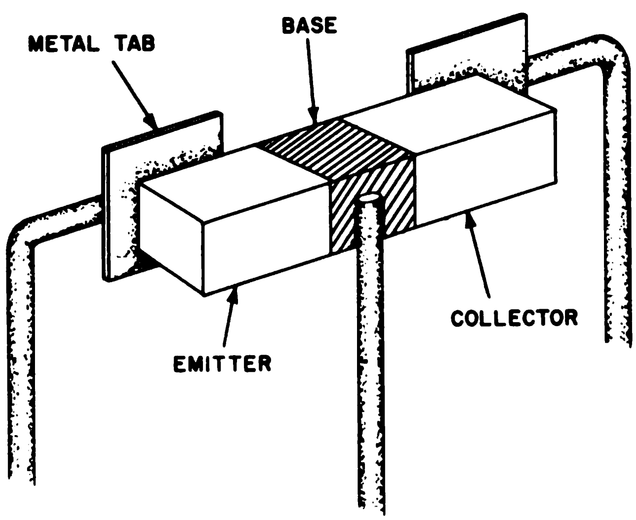


Figure 36. Simultaneous application of forward bias between emitter and base and reverse bias between base and collector of an NPN transistor.



A. TRIODE ELECTRON TUBE



B. JUNCTION TRANSISTOR

TM690-25

Figure 37. Internal structure of triode electron tube and junction transistor.

its output resistance and similarly the input resistance of a transistor is much lower than its output resistance.

35. Summary

a. All matter is composed of one or more elements.

b. The smallest part of an element which can take part in ordinary chemical changes is called an atom.

c. Atoms are composed of positively charged particles called protons, negatively charged particles called electrons, and uncharged particles called neutrons.

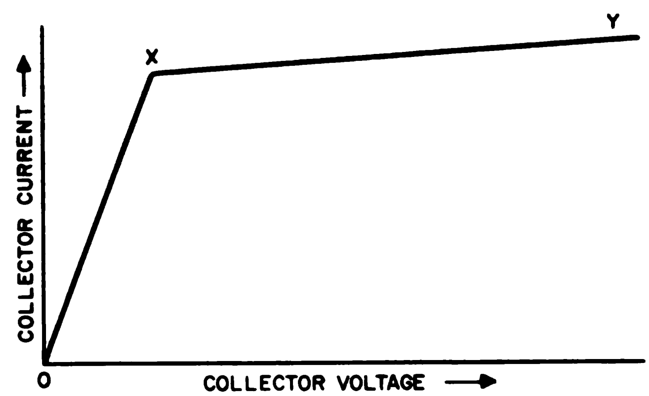
d. A conductor is a material that has many loosely held electrons. Examples are silver, copper, and aluminum.

e. An insulator is a material that has few loosely held electrons. Examples are rubber, glass, and porcelain.

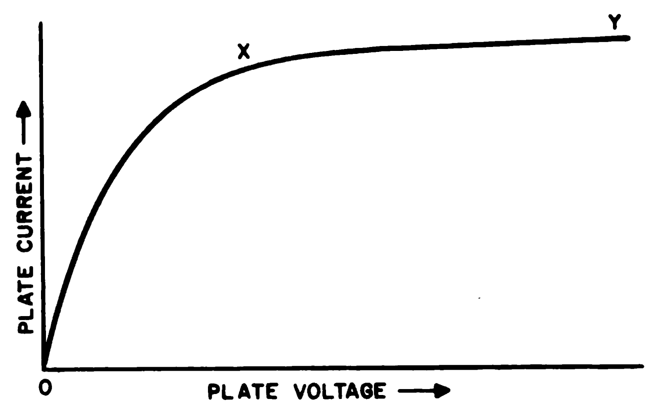
f. A semiconductor is a material the resistivity of which is between those of conductor and insulators. Examples are germanium, silicon, and carbon (in diamond form).

g. A crystal is a material with atoms arranged in a specific pattern.

h. The properties and characteristics of *polycrystalline* materials such as copper, and silver are quite different from those of *single* crystal



A. JUNCTION TRANSISTOR



B. PENTODE ELECTRON TUBE

TM690-26

Figure 38. Collector characteristic for junction transistor, and plate characteristic for pentode electron tube.

materials; single crystal materials are prepared for transistor use. Germanium and silicon may be processed as single crystal materials.

i. Electrons shared by adjacent atoms in a crystal form electron-pair bonds.

j. N-type germanium contains donor impurities. Donor impurities are materials which have *five* valence electrons, one of which cannot form an electron-pair bond. This electron is called an excess electron.

k. Arsenic, antimony, and boron are examples of donor materials.

l. P-type germanium contains acceptor impurities. Acceptor impurities are materials which have three valence electrons. Because *four* valence electrons are required to form and complete all adjacent electron-pair bonds, a hole is created.

m. Aluminum, gallium, and indium are examples of acceptor impurities.

n. A hole can be considered a positive charge which diffuses or drifts through a crystal. The drift of holes constitutes a current.

o. A depletion (space charge) region occurs at a PN junction. The potential difference across the depletion region is called a barrier. The width of the barrier is the width of the depletion region. The potential difference is called the height of the barrier.

p. Forward bias of a PN junction causes heavy current (flow of majority carriers). Reverse bias causes very low current (flow of minority carriers).

q. A junction transistor is a *sandwich* of P-, N-, and P-type germanium, or N-, P-, and N-type germanium.

r. Holes constitute the main current through the PNP transistor. Electrons constitute the main current through the NPN transistor.

s. The emitter, base, and collector of the transistor are comparable to the cathode, grid, and plate respectively of the vacuum tube.

t. The emitter-base junction is normally biased in the forward (low-resistance) direction.

u. The base-collector junction is normally biased in the reverse (high-resistance) direction.

v. Collector current depends on the emission of carriers from the emitter-base barrier.

CHAPTER 3

TRANSISTOR AMPLIFIER FUNDAMENTALS

36. Reference Designations and Graphical Symbols

The reference designations and graphical symbols for transistors are shown in B and C, figure 39. For comparison purposes, those of a triode electron tube are also included (A, fig. 39).

a. Reference Designations. The letter portion of the reference designation for an electron tube is *V*. The number portion may be any whole number. For the transistor, the letter portion of the reference designation is *Q* and the number portion may be any whole number.

b. Graphical Symbols.

(1) The three element (triode) electron tube requires only one graphical symbol (A, fig. 39), because conduction within an electron tube occurs only through the movement of electrons. As a result, the plate must be at a positive potential with respect to the cathode for current to flow. The electrons in the external circuit will flow only from plate to cathode.

(2) In the PNP transistor (B, fig. 39) the emitter to collector current carrier in the crystal is the hole (par. 32). For holes to flow internally from emitter to collector, the collector must be negative with respect to the emitter. In the external circuit, electrons flow from emitter (*opposite* to direction of the emitter arrow) to collector.

(3) In the NPN transistor (C, fig. 39), the emitter to collector current carrier in the crystal is the electron (par. 33). For electrons to flow internally from emitter to collector, the collector must be positive with respect to the emitter. In the external circuit, electrons flow from the collector to the emitter (*opposite* to direction of the emitter arrow).

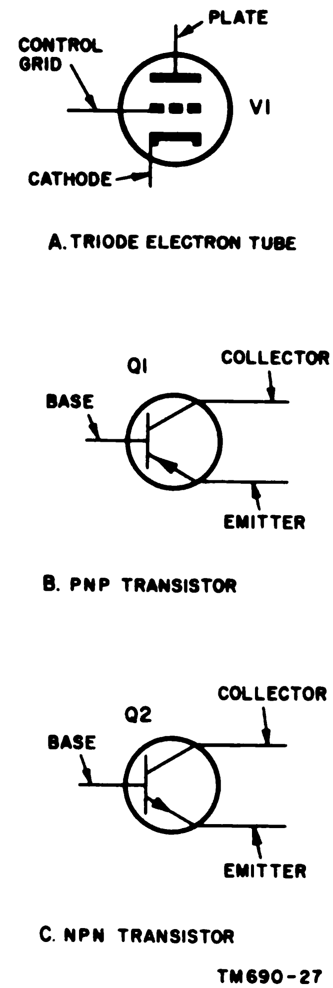


Figure 39. Transistor and triode electron tube reference designations and graphical symbols.

37. Transistor Amplifier Generalizations

The following generalizations are extremely helpful in analyzing the qualitative (nonmathematical) behavior of transistor circuitry. These generalizations apply to a transistor circuit that is operated as a *Class A* amplifier.

a. The first letter of the type of transistor indicates the polarity of the emitter voltage with *respect to the base.* A PNP transistor has positive dc voltage applied to the emitter. An NPN transistor has negative dc voltage applied to the emitter.

b. The second letter of the type of transistor indicates the polarity of the collector with respect to the base. A PNP transistor has negative dc voltage applied to the collector. An NPN transistor has positive dc voltage applied to the collector.

c. The first and second letter of the type of transistor indicate the relative polarities between the emitter and the collector respectively. In a PNP transistor, the emitter is positive with respect to the collector; the collector is negative with respect to the emitter. In an NPN transistor, the emitter is negative with respect to the collector; the collector is positive with respect to the emitter.

d. The dc electron current direction is always against the direction of the arrow on the emitter (B and C, fig. 39).

e. If the electrons flow into or out from the emitter, the electrons flow out from or into the collector, respectively.

f. The base-emitter junction is always forward biased.

g. The collector-base junction is always reversed biased.

h. An input voltage that aids (increases) the forward bias increases the emitter and collector currents.

i. An input voltage that opposes (decreases) the forward bias decreases the emitter and collector currents.

38. Transistor Arrangements

Several circuit arrangements for introducing a signal into the transistor and extracting the signal from the transistor are possible.

a. Common-Base Amplifier.

(1) In A, figure 40, the signal is introduced into the emitter-base circuit, and extracted from the collector-base circuit. The base element of the transistor is common to the input circuit and the output circuit. For this reason, the configuration is referred to as a *common-base (CB) amplifier.*

(2) In equipments employing electronic circuits, a large number of points may be at the same potential. To simplify the wiring, all these points are usually connected to one wire (a bus bar) or to the metallic chassis, which is referred to as *ground* or *chassis ground.* Usually the common element of a three-element device is at ground potential and is connected to chassis directly (dc and ac ground) or through a low-impedance capacitor (ac ground only). The common-base circuit therefore is also referred to as the *grounded-base amplifier.*

(3) A similar circuit arrangement employing the triode electron tube is shown in B, figure 40. This circuit is called the grounded-grid amplifier. Note that the comparable elements (par. 34) of the transistor and the electron vacuum tube assume the same relative positions within their respective circuits. The two circuits are similar to the extent that the input impedance is low and the output impedance is high. For typical values of input and output impedances and voltage, current and power gains of the CB amplifier, refer to paragraph 105.

b. Common-Emitter Amplifier.

(1) In C, figure 40, the signal is introduced into the base-emitter circuit and extracted from the collector-emitter circuit. The emitter element of the transistor is common to the input circuit and the output circuit; this configuration is referred to as the *common-emitter (CE) amplifier* or the *grounded-emitter amplifier.*

(2) A similar circuit arrangement employing the triode electron tube is shown in D, figure 40. Although this is a *common-* or *grounded*-cathode amplifier, it is not usually referred to as such. This circuit is considered the standard, or conventional, triode amplifier. Other circuit arrangements (B and F, fig. 40) were subsequently designed and given the names used in this manual. For detailed coverage of the conventional (grounded cathode) amplifier, refer to TM 11–662. For typical values of input and output impedances, voltage, current, and power gains of the common-emitter amplifier, refer to paragraph 105.

(3) If an NPN transistor were to be used in the circuit (C, fig. 40), the polarities of the biasing batteries would have to be opposite to those shown to maintain forward bias in the base-emitter circuit and reverse bias in the collector-base circuit.

c. Common-collector Amplifier.

(1) In E, figure 40, the signal is introduced into the base-collector circuit and extracted from the emitter-collector circuit. The collector element of the transistor is common to the input circuit and the output circuit; this configuration is referred to as the *common-collector (CC) amplifier* or the *grounded-collector amplifier*.

(2) A similar circuit arrangement employing the triode electron tube is shown in F, figure 40. This circuit is called the *cathode follower*. If the circuit does not look familiar, it is because it is drawn so that the comparable elements of the transistor and the electron tube are in the same relative positions in their respective circuits. For detailed coverage of the cathode follower, refer to TM 11–670. For typical values of input and output impedances, voltage, current, and power gains of the common-collector circuit, refer to paragraph 105.

39. Additional Bias Arrangements

Proper biasing of a class A transistor amplifier, regardless of the circuit configuration used (par. 38), consists of forward bias on the emitter-base circuit and reverse bias in the collector-base circuit. One method for achieving proper

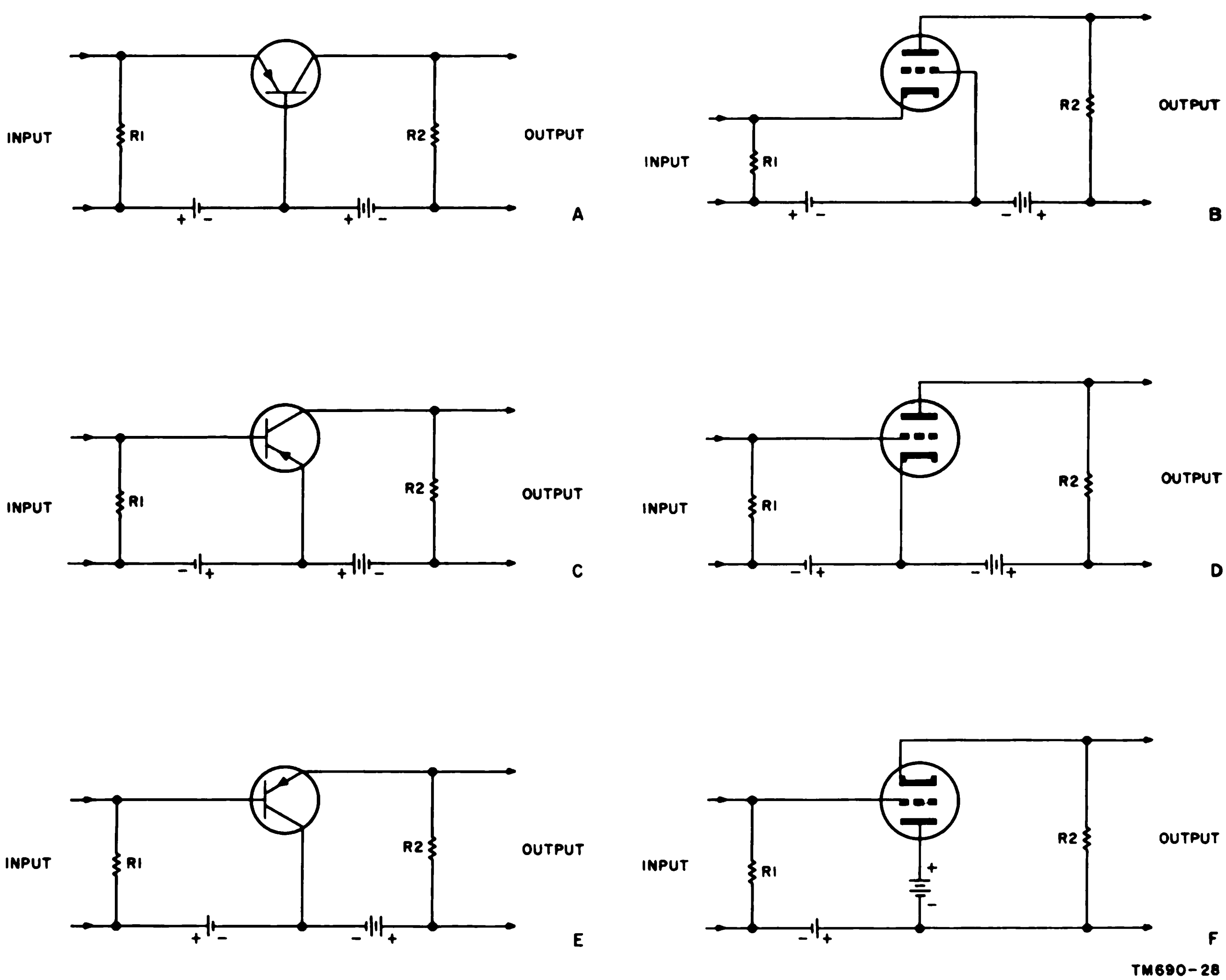


Figure 40. Basic amplifier configurations using transistors or triode vacuum tubes.

bias for each circuit configuration is shown in A, C, and E, figure 40. Other methods are discussed in *a*, *b*, and *c* below. A detailed study of bias considerations is covered in chapter 5.

a. Common-Base Amplifier. In order to operate the common-base amplifier with a single battery (fig. 41), a voltage-divider network is required. The collector-base bias is achieved directly by the battery in the collector-base circuit. Since the transistor shown is a PNP transistor, reverse bias is achieved by making the collector negative with respect to the base, as shown. Forward bias in the emitter-base circuit (PNP transistor) requires that the emitter be positive with respect to the base. This condition is achieved by the voltage divider network consisting of resistors R3 and R4. The electron current (*I*) from the battery and through the voltage divider is in the direction shown. This current flow causes a voltage drop across resistor R3 with polarity as indicated. This voltage drop places the emitter at a positive potential with respect to the base. If an NPN transistor were used, the battery only would have to be reversed.

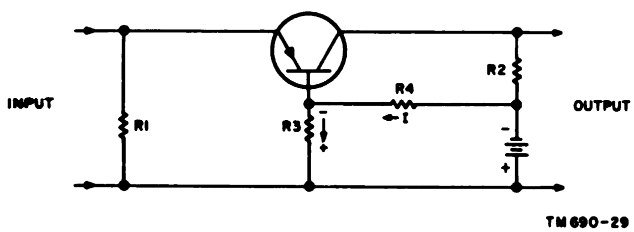


Figure 41. Common-base amplifier with single battery bias.

b. Common-Emitter Amplifier

(1) There are two methods of introducing bias voltages from separate batteries into the common-emitter amplifier. The first method (C, fig. 40) is used if it is desired to place the emitter at ac *and* dc ground potential in an equipment. The second method (A, fig. 42) is used if it is desired to have the base-emitter bias battery series aid the collector-emitter battery. If an NPN transistor were used, both batteries would have to be reversed in direction.

(2) The common emitter amplifier may also be biased with a single battery (B, fig. 42). The single battery directly produces the required reverse bias voltage in the collector-base circuit. To understand the method by which the forward bias between the emitter and the base is produced by the single battery, a knowledge of the internal structure of the transistor is required. It has been stated previously that forward bias for the PNP transistor requires the base to be negative with respect to the emitter. In a PNP transistor, the collector is at the highest negative potential; the emitter is at the highest positive potential. Structurally the base is between the two, and therefore must assume a voltage between the two. Thus the base must be less positive than the emitter or, in other words, negative with respect to the emitter. This condition satisfies the requirement of polarity necessary to produce a forward bias. Current flow through the base lead is covered in paragraph 43. The magnitude of the voltage between the emitter and the base must be very small compared to that between the collector and the base. Internally, the two PN junctions act as a voltage divider. The PN junction between the collector and the base represents a high resistance and develops the larger voltage drop. The PN junction between emitter and base represents a low resistance and develops a low volt-

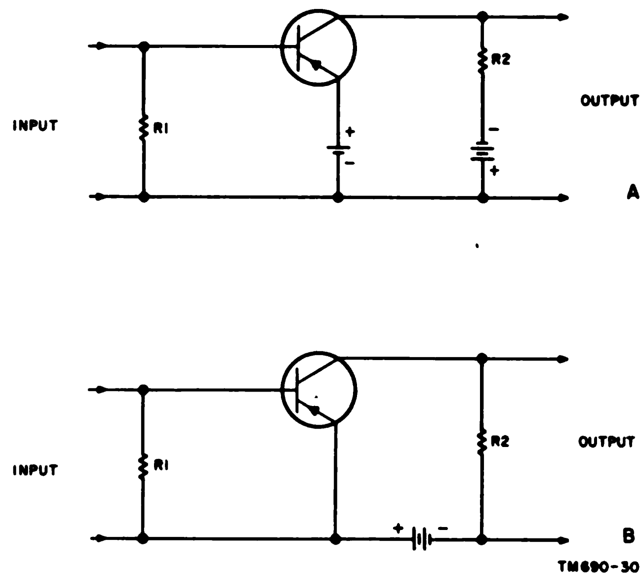


Figure 42. Common-emitter amplifier using two bias batteries, or a single bias battery.

age as required. Use of a single battery with *external* voltage dividers is covered in chapter 5.

c. Common-Collector Amplifier.

(1) There are three methods of introducing bias voltages from separate batteries into the common-collector amplifier. One method is shown in E, figure 40. A second and a third method are shown in A and B, figure 43. In each case the batteries establish the proper forward bias (base-emitter) and the reverse bias (base-collector). If an NPN transistor were to be substituted, the polarities of both batteries in each circuit would have to be reversed.

(2) The common-collector amplifier can also be biased with a single battery (C, fig. 43). The reverse bias (base-collector) is established directly by the battery. The forward bias (base-emitter) depends on the internal structure of the transistor. The internal mechanism for establishing the forward bias is the same as that covered in *b*(2) above.

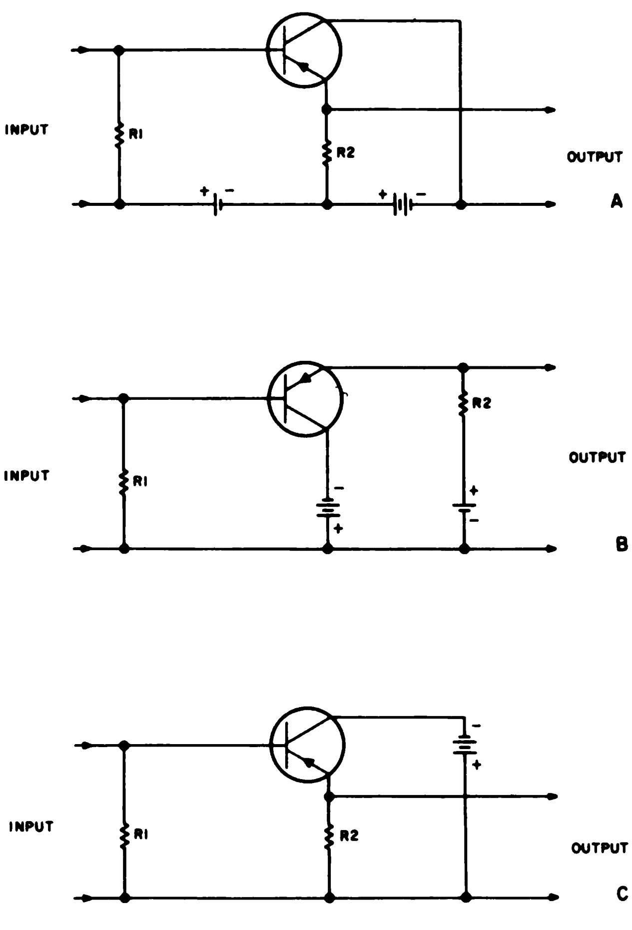


Figure 43. Common-collector amplifiers using two bias batteries or a single bias battery.

40. Current Flow and Voltage Phase Relations, CB Amplifier

a. Current Flow. Electron current flow through an NPN CB amplifier is indicated by the direction of the arrows in figure 44. To simplify the circuit explanation, the flow of the minority carriers across the reverse-biased base collector junction has been ignored. In paragraph 34*a* it was stated that most of the current from the emitter flows toward the collector. In practical transistors, from 92 to 98 percent of the emitter current reaches the collector; the remainder flows through the base. In the figure, the total emitter current is represented by the letter *I*. For discussion purposes it is assumed that 95 percent (or 0.95*I*) of the current reaches the collector; 5 percent (or 0.05*I*) flows to the base. If a PNP transistor were to be used, the polarities of the batteries would have to be reversed, but the voltage phase relationship would be the same.

b. Voltage Phase Relations.

(1) The wave forms on the illustration represent *voltage* wave forms. The input signal produced by the signal generator is on the left and the output signal developed across resistor R1 is on the right. Consider an instant of time when the voltage (*AB*) from the signal generator *opposes* the forward bias produced by the base-emitter battery. The resultant forward bias at this instant has been reduced, thereby reducing the total current (*I*) flowing through the emitter. By corresponding amounts, the collector and the base currents have been reduced. The reduced current flow through resistor R1 causes the top point of the resistor to become less negative (or more positive) with respect to the lower point . This effect is shown by AB on the output wave form. For the entire half cycle that the input signal *goes positive* and *opposes* the forward bias, the output signal goes positive.

(2) Consider a second instant of time when the voltage (*CD*) from the input signal *aids* the forward bias produced by the

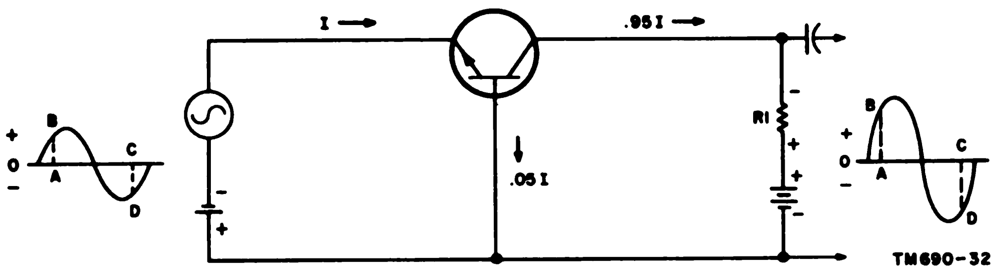


Figure 44. Common base (CB) amplifier, current flow and voltage wave forms.

base-emitter battery. The resultant forward bias at this instant has been increased, thereby increasing the total current (I) flowing through the emitter. By corresponding amounts, the collector and the base currents have been increased. The increased current flow through resistor R1 causes the top point of the resistor to become more negative (less positive) with respect to the lower point. This effect is shown by (CD) on the output wave form. For the entire half cycle that the input signal *goes negative* and *aids* the forward bias, the output signal goes negative.

(3) From the above discussion, it can be concluded that there is no voltage phase reversal between the input and the output signal of a CB amplifier.

41. Current Flow and Voltage Phase Relations, CE Amplifier

a. Current Flow. The electron current flow through an NPN CE amplifier is indicated by the direction of the arrows in figure 45. To simplify the circuit explanation, the flow of the minority carriers across the reverse-biased base-collector junction has been ignored. The portions of emitter current that flow through the collector and the base were discussed in paragraph 40*a*. If a PNP transistor were to be used, the polarities of the batteries would have to be reversed but the phase relationship would be the same.

b. Voltage Phase Relations.

(1) Consider an instant of time when the input voltage (AB) from the signal generator *aids* the forward bias produced by the base-emitter battery. The resultant forward voltage at this instant is increased, thereby increasing the total current (I) flowing through the emitter. By corresponding amounts the collector and the base currents are increased. The increased current flow through load resistor R1, causes the top part of the resistor to become more negative (less positive) with respect to the lower part. This effect is shown by AB on the output waveform. For the entire half cycle that the input signal goes *positive* and *aids* the forward bias, the output signal goes negative.

(2) Consider a second instant of time when the voltage (CD) from the signal generator *opposes* the forward bias produced by the emitter-base battery. The resultant forward voltage is decreased, thereby decreasing the emitter current. By corresponding amounts the collector and the base currents are decreased. The decreased current through load resistor R1 causes the top point of the resistor to become less negative (or more positive) with respect to the lower point.

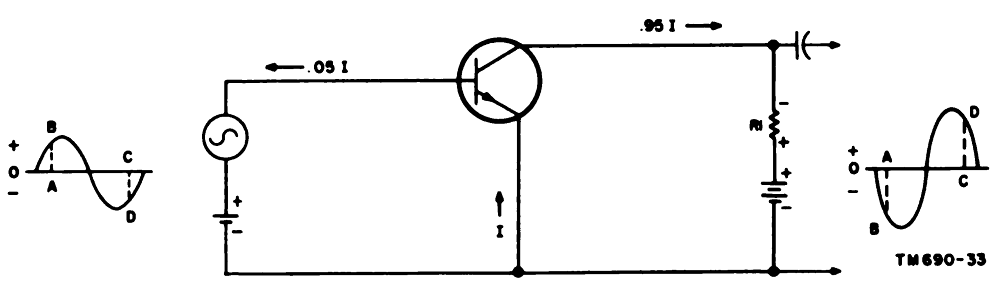


Figure 45. Common-emitter (CE) amplifier, current flow and voltage wave forms.

This effect is shown by CD on the output waveform. For the entire half cycle that the input signal *goes negative* and *opposes* the forward bias, the output signal goes positive.

(3) From the above discussion, it can be concluded that the input signal voltage is reversed 180° in phase in going through the CE amplifier.

42. Current Flow and Voltage Phase Relations, CC Amplifier

a. Current Flow. The electron current flow through an NPN CC amplifier is indicated by the direction of the arrows in figure 46. To simplify the circuit explanation the flow of minority carriers across the reverse-biased collector-base junction has been ignored. The portions of emitter current that flow through the collector and the base were discussed in paragraph 40*a*. If a PNP transistor were to be used, the polarities of the batteries would have to be reversed but the voltage phase relationship would be the same.

b. Voltage Phase Relations.

(1) Consider an instant of time when the input voltage (*AB*) from the signal generator *aids* the forward bias voltage. The resultant forward voltage at this instant increases, thereby increasing the total emitter current (*I*). The increased current through load resistor R1 causes the top point of the resistor to become more positive with respect to the lower part. This effect is shown by *AB* on the output waveform. For the entire half cycle that the input signal is positive, the output signal is also positive.

(2) When the input signal is negative (*CD*), the forward bias is reduced and the emitter current is correspondingly reduced. The reduction in emitter current causes the top point of load resistor R1 to become less positive (more negative) with respect to the lower part. This effect is shown by *CD* on the output waveform. For the entire half cycle that the input signal is negative, the output signal is also negative.

(3) From the above discussion, it can be concluded that there is no phase reversal of the signal amplified by the CC amplifier.

43. Base Lead Current

a. General.

(1) In figures 44, 45, and 46 the electron current in the base lead for an NPN transistor is shown to flow away from the base region. This direction of electron current flow may, or may not, be the case for an NPN transistor employed in a specific circuit application. Figures 44, 45, and 46 neglect the reverse current flow between the collector and the base (par. 30).

(2) For a PNP transistor, the electron current flow in the base lead would be toward the base region at all times, if the reverse bias collector-base current were to be ignored.

(3) Actually for either type transistor, the direction of electron current flow in the base region depends on the percentage of emitter current that enters the base lead and the magnitude of the reverse bias collector-base current that enters the base lead. These values in turn depend on the operating point of the transistor and the characteristics of the specific transistor. A discussion of the

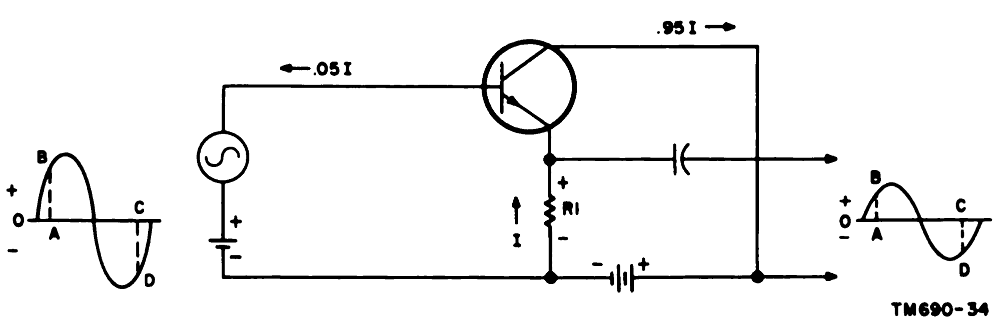


Figure 46. Common-collector (CC) amplifier, current flow and voltage waveforms.

current flow in an NPN transistor is given in *c* below; the PNP transistor is discussed in *d* below. The letter symbols used in the discussion are explained in *b* below.

b. Letter Symbols.

(1) The current gain of the CB amplifier is the ratio of the collector current to the emitter current. This value is important in transistor electronics and is referred to as alpha$_{fb}$ (α_{fb}). With *f* being the current from the emitter to the collector (in a *f*orward direction), and *b* indicating that a common-*b*ase amplifier is being used. The exact formula for and mathematical derivation of α_{fb} is given in paragraph 62*c*(3). Roughly, however, α_{fb} represents the percentage of emitter current that reaches the collector.

(2) The total emitter current is designated I_e.

(3) The reverse bias current flow between base and collector (par. 74*a*), also referred to as *saturation* or *leakage* current, is designated I_{CBO}. The subscript letters indicate a *c*ollector-*b*ase current, measured with the emitter *o*pen.

c. NPN Transistor.

(1) Figure 47 shows an NPN transistor with proper biases applied. To the minority carriers (electrons) in the base region (par. 30), the base region acts as N-type germanium. This is indicated by the circled *N*. To the minority carriers (holes) in the collector region, the collector region acts as P-type germanium. This is indicated by the circled *P*.

(2) The emitter current (I_e) consists of electrons; 92 to 99 percent of the electrons ($\alpha_{fb}\ I_e$) go through the base region and the collector region into the positive terminal of the collector-base battery. About 1 to 8 percent of the emitter electrons ($I_e-\alpha_{fb}\ I_e$) (total emitter current less current that flows to collector) combine in the base-emitter junction with holes from the base region. These holes are generated in the base region and near the base lead. Generation of the holes causes an electron current to flow in the base lead and *away from the base.*

(3) In addition, reverse-bias current I_{CBO} flows between base and collector. I_{CBO} consists of a flow of electrons in the base region that combine with holes from the collector region at the base-collector junction. The reverse bias current causes a flow of electrons in the base lead and *toward the base.*

(4) The direction of electron flow ((2) and (3) above), in the base lead depends on the relative magnitudes of ($I_e-\alpha_{fb}I_e$) and I_{CBO}. For clarification, study the following examples:

Example 1:

Given: $I_e=1$ ma
$I_{CBO}=0.01$ ma
$\alpha_{fb}=0.92$

Determine emitter-base current:

$$I_e-\alpha_{fb}I_e=1-0.92\times 1$$
$$=0.08 \text{ ma}$$

Conclusion: The emitter-base current (0.08 ma) is larger than the collector-base current ($I_{CBO}=0.01$ ma), so that the base lead current (fig. 47), 0.07 ma, is *flowing away from the base.*

Example 2:

Given: $I_e=1$ ma
$I_{CBO}=0.02$ ma
$\alpha_{fb}=0.99$

Determine emitter-base current:

$$I_e-\alpha_{fb}I_e=1-0.99\times 1$$
$$=0.01 \text{ ma}$$

Conclusion: The collector-base current ($I_{CBO}=0.02$ ma) is larger than the emitter-base current (0.01 ma) so that the base lead current (fig. 47) 0.01 ma, is *flowing toward the base.*

d. PNP Transistor. The discussion in *c* above, applies also to the PNP transistor (fig. 48). Internally, however, the electron and the hole currents are interchanged. Externally the polarity of the biases and the directions of electron current flow are reversed.

44. Summary

a. Because of the existence of PNP and NPN transistors, a graphical symbol is required to represent each type transistor.

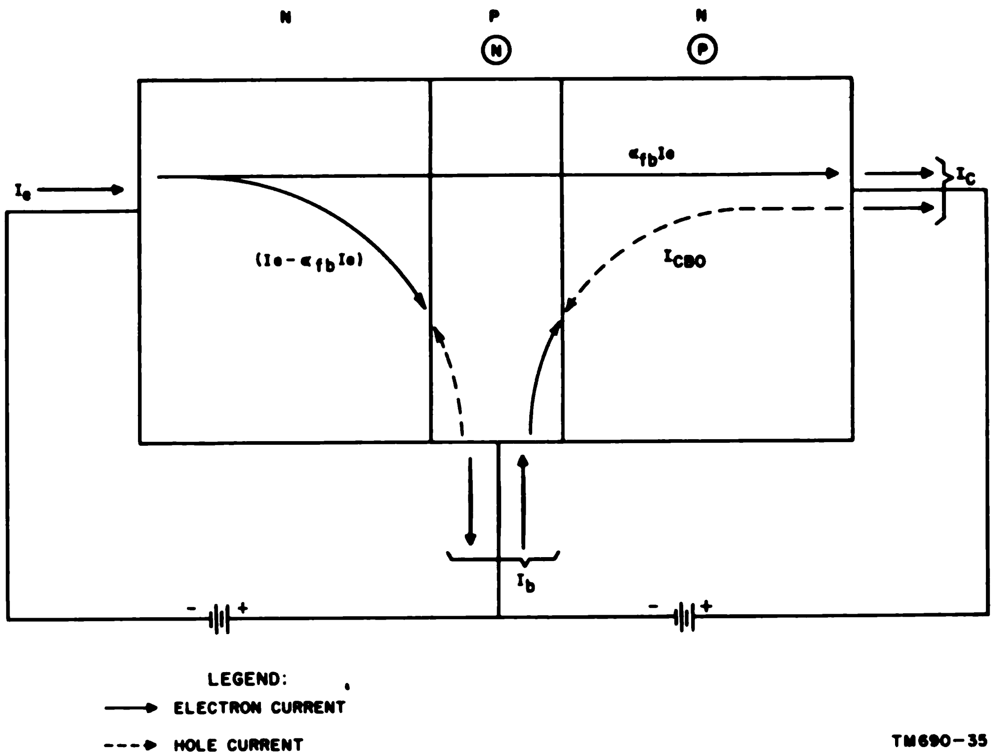


Figure 47. Current flow in NPN transistor.

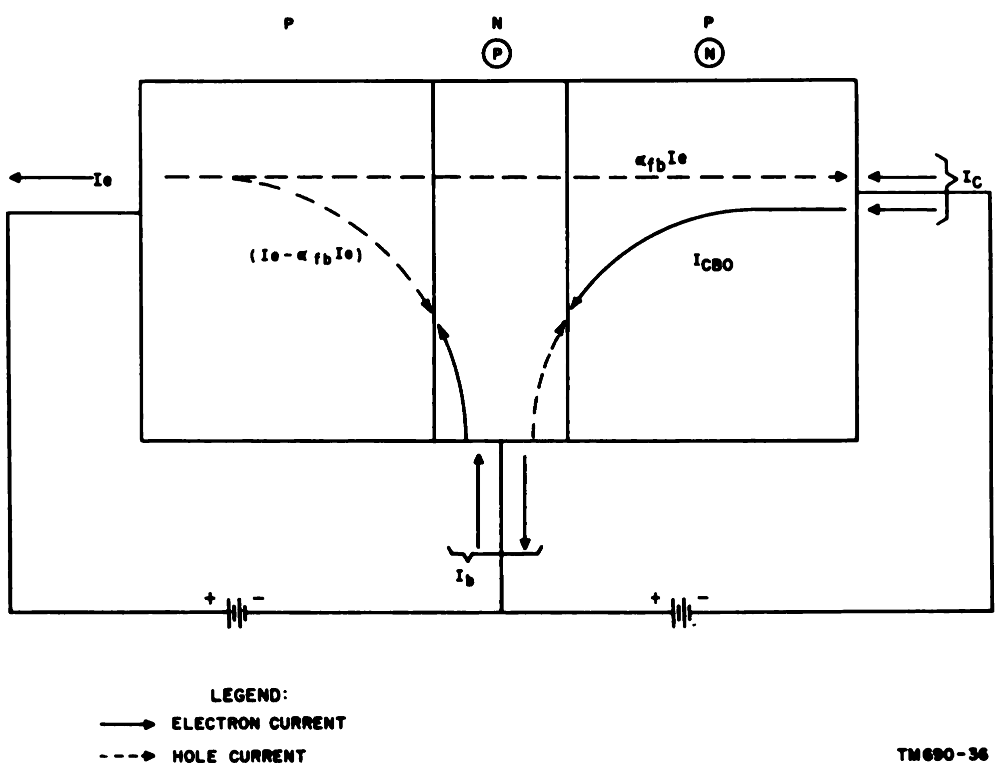


Figure 48. Current flow in PNP transistor.

b. The transistor can be connected as a common base, a common collector, or a common-emitter amplifier.

c. The circuit configuration of the common emitter amplifier is similar to that of the conventional (common-cathode) amplifier.

d. The circuit configuration of the common-base amplifier is similar to that of the grounded (common) grid amplifier.

e. The circuit configuration of the common-collector amplifier is similar to that of the cathode follower (common plate) amplifier.

f. Each transistor configuration may be biased with two batteries or a single battery.

g. There is no voltage phase reversal of a signal amplified by a common-base or a common-collector amplifier.

h. There is a 180° voltage phase reversal of a signal amplified by a common-emitter amplifier.

i. The dc electron current through the base lead of a transistor (NPN or PNP) may flow toward or away from the base region, depending on the relative magnitudes of the base-emitter current ($I_e - \alpha_{fb} I_e$) and the saturation current (I_{CBO}).

CHAPTER 4

PARAMETERS, EQUIVALENT CIRCUITS, AND CHARACTERISTIC CURVES

Note. **The mathematical analysis of the transistor as contained in this chapter requires the reader to have a knowledge of algebraic equations. Chapters 5 through 13 may be comprehended without covering chapter 4.**

Section I. INTRODUCTION

45. General

a. The circuit behavior of any electrical element or device can be expressed in terms of mathematical equations. The mathematical equations, however, only can show relationships among currents, voltages, resistances, inductances, and capacitances. These quantities, or dimensions, including their mathematical combinations, can be measured directly or indirectly and are called parameters. Under most operating conditions, a circuit element or device such as a resistor, a capacitor, an inductor, a current generator, or a voltage generator exhibits a single characteristic and may be represented in an equation by a single parameter. A circuit element or device such as a quartz crystal, an electron tube, or a transistor (A, C, or E, fig. 49) is more complex (exhibits characteristics of more than one parameter) and must be represented by an equivalent circuit (B, D, or F, fig. 49) which may consist of resistance, capacitance, inductance, and possibly voltage, and current generators.

b. Equivalent circuits of quartz crystals, electron tubes, and transistors are discussed in (1), (2), and (3) below.

(1) An equivalent circuit of a quartz crystal is shown in B, figure 49. The effect of the quartz crystal in a given circuit is that of two capacitors, an inductor, and a resistor connected as shown. Once the equivalent circuit has been determined, the behavior of the crystal over a wide range of applied voltages at varying frequencies can be computed by expressing the relationships of the parameters (within the dashed line) in mathematical equations. Such computations permit the prediction of the behavior of the crystal in a particular circuit without actually constructing the circuit.

(2) The equivalent circuit for an electron tube is shown in D, figure 49. This equivalent circuit shows that an electron tube can be represented by a voltage generator (μv_g) in series with a resistor (r_p). The procedure for developing this equivalent circuit for the electron tube is covered in TM 11–662. Using this equivalent circuit, it is possible to compute the input and output resistances, the voltage, current, and power gains, under many operating conditions without making actual measurements for each operating condition.

(3) An equivalent circuit for the transistor is shown in F, figure 49. This equivalent circuit indicates that the transistor can be represented by two resistors, a voltage generator, and a current generator, connected as indicated. The procedure for developing this equivalent circuit is covered in paragraphs 47 through 53. Using the equivalent circuit, the formulas for calculation of input and output resistances, and current, voltage, and power gain may be developed (pars. 54–58).

46. CE Amplifier

a. In paragraph 38*b* the transistor CE amplifier was compared to the electron tube common-

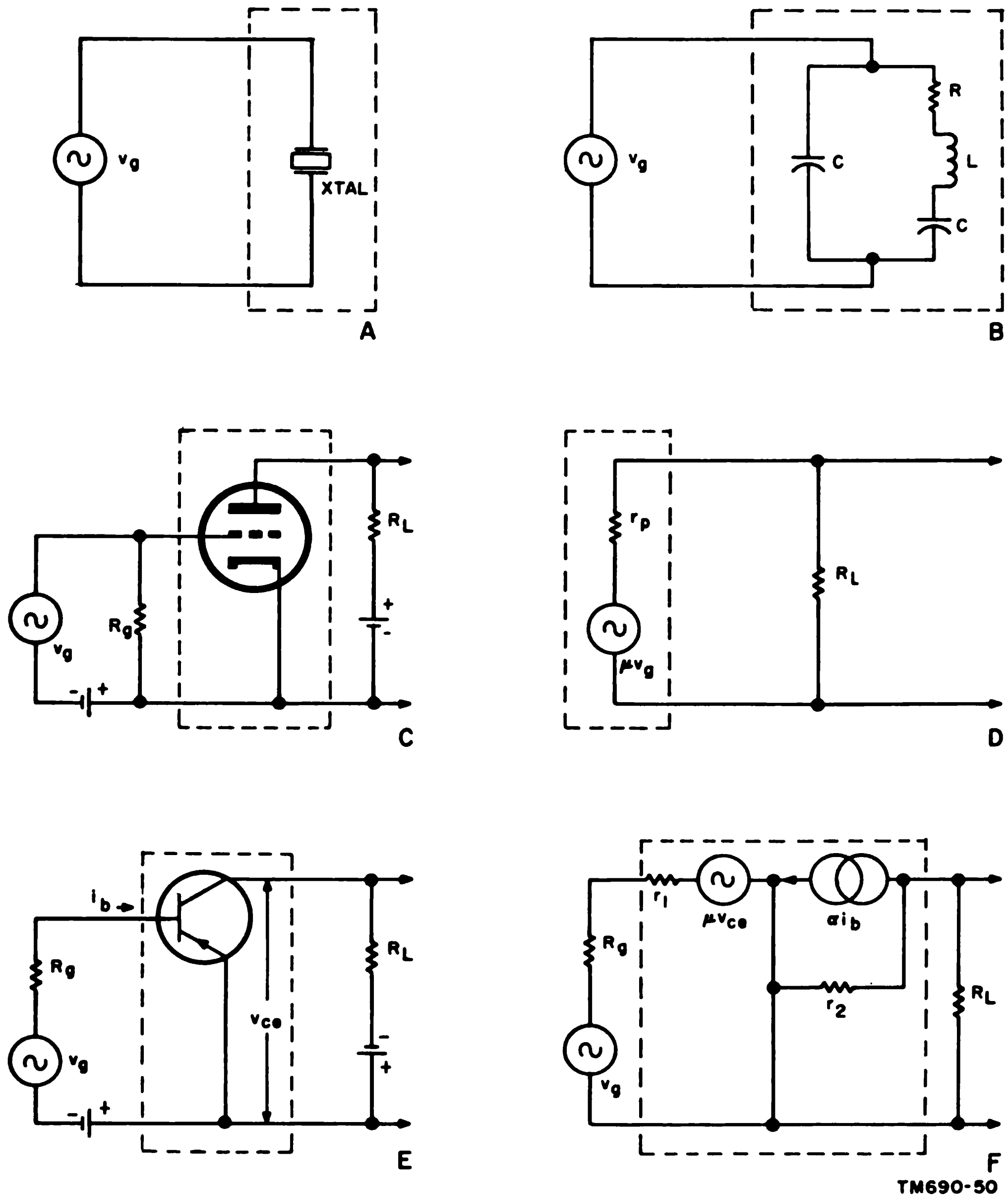


Figure 49. Crystal, electron tube, and transistor circuits and their equivalent circuits.

cathode (conventional) amplifier. In equipments containing electron tubes, the conventional amplifier is most often used. In equipments containing transistors, the CE amplifier is most often used because it provides current, voltage, and power gain while the CB amplifier current gain is always less than 1 and the CC amplifier voltage gain is always less than 1. The CB amplifier is used most often to match a low-impedance circuit to a high-impedance circuit. The CC amplifier is used most often to match a high-impedance circuit to a low-impedance circuit.

b. Because the CE amplifier is most often used, the procedure for the development of its equivalent circuit is covered in detail in paragraphs 49 through 53. The equivalent circuits of the CB and CC amplifiers and the resultant formulas for voltage, current, and power gain and input and output resistances are explained briefly in paragraphs 61 through 64.

47. Choice of Dependent and Independent Variables

If a device, for which an equivalent circuit must be developed, is considered as an unknown object, such as a black box (fig. 50), the only information available would be the currents (i_b, i_c) and voltages (v_{be}, v_{ce}) measured at the input and output terminals. Before equations can be established to show the relationships among the currents and voltages, it must be decided which quantities shall be considered dependent variables and which quantities, independent variables. Any two quantities may be designated as dependent variables with the other two becoming independent variables. The choice is arbitrary, but each choice results in different equivalent circuits. This means that there are several possible combinations of circuit elements (resistors, capacitors, inductors) and devices (generators) which can electrically represent the unknown object. Regardless of the chosen equivalent circuit, the same results are obtained for the input and output resistances, and current, voltage and power gain.

Note. **Lower-case letters are used for ac current and voltage values; upper-case letters are used for dc current and voltage values.**

a. When the two voltages (v_{be}, v_{ce}) are chosen as the dependent variables, and the two currents (i_b, i_c) are chosen as the independent variables, the following general equations can be written:

$$v_{be}=f_1\ (i_b,\ i_c) \text{ and}$$
$$v_{ce}=f_2\ (i_b,\ i_c)$$

The first equation states that base voltage v_{be} (dependent variable) is determined by (or (f) is a function of) base current i_b and collector current i_c (independent variables). The second equation states that collector voltage v_{ce} (dependent variable) is determined by base current i_b, and collector current i_c (independent variables). This choice of dependent and independent variables results in an equivalent circuit whose parameters are called *open-circuit parameters*. This equivalent circuit is discussed in paragraphs 68 and 69.

b. When the two currents (i_b, i_c) are chosen as the dependent variables, and the two voltages (v_{be}, v_{ce}) are chosen as the independent variables, the following general equations can be written:

$$i_b=f_1\ (v_{be},\ v_{ce}) \text{ and}$$
$$i_c=f_2\ (v_{be},\ v_{ce}).$$

These equations are interpreted in the same manner as those in *a* above. This choice of dependent and independent variables results in an equivalent circuit whose parameters are called *short-circuit parameters*. This equivalent circuit is discussed in paragraphs 70 and 71.

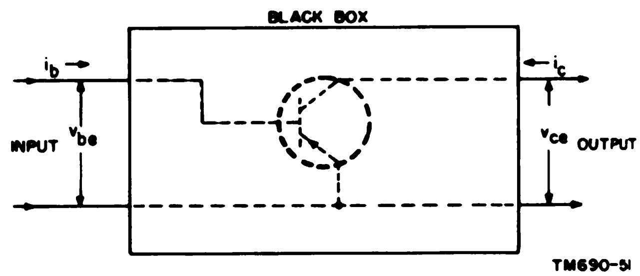


Figure 50. ***Black box containing device for which equivalent circuit is to be developed.***

c. When base voltage v_{be} and collector current i_c are chosen as the dependent variables, and base current i_b and collector voltage v_{ce} are chosen as the independent variables, the following general equations can be written:

$$v_{be}=f_1\ (i_b,\ v_{ce}) \text{ and}$$
$$i_c=f_2\ (i_b,\ v_{ce})$$

These equations are interpreted in the same manner as those in *a* above. This choice of dependent and independent variables results in an equivalent circuit whose parameters are called *hybrid parameters*. The reason for the term *hybrid parameters* is covered in paragraph 53*a*. Since this equivalent circuit is in most common use, its development will be discussed in detail in paragraphs 49 through 58.

d. Other choices of dependent and independent variables are theoretically possible, but have no practical application and are not discussed.

48. Base Current and Collector Voltage as Independent Variables

a. The most commonly used equivalent circuit (par. 47*c*) is that using hybrid parameters. Hybrid parameters may be developed by considering the base current and the collector voltage as the independent variables. Some practical considerations (*b* and *c* below), of the transistor indicate that base current and collector voltage are most easily controlled externally and would therefore naturally be selected as the independent variables. When these two quantities are determined, then the dependent variables, base voltage, and collector current, may then be established.

b. Consider the input circuit of an electron tube (where the grid voltage is the independent variable) as opposed to the input circuit of a transistor (where the base-emitter current is the independent variable):

(1) In an electron tube, the diode formed by the grid and the cathode is actually a reverse-biased diode. Because of the high resistance, the grid voltage may be easily measured and controlled. The input resistance of an electron tube is usually much greater than the resistance of the signal source, so that the signal source appears as a *constant voltage source*. Grid current rarely flows and when it does it is usually negligible in magnitude.

(2) The input circuit of a transistor (base-emitter for all configurations) is a forward-biased diode. Because of the low resistance, the base-emitter current is easily measured and controlled. The input resistance of the transistor is usually much smaller than the resistance of the signal source, so that the signal source appears as a *constant current source*. Base-emitter voltage is usually very small, measured in tenths of a volt. Under these conditions, consideration of the base-emitter current as an independent variable is desirable.

c. In the collector circuit, the collector voltage is considered the independent variable; the collector-base circuit exhibits a high resistance. Because of this high resistance, the collector voltage is easily measured and controlled. The collector base resistance is much higher than the internal resistance of the power source feeding the collector, so that the power source appears as a constant voltage generator. Under these conditions, consideration of the collector voltage as an independent variable is desirable.

49. Development of Input Circuit Equation

a. In paragraph 47*c*, the general equation relating base voltage v_{be} (dependent variable) to base current i_b and collector voltage v_{ce} (independent variables) was expressed as follows:

$$v_{be}=f_1\ (i_b,\ v_{ce})$$

Since base voltage v_{be} is the input voltage (fig. 50), this equation is the input circuit equation. The expansion and application of this equation are discussed in *b* and *c* below:

b. Expansion of the equation (*a* above), follows:

(1) Determine the effect of each of the independent variables (i_b, v_{ce}) on the dependent variable (v_{be}). Then, add the separate effects to obtain the total effect of the independent variables on the dependent variable.

(2) This procedure is comparable to determining the speed of a ship operating under its own power in a current (moving body) of water. The total speed (dependent variable) of the ship is the sum of the speed of the ship due to its own power (first independent variable) and the speed of the ship due to the current (second independent variable). To calculate the speed of the ship, the speed of the ship operating under its own power is still water (second independent variable=0) and the speed of the ship in the moving body of water with its power turned off (first independent variable=0) must be known. The speed of the ship will then be the sum of these two values.

(3) Similarly in the case of the general equation (*a* above), the effect of the ac base current (first independent variable) on the ac base voltage (dependent variable) must be obtained while the collector voltage (second independent variable) is kept at zero ($v_{ce}=0$). This is discussed in (4) below. Next, the effect of the ac collector voltage (second independent variable) on the ac base voltage (dependent variable) must be obtained while the ac base current (first independent variable) is kept at zero ($i_b=0$). This is discussed in (5) below.

Note. In practice the ac voltage across the transistor input and output terminals is made zero, while maintaining a dc bias voltage, by placing a low-reactance capacitor across the terminals. The ac current to the transistor input or output terminals is made zero, while maintaining a dc bias current, by placing a high-reactance coil in series with the terminals.

(4) The effect of base current (i_b) on base voltage (v_{be}) while collector voltage (v_{ce}) is at zero ($v_{ce}=0$) can be determined by use of the circuit shown in A, figure 51. To keep ac collector voltage at zero, the output circuit is ac short-circuited. A current generator varies base current i_b (measured by ammeter I), while the variations of base voltage v_{be} are measured by voltmeter V. The ratio of base voltage v_{be} to base current i_b while collector voltage is zero ($v_{ce}=0$) is expressed as follows:

$$\left(\frac{v_{be}}{i_b}\right)_{v_{ce}=0}$$

Once this ratio is obtained, the effect of a given value of base current on base voltage can be expressed as the product of this ratio and the given value of base current:

$$\left(\frac{v_{be}}{i_b}\right)_{v_{ce}=0} i_b$$

This quantity forms the first term of the expanded equation (c below), of the general equation (a above). The numerical values of the ratio of base voltage to base current while collector voltage is constant ($v_{ce}=0$), are obtained from characteristic curves (par. 51).

Note. This ratio and subsequent ratios of ac currents and voltages used in this chapter are valid for ***small signals only***. This limitation is required because of the nonlinear current and voltage characteristics of the transistor. That is, the ratio is valid only over a *linear portion* of the characteristic curve under consideration. The greater the nonlinearity the smaller the signal required.

(5) The effect of collector voltage (v_{ce}) on base voltage (v_{be}) while base current (i_b) is kept at zero ($i_b=0$) can be determined by use of the circuit shown in B, figure 51. To keep the ac base current at zero, the input circuit is ac open-circuited. A voltage generator varies collector voltage v_{ce} (measured by voltmeter $V2$), while the variations of base voltage v_{be} are measured by voltmeter $V1$. The ratio of base voltage v_{be} to collector voltage v_{ce} while base current is zero ($i_b=0$) is expressed as follows:

$$\left(\frac{v_{be}}{v_{ce}}\right)_{i_b=0}$$

Once this ratio is obtained, the effect of a given value of collector voltage on base voltage can be expressed as the product of this ratio and the given value of collector voltage:

$$\left(\frac{v_{be}}{v_{ce}}\right)_{i_b=0} v_{ce}$$

This quantity forms the second term of the expanded equation (c below), of the general equation (a above). The numerical values of the ratio of base voltage to collector voltage while base current is zero ($i_b=0$) are obtained from characteristic curves (par. 51).

c. Using the terms developed in *b* (4) and (5) above, the general equation (a above), of the input circuit can be expressed as follows:

$$v_{be}=\left(\frac{v_{be}}{i_b}\right)_{v_{ce}=0} i_b+\left(\frac{v_{be}}{v_{ce}}\right)_{i_b=0} v_{ce}$$

Paragraph 53 explains the use of this equation in determining the equivalent circuit of the transistor.

50. Development of Output Circuit Equation

a. In paragraph 47*c*, the general equation relating collector current i_c (dependent variable) to base current i_b and collector voltage v_{ce} (independent variables) was expressed as follows:

$$i_c=f_2(i_b,v_{ce})$$

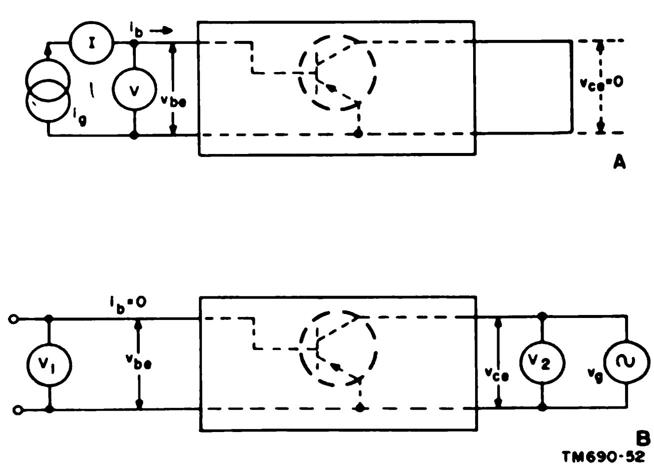


Figure 51. Circuit (black box) arrangements for measuring effect of base current (i_b) and collector voltage (v_{ce}) on base voltage (v_{be}).

Since collector current i_c is the output current (fig. 50), this equation is the output circuit equation. The expansion of this equation is discussed in *b* and *c* below.

b. As in the case of expanding the input circuit equation (par. 49), the technique employed is to determine the effect of each of the independent variables (i_b, v_{ce}) on the dependent variable (i_c).

(1) The effect of base current (i_b) on collector current (i_c), while collector voltage (v_{ce}) is zero ($v_{ce}=0$), can be determined by use of the circuit shown in A, figure 52. To keep the ac collector voltage zero, the output circuit is ac short-circuited (ammeter *I*2 has zero resistance). A current generator varies base current i_b (measured by ammeter *I*1), while the variations of collector current i_c are measured by ammeter *I*2. The ratio of collector current i_c to base current i_b while collector voltage is zero ($v_{ce}=0$) is expressed as follows:

$$\left(\frac{i_c}{i_b}\right)_{v_{ce}=0}$$

Once this ratio is obtained, the effect of a given value of base current on collector current can be expressed as the product of this ratio and the given value of base current:

$$\left(\frac{i_c}{i_b}\right)_{v_{ce}=0} i_b$$

This quantity forms the first term of the expanded equation (*c* below), of the general equation (*a* above). The numerical values of the ratio of collector current with base current while collector voltage is zero ($v_{ce}=0$) are obtained from characteristic curves (par. 51).

(2) The effect of collector voltage (v_{ce}) on collector current (i_c) while base current (i_b) is zero ($i_b=0$) can be determined by use of the circuit shown in B, figure 52. To keep the base current zero ($i_b=0$), the input circuit is ac open-circuited. A voltage generator varies collector voltage v_{ce} (measured by voltmeter *V*) while variations of collector current (i_c)

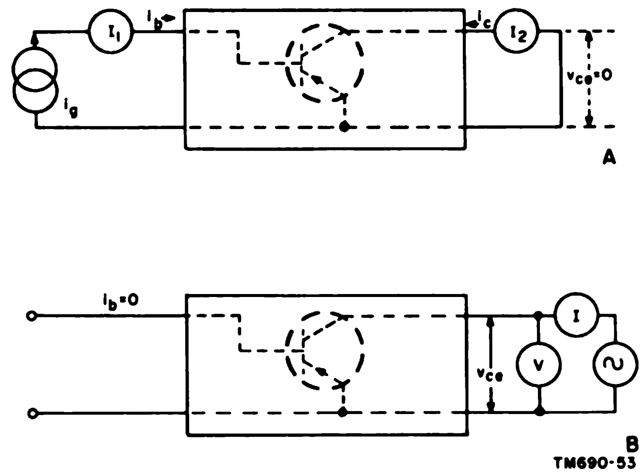


Figure 52. Circuit (black box) arrangements for measuring effect of base current (i_b) *and collector voltage* (v_{ce}) *on collector current* (i_c).

are measured by ammeter *I*. The ratio of collector current (i_c) to collector voltage v_{ce} while base current is zero ($i_b=0$) is expressed as follows:

$$\left(\frac{i_c}{v_{ce}}\right)_{i_b=0}$$

Once this ratio is obtained, the effect of a given value of collector voltage on collector current can be expressed as the product of this ratio and the given value of collector voltage:

$$\left(\frac{i_c}{v_{ce}}\right)_{i_b=0} v_{ce}$$

This quantity forms the second term of the expanded equation (*c* below), of the general equation (*a* above). The numerical values of the ratio of collector current to collector voltage while base current is zero ($i_b=0$) are obtained from characteristic curves (par. 51).

c. Using the terms developed in *b* (1) and (2) above, the general equation (*a* above), can be expressed as follows:

$$i_c=\left(\frac{i_c}{i_b}\right)_{v_{ce}=0} i_b+\left(\frac{i_c}{v_{ce}}\right)_{i_b=0} v_{ce}$$

Use of this equation for determining the equivalent circuit of the transistor is covered in paragraph 53.

51. Static Characteristic Curves, Input Circuit

The coefficients of the terms in the input circuit equation (par. 49*c*) are obtained from static characteristic curves. The static characteristic curves are plotted from dc values of current and voltage. The *form* of the static characteristic curves is determined by the particular coefficient, the value of which must be obtained. That is—the *ordinate* (vertical plot) is the dependent variable; the *abscissa* (horizontal plot) is the independent variable; and the *running parameter*, for a family of curves, is the independent variable that is kept constant.

a. First Term af Input Circuit Equation.

(1) The coefficient of the first term of the input circuit equation (par. 49*c*) is the ratio of base voltage to base current while collector voltage is zero

$$\left(\frac{v_{be}}{i_b}\right)_{v_{ce}=0}.$$

The input (base) static characteristic curves of a typical transistor for this coefficient are shown in figure 53. Note that the base voltage (dependent variable) is plotted vertically, the base current (independent variable) is plotted horizontally, and the collector voltage (independent variable) is the *running* parameter (the fixed dc value at which each curve is obtained). The ratio of base voltage to base current at particular ac values of base current and collector voltage is the slope of the curve at those particular values. Since the slope varies, the slope (the coefficient) can be determined at a particular point by taking incremental (small values) of base current and base voltage. Incremental values are indicated by the use of the Greek letter delta (Δ) before the current or voltage symbol. Expressing this concept mathematically, the following statement can be made:

$$\left(\frac{v_{be}}{i_b}\right)_{v_{ce}=0}=\left(\frac{\Delta V_{BE}}{\Delta I_B}\right)_{V_{CE}=C}.$$

The left hand side of the equation represents small ac values, the right hand side, incremental dc values. The ac values must be small to insure operation on the straight-line portion of the characteristic curve. $V_{CE}=C$ indicates that the dc collector voltage is kept constant. The expression on the right hand side of the equation can be directly related to the static characteristic curves. The expression is a ratio of voltage to current and represents a *resistance*. This resistance is called the *input resistance with output ac short-circuited* and is designated h_{ie}; *h* for hybrid, *i* for input, and *e* for common emitter configuration; thus—

$$h_{ie}=\left(\frac{\Delta V_{BE}}{\Delta I_B}\right)_{V_{CE}=C}$$

(2) To calculate h_{ie} for a particular transistor, the input characteristics for that transistor may be used (fig. 53). Since the value of h_{ie} depends on the operating point (dc bias), it is necessary to first establish the dc bias point. The dc bias conditions are established by specifying the dc values of the independent variables (base current and collector voltage). Assume in this case that it is desired to operate with a base current of 150 microamperes (μa) and a collector voltage of 7.5 volts. This point is located and marked X on the 7.5-volt input characteristic curve. A right-angle triangle (ABC) is drawn so that point X is approximately at the center of the hypotenuse of the triangle. The slope of the curve at point X is equal to $\frac{BC}{AC}$. Line BC, however, equals ΔV_{BE} and line AC equals ΔI_B; so that:

$$h_{ie}=\frac{BC}{AC}$$

$$=\left(\frac{\Delta V_{BE}}{\Delta I_B}\right)_{V_{CE}=C}$$

To determine ΔV_{BE}, the value of base voltage at point C (0.23 volt) must be subtracted from the value of base voltage at point B (0.3 volt); thus—

$$\Delta V_{BE}=0.3-0.23=0.07 \text{ volt}.$$

To determine ΔI_B, the value of base current at point A ($95\mu a$) must be subtracted from the value of base current at point C ($210\mu a$); thus—

$$\Delta I_B = 210 - 95$$
$$= 115\mu a$$
$$= 115 \times 10^{-6} \text{ amp}$$

Placing these values in the formula for h_{ie}:

$$h_{ie} = \left(\frac{\Delta V_{BE}}{\Delta I_B}\right)_{V_{CE}=C}$$
$$= \frac{0.07}{115 \times 10^{-6}}$$
$$= 608 \text{ ohms}$$

This is a typical value for h_{ie}. Depending on the transistor, h_{ie} may vary from 300 to 800 ohms.

b. Second Term of Input Circuit Equation.

(1) The coefficient of the second term of the input circuit equation (par. 49*c*) is the ratio of base voltage to collector voltage while ac base current is zero $\left(\frac{v_{be}}{v_{ce}}\right)_{i_b=0}$. The static characteristic curves of a typical transistor for this coefficient are shown in figure 54. Note that base voltage (dependent variable) is plotted vertically, collector voltage (independent variable) horizontally, and base current (constant independent variable) is the *running* parameter. The value of the coefficient is the slope of the curve at a given point. Since the slope is not constant, incremental values of current and voltage must be obtained. Equating the coefficient $\left(\frac{v_{be}}{v_{ce}}\right)_{i_b=0}$ to the incremental values, the following is obtained:

$$\left(\frac{v_{be}}{v_{ce}}\right)_{i_b=0} = \left(\frac{V_{BE}}{V_{CE}}\right)_{I_B=C}$$

Note that $I_B = C$ indicates that the dc base current is kept constant. The expression on the right hand side can be directly related to the static characteristic curves. The expression is the ratio of two voltages, and is therefore a pure number. Since the expression indicates the effect of the output voltage on the input voltage, it is referred to as the

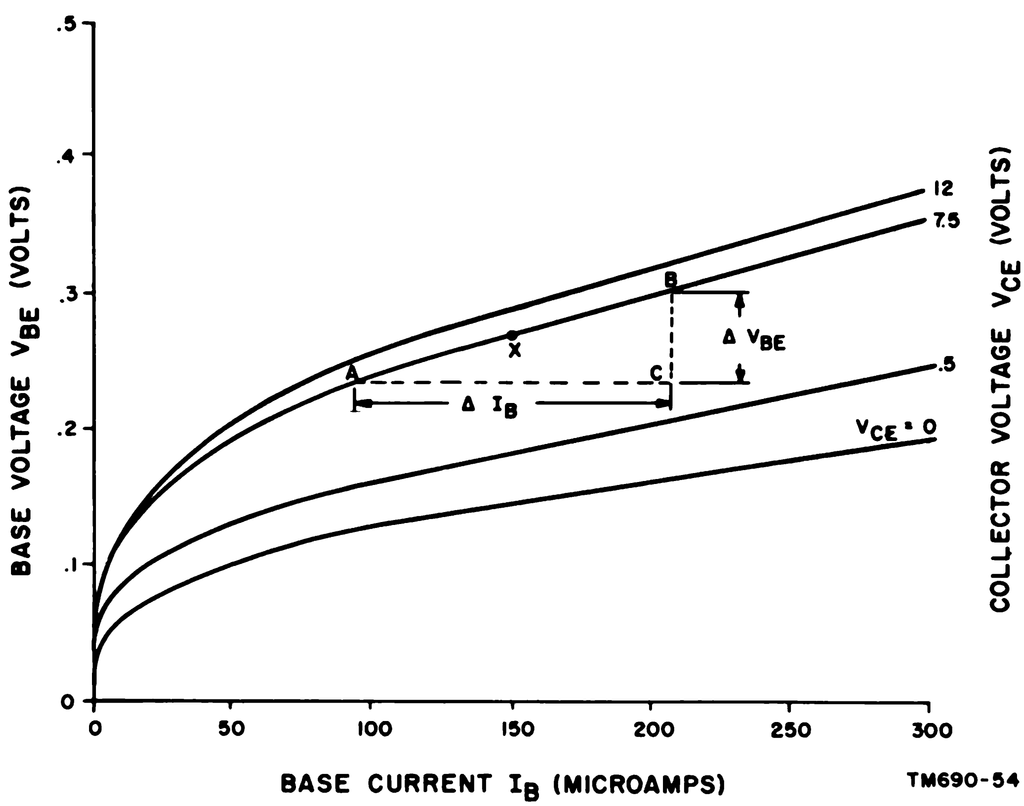


Figure 53. Typical input (base) static characteristic curves for common-emitter configuration of hybrid equivalent circuit.

reverse (feedback) open circuited voltage amplification factor, and is designated μ_{re}; μ for voltage amplification (as in the case of electron tubes), *r* for reverse, and *e* for common emitter; thus—

$$\mu_{re}=\left(\frac{\Delta V_{BE}}{\Delta V_{CE}}\right)_{I_B=C}$$

(2) To calculate μ_{re} for a particular transistor, the reverse-transfer characteristics for that transistor may be used (fig. 54). Assume dc bias conditions of 150 μa of base current and 7.5 volts collector voltage. This point is marked X on the appropriate curve. A right-angle triangle is drawn so that point X is near the center of the hypotenuse. The slope of the curve at point X is equal to $\frac{BC}{AC}$. Line BC equals ΔV_{BE} and line AC equals ΔV_{CE}, so that—

$$\mu_{re}=\frac{BC}{AC}$$

$$=\left(\frac{\Delta V_{BE}}{\Delta V_{CE}}\right)_{I_B=C}$$

To determine ΔV_{BE}, the value of base voltage at point C (0.22 volt) must be subtracted from the value of base voltage at point B (0.23 volt); thus—

$$\Delta V_{BE}=0.23-0.22=0.01 \text{ volt.}$$

To determine ΔV_{CE}, the value of collector voltage at point A (3 volts) must be subtracted from the value of collector voltage at point C (10 volts); thus—

$$\Delta V_{CE}=10-3$$

$$=7 \text{ volts.}$$

Placing these values in the formula for μ_{re}:

$$\mu_{re}=\left(\frac{\Delta V_{BE}}{\Delta V_{CE}}\right)_{I_B=C}$$

$$=\frac{0.01}{7}$$

$$=0.0014$$

This is a typical value for μ_{re}. Depending on the transistor, μ_{re} may vary from 0.0010 to 0.0030.

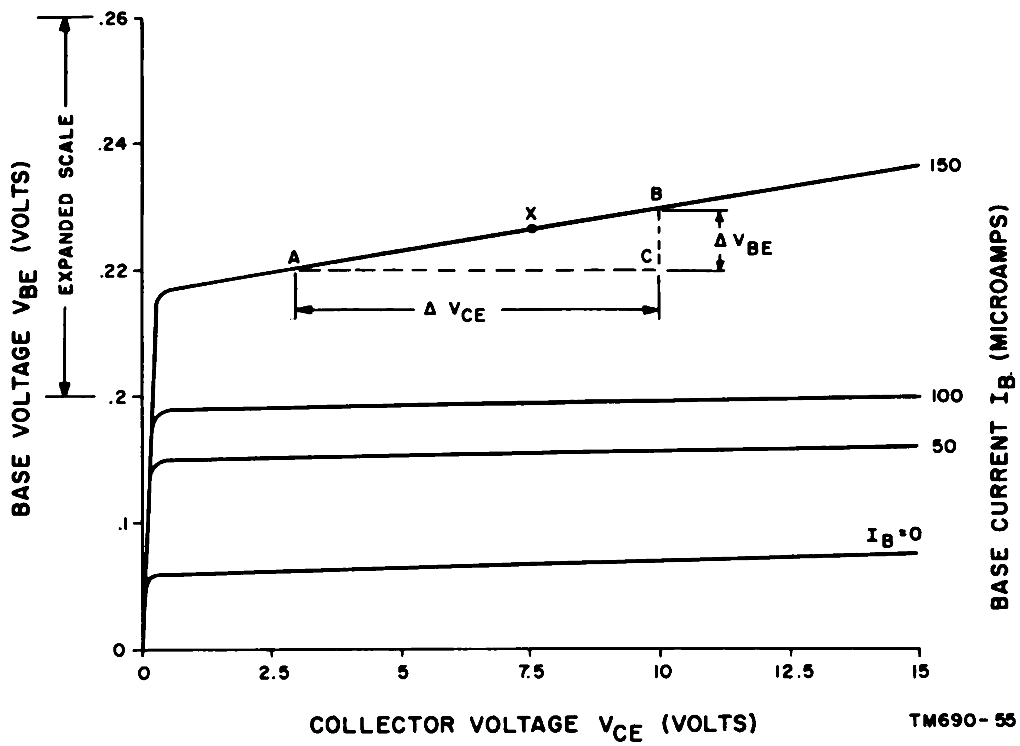


Figure 54. Typical reverse transfer (feedback) static characteristic curves for common-emitter configuration of hybrid equivalent circuit.

52. Static Characteristic Curves, Output Circuit

a. First Term of Output Circuit Equation.

(1) The coefficient of the first term of the output circuit equation (par. 50*c*) is the ratio of collector current to base current while collector voltage is zero, $\left(\frac{i_c}{i_b}\right)_{v_{ce}=0}$. The static characteristic curves of a typical transistor for this coefficient are shown in figure 55. Note that collector current is plotted vertically, base current horizontally, and collector voltage is the running parameter. The value of the coefficient is the slope of the curve at a given point. Since the slope is not constant, incremental values of current and voltage must be used. Equating the coefficient $\left(\frac{i_c}{i_b}\right)_{v_{ce}=0}$ to the incremental values, the following is obtained:

$$\left(\frac{i_c}{i_b}\right)_{v_{ce}=0}=\left(\frac{\Delta I_C}{\Delta I_B}\right)_{V_{CE}=C}$$

Note that $V_{CE}=C$ indicates that the dc collector voltage is kept constant. The expression on the right hand side can be directly related to the static characteristic curves. The expression is the ratio of output current to input current. The expression is called *the forward short-circuit current amplification factor* and is designated α_{fe}; α is arbitrarily used for current amplification, *f* for forward (effect of input on output), and *e* for common emitter; thus—

$$\alpha_{fe}=\left(\frac{\Delta I_C}{\Delta I_B}\right)_{V_{CE}=C}$$

(2) To calculate α_{fe} for a particular transistor, the forward transfer characteristics for that transistor may be used (fig. 55). Assume dc bias conditions of 150 μa of base current and 7.5 volts of collector voltage. This point is marked X on the appropriate curve. A right-angle triangle is drawn so that point X is near the center of the hypotenuse. The slope of the curve at point X is equal to $\frac{BC}{AC}$. Line BC equals ΔI_C and line AC equals ΔI_B, so that—

$$\alpha_{fe}=\frac{BC}{AC}$$

$$=\left(\frac{\Delta I_C}{\Delta I_B}\right)_{V_{CE}=C}$$

To determine ΔI_C, the value of collector current at point C (4 ma) must be subtracted from the value of collector current at point B (5.5 ma); thus—

$$\Delta I_C=5.5-4$$
$$=1.5 \text{ ma}$$
$$=1.5\times10^{-3} \text{ amp}$$

To determine the value of ΔI_B, the value of base current at point A (120 μa) must be subtracted from the value of base current at point C (175 μa); thus—

$$I_B=175-120$$
$$=55\ \mu a$$
$$=55\times10^{-6} \text{ amp}$$

Placing these values in the formula for α_{fe}:

$$\alpha_{fe}=\left(\frac{\Delta I_C}{\Delta I_B}\right)_{V_{CE}=C}$$

$$=\frac{1.5\times10^{-3}}{55\times10^{-6}}$$

$$=27$$

This is a typical value for α_{fe}. Depending on the transistor, α_{fe} may vary from 20 to 80.

b. Second Term of Output Circuit Equation.

(1) The coefficient of the second term of the output circuit equation (par. 50*c*) is the ratio of collector current to collector voltage while ac base current is zero, $\left(\frac{i_c}{v_{ce}}\right)_{i_b=0}$. The static characteristic curves of a typical transistor for this coefficient is shown in figure 56. Note that collector current is plotted vertically, collector voltage horizontally, and base current is the running parameter. Since the slope (value of coefficient) of the curve is not constant, incremental values

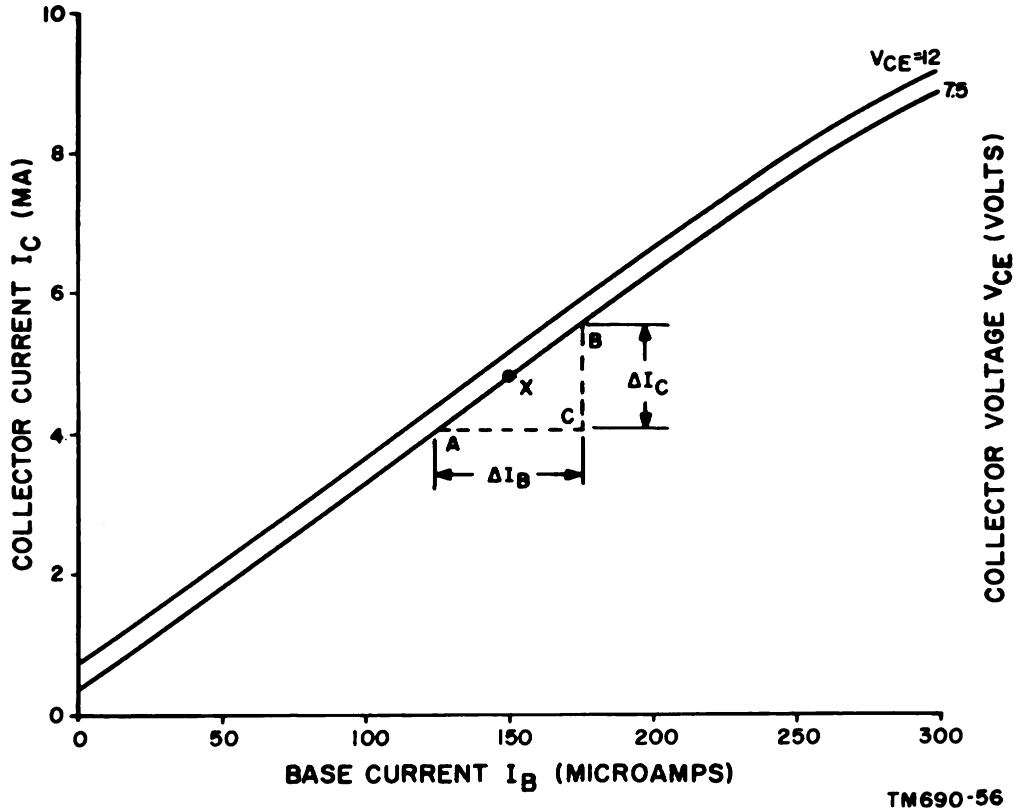


Figure 55. Typical forward transfer static characteristic curves for common emitter configuration of hybrid equivalent circuit.

of current and voltage must be used. Equating the coefficient to the incremental values, the following results:

$$\left(\frac{i_c}{v_{ce}}\right)_{i_b=o}=\left(\frac{\Delta I_C}{\Delta V_{CE}}\right)_{I_B=C}$$

The expression on the right can be directly related to the static characteristic curves. The expression is the ratio of output current to output voltage (a conductance). The expression is called *the output conductance with input ac open circuited* and is designated h_{oe}; h for hybrid, o for output, e for common-emitter configuration; thus —

$$h_{oe}=\left(\frac{\Delta I_C}{\Delta V_{CE}}\right)_{I_B=C}$$

(2) To calculate h_{oe} for a particular transistor, the output static characteristics can be used (fig. 56). Assume dc bias conditions of 150 μa of base current and 7.5 volts of collector voltage. This point is marked X on the appropriate curve and a right-angle triangle is drawn so that point X is near the center of the hypotenuse. The slope of the curve is equal to $\frac{BC}{AC}$. Line BC equals ΔI_C and line AC equals ΔV_{CE} so that—

$$h_{oe}=\frac{BC}{AC}$$

$$=\left(\frac{\Delta I_C}{\Delta V_{CE}}\right)_{I_B=C}$$

To determine ΔI_C, the value of collector current at point C (4.2 ma) must be subtracted from the value of collector current at point B (4.6 ma); thus—

$$\Delta I_C=4.6-4.2$$

$$=0.4 \text{ ma}$$

$$=0.4\times10^{-3} \text{ amp}$$

To determine ΔV_{CE}, the value of collector voltage at point A (2.5 volts) must be subtracted from the value of collector voltage at point C (12.5 volts); thus—

$$\Delta V_{CE} = 12.5 - 2.5$$
$$= 10 \text{ volts}$$

Placing these values in the formula for h_{oe}:

$$h_{oe} = \left(\frac{\Delta I_C}{\Delta V_{CE}}\right)_{I_B = C}$$
$$= \frac{0.4 \times 10^{-3}}{10}$$
$$= 0.04 \times 10^{-3} \text{ mho, or}$$
$$= 40 \mu\text{mho}$$

This is a typical value for h_{oe}. Depending on the transistor, h_{oe} may vary from 10 μmhos to 50 μmhos. Note that these values are equivalent to an output *resistance* (with input ac open-circuited) from 100,000 to 20,000 ohms respectively.

53. Construction of Hybrid Equivalent Circuit

a. General. The important factors obtained from consideration of static characteristic curves (pars. 51 and 52) are the following:

(1) $h_{ie} = \left(\frac{\Delta V_{BE}}{\Delta I_B}\right)_{V_{CE} = C}$; input resistance, output shorted.

(2) $\mu_{re} = \left(\frac{\Delta V_{BE}}{\Delta V_{CE}}\right)_{I_B = C}$; reverse open-circuit voltage amplification factor.

(3) $\alpha_{fe} = \left(\frac{\Delta I_C}{\Delta I_B}\right)_{V_{CE} = C}$; forward short-circuit current amplification factor.

(4) $h_{oe} = \left(\frac{\Delta I_C}{\Delta V_{CE}}\right)_{I_B = C}$; output conductance, input open.

Note. The factors ((1)–(4) above), consist of two pure numbers μ_{re} and α_{fe}), a resistance (h_{ie}), and a conductance (h_{oe}). This *mixture* of parameters is referred to as *hybrid* parameters; hybrid meaning mixture.

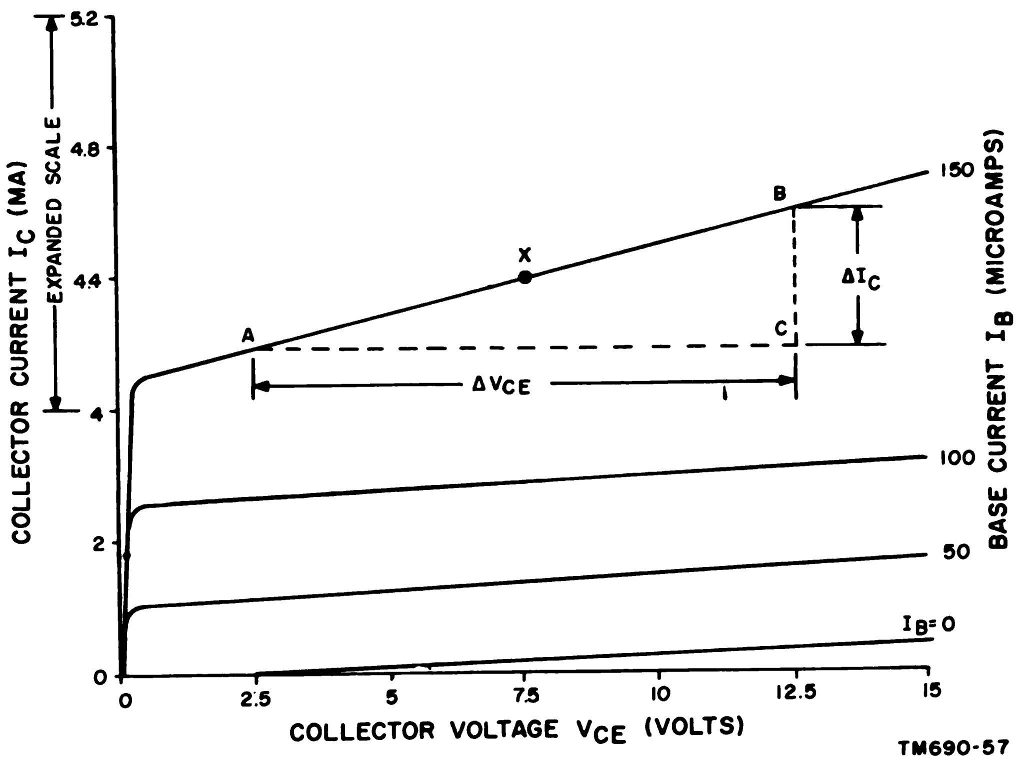


Figure 56. Output (collector) static characteristic curves for common-emitter configuration of hybrid equivalent circuit.

b. Input Circuit Equation. If h_{ie} and μ_{re} are substituted in the input circuit equation (par. 49*c*), the following equation results:

$$v_{be}=h_{ie}i_b+\mu_{re}v_{ce}$$

This equation suggests a series circuit (A, fig. 57) consisting of a resistor (h_{ie}) in series with a voltage generator ($\mu_{re}\ v_{ce}$).

c. Output Circuit Equation. If α_{fe} and h_{oe} are substituted in the output circuit equation (par. 50*c*), the following equation results:

$$i_c=\alpha_{fe}i_b+h_{oe}v_{ce}$$

This equation suggests a current generator ($\alpha_{fe}i_b$) in parallel with a conductance (h_{oe}). This circuit is shown in B, figure 57.

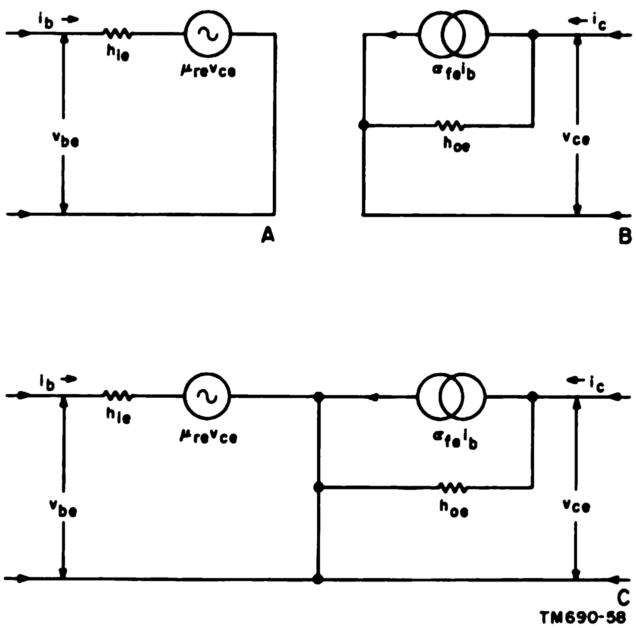


Figure 57. Hybrid parameters used for input, output, and single equivalent circuit of transistor.

d. Combined Input and Output. If the input circuit and the output circuit are combined (C, fig. 57), a single equivalent circuit is obtained to represent the transistor. Use of this circuit to develop formulas for gains and input and output resistances is covered in paragraphs 54 through 58.

54. Derivation of Current-Gain Formula

a. General. The current-gain formula for a CE amplifier can be derived by using the circuits shown in figure 58. The dc biasing circuits for the amplifier are not shown because these circuits are not reflected in the equations required to determine the ac current gain. To derive the formula, the transistor (shown in dashed lines A, fig. 58) must be replaced by its equivalent circuit (shown in dashed lines B, fig. 58). The notations used for the equivalent circuit are explained in paragraph 53*a*. The additional notations are as follows:

(1) v_g, voltage of source.
(2) R_g, internal resistance of voltage source.
(3) i_b, base current.
(4) v_{be}, base voltage.
(5) i_c, collector current.
(6) v_{ce}, collector voltage.
(7) R_L, load resistance.
(8) A_i=current gain.

Note. Input resistance r_i is not used to derive the current-gain formula. Derivation of the formula for r_i is covered in paragraph 57.

b. Derivation. To derive the formula, proceed as follows:

(1) By definition, current gain is the ratio of collector current to base current:

$$A_i=\frac{i_c}{i_b}$$

(2) Consideration of the output circuit will result in an expression containing both i_c i_b. Using Kirchoff's current law at point X (sum of all currents approaching or leaving a given point equals zero):

$$0=\alpha_{fe}i_b+h_{oe}v_{ce}+i_c$$

Note. In establishing this equation, all currents are *assumed* to flow *toward* point X. This assumption will result in a negative sign in the final answer ((6) below).

(3) Note that by Ohm's Law

$$v_{ce}=R_Li_c$$

(4) Substitute R_Li_c for v_{ce} in the equation ((2) above):

$$0=\alpha_{fe}i_b+(h_{oe}R_Li_c+i_c)$$

(5) Factor i_c:

$$0=\alpha_{fe}i_b+(h_{oe}R_L+1)i_c$$

(6) Solve for $\frac{i_c}{i_b}$ (that is, A_i):

$$A_i=\frac{i_c}{i_b}$$

$$=\frac{-\alpha_{fe}}{h_{oe}R_L+1}$$

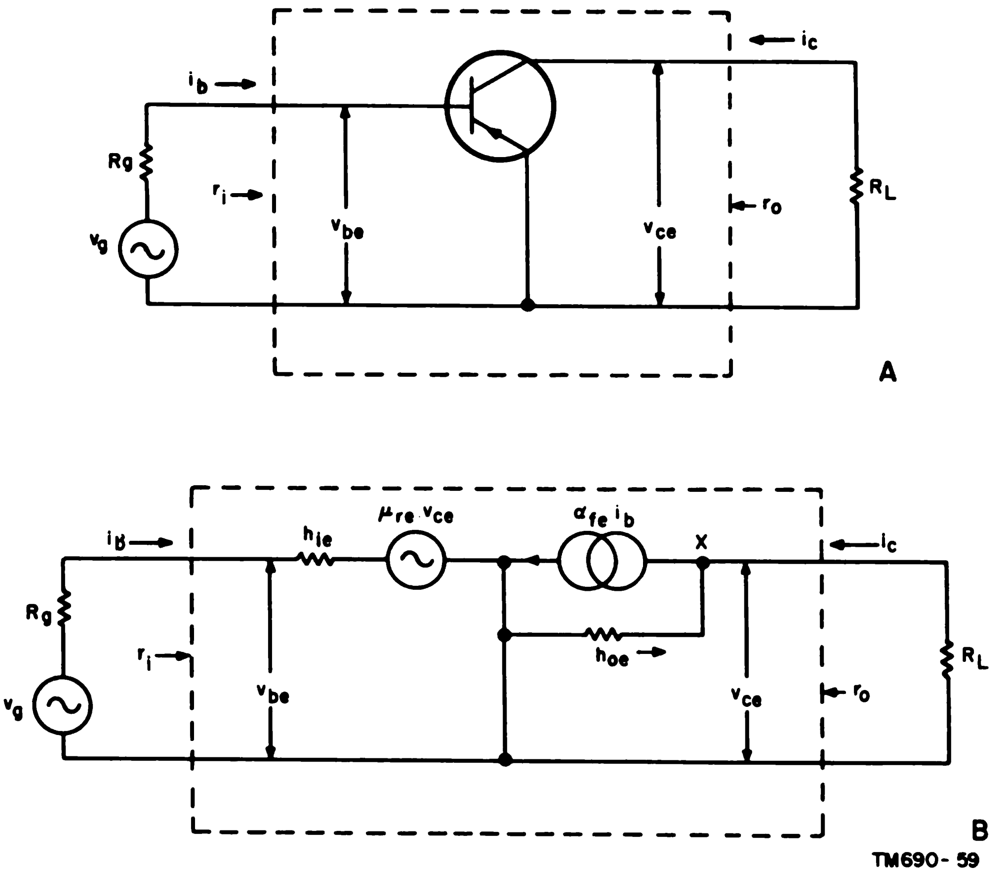


Figure 58. CE amplifier and its equivalent circuit.

The negative sign before α_{fe} indicates that there is 180° phase reversal between the input current and the output current; that is, a clockwise or counterclockwise flow of current in the input circuit results in a counterclockwise or clockwise flow of current (respectively) in the output circuit. Use of this formula in a numerical example is given in paragraph 59*b*.

55. Derivation of Voltage-Gain Formula

a. General. The voltage-gain formula for a CE amplifier can be derived by using equivalent circuit (B, fig. 58). The notations used in the derivation (*b* below), are the same as those given in paragraph 54*a*. In addition, the notation A_v is used for voltage gain.

b. Derivation. To derive the formula, proceed as follows:

(1) By definition, voltage gain is the ratio of collector voltage to base voltage:

$$A_v=\frac{v_{ce}}{v_{be}}$$

(2) Repeat the equation for the output circuit (par. 54*b*(2)):

$$0=\alpha_{fe}\, i_b+h_{oe}\, v_{ce}+i_c$$

(3) Note that by Ohm's Law:

$$i_c=\frac{v_{ce}}{R_L}$$

(4) Substitute (3) above in (2) above:

$$0=\alpha_{fe}i_b+h_{oe}v_{ce}+\frac{v_{ce}}{R_L}$$

(5) To eliminate i_b in (4) above, write the equation for v_{be}:

$$v_{be}=h_{ie}i_b+\mu_{re}v_{ce}$$

(6) Solve (5) above for i_b:

$$i_b=\frac{v_{be}}{h_{ie}}-\frac{\mu_{re}v_{ce}}{h_{ie}}$$

(7) Substitute this expression for i_b in equation (4) above, and expand:

$$0=\frac{\alpha_{fe}\, v_{be}}{h_{ie}}-\frac{\alpha_{fe}\, \mu_{re}\, v_{ce}}{h_{ie}}+h_{oe}\, v_{ce}+\frac{v_{ce}}{R_L}$$

495909 O—59——5

(8) Rearrange (7) above, and factor v_{ce}:

$$\left(h_{oe}+\frac{1}{R_L}-\frac{\alpha_{fe}\,\mu_{re}}{h_{ie}}\right)v_{ce}=\frac{-\alpha_{fe}\,v_{be}}{h_{ie}}$$

(9) Multiply both sides by $\frac{1}{V_{be}}$ and

$$\frac{1}{h_{oe}+\frac{1}{R_L}-\frac{\alpha_{fe}\,\mu_{re}}{h_{ie}}}$$ and:

$$\frac{v_{ce}}{v_{be}}=\frac{-\alpha_{fe}}{h_{ie}\,(h_{oe}+\frac{1}{R_L}-\frac{\alpha_{fe}\,\mu_{re}}{h_{ie}}}$$

(10) This equation ((9) above), is the voltage gain, but it is normally rearranged by expanding the denominator and multiplying the numerator and denominator by R_L:

$$A_v=\frac{-\alpha_{fe}\,R_L}{(h_{ie}\,h_{oe}-\alpha_{fe}\,\mu_{re})\,R_L+h_{ie}}$$

The negative sign indicates 180° phase reversal between input and output voltage. Use of this formula in a numerical example is given in paragraph 59*c*.

56. Derivation of Power-Gain Formula

a. General. The power-gain formula for a *CE* amplifier can be derived by using the equivalent circuit (B, fig. 58). The notations used in the derivation are the same as those given in paragraph 54*a*. In addition, the notation G is used for power gain.

b. Derivation. To derive the power-gain formula, proceed as follows:

(1) By definition:

$$G=\frac{\text{Power delivered to load}}{\text{Power delivered to input}}$$

$$=\frac{i_c\,v_{ce}}{i_b\,v_{be}}$$

(2) Rearrange (1) above:

$$G=\left(\frac{i_c}{i_b}\right)\times\left(\frac{v_{ce}}{v_{be}}\right)$$

(3) Note that power gain is the product of current gain $\left(\frac{i_c}{i_b}\right)$ and voltage gain $\left(\frac{v_{ce}}{v_{be}}\right)$.

Substitute in (2) above the formulas for current gain (par. 54*b*(6)) and voltage gain (par. 55*b*(10)):

$$G=\left(\frac{-\alpha_{fe}}{h_{oe}\,R_L+1}\right)\left(\frac{-\alpha_{fe}\,R_L}{h_{ie}\,h_{oe}-\alpha_{fe}\,\mu_{re}\,R_L+h_{ie}}\right)$$

(4) Expand (3) above:

$$G=\frac{(\alpha_{fe})^2R_L}{(h_{oe}R_L+1)[(h_{ie}h_{oe}-\alpha_{fe}\mu_{re})R_L+h_{ie}]}$$

Use of this formula in a numerical example is given in paragraph 59*d*.

57. Derivation of Input-Resistance Formula

a. General. The input resistance formula for a *CE* amplifier can be derived by using the equivalent circuit (B, fig. 58). See paragraph 54*a* for notations used.

b. Derivation. To derive the formula, proceed as follows:

(1) By definition, input resistance is the ratio of base voltage to base current:

$$r_i=\frac{v_{be}}{i_b}$$

(2) Repeat equation for v_{be} (par. 55*b*(5)):

$$v_{be}=h_{ie}i_b+\mu_{re}v_{ce}$$

(3) Solve (2) above for $\frac{v_{be}}{i_b}$:

$$\frac{v_{be}}{i_b}=h_{ie}+\mu_{re}\,\frac{v_{ce}}{i_b}$$

(4) To eliminate $\frac{v_{ce}}{i_b}$ from the equation ((3) above), rewrite the output circuit equation (par. 54*b*(2)) as follows:

$$O=\alpha_{fe}i_b+\left(h_{oe}+\frac{1}{R_L}\right)v_{ce}$$

(5) Solve the equation ((4) above), for $\left(\frac{v_{ce}}{i_b}\right)$

$$\frac{v_{ce}}{i_b}=\frac{-\alpha_{fe}}{h_{oe}+\frac{1}{R_L}}$$

(6) Substitute the expression ((5) above), for $\left(\frac{v_{be}}{i_b}\right)$ in (3) above:

$$\frac{v_{be}}{i_b}=h_{ie}+\mu_{re}\left(\frac{-\alpha_{fe}}{h_{oe}+\frac{1}{R_L}}\right)$$

(7) This expression ((6) above), is the formula for r_i, but by performing several algebraic operations, it is normally rearranged, as follows:

$$r_i=\frac{h_{ie}+(h_{oe}h_{ie}-\alpha_{fe}\mu_{re})(+R_L)}{1+h_{oe}R_L}$$

Use of this formula in a numerical example is given in paragraph 59*e*.

58. Derivation of Output-Resistance Formula

a. General.

(1) The output resistance formula for the *CE* amplifier can be derived by using the circuits shown in figure 59. For notations used, refer to paragraph 54*a*. The only additional notation used is r_o for output resistance.

(2) The technique used to determine the output resistance formula is to remove the signal source from the input circuit (B, fig. 58) and place it in the output circuit (A, fig. 59). The equivalent of the signal source resistance (R_g) remains in the input circuit. In the output circuit, the signal source is considered to have zero resistance.

b. Conversion of Current Generator.

(1) To analyze the output circuit with voltage generators alone, the current generator (in dashed lines) in the output circuit must be replaced by a voltage generator. According to network theory, a current generator and its shunt conductance (B, fig. 59) can be replaced by a voltage generator and a series conductance (C, fig. 59). It is required, however, that the voltage of the voltage generator be made equal to the current ($\alpha_{fe}i_b$) of the current generator divided by its shunt conductance (h_{oe}). The voltage output of the generator then is $\frac{\alpha_{fe}i_b}{h_{oe}}$. Furthermore a conductance equal to that of the shunt conductance must be placed in series with the voltage generator as indicated.

(2) A numerical example of the conversion ((1) above), is given by the numbers shown in parentheses (B and C, fig. 59). Assume a current generator with 5 ma of output ($\alpha_{fe}i_b$=5 ma), shunted by a conductance of 5 mhos (h_{oe}=5 mhos). The generator voltage output would then be:

$$\frac{\alpha_{fe}i_b}{h_{oe}}=\frac{5\text{ ma}}{5\text{ mhos}}$$
$$=1\text{ mv}$$

The conductance in series with the voltage generator would have to be 5 mhos (h_{oe}=5 mhos). If the two circuits are equivalent, then the short circuit and open circuit results obtained from each must be the same. A short circuit of the current generator results in a current flow of 5 ma (all the current generator output). A short circuit of the voltage generator output results also in 5 ma output as follows:

Voltage (mv) × conductance (mhos) = current (ma).

1 mv × 5 mhos = 5 ma.

An open circuit of the current generator (all current through shunt) results in a terminal voltage of 1 mv obtained as follows:

$$\frac{\text{current (mv)}}{\text{conductance (mhos)}}=\text{voltage (mv)}.$$

$$\frac{5\text{ ma}}{5\text{ mhos}}=1\text{ mv}$$

An open circuit of the voltage generator results in all the generator voltage (1 mv) being available at the terminals. It may be stated then that the current generator and the voltage generator as indicated are equivalent. With the current generator and the shunt conductance of the equivalent circuit (A, fig. 59) replaced by a voltage generator and a series conductance, the equivalent circuit of the transistor appears as shown in D of figure 59.

c. Derivation. To derive the output-resistance formula, proceed as follows:

(1) The current and voltage relationship in the input circuit (D, fig. 59) is expressed as follows:

$$\mu_{re}v_{ce}=h_{ie}i_b+R_g i_b$$

(2) The current and voltage relationship in the output circuit is expressed as follows:

$$v_g=R_L i_c+\frac{i_c}{h_{oe}}+\frac{\alpha_{fe}i_b}{h_{oe}}$$

(3) The output resistance (r_o) is the resistance *looking* into the output of the transistor (D, fig. 59). The signal source (v_g) *sees* a series circuit consisting of R_L and r_o (E, fig. 59). The equation for this series circuit is as follows:

$$v_g=R_L i_c+r_o i_c$$

(4) Substitute equation (3) above, for v_g in (2) above.

$$R_L i_c+r_o i_c=R_L i_c+\frac{i_c}{h_{oe}}+\frac{\alpha_{fe}}{h_{oe}}\,i_b$$

(5) Subtract $R_L i_c$ from both sides and divide both sides by i_c:

$$r_o=\frac{1}{h_{oe}}+\frac{\alpha_{fe}}{h_{oe}}\frac{i_b}{i_c}$$

(6) Solve equation (1) above, for i_b:

$$i_b=\frac{\mu_{re}v_{ce}}{h_{ie}+R_g}$$

(7) Note that—

$$i_c=\frac{v_{ce}}{r_o}$$

(8) Substitute the expressions for i_b ((6) above), and i_c ((7) above), in (5) above:

$$r_o=\frac{1}{h_{oe}}+\left(\frac{\alpha_{fe}}{h_{oe}}\right)\left(\frac{\mu_{re}v_{ce}}{h_{ie}+R_g}\Big/\frac{v_{ce}}{r_o}\right)$$

(9) Simplify the complex fraction in (8) above:

$$r_o=\frac{1}{h_{oe}}+\left(\frac{\alpha_{fe}}{h_{oe}}\right)\left(\frac{\mu_{re}\,r_o}{h_{ie}+R_g}\right)$$

(10) By performing several algebraic operations on the equtation ((9) above), r_o can be isolated to the left hand side to give:

$$r_o=\frac{h_{ie}+R_g}{h_{oe}h_{ie}-\mu_{re}\alpha_{fe}+h_{oe}R_g}$$

Use of this formula in a numerical example is covered in paragraph 59*f*.

59. Numerical Example Using Hybrid Parameters

a. General.

(1) A, figure 60 shows an amplifier driven by a signal source (v_g) which has an internal resistance R_g. Capacitor $C1$ blocks the dc bias voltage from the signal source and couples the signal to the base of the transistor. Resistors $R1$ and $R2$ form a voltage divider to establish the base-emitter forward bias. Resistor $R3$ is the emitter-swamping resistor, used to temperature-stabilize the emitter current (par. 74*b*(2)). Capacitor $C2$ bypasses the signal around resistor $R3$. Resistor R_L is the collector load resistor. Capacitor $C3$ blocks dc voltage from and couples the signal to the following stage.

(2) Assume it is desired to calculate the input resistance (r_i), output resistance (r_o), current gain (A_i), voltage gain (A_v), and power gain (G) of the amplifier ((1) above). The first step is to eliminate the parts from the circuit that do not directly affect the computations. These parts are the low-impedance coupling capacitors ($C1$ and $C2$), the high-resistance voltage divider ($R1$ and $R2$), the swamping resistor ($R3$) with its bypass capacitor ($C2$), and the battery. The resultant circuit is shown in B, figure 60. The second step is to replace the transistor (in dashed lines) with its equivalent circuit (C, fig. 60). At the operating point (emitter current of 1 ma, and collector-emitter voltage of 8 volts), h_{ie}=1,500 ohms, $\mu_{re}=5\times10^{-4}$, $\alpha_{fe}=50$, and $h_{oe}=20\times10^{-6}$ mhos.

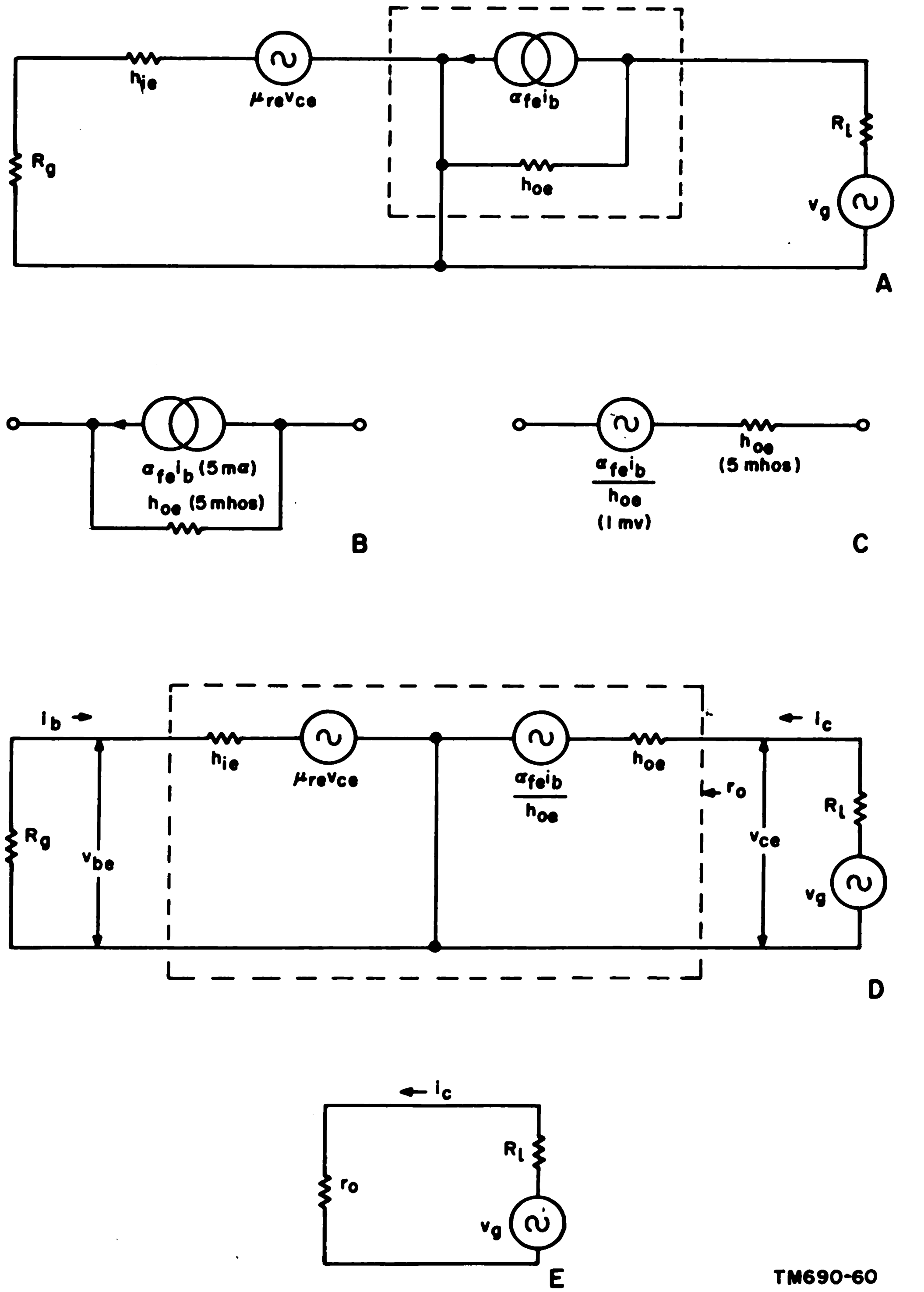


Figure 59. Equivalent circuits and conversions used to determine the output resistance formula.

b. Current Gain. To determine the current gain, proceed as follows:

(1) Express the formula for current gain (par. 54*b*(6)):

$$A_i=\frac{-\alpha_{fe}}{h_{oe}R_L+1}$$

(2) Substitute known values (*a* above), in this formula.

$$A_i=\frac{-50}{(20\times10^{-6})(15\times10^{3})+1}$$

$$=\frac{-50}{1.3}$$

$$=-38.4$$

The answer indicates that the input current is amplified 38.4 times and is reversed 180° in phase when going through the transistor.

c. Voltage Gain. To determine the voltage gain, proceed as follows:

(1) Express the formula for voltage gain (par. 55*b*(10)):

$$A_v=\frac{-\alpha_{fe}R_L}{(h_{ie}h_{oe}-\alpha_{fe}\mu_{re})R_L+h_{ie}}$$

(2) Substitute known values (*a* above), in this formula.

$$A_v=\frac{-50\times15{,}000}{(1{,}500\times20\times10^{-6}-50\times5\times10^{-4})15{,}000+1{,}500}$$

$$=\frac{-75\times10^{4}}{1{,}575}$$

$$=-476$$

The answer indicates that the input voltage is amplified 476 times and is reversed 180° in phase when going through the transistor.

d. Power Gain. To determine the power gain, proceed as follows:

(1) Express the formula for power gain gain (par. 56*b*(4)):

$$G=\frac{(\alpha_{fe})2R_L}{(h_{oe}R_L+1)[(h_{ie}h_{oe}-\alpha_{fe}\mu_{re})R_L+h_{ie}]}$$

(2) By substituting the known values in this formula, the power gain can be obtained. However, the power gain is the product of voltage gain and current gain (*b* and *c* above); thus—

$$G=A_iA_v$$

(3) Substitute the known values for A_i and A_v:

$$G=(38.4)(476)$$

$$=18{,}278$$

The power input is increased 18,278 times when going through the transistor.

(4) The power gain in db is:

$$G=10\log 18{,}278$$

$$=10\times4.26$$

$$=42.6\text{ db}$$

e. Input Resistance. To determine the input resistance, proceed as follows:

(1) Express the formula for input resistance (par. 57*b*(7)):

$$r_i=\frac{h_{ie}+(h_{oe}h_{ie}-\alpha_{fe}\mu_{re})R_L}{1+h_{oe}R_L}$$

(2) Substitute the known values in the formula:

$$r_i=\frac{1{,}500+(20\times10^{-6}\times1{,}500-50\times5\times10^{-4})15{,}000}{1+20\times10^{-6}\times15{,}000}$$

$$r_i=\frac{1{,}575}{1.3}$$

$$r_i=1{,}212\text{ ohms}$$

f. Output Resistance. To determine the output resistance, proceed as follows:

(1) Write the formula for output resistance (par. 58*c*(10)):

$$r_o=\frac{h_{ie}+R_g}{h_{oe}h_{ie}-\mu_{re}\alpha_{fe}+h_{oe}R_g}$$

(2) Substitute known values in the formula.

$$r_o=\frac{1{,}500+1{,}500}{20\times10^{-6}\times1{,}500-5\times10^{-4}\times50+20\times10^{-6}\times1{,}500}$$

$$r_o=\frac{3{,}000}{.035}$$

$$r_o=85{,}714\text{ ohms}$$

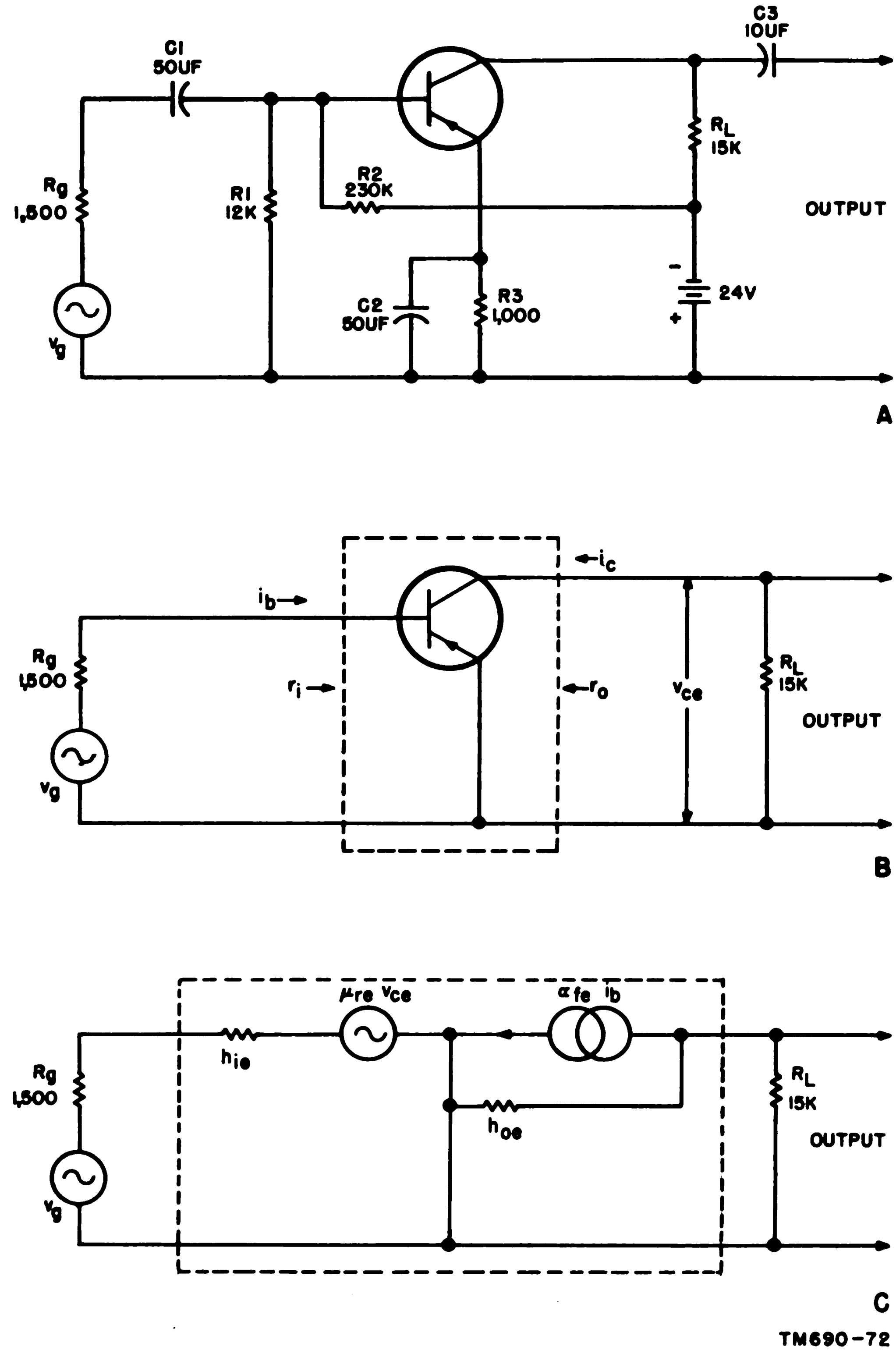


Figure 60. Common emitter amplifier and its preparation for analysis.

60. Transducer Gain and Matched Amplifier Formulas

The following formulas are useful in many applications. The formulas are written for the *CE* configuration; they can be made to apply to the *CB* and the *CC* configuration by changing subscript *e* to *b* or *c* respectively.

a. Transducer Gain. The formula for power gain (par. 56) compares the power delivered to the output of the transistor to that delivered to the input of the transistor. The formula for comparison of the power delivered to the load to that generated at the transducer (signal voltage source) is as follows:

$$G_t=\frac{4R_g}{R_L}\left[\frac{\alpha_{fe}}{(h_{ie}+R_g)\left(h_{oe}+\frac{1}{R_L}\right)-\alpha_{fe}\mu_{re}}\right]^2$$

b. Matched Power Gain. If the load resistance (R_L) equals (matches) the amplifier output resistance (r_o), maximum power output from the amplifier is realized. The formula for maximum power gain (G_m) is as follows:

$$G_m=\frac{\alpha_{fe}^2}{h_{ie}h_{oe}\left(1+\sqrt{1-\frac{\alpha_{fe}\mu_{re}}{h_{ie}h_{oe}}}\right)^2}$$

c. Signal Source Resistance (R_g) *for Matched Condition.* The formula for calculation of the signal source resistance under matched conditions ($R_g=r_i$, and $R_L=r_o$) is as follows:

$$R_g=h_{ie}\sqrt{1-\frac{\alpha_{fe}\mu_{re}}{h_{ie}h_{oe}}}$$

d. Load Resistance for Matched Condition. The formula for calculation of the load resistance under matched conditions ($R_g=r_i$, and $R_L=r_o$) is as follows:

$$\frac{1}{R_L}=h_{oe}\sqrt{1-\frac{\alpha_{fe}\mu_{re}}{h_{ie}h_{oe}}}$$

Section III. HYBRID PARAMETERS, CB AND CC AMPLIFIERS

61. General

a. The detailed procedures for the development of the hybrid parameters static characteristic curves, equivalent circuits, and the derivation of circuit formulas for the CB and the CC amplifiers are similar to those for the CE amplifiers (pars. 47–58). Because of the similarity, only a brief discussion of the developments and the derivations with the complete final results is presented for the CB amplifier (par. 62) and the CC amplifier (par. 63).

b. In addition, a table (par. 65) is provided for the conversion of one set of hybrid parameters of one configuration to the hybrid parameters of the other configurations.

62. Hybrid Parameters, Static Characteristic Curves, and Equivalent Circuit for CB Configuration

a. General. Figure 61 shows the transistor considered as a black box and so connected that the *CB* configuration results. The directions of current flow are arbitrarily assumed as shown. Consideration of the input current (emitter current i_e) and the output voltage (collector-base voltage v_{cb}) as the independent variables results in hybrid parameters. The remaining quantities (emitter-base voltage v_{eb} and collector current i_c) are the dependent variables. The general equations relating the independent and the dependent variables may be written as follows:

(1) Input Circuit:

$$v_{eb}=f_1(i_e,\ v_{cb})$$

(2) Output Circuit:

$$i_c=f_2(i_e,\ v_{cb})$$

b. Expansion of General Equations. Expansion of the general equations (*a* above), results in the following:

(1) Input Circuit:

$$v_{eb}=\left(\frac{v_{eb}}{i_e}\right)_{v_{cb}=0}^{i_e}+\left(\frac{v_{eb}}{v_{cb}}\right)_{i_e=0}^{v_{cb}}$$

(2) Output Circuit:

$$i_c=\left(\frac{i_c}{i_e}\right)_{v_{cb}=0}^{i_e}+\left(\frac{i_c}{v_{cb}}\right)_{i_e=0}^{v_{cb}}$$

c. Static Characteristic Curves. The values for the coefficients of the currents and voltages of the input and output circuit equations (*b* (1) and (2) above), can be obtained from static characteristic curves. The numerator of the coefficient indicates the corresponding dc quantity to be plotted vertically (dependent variable). The denominator indicates the corresponding dc quantity to be plotted horizontally (independent variable). The second independent variable, which is not permitted to vary, indicates the corresponding dc quantity to be used as the *running parameter*. The resultant characteristic curves for the four coefficients are shown in figure 62. As in the case of the common emitter configuration (pars. 51 and 52), the coefficients are given specific designations and are equated to the corresponding incremental dc values.

(1) The first coefficient of the equation (b(1) above), is the *input resistance with output short-circuited* and is designated h_{ib}; subscript b indicates common base. Relating this factor to the slope of the input characteristic curves (A, fig. 62):

$$h_{ib}=\left(\frac{v_{eb}}{i_e}\right)_{v_{cb}=0}$$

$$=\left(\frac{\Delta V_{EB}}{\Delta I_E}\right)_{V_{CB}=C}$$

This value is in ohms.

(2) The second coefficient of the equation (b(1) above), is the *reverse (feedback) open circuit voltage amplification factor* and is designated μ_{rb}. Relating this factor to the slope of the reverse transfer characteristic curves (B, fig. 62):

$$\mu_{rb}=\left(\frac{v_{eb}}{v_{cb}}\right)_{i_e=0}$$

$$=\left(\frac{\Delta V_{EB}}{\Delta V_{CB}}\right)_{I_E=C}$$

This is a pure number.

(3) The first coefficient of the equation (b(2) above), is the *forward short circuit current amplification factor* and is designated α_{fb}. Relating this factor to the slope of the forward transfer characteristic curves (C, fig. 62):

$$\alpha_{fb}=\left(\frac{i_c}{i_e}\right)_{v_{cb}=0}$$

$$=\left(\frac{\Delta I_C}{\Delta I_E}\right)_{V_{CB}=C}$$

This value is a pure number and is negative. The number is negative because the direction of current flow (arbitrarily assumed (fig. 61)) in the output circuit is opposite to that shown.

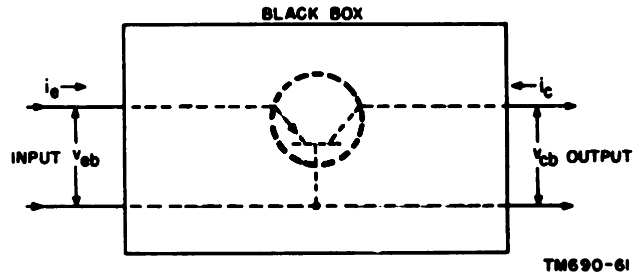


Figure 61. Black box representation of transistor connected so that base is common to input and output circuits.

(4) The second coefficient of the equation (b(2) above), is the *output conductance with input open* and is designated h_{ob}. Relating this factor to the slope of the output static characteristics (D, fig. 62):

$$h_{ob}=\left(\frac{i_c}{v_{cb}}\right)_{i_e=0}$$

$$=\left(\frac{\Delta I_C}{\Delta V_{CB}}\right)_{I_E=C}$$

This value is in ohms.

d. Input and Output Equivalent Circuits. If the factors derived in c above, are substituted in the equations (b above), the following equations result:

(1) Input Circuit:

$$v_{eb}=h_{ib}i_e+\mu_{rb}v_{cb}$$

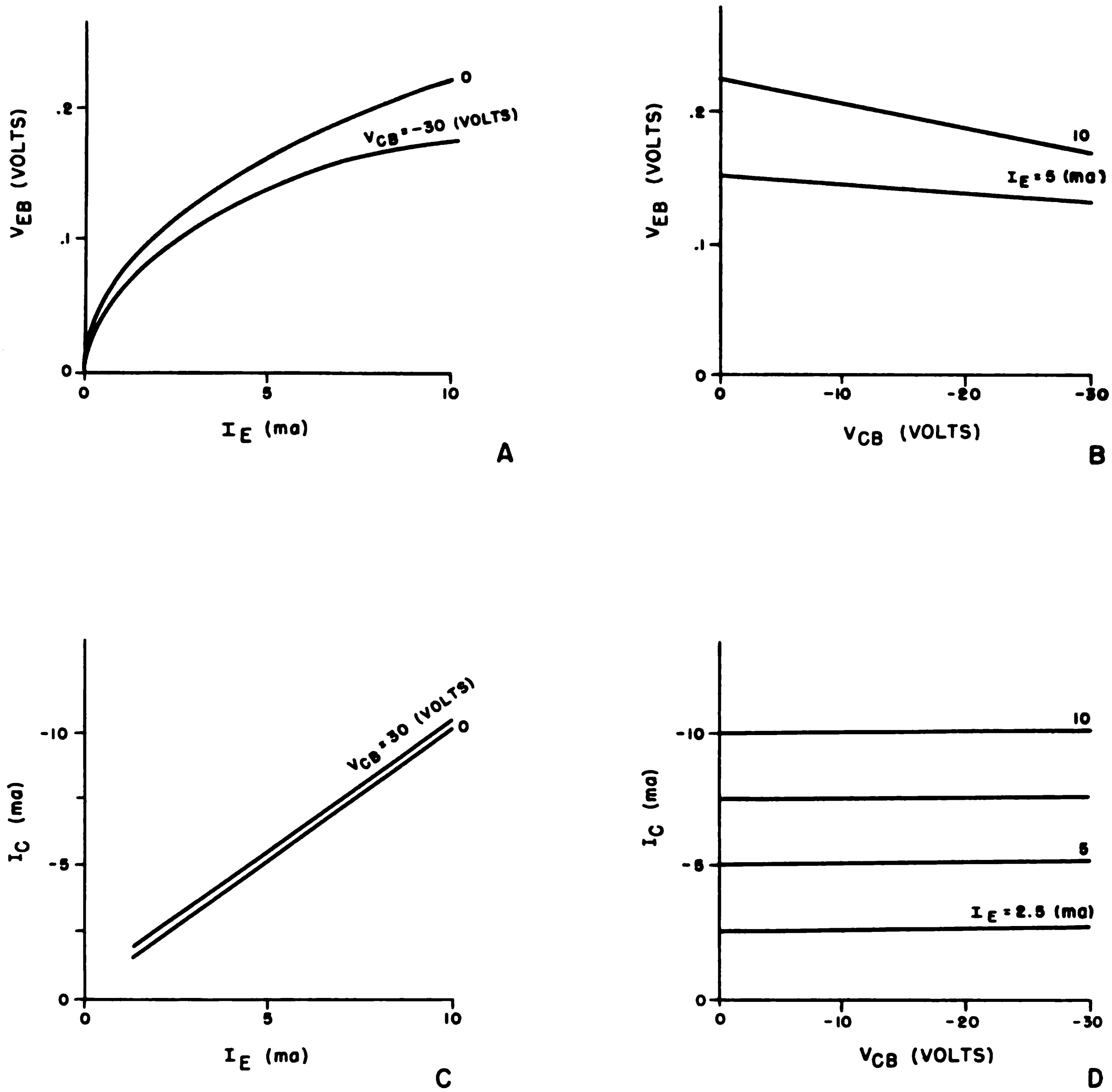
This equation represents the circuit shown in A, figure 63.

(2) Output Circuit:

$$i_c=\alpha_{fb}i_e+h_{ob}v_{cb}$$

This equation represents the circuit shown in B, figure 63.

e. Complete Equivalent Circuit. The input circuit and the output circuit can be combined as shown in C, figure 63. This circuit can be used to represent the behavior of the *CB* amplifier.



TM690-62

Figure 62. Static characteristic curves of CB configuration for determination of hybrid parameters.

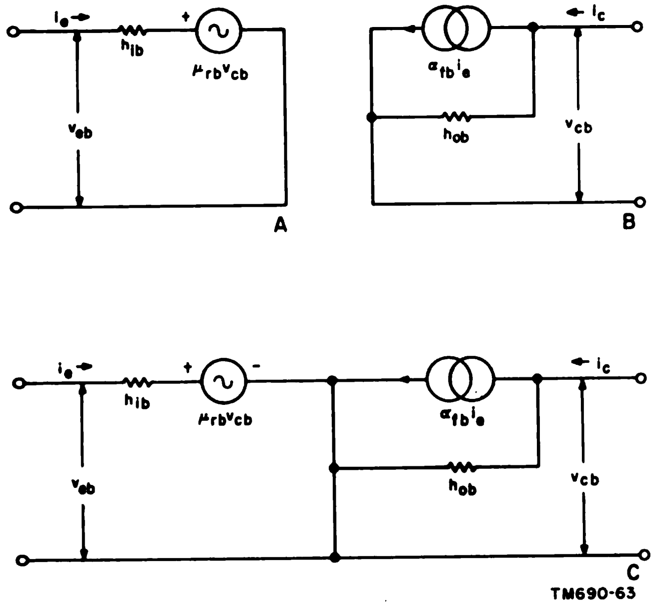


Figure 63. Input, output, and complete equivalent circuit for CB configuration using hybrid parameters.

63. Hybrid Parameters, Static Characteristic Curves, and Equivalent Circuits for CC Configuration

a. General. Figure 64 shows the transistor connected as a black box and so connected that the *CC* configuration results. The directions of current flow are arbitrarily assumed as shown. Considering the input current (i_b) and output voltage (v_{ec}) as independent variables, and input voltage (v_{bc}) and output current (i_e) as dependent variables, the following general equations may be written:

(1) Input Circuit:

$$v_{bc}=f_1(i_b,\ v_{ec})$$

(2) Output Circuit:

$$i_e=f_2(i_b, v_{ec})$$

b. Expansion of General Equations. Expansion of the general equations give:

(1) Input Circuit:

$$v_{bc}\left(\frac{v_{bc}}{i_b}\right)_{v_{ec}=0} i_b+\left(\frac{v_{bc}}{v_{ec}}\right)_{i_b=0} v_{ec}$$

(2) Output Circuit:

$$i_e\left(\frac{i_e}{i_b}\right)_{v_{ec}=0} i_b+\left(\frac{i_e}{v_{ec}}\right)_{i_b=0} v_{ec}$$

c. Static Characteristic Curves. The static characteristic curves required for the coefficients (*b*) above, are shown in figure 65.

(1) From the input characteristic curves (A, fig. 65), the *input resistance with output shorted* (h_{ic}) can be determined:

$$h_{ic}=\left(\frac{v_{bc}}{i_b}\right)_{v_{ec}=0}$$

$$=\left(\frac{\Delta V_{BC}}{\Delta I_B}\right)_{V_{EC}=C}$$

This value is in ohms.

(2) From the reverse transfer characteristic curves (B, fig. 65) the *reverse (feedback) open circuit voltage amplification factor* (μ_{rc}) can be determined:

$$\mu_{rc}=\left(\frac{v_{bc}}{v_{ec}}\right)_{i_b=0}$$

$$=\left(\frac{\Delta V_{BC}}{\Delta V_{BC}}\right)_{I_B=C}$$

This value is a pure number.

(3) From the forward transfer characteristic curves (C, fig. 65), the forward short circuit current amplification factor (α_{fc}) can be determined:

$$\alpha_{fc}=\left(\frac{i_e}{i_b}\right)_{v_{ec}=0}$$

$$=\left(\frac{\Delta I_E}{\Delta I_B}\right)_{V_{EC}=C}$$

This value is a pure number and is negative. The number is negative because the direction of current flow (arbitrarily assumed (fig. 64)) in the output circuit is opposite to that shown.

(4) From the output static characteristic curves (D, fig. 65) *the output conductance*

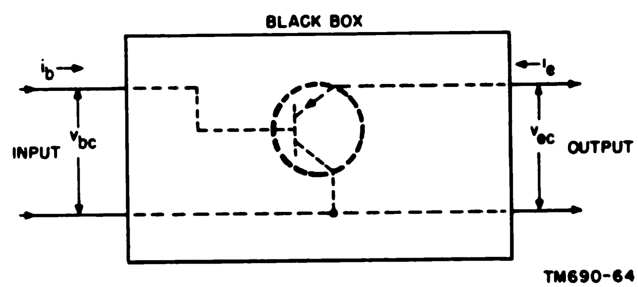


Figure 64. Black box representation of transistor connected so that collector is common to input and output circuits.

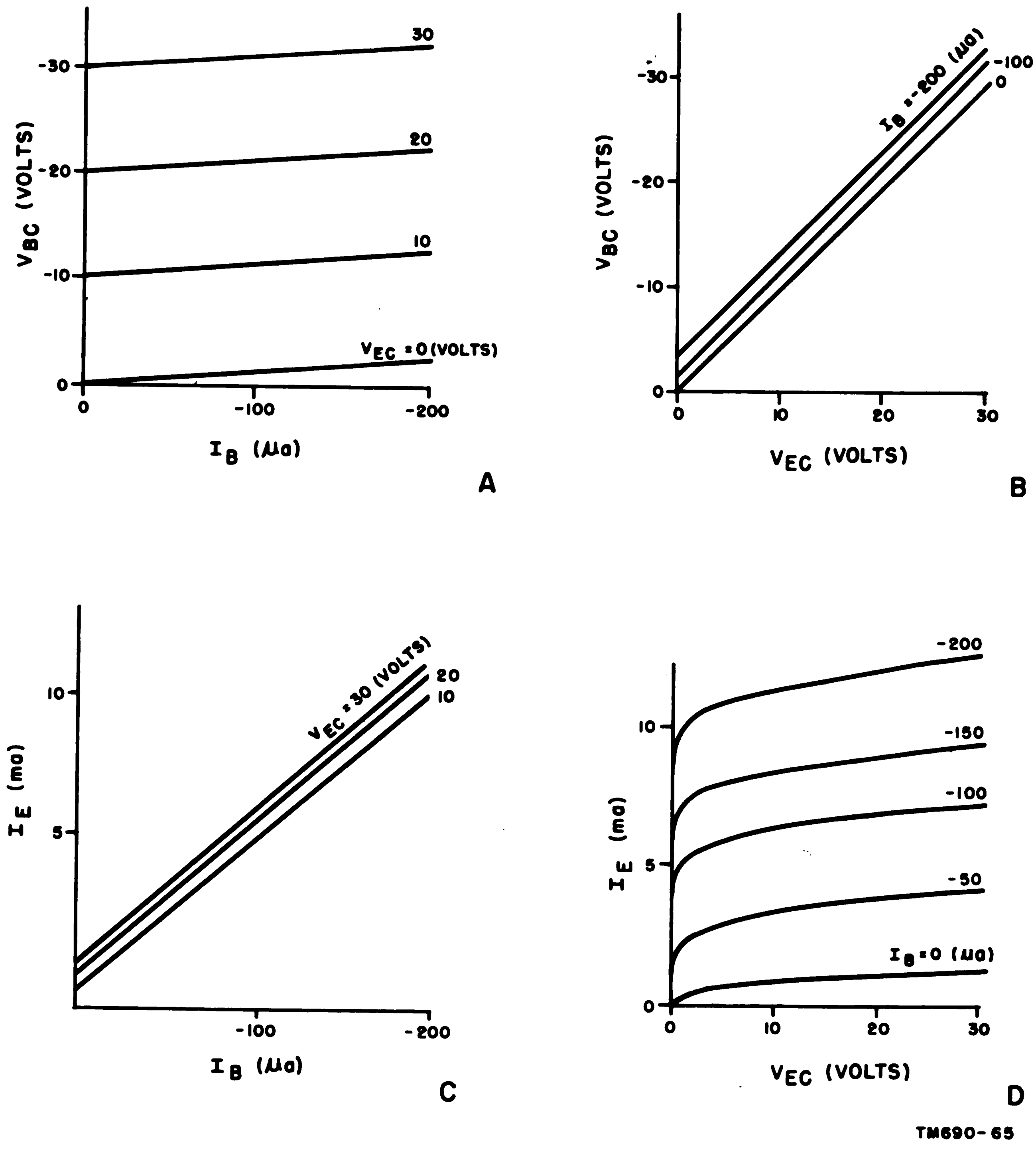


Figure 65. Static characteristic curves of* CC *configuration for determination of hybrid parameters.

with input open (h_{oc}) can be determined:

$$h_{oc}=\left(\frac{i_e}{v_{ec}}\right)_{i_b=0}$$

$$=\left(\frac{\Delta I_E}{\Delta V_{EC}}\right)_{I_B=C}$$

d. Input and Output Equivalent Circuits. If the factors derived in *c* above, are substituted in the equations (*b* above), the following equations result:

(1) Input Circuit:

$$v_{bc}=h_{ic}i_b+\mu_{rc}v_{ec}$$

This equation represents the series combination of resistor h_{ic} and voltage generator $\mu_{rc}v_{ec}$ (B, fig. 67).

(2) Output Circuit:

$$i_e=\alpha_{fc}i_b+h_{oc}v_{ec}$$

This equation represents the parallel combination of conductance h_{oc} and current generator $\alpha_{fc}i_b$ (B, fig. 67).

e. Complete Equivalent Circuit. Combination of the input and output equivalent circuits (*d* above), results in the complete equivalent circuit for the *CC* configuration. The equivalent circuit is shown within the dashed lines (B, fig. 67).

64. Input Resistance, Output Resistance, and Gains

a. CE *Amplifier.* The formulas for input resistance, output resistance, and gains for the *CE* amplifier are as follows:

(1) Current gain (par. 54):

$$A_i=\frac{-\alpha_{fe}}{h_{oe}R_L+1}$$

(2) Voltage gain (par. 55):

$$A_v=\frac{-\alpha_{fe}R_L}{(h_{ie}h_{oe}-\alpha_{fe}\mu_{re})R_L+h_{ie}}$$

(3) Power gain (par. 56):

$$G=\frac{(\alpha_{fe})^2R_L}{(h_{oe}R_L+1)[(h_{ie}h_{oe}-\alpha_{fe}\mu_{re})R_L+h_{ie}]}$$

(4) Input resistance (par. 57):

$$r_i=\frac{h_{ie}+(h_{oe}h_{ie}-\alpha_{fe}\mu_{re})R_L}{1+h_{oe}R_L}$$

(5) Output resistance (par. 58):

$$r_o=\frac{h_{ie}+R_g}{h_{oe}h_{ie}-\mu_{re}\alpha_{fe}+h_{oe}R_g}$$

b. CB and CC Amplifier.

(1) A, figure 66 shows a transistor connected in a common base configuration. In B, figure 66, the transistor is replaced by its equivalent circuit (in dashed lines). This circuit is similar to that of the *CE* amplifier (B, fig. 58). The equations for the derivation of the desired formulas (*a* (1) through (5) above), for the *CB* amplifier are the same as those for the *CE* amplifier; only the subscript letters change. The formulas in *a* (1) through (5) above, apply equally to the *CB* amplifier provided the subscript letter *e* is changed to *b* wherever *e* appears.

(2) A, figure 67 shows a transistor connected in a common collector configuration. In B, figure 67, the transistor is replaced by its equivalent circuit (in dashed lines). The circuit is similar to that of the CE amplifier (B, fig. 58). The relationships between the CE and the CB amplifier stated in (1) above, also exist between the CE and the CC amplifiers. The formulas in *a*(1) through (5) above, also apply to the CC amplifier provided the subscript letter *e* is changed to *c* wherever *e* appears.

65. Hybrid Parameters, Typical Values

Paragraph 64 states that the formulas for input and output resistances and for gains are the same for the three basic configurations except that the subscript letter (*e*, *b*, or *c*) designating the configurations must be changed. It must not be assumed that the values for a given quantity (such as input resistance) is the *same* for each configuration. The parameter values vary for each configuration. The following chart presents a set of values for the hybrid parameters of a typical transistor. Actually the

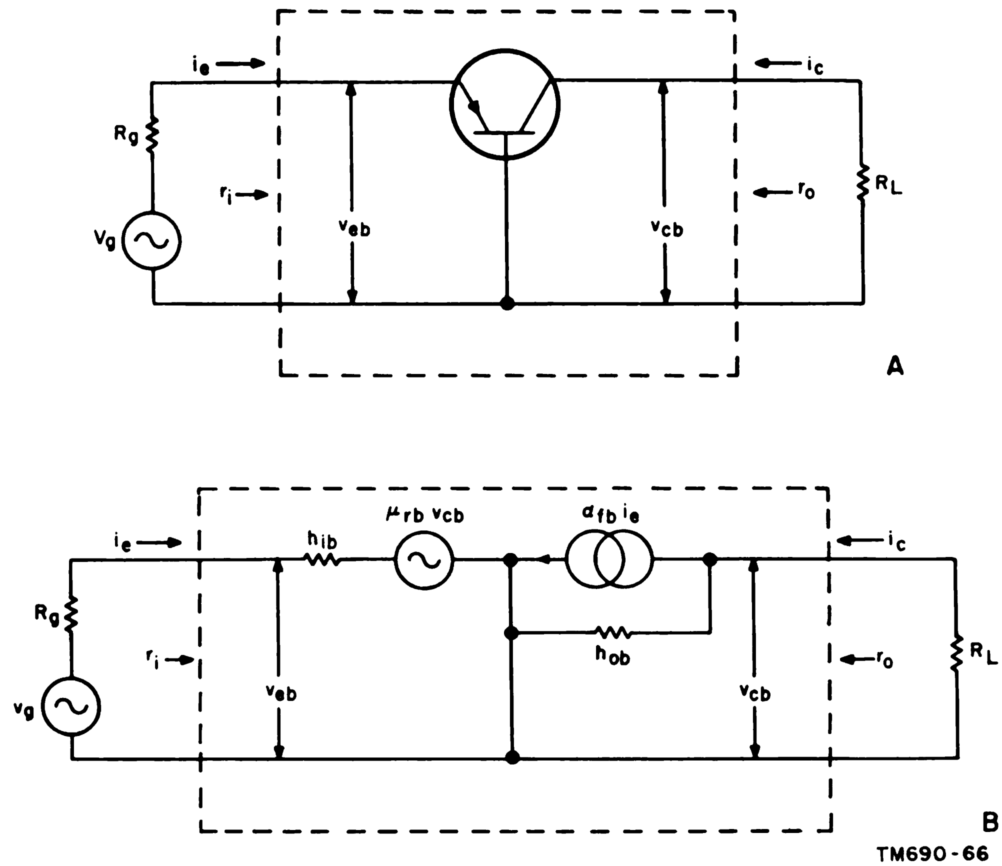


Figure 66. CB amplifier and its equivalent circuit.

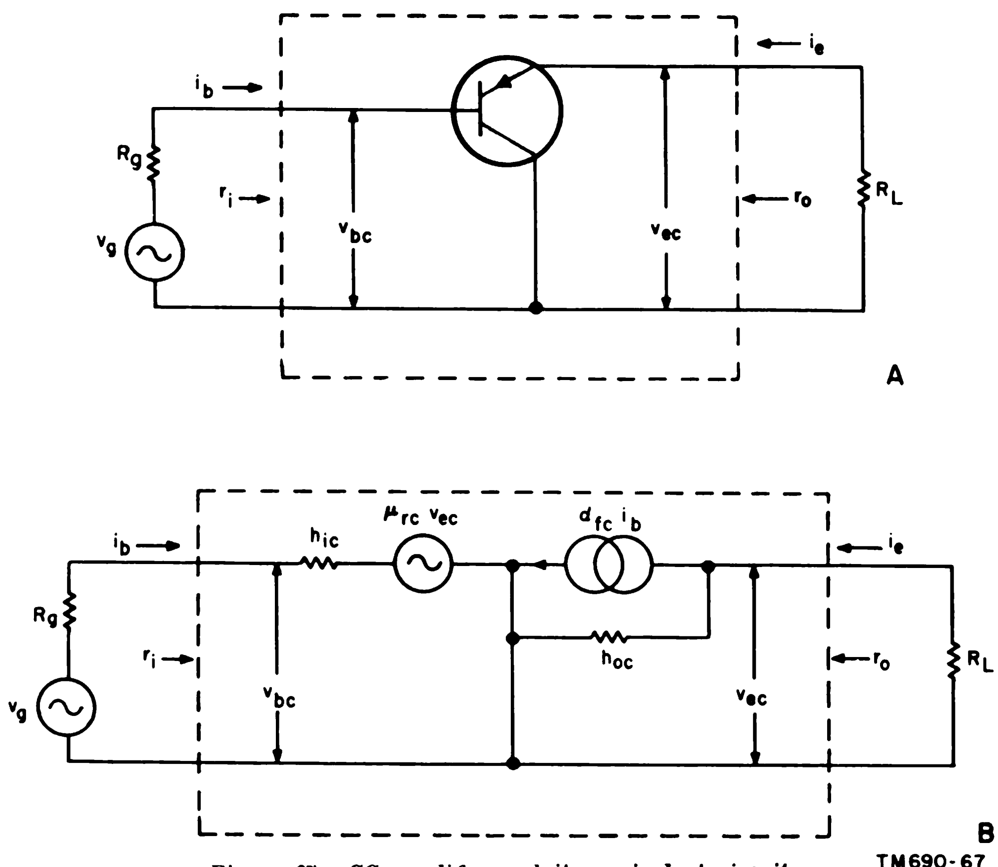


Figure 67. CC amplifier and its equivalent circuit.

parameter values for a given configuration vary over a very wide range, depending on the transistor types. The value of the following table, therefore, is in showing the relative magnitudes in parameter values among the three configurations.

Common emitter	Common base	Common collector
$h_{ie}=1{,}950$ ohms	$h_{ib}=39$ ohms	$h_{ic}=1{,}950$ ohms.
$\mu_{re}=575\times10^{-6}$	$\mu_{rb}=380\times10^{-6}$	$\mu_{rc}=1$
$\alpha_{fe}=49$	$\alpha_{fb}=-0.98$	$\alpha_{fc}=-50.$
$h_{oe}=24.5\ \mu$ mhos	$h_{ob}=0.49\ \mu$ mhos	$h_{oc}=24.5\ \mu$ mhos.

66. Hybrid Parameters, Conversion Formulas

a. Conversion Formulas. Manufacturer's data sheets on particular transistors usually present hybrid parameter values for the CE configuration, the CB configuration, or a portion of each. The first two columns of the following table list formulas for converting hybrid parameters of the CE configuration to those of the CB and the CC configurations. The last two columns list formulas for converting hybrid parameters of the CB configuration to those of the CE and the CC configurations. Conversion formulas from the CC configurations to the CE and the CB configurations are not given because the CC configuration parameters are not usually given by the manufacturer.

From *CE* to *CB*	From *CE* to *CC*	From *CB* to *CE*	From *CB* to *CC*
$h_{ib}=\dfrac{h_{ie}}{1+\alpha_{fe}}$	$h_{ic}=h_{ie}$	$h_{ie}=\dfrac{h_{ib}}{1+\alpha_{fb}}$	$h_{ic}=\dfrac{h_{ib}}{1+\alpha_{fb}}$
$\mu_{rb}=\dfrac{h_{ie}h_{oe}}{1+\alpha_{fe}}-\mu_{re}$	$\mu_{rc}=1-\mu_{re}\cong1$	$\mu_{re}=\dfrac{h_{ib}h_{ob}}{1+\alpha_{fe}}-\mu_{rb}$	$\mu_{rc}=\dfrac{1=h_{ib}h_{ob}}{1+\alpha_{fb}}+\mu_{rb}$
$\alpha_{fb}=\dfrac{-\alpha_{fe}}{1+\alpha_{fe}}$	$\alpha_{fc}=-(1+\alpha_{fe})$	$\alpha_{fe}=\dfrac{-\alpha_{fb}}{1+\alpha_{fb}}$	$\alpha_{fc}=\dfrac{-1}{1+\alpha_{fb}}$
$h_{ob}=\dfrac{h_{oe}}{1+\alpha_{fe}}$	$h_{oc}=h_{oe}$	$h_{oe}=\dfrac{h_{ob}}{1+\alpha_{fb}}$	$h_{oc}=\dfrac{h_{ob}}{1+\alpha_{fb}}$

b. Examples. Assume that the hybrid parameters of the *CB* configuration are given (second column of table in par. 65), and it is desired to find the hybrid parameters of the CE configuration. These can be found by using the formulas in the third column of the table (*a* above), and substituting the known values.

(1) Example 1:

$$h_{ie}=\frac{h_{ib}}{1+\alpha_{fb}}$$

$$=\frac{39}{1+(-0.98)}$$

$$=\frac{39}{0.02}$$

$=1{,}950$ ohms (See first col. of table, par. 65)

(2) Example 2:

$$\mu_{re}=\frac{h_{ib}h_{ob}}{1+\alpha_{fb}}-\mu_{rb}$$

$$=\frac{(39)(0.49\times10^{-6})}{1+(-0.98)}-380\times10^{-6}$$

$$=\frac{19.11\times10^{-6}}{0.02}-380\times10^{-6}$$

$$=955\times10^{-6}-380\times10^{-6}$$

$=575\times10^{-6}$ (See par. 65)

(3) Example 3:

$$\alpha_{fe}=-\frac{\alpha_{fb}}{1+\alpha_{fb}}$$

$$=-\frac{-0.98}{1+(-0.98)}$$

$$=\frac{0.98}{0.02}$$

$=49$ (See par. 65)

(4) Example 4:

$$h_{oe} = \frac{h_{ob}}{1+\alpha_{fb}}$$

$$= \frac{0.49 \times 10^{-6}}{1+(-0.98)}$$

$$= \frac{0.49 \times 10^{-6}}{0.02}$$

$$= 24.5 \times 10^{-6} \text{ mhos}$$

$$= 24.5\ \mu\text{mhos (par. 65)}$$

67. Additional Notations, Hybrid Parameters

Hybrid-parameter notations other than those used in this manual may be used in transistor data sheets prepared by manufacturers. The following table lists the hybrid parameter notations used by manufacturers and those used in this manual (col. No. 1 under the particular configuration):

Parameter	Configuration														
	Common emitter					Common base					Common collector				
	1	2	3	4	5	1	2	3	4	5	1	2	3	4	5
1. Input resistance with output shorted	h_{ie}	h_{11e}	$\frac{1}{r_{11e}}$	h_{bb}	h_{ie}	h_{ib}	h_{11b}	$\frac{1}{r_{11b}}$	h_{ee}	h_{ib}	h_{ic}	h_{11c}	$\frac{1}{r_{11c}}$	h_{bb}	h_{ic}
2. Reverse open-circuit voltage amplification factor	μ_{re}	h_{12e}	μ_{bc}	h_{bc}	h_{re}	μ_{rb}	h_{12b}	μ_{ec}	h_{ec}	h_{rb}	μ_{rc}	h_{12c}	μ_{be}	h_{be}	h_{rc}
3. Forward short-circuit amplification factor	α_{fe}	h_{21e}	α_{cb}	h_{cb}	h_{fe}	α_{fb}	h_{21b}	α_{ce}	h_{ce}	h_{fb}	α_{fc}	h_{21c}	α_{eb}	h_{eb}	h_{fc}
4. Output conductance with input open	h_{oe}	h_{22e}	$\frac{1}{r_{22e}}$	h_{cc}	h_{oe}	h_{ob}	h_{22b}	$\frac{1}{r_{22b}}$	h_{cc}	h_{ob}	h_{oc}	h_{22c}	$\frac{1}{r_{22c}}$	h_{ee}	h_{oc}

Section IV. OPEN-CIRCUIT AND SHORT-CIRCUIT PARAMETERS

68. Open-Circuit Parameters

a. If the input and the output currents (fig. 50) were chosen as the independent variables, the general equations relating the dependent variables and the independent variables would be those given in paragraph 47*a*. If these equations are expanded, the following will result:

(1) Input circuit

$$v_{be} = r_{ie} i_b + r_{re} i_c$$

(2) Output circuit

$$v_{ce} = r_{fe} i_b + r_{oe} i_c$$

b. The equivalent circuit (input and output circuits combined) represented by these equations is shown in A, figure 68. The definitions for the coefficient of the equations are as follows:

(1) r_{ie} = input resistance with output open.

(2) r_{re} = reverse transfer resistance with input open.

(3) r_{fe} = forward transfer resistance with output open.

(4) r_{oe} = output resistance with input open.

c. The circuit setups required for measuring the coefficients (r_{ie}, r_{re}, r_{fe}, and r_{oe}) are shown in A, B, C, and D respectively of figure 69. Note that each measurement requires either the input or the output circuit to be open for ac; thus the term *open circuit parameters*. This condition might have been expected with two currents as the independent variables; to keep one constant while measuring the effect of the other, the circuit of the one to be kept constant must be ac open-circuited. Note that the coefficients are all resistance values; these parameters are also referred to as *resistance* (R) parameters. At high frequencies, reactive components change the resistance values to resistance and reactance, or impedance (Z); the parameters may also be referred to as *impedance* (Z) parameters.

d. For a better correlation of the physical structure of the transistor with its equivalent circuit, the equivalent circuit (A, fig. 68) can be transformed to that shown in B, figure 68. In this circuit r_b is the base resistance, r_e the emitter resistance, r_d the collector resistance, and r_m is called the mutual resistance (effect of base current on output voltage). The relationships between the parameters of the first equivalent circuit (A, fig. 68) and those of the second equivalent circuit (B, fig. 68) are as follows:

(1) $r_b = r_{ie} - r_{re}$.
(2) $r_e = r_{re}$.
(3) $r_d = r_{oe} - r_{re}$.
(4) $r_m = r_{re} - r_{fe}$.

Note. Equations relating the *open-circuit* parameters to the hybrid parameters are given in paragraph 72.

e. Typical values for the *open-circuit* parameters are as follows:

(1) r_b=370 ohms.
(2) r_e=23 ohms.
(3) r_d=58 kilohms.
(4) r_m=3 megohms.

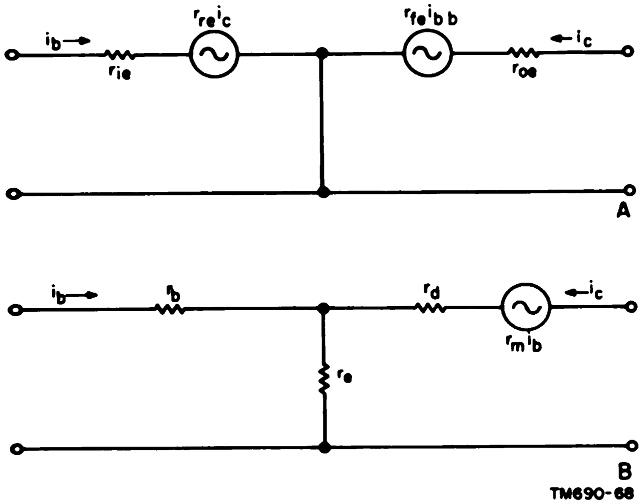


Figure 68. ***Equivalent circuits (common emitter configuration) using open-circuit parameters.***

69. Disadvantages of Open-Circuit Parameters

a. Measurement. To measure the input resistance with output open (A, fig. 69) and the forward transfer resistance with output open (C, fig. 69), the output circuit must be ac open-circuited. A dc path in the output circuit must be maintained to supply dc bias current and voltage. In practice, this can be achieved by placing a high-inductance coil in series with the output lead. However, at the frequency (270 cps) normally used to make the measurements, the required coil is large and cumbersome because its reactance must be much higher than the resistance of output circuit. The output circuit resistance varies from 30 kilohms to 4 megohms depending on the configuration. Accuracy under these conditions is very difficult to achieve. Measurement of the hybrid parameters (figs. 51 and 52) does not require open-circuiting the high-resistance output circuit. It requires ac short-circuiting the high-resistance output, which is easily achieved by a small capacitor. However, open-circuit parameters are easily measured at a high frequency because the capacitance (neglected at low frequencies) between the collector and the other elements substantially reduces the output impedance. For this reason, the equivalent circuit employing open-circuit parameters is sometimes used for analysis of high-frequency circuits.

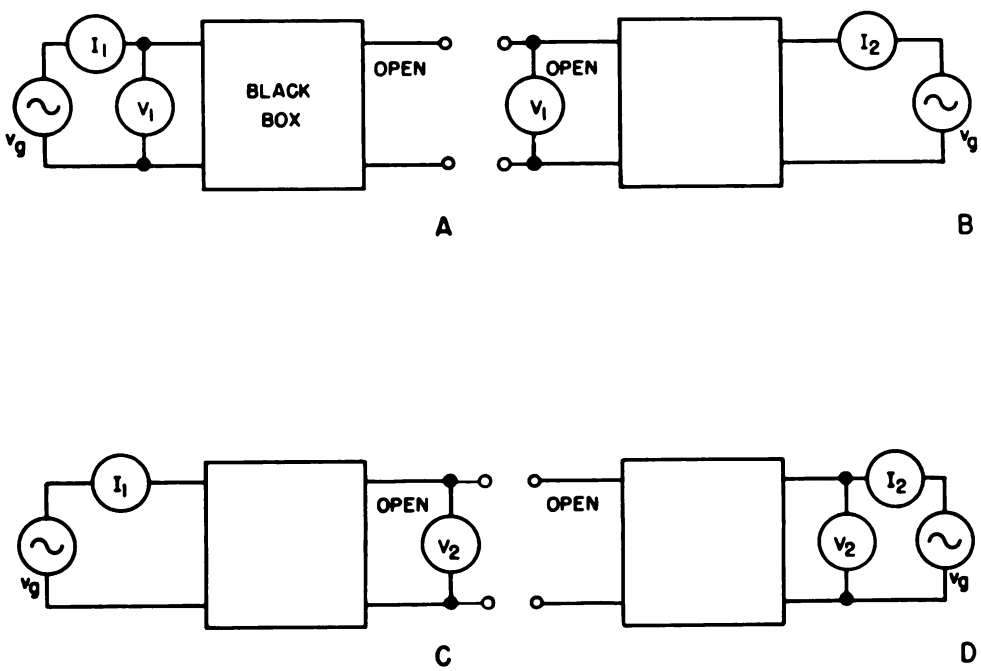


Figure 69. ***Test setups employing black box for measurement of open-circuit parameters.***

495900 O—59——6

b. Current Amplification Factor. The open-circuit parameters do not produce the current amplification factor directly. The hybrid parameters produce this factor (α_{fe}, α_{fb}, and α_{fc}) directly.

70. Short-Circuit Parameters

a. If the input and the output voltages (fig. 50) were chosen as the independent variables, the general equations relating the dependent variables and the independent variables would be those given in paragraph 47*b*. If these equations are expanded, the following equations result:

(1) Input circuit:

$$i_b = g_{ie}v_{be} + g_{re}v_{ce}$$

(2) Output circuit:

$$i_c = g_{fe}v_{be} + g_{oe}v_{ce}$$

b. The equivalent circuit (input and output circuits combined) represented by these equations is shown in A, figure 70. The definitions for the coefficients of the equations are as follows:

(1) g_{ie}=input conductance with output shorted.
(2) g_{re}= reverse transfer conductance with input shorted.
(3) g_{fe}=forward transfer conductance with output shorted.
(4) g_{oe}=output conductance with input shorted.

c. The circuit setups required for measuring the coefficients (g_{ie}, g_{re}, g_{fe}, and g_{oe}) are shown in A, B, C, and D respectively of figure 71. Note that each measurement requires that either the input or the output circuit be ac shorted; thus the term *short-circuit* parameters. This condition might have been expected with two voltages as the independent variables; to keep one constant while measuring the effect of the other, the circuit of the one to be kept constant must be ac short-circuited. Note that the coefficients are all conductance (reciprocal of resistance) values. These parameters are also referred to as *conductance* (*G*) *parameters.* At high frequencies, reactive components change the conductance values to conductance and susceptance (reciprocal of reactance), or to admittance (reciprocal of impedance); the parameters are also referred to as *admittance* (*Y*) *parameters.*

d. For a better correlation of the physical structure of the transistor to its equivalent circuit, the equivalent circuit (A, fig. 70) can be transformed to that shown in B, figure 70. In this circuit, g_{be} is the base-emitter conductance, g_{bc} is the base-collector conductance, g_{ce} is the collector-emitter conductance, and g_m is the mutual conductance (effect of base voltage on output current). The relationships of the parameters of the first equivalent circuit (A, fig. 70) and those of the second equivalent circuit (B, fig. 70) are as follows:

(1) $g_{be} = g_{ie} + g_{re}$.
(2) $g_{bc} = -g_{re}$.
(3) $g_{ce} = g_{oe} + g_{re}$.
(4) $g_m = g_{fe} + g_{re}$.

Equations relating the *short-circuit* parameters to the hybrid parameters are given in paragraph 72.

e. Typical values for the *short-circuit* parameters are as follows:

(1) $g_{be} = 0.64 \times 10^{-3}$ mho.
(2) $g_{bc} = 0.25 \times 10^{-6}$ mho.
(3) $g_{ce} = 0.40 \times 10^{-6}$ mho.
(4) $g_m = 0.033$ mho.

71. Disadvantages of Short-Circuit Parameters

a. Measurement. To measure the reverse transfer conductance with input shorted (B, fig. 71) and the output conductance with input shorted (D, fig. 71), the input circuit must be ac short-circuited. Connecting the base lead to the emitter lead to achieve the short is not practical because dc bias current and voltage must be supplied to the input circuit. In practice, the input circuit can be ac short-circuited by placing a capacitor across the input. However, at the frequency (270 cps) normally used to make the measurements, the required capacitor must be large and cumbersome because its reactance must be much lower than the resistance of the input circuit. The input circuit resistance is usually very low (30 ohms to 2 kilohms). Accuracy under these conditions is very difficult to achieve. Measurement of the hybrid parameters (figs. 51 and 52) does not require short-circuiting the low-resistance input circuit. They require ac open-circuiting the low-resistance input, which is easily achieved with a small inductive coil or resistor.

b. Current Amplification Factor. The short-circuit parameters do not produce the current amplification factor directly. The hybrid parameters produce this factor (α_{fe}, α_{fb}, α_{fc}) directly.

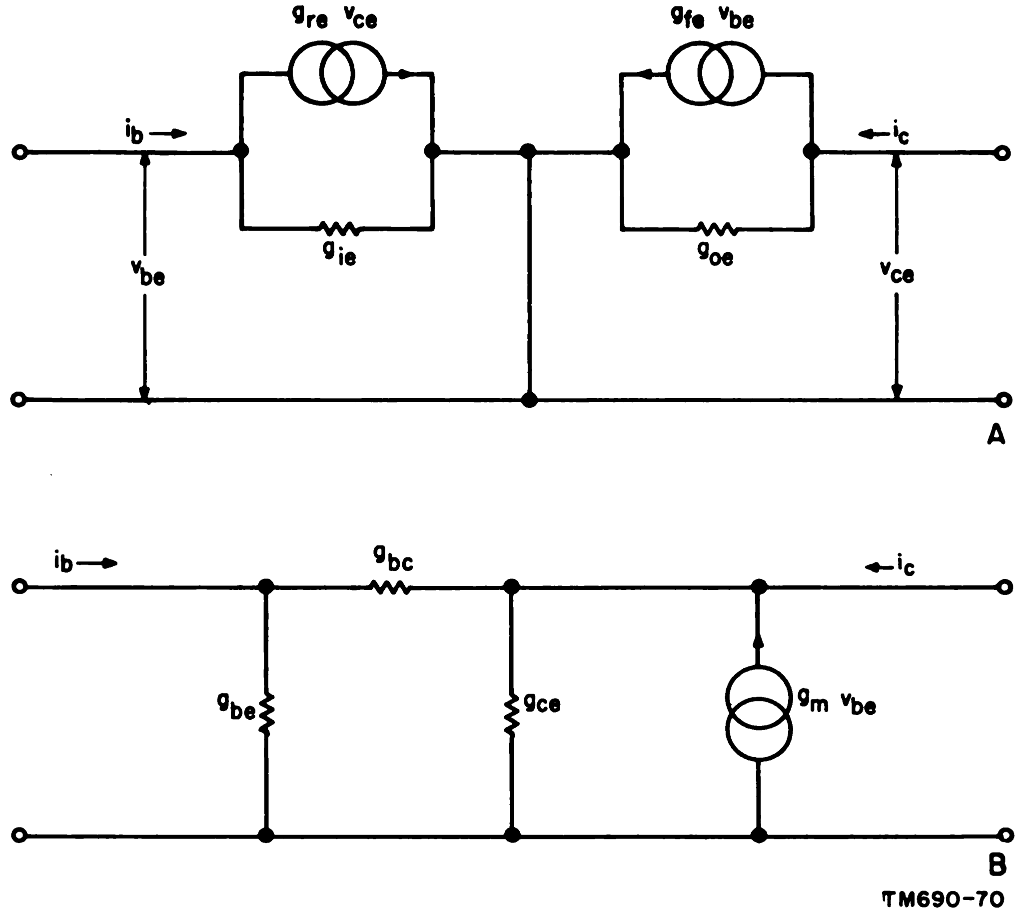


Figure 70. Equivalent circuits (common-emitter configuration) using short-circuit parameters.

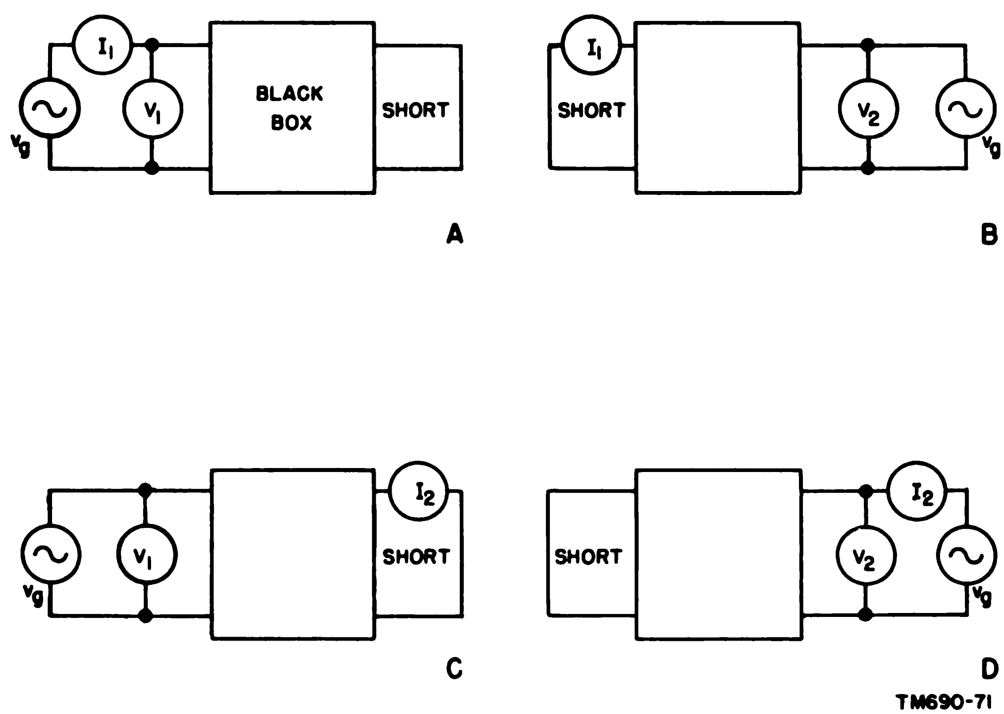


Figure 71. Test setup using black box for measurement of short-circuit parameters.

72. Open-Circuit, Short-Circuit, and Hybrid Parameter Conversion Formulas

In most instances, the values of the hybrid parameters will be given for a particular transistor. To determine the open-circuit parameters and the short-circuit parameters from the hybrid parameters, use the formulas in the first and second columns respectively of the following chart. Occasionally the open-circuit parameters are given; to determine the hybrid parameters from the open-circuit parameters, use the formulas in the third column.

From hybrid parameters to open-circuit parameters	From hybrid parameters to short-circuit parameters	From open-circuit parameters to hybrid parameters
$r_{ie}=h_{ie}-\frac{\mu_{re}\alpha_{fe}}{h_{oe}}$	$g_{ie}=\frac{1}{h_{ie}}$	$h_{ie}=r_{ie}-\frac{r_{re}r_{fe}}{r_{oe}}$
$r_{re}=\frac{\mu_{re}}{h_{oe}}$	$g_{re}=\frac{-\mu_{re}}{h_{ie}}$	$\mu_{re}=\frac{r_{re}}{r_{oe}}$
$r_{fe}=\frac{-\alpha_{fe}}{h_{oe}}$	$g_{fe}=\frac{\alpha_{fe}}{h_{ie}}$	$\alpha_{fe}=-\frac{r_{fe}}{r_{oe}}$
$r_{oe}=\frac{1}{h_{oe}}$	$g_{oe}=h_{oe}-\frac{\mu_{re}\alpha_{fe}}{h_{ie}}$	$h_{oe}=\frac{1}{r_{oe}}$

Note. The above formulas apply directly to the *CE* configuration. These formulas can be made to apply to the *CB* or the *CC* configuration by changing each subscript letter *e* to *b* or *c* respectively.

73. Summary

a. For purposes of mathematical analysis, devices such as crystals, electron tubes, and transistors, can be represented by equivalent circuits consisting of resistors, inductors, capacitors, and current and voltage generators.

b. The electrical quantities used in the equivalent circuit of a device (*a* above), are called the *parameters* of the device.

c. If the input current and the output voltage of the transistor are considered the independent variables when developing the transistor equivalent circuit, the resultant parameters (fig. 57) are a resistance (h_{ie}), a voltage generator ($\mu_{re}v_{ce}$), a current generator ($\alpha_{fe}i_b$), and a conductance (h_{oe}). Because of the *mixture* in types of parameters, these are called *hybrid* parameters.

d. The values of the parameters (*c* above), may be obtained from the related static characteristic curves (pars. 49–52).

e. By substituting the equivalent circuit (fig. 58) for the transistor, formulas for the following desired values may be derived:

(1) Current gain (par. 54*b*(6)).
(2) Voltage gain (par. 55*b*(10)).
(3) Power gain (par. 56*b*(4)).
(4) Input resistance (par. 57*b*(7)).
(5) Output resistance (par. 58*c*(10)).

f. If the hybrid parameters for a given configuration are known, the hybrid parameters for the other two configurations can be obtained (par. 66).

g. If the input and the output current of the transistor are considered the independent variables, an equivalent circuit (fig. 68) consisting of *open-circuit* (resistance or impedance) parameters is derived (par. 68).

h. The main disadvantage of the open-circuit parameters is that their measurement requires the ac open-circuiting of the *high* output resistance of the transistor (fig. 69).

i. If the input and the output voltages of the transistor are considered the independent variables, an equivalent circuit (fig. 70) consisting of *short-circuit* (conductance or admittance) parameters is derived (par. 70).

j. The main disadvantage of the short-circuit parameters is that their measurement requires the ac short-circuiting of the *low* input resistance of the transistor (fig. 71).

k. If the hybrid parameters for a given transistor are known, the open-circuit and the short-circuit parameters can be derived (par. 72).

CHAPTER 5

BIAS STABILIZATION

Section I. INTRODUCTION

74. General

Bias (operating point) is established for a transistor by specifying the quiescent (dc, no-signal) values of collector voltage and emitter current. Reliable operation of a transistor over a wide range of temperatures requires that bias voltage and current remain stable. However, variations of reverse-bias collector current (*a* below), and emitter-base junction resistance (*b* below), with temperature, preclude stable bias unless external compensating circuits (par. 75) are employed.

a. Reverse-Bias Collector Current (I_{CBO}).

(1) A detailed discussion of the internal mechanism that produces the reverse-bias collector (saturation) current is given in paragraph 43*b*(3). Variation of the saturation current with the temperature of the base-collector junction is shown in figure 72. The saturation current value varies from almost zero at 10° C to well over 1 milliampere at 125° C. *Note that at temperatures below 10° C, the saturation current causes no problem.*

(2) In an NPN transistor (fig. 47), saturation current consists of a flow of holes (minority carriers) from the collector region toward the base region. If the resistivity of the base region is high, or if external resistors connected to the base are high in value, holes from the collector can accumulate in the base region. Such an accumulation of holes will cause an increase of *emitter* electron flow into and through the base, increasing the collector current. Increased collector current would raise the temperature of the collector-base junction, and increase the saturation current (fig. 72). The cycle would continue until severe distortion occurs, the transistor becomes inoperative, or it destroys itself. *This condition can be minimized by avoiding the use of high-valued resistors in the base lead.*

(3) In a PNP transistor (fig. 48), the condition described in (2) above, is the same, except that the roles of holes and electrons are interchanged.

b. Emitter-Base Junction Resistance.

(1) Figure 73 indicates the variation of collector current with temperature. A *family* of curves is presented. Each curve is plotted with a *fixed* collector-base voltage (V_{CB}) and a *fixed* emitter-base voltage (V_{EB}). If the collector current variation with temperature were caused only by the saturation current (*a* above), then

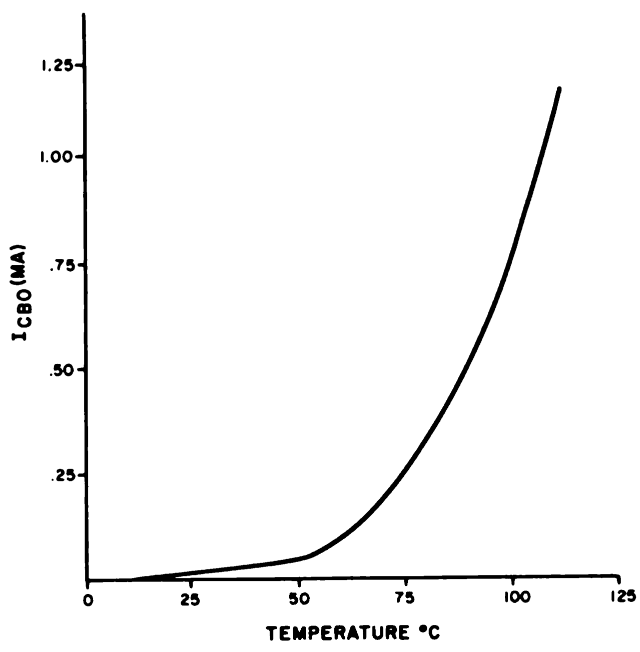


Figure 72. Graph showing variation of saturation current with junction temperature.

the collector current variation at temperatures below 10° C (V_{BB}=200 mv and V_{BB}=300 mv) should not accur (*a*(1) above). However, collector current does vary with temperature even when the saturation current is near zero. This variation is caused by the decrease in emitter-base junction resistance when the temperature is *increased*. That is, the emitter-base junction resistance has a *negative temperature-coefficient of resistance.*

(2) One method of reducing the effect of the negative temperature coefficient of resistance is to place a large valued resistor in the emitter lead. Essentially, this causes the variation of the emitter-base junction resistance to be a small percentage of the *total* resistance in the emitter circuit. The external resistor *swamps* (overcomes) the junction resistance; the resistor is referred to as a *swamping resistor.*

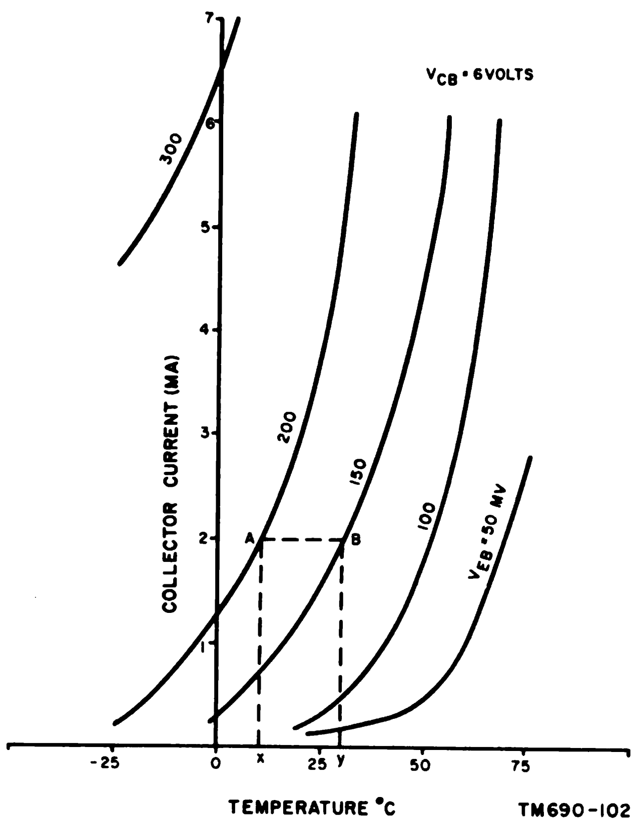


Figure 73. Graph showing variation of collector current with transistor temperature.

(3) A second method of reducing the effect of the negative temperature coefficient of resistance is to reduce the emitter-base forward bias as the temperature increases. For instance, to maintain the collector current at 2 ma (fig. 73) while the transistor temperature varies from 10° C (at *X*) to 30° C (at *Y*), the forward bias must be reduced from 200 mv (at *A*) to 150 mv (at *B*). The temperature difference (30−10) is 20° C; the voltage difference (200 mv−150 mv) is 50 mv. The variation in forward bias per degree centigrade is calculated as follows:

$$\frac{\text{Difference in forward bias}}{\text{Difference in temperature}} = \frac{50 \text{ mv}}{20° \text{ C}} = 2.5 \text{ mv/° C}$$

This calculation indicates that collector current will not vary with emitter-base junction resistance if the forward bias is reduced 2.5 mv/° C for increasing temperature, or increased 2.5 mv/° C for decreasing temperature.

75. Bias Stabilizing Circuit

Bias stabilizing circuits may employ a variety of circuit elements:

a. Common resistors (pars. 76–79). The effectiveness of external resistors in bias stabilization of transistor amplifiers is indicated by an analysis of the general bias circuit (par. 76). This analysis results in mathematical expressions for current and voltage stability factors (S_I and S_V respectively) for each amplifier configuration.

b. Thermistors (pars. 81–84).

c. Junction diodes (pars. 85–88).

d. Transistors (pars. 89–92).

e. Breakdown diodes (pars. 93–96).

Section II. RESISTOR STABILIZING CIRCUITS

76. General Bias Circuit

In paragraph 38, the basic transistor configurations (common base, common collector, and common emitter amplifiers) were discussed. These circuits are special cases of the general transistor circuit shown in A, figure 74. Depending upon the points of input and output of the signal into the general transistor circuit and the values assigned to the resistors, the three basic configurations can be derived.

Note. For a discussion of the currents shown flowing through the transistor leads, refer to paragraphs 40 through 43.

a. CB *Amplifier.* The *CB* amplifier (B, fig. 74) can be derived from the general transistor circuit by opening resistor R_F ($R_F=\infty$), and shorting resistor R_B ($R_B=0$). The input signal is introduced into the emitter-base circuit and extracted from the collector-base circuit.

b. CE *Amplifier.* The *CE* amplifier (C, fig. 74) can be derived by opening resistor R_F and shorting resistor R_B. In this case, the signal is introduced into the base-emitter circuit and extracted from the collector-emitter circuit.

c. CC *Amplifier.* The *CC* amplifier (D, fig. 74) can be derived by opening resistor R_F and shorting resistor R_C. The collector is common to the input and output circuits.

Note. Circuits in which R_F is not open are discussed in paragraph 79.

77. Current Stability Factor

(E, fig. 74)

a. The ratio of a *change in emitter current* (ΔI_E to a *change in saturation current* (ΔI_{CBO}) is a measure of the bias current stability of the transistor. This ratio indicates the effect on the emitter current of a change in saturation current and is called the current stability factor (S_I). Mathematically, it is expressed as follows:

$$S_I=\frac{\Delta I_E}{\Delta I_{CBO}}$$

Because the current stability factor is a ratio of two currents, it is expressed as a pure number. Under ideal conditions, the current stability factor would be equal to zero; in other words, the emitter current would not be affected by a change in the saturation current.

b. By using the resistors and the indicated expressions for current flow (A, fig. 74), current I_E can be expressed in terms of saturation current I_{CBO}, and the resistors used in the circuit. Analy-

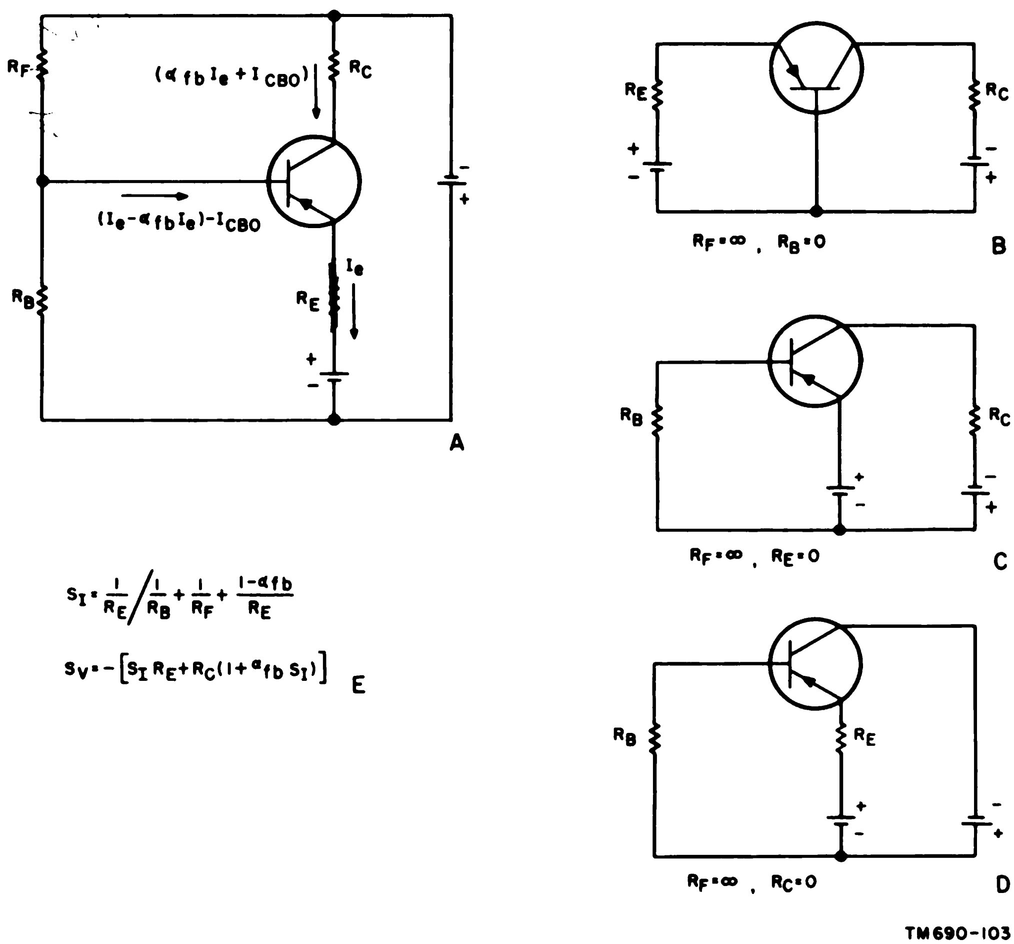


Figure 74. General transistor circuit and derived configurations including current and voltage stability factor formulas.

sis of such an expression to indicate how current I_E varies with saturation current I_{CBO} results in the expression for current stability factor (S_I).

$$S_I=\frac{1}{R_E}\Big/\left(\frac{1}{R_B}+\frac{1}{R_F}+\frac{1-\alpha_{fb}}{R_E}\right)$$

Note that the current stability factor evaluates the effectiveness of the external circuit to minimize emitter current variations.

c. To derive the *CB* amplifier (B, fig. 74), resistor R_F is opened $(=\infty)$ and resistor R_B is shorted $(=0)$. To obtain the stability factor (*b* above), for this *CB* amplifier, substitute these values in the equation for S_I (note that $\frac{1}{\infty}=0$, and $\frac{1}{0}=\infty$):

$$S_I=\frac{1}{R_E}\Big/\left(\infty+0+\frac{1-\alpha_{fb}}{R_E}\right)$$

With an infinite denominator, the expression for S_I equals zero. As expected, this is the ideal condition since it minimizes the accumulation of minority carriers in the base region (par. 74*a*(2)), and permits the use of an emitter swamping resistor (par. 74*b*(2)).

d. Directly substituting (for the *CE* amplifier (C, fig. 74)) the values of resistor R_F $(=\infty)$ and resistor R_B $(=0)$ in the equation for S_I (*b* above), the ratio of infinity (∞) divided by infinity (∞) is obtained. This ratio results in an indeterminate value which cannot be interpreted. To avoid reducing the results to an indeterminate value, multiply the numerator and denominator by R_E before substituting the values, to obtain:

$$S_I=\frac{1}{\frac{R_E}{R_B}+\frac{R_E}{R_F}+1-\alpha_{fb}}$$

Substituting the values of resistor R_B $(=0)$ and resistor R_F $(=\infty)$, in the equation, then:

$$S_I=\frac{1}{\frac{0}{RB}+\frac{0}{\infty}+1-\alpha_{fb}} \text{ and}$$

$$S_I=\frac{1}{1-\alpha_{fb}}$$

Because α_{fb} is close to unity, the value of S_I is high, indicating very poor current stability. This condition might have been expected because there is no swamping resistor $(R_E=0)$, and resistance is used in the base lead (R_B not zero).

e. Substituting (for the *CC* amplifier (D, fig. 74)) the values of resistor R_F $(=\infty)$ into the equation (*b* above).

$$S_I=\frac{1}{R_E}\Big/\left(\frac{1}{R_B}+0+\frac{1-\alpha_{fb}}{R_E}\right)$$

Since $1-\alpha_{fb}$ is approximately equal to $(\cong)$ zero:

$$S_I\cong\frac{1}{R_E}\Big/\frac{1}{R_B}$$

$$S_I\cong\frac{R_B}{R_E}$$

This expression indicates that the current stability of the *CC* amplifier depends upon the ratio of base resistance to emitter resistance. Again it is shown that the higher the base resistance, the poorer the current stability factor (par. 74*a*(2)), and the higher the emitter resistance, the better the current stability factor (par. 74*b*(2)).

78. Base Resistance and Emitter Resistance

The effect on collector current stability of various combinations of base and emitter resistance values is shown by the graph in figure 75. The worst condition occurs if the emitter and the base resistances are both zero (curve AA). Only slight improvement occurs with the base resistance equal to 40K ohms and the emitter resistance equal to zero (curve BB). The best condition occurs, assuming linear resistors are used, if the emitter resistance is greater than zero (2K ohms in this case), and if the base resistance is zero (curve CC). This series of curves further substantiates the statements derived from a purely physical consideration of the transistor (par. 74*a* (2) and *b*(2)).

79. Additional Bias Techniques

a. One method of obtaining good current stability in a *CE* amplifier is by using near-zero base resistance and an emitter swamping resistor as shown in A, figure 76. Resistor R_E is the swamping resistor; the secondary of transformer T1 offers a very low dc resistance in the base circuit. The collector current stability curve for this amplifier is similar to that of curve *CC*, figure 75. The current stability factor for this circuit is equal to zero (ideal) as discussed in paragraph 77*c*.

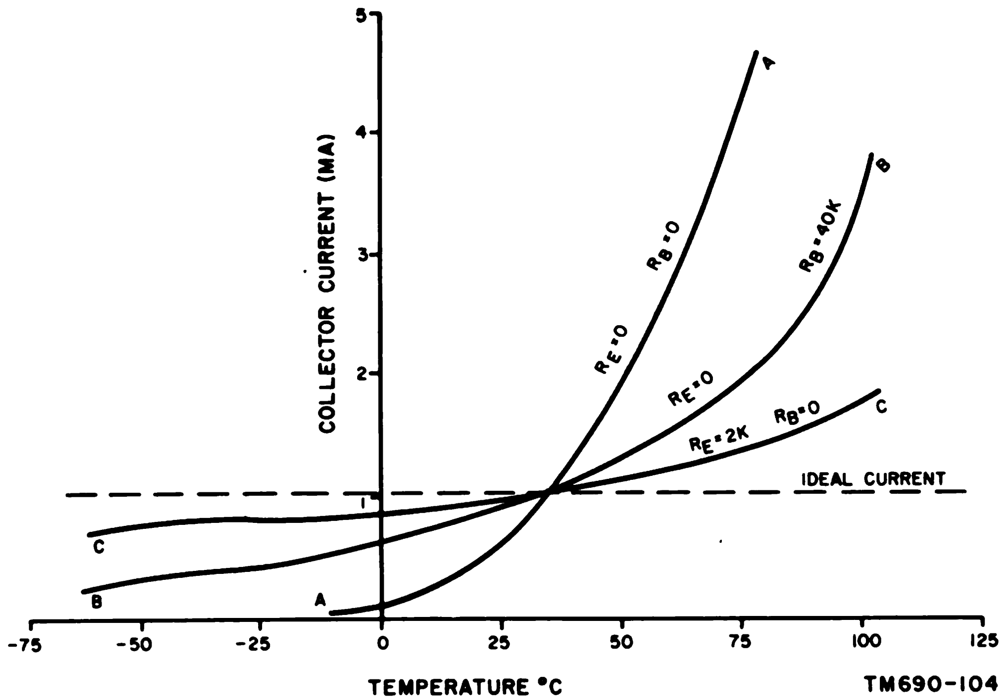


Figure 75. Graph of variation of collector current with temperature, using different values of emitter resistance and base resistance.

b. Use of a fixed emitter-base bias can be obtained in a single-battery *CE* amplifier by means of a voltage divider network (B, fig. 76). The voltage divider network consists of resistors R_F and R_B. The voltage developed across resistor R_B contributes part of the emitter-base voltage. Consider this circuit with reference to the currrent stability factor:

$$S_I=\frac{1}{R_B}\Big/\left(\frac{1}{R_B}+\frac{1}{R_F}+\frac{1-\alpha_{fb}}{R_E}\right)$$

Since $1-\alpha_{fb}$ can be considered equal to zero ($\alpha=$ 0.92 to 0.99), then:

$$S_I=\frac{1}{R_E}\Big/\left(\frac{1}{R_B}+\frac{1}{R_F}\right), \text{ and}$$

$$S_I=\frac{R_BR_F}{R_B+R_F}\Big/R_E$$

This formula states that the current stability factor in this circuit is equal to the ratio of the parallel resistance value of R_B and R_F (base ground-return resistance) to the emitter resistance. This formula again substantiates the statement that the lower the base ground-return resistance and the higher the emitter resistance, the better the current stability (par. 74).

c. A circuit employing negative feedback voltage to improve current stability is shown in C of figure 760. If the collector current (I_C) rises, the collector becomes less negative because of the larger dc drop in resistor R_C. As a result, less forward bias (negative base to positive emitter) is coupled through resistor R_F to the base. Reduced forward bias then reduces the collector current.

80. Voltage Stability Factor

(E, fig. 74)

a. The ratio of an *increment in collector voltage* (ΔV_{CB}) to an *increment in saturation current* (ΔI_{CBO}) is a measure of the collector voltage stability of the transistor. This ratio indicates the effect on the collector voltage of a change in saturation current and is called the *voltage stability factor* (S_V). Mathematically it is expressed as follows:

$$S_V=\frac{\Delta V_{CB}}{\Delta I_{CBO}}$$

Because the voltage stability factor is a ratio of a voltage and a current, the voltage stability factor is expressed as a *resistance* (ohms). Under

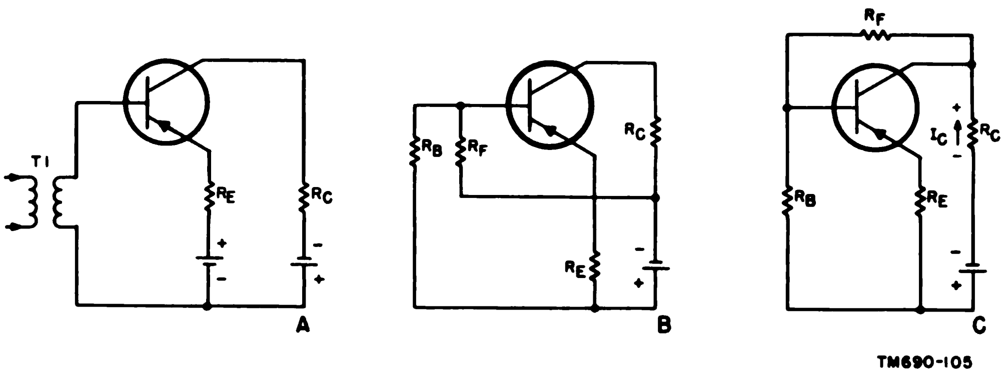


Figure 76. CE amplifiers employing transformer input, fixed base voltage, and negative feedback.

ideal conditions, the voltage stability factor would be equal to zero; in other words, the collector voltage would not be affected by a change in the saturation current.

b. By using the resistors and the indicated expressions for current flow (A, fig. 74):

$$S_V = -[S_I R_B + R_C(1+\alpha_{fb} S_I)]$$

Note that if the current stability factor (S_I) is zero (pars. 77*c* and 79*a*), the voltage stability factor (S_V) equals the collector resistance (R_C). Use of a transformer with a low-resistance primary in the collector circuit tends to reduce the voltage stability factor to near zero. Circuits designed to minimize collector voltage variations are covered in paragraphs 94 and 96.

Section III. THERMISTOR STABILIZING CIRCUITS

81. General

It has been established that the bias current of the transistor is temperature-sensitive (par. 74). Specifically, emitter current increases with an increase in temperature. Emitter current stabilization can be achieved by use of external circuits using temperature-sensitive elements. There are several temperature-sensitive electrical elements. One such element is the thermistor (contraction of the words *thermal* and *resistor*). The thermistor, as used in this chapter, has a *negative* temperature coefficient of resistance; that is, its resistance value *decreases* with an *increase* in temperature. The use of thermistors in transistor temperature-stabilizing circuits is discussed in paragraphs 82 and 83.

82. Emitter Voltage Control

a. The circuit shown in figure 77 employs a thermistor to vary the emitter voltage with temperature to minimize temperature variations in emitter current. This circuit contains two voltage dividers, the first consisting of resistors R4 and R1, and the second consisting of resistor R2 and thermistor RT1. The first voltage divider permits the application of a portion of battery V_C voltage to the base terminal and ground (common return). The base terminal voltage is developed across resistor R1 and is in the *forward* bias direction. The second voltage divider applies a portion of battery V_C voltage to the emitter terminal. The emitter terminal voltage is developed across resistor R2 and is in the reverse bias direction. The forward bias voltage applied to the base terminal is larger than the reverse bias applied to the emitter terminal, so that the resultant base-emitter bias is always in the forward direction.

b. With an increase in temperature, the collector current would normally increase if the transistor were not stabilized. The increase in collector current can be prevented by reducing the forward bias. This is accomplished by the voltage divider consisting of resistor R2 and thermistor RT1. As the temperature increases, the resistance of thermistor RT1 is decreased, causing more current to flow through the voltage divider. The increased current raises the negative potential at the emitter connection of resistor R2. This action increases the *reverse* bias applied to the emitter and decreases the net emitter-base forward bias. As a result, the collector current is reduced. Similarly, decreasing the tempera-

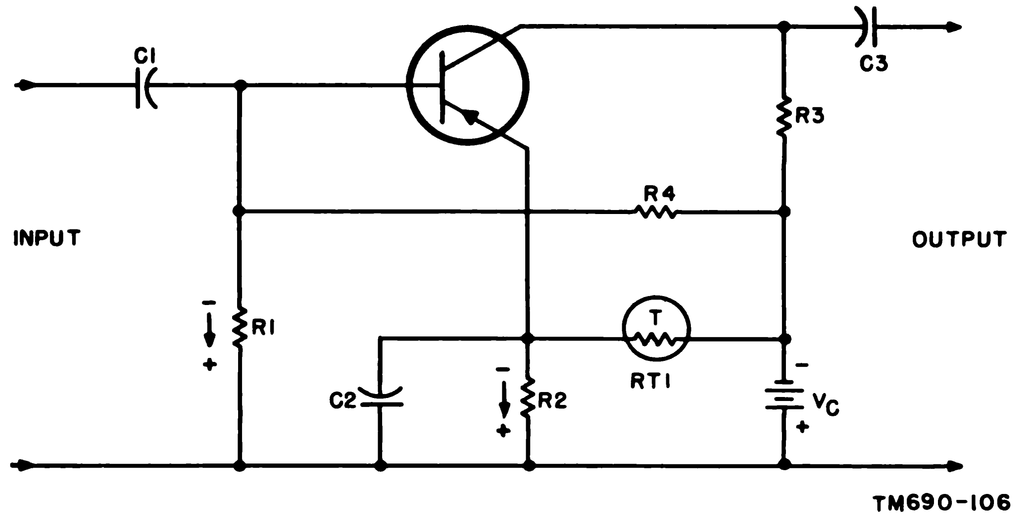


Figure 77. Transistor amplifier with thermistor control of emitter bias voltage.

ture would cause the reverse actions, and prevent the decrease of collector current.

c. Capacitor C1 blocks the dc voltage of the previous stage and couples the ac signal into the base-emitter circuit. Capacitor C2 bypasses the ac signal around resistor R2. Resistor R3 is the collector load resistor and develops the output signal. Capacitor C3 blocks the dc collector voltage from and couples the ac signal to the following stage.

83. Base Voltage Control

a. The circuit shown in figure 78 employs a thermistor to vary the base voltage with temperature to minimize temperature variations in emitter current. This current contains a voltage divider consisting of resistor $R1$ and thermistor $RT1$. The voltage divider applies a portion of battery V_C voltage to the base-emitter circuit. Electron current flow through the voltage divider is in the direction of the arrow. This current produces a voltage of the polarity indicated across thermistor $RT1$. This circuit produces a forward bias.

b. If the temperature of the transistor rises, the emitter current would tend to rise. However, the resistance of thermistor $RT1$ decreases with increase in temperature, causing more current to flow through the voltage divider. Increased current through the voltage divider causes a larger portion of battery V_C voltage to be dropped across resistor $R1$. The available voltage for forward bias, that cross thermistor $RT1$, is reduced, thereby reducing the emitter current.

c. Transformer $T1$ couples the ac signal into the base-emitter circuit. Capacitor $C1$ bypasses the ac signal around thermistor $RT1$. Transformer $T2$ primary acts as the collector load and develops the output signal which is coupled to transformer $T2$ secondary.

84. Thermistor Limitations

The ability of the thermistor (figs. 77 and 78) to limit the variation of collector current with temperature is represented in figure 79. One curve shows the variation of collector current for a transistor circuit that is not stabilized. Another curve shows the variation of collector current for a transistor circuit that is thermistor-stabilized. There is improvement in the stability of the transistor circuit employing the thermistor. However, thermistor stabilization achieves *ideal current* at only three points (A, B, and C, fig. 79) because the thermistor resistance variation is not equal to (does not *track*) the variation in emitter-base junction resistance (par. 74*b*).

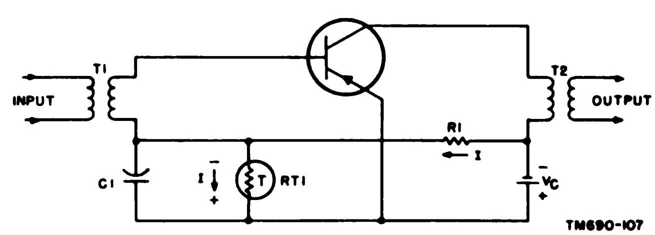


Figure 78. Transistor amplifier with thermistor control of base bias voltage.

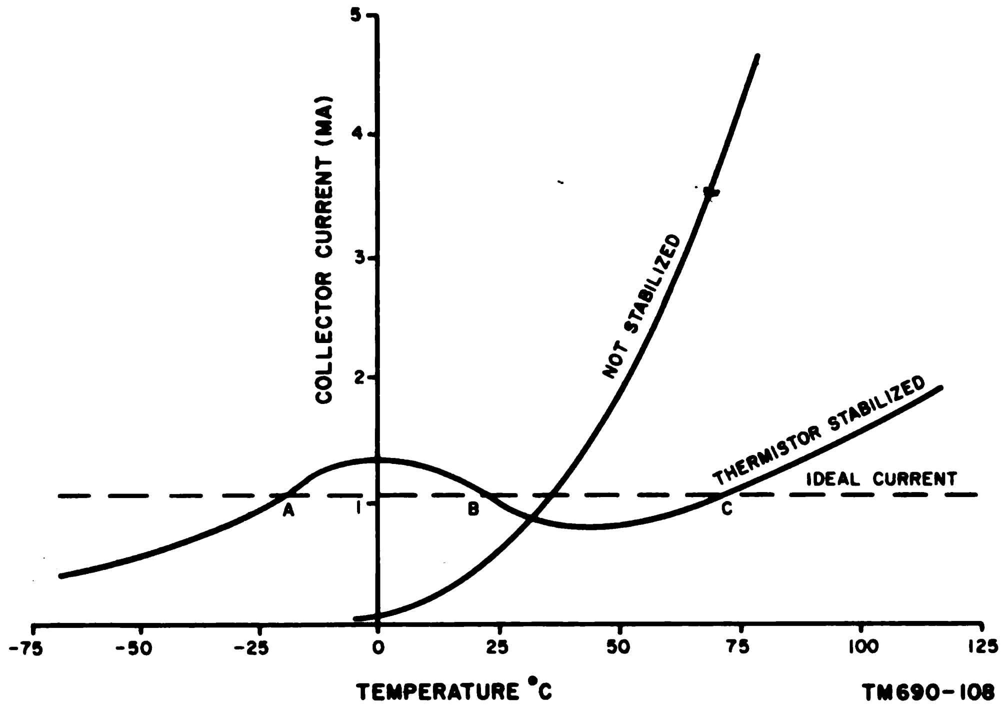


Figure 79. Graph showing variation of collector current with temperature for nonstabilized and thermistor-stabilized transistor circuits.

Section IV. DIODE STABILIZING CIRCUITS

85. General

a. Variation of collector current with temperature is caused by the variation with temperature of the emitter-base junction resistance and the saturation (reverse bias) current (par. 74). Variation with temperature of the resistance and the reverse bias current of a PN junction occurs whether the PN junction is part of a transistor or part of a junction diode. The junction diode, then like the transistor, can be used in bias stabilizing circuits. The main advantage of using the junction diode as a temperature-sensitive element is that it can be made of the same material as that of the transistor. The temperature coefficient of resistance of the diode and that of the transistor of the same material are the same. This condition permits a more constant collector current over a wide range of temperatures because of better *tracking* (par. 84).

b. Junction diodes have a negative temperature coefficient of resistance whether they are forward or reverse biased. The use of junction diodes in transistor temperature-stabilizing circuits is discussed in paragraphs 86 through 88.

86. Forward-Biased Single Diode Stabilization

a. The circuit shown in figure 80 employs forward-biased junction diode CR1 as a temperature-sensitive element to compensate for variations of emitter-base junction resistance. Consider the voltage divider consisting of resistor R1 and junction diode CR1. The current (I) through the voltage divider flows in the direction shown, and develops a voltage across diode CR1 with polarity as indicated. This voltage is a *forward* bias. With increased temperature the collector current would tend to increase. However, increased temperature decreases the resistance of diode CR1, causing more current to flow through the voltage divider. As a result there is an increased voltage drop across resistor R1. The voltage drop across diode CR1 is correspondingly decreased, thereby reducing the forward bias and the collector current.

b. The ac signal is coupled into the transistor amplifier by transformer T1. Capacitor C1 bypasses the ac signal around diode CR1. Collector load resistor R2 develops the output signal. Ca-

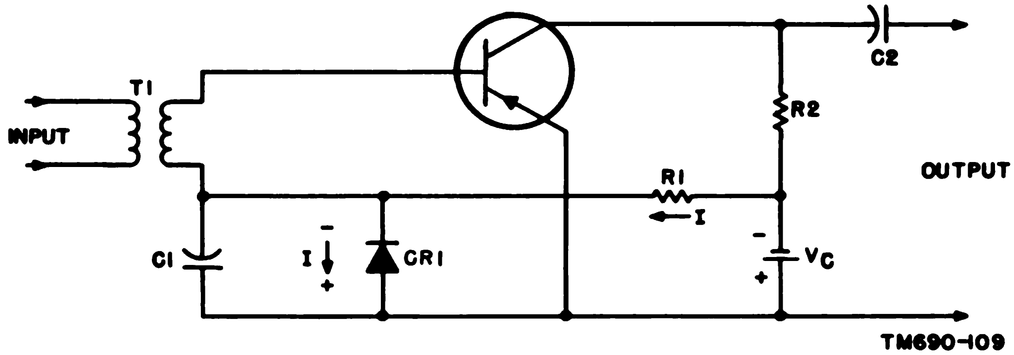


Figure 80. Transistor amplifier using single forward-biased junction diode for compensation of temperature variations of emitter-base junction resistance.

pacitor C2 blocks the dc voltage from and couples the ac signal to the following stage.

c. The effectiveness of this circuit to stabilize collector current with temperature is indicated by curve *BB* of figure 82. Compare this curve with curve *AA*, for which the transistor is not stabilized, and with the curve for ideal current. Curve *BB* shows that a marked improvement in the collector current stability occurs for temperature below 50° C. This indicates that the variations with temperature of the junction diode resistance tracks and compensates for the variations of emitter-base junction resistance (par. 74*b*). The sharp increase in collector current (curve *BB*) at temperature above 50° C indicates that junction diode *CR*1 *does not* compensate for the increase in saturation current (I_{CBO}) (par. 74*a*). This condition might have been anticipated since the saturation current (collector-base reverse bias current) flows out of the base, through *T*1 primary, through diode *CR*1, battery *VC*, and back to the collector. Because the saturation current is a small percentage of the total current through diode *CR*1 (forward biased and low in resistance) it causes no appreciable voltage drop across diode *CR*1. A circuit using a second junction diode (reverse biased) is necessary to compensate for the saturation current and is discussed in paragraph 87.

87. Double Diode Stabilization

a. The circuit shown in figure 81 employs two junction diodes as temperature sensitive elements. One junction diode compensates for the temperature variations of emitter-base junction resistance; the other compensates for the temperature variations of saturation current. The circuit is similar to that shown in figure 80 and discussed in paragraph 86. Resistor R3 and junction diode CR2 (reverse biased) have been added (fig. 81). Resistor R1 and junction diode CR1 (forward biased) compensate for the change in emitter-base junction resistance at temperatures below 50° C. (See curve BB, fig. 82.)

b. Reverse-biased junction diode *CR*2 can be considered an open circuit at low temperatures. At room temperature, diode *CR*2 reverse bias current I_s flows through junction diode *CR*2 in the direction indicated. Diode *CR*2 is selected so that its reverse bias current (I_s) is larger than that (I_{CBO}) of the transistor. Diode reverse-bias current I_s consists of transistor reverse bias current I_{CBO} and a component of current (I_1) drawn from

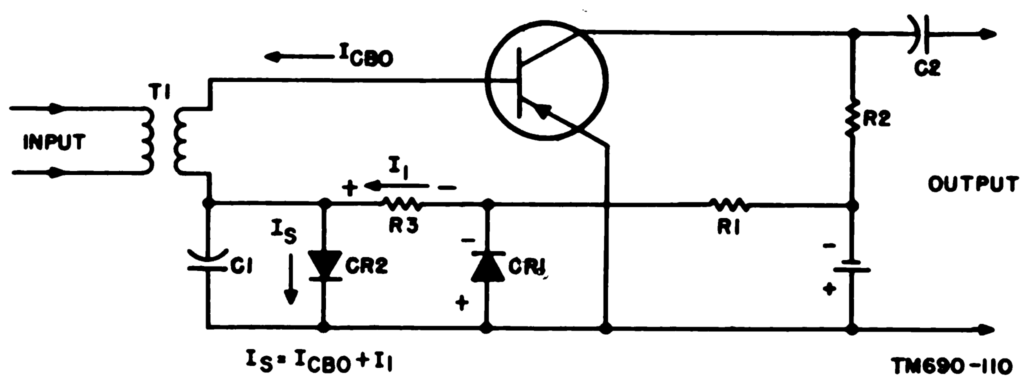


Figure 81. Transistor amplifier using two junction diodes for compensation of temperature variations of both emitter-base junction resistance and saturation current.

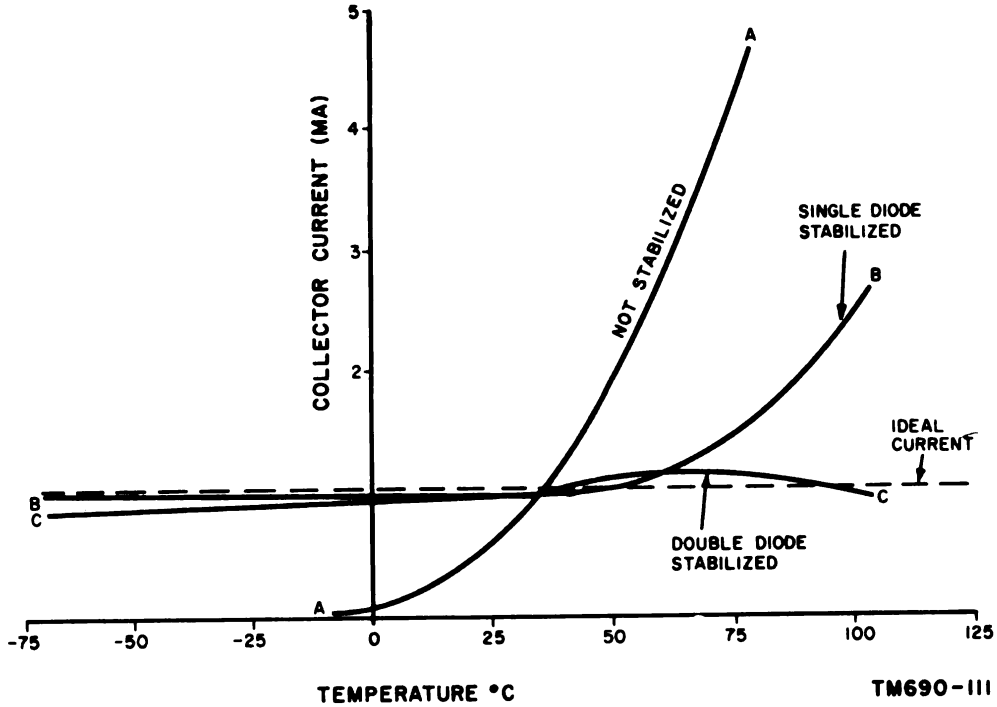


Figure 82. Graph showing variation of collector current with temperature for nonstabilized single-diode stabilized, and double-diode stabilized transistor circuits.

the battery. The voltage polarity developed by current I_1 across resistor $R3$ is indicated. Note that the emitter-base bias voltage is the sum of the opposing voltages across resistor $R3$ and junction diode $CR1$, assuming negligible resistance in transformer $T1$ secondary. As the temperature increases, current I_{CBO}, I_s, and I_1 increase. The resultant *reverse-bias* voltage developed across resistor $R3$ by current I_1 increases. The total *forward* bias (voltage across diode $CR1$ and resistor $R3$) decreases with increasing temperature to stabilize the collector current.

c. The functions of transformer T1, capacitors C1 and C2, and resistor R2, are the same as those for the corresponding elements of figure 80.

d. The effectiveness of this circuit to stabilize collector current at high and low temperatures is indicated by curve CC of figure 82.

88. Reverse-Biased Single-Diode Stabilization

a. The circuit shown in figure 83 employs reverse-biased junction diode $CR1$ as a temperature-sensitive element. This circuit provides two separate paths for the two components of the base current (par. 43). The base-emitter current $(I_e - \alpha_{fb} I_e)$ flows through the base region to the emitter, through resistor $R2$, battery V_C, and resistor $R1$ to the base lead. The saturation current (I_{CBO}) flows from the base lead through junction diode $CR1$, battery V_B, battery V_C, resistor $R3$, and the collector region to the base region. The diode is selected so that its saturation (reverse bias) current equals that of the transistor over a wide range of temperature.

b. As the temperature increases, I_{CBO} increases. However, the saturation current of diode $CR1$ increases by an equal amount, so that there is no accumulation of I_{CBO} current carriers in the base region. Such an accumulation would cause an increase in emitter current (par. 74*a*). Diode $CR1$ acts as a gate, opening wider to accommodate the increase in I_{CBO} with temperature. This circuit is employed if resistance-capacitance coupling is used between it and the previous stage. The reverse-biased diode provides high resistance in the input.

c. Resistor $R2$ swamps the emitter-base junction resistance, and prevents a large increase in emitter current, particularly at the low temperatures (par. 74*b*). Resistor $R3$ is the collector load resistor and develops the output ac signal.

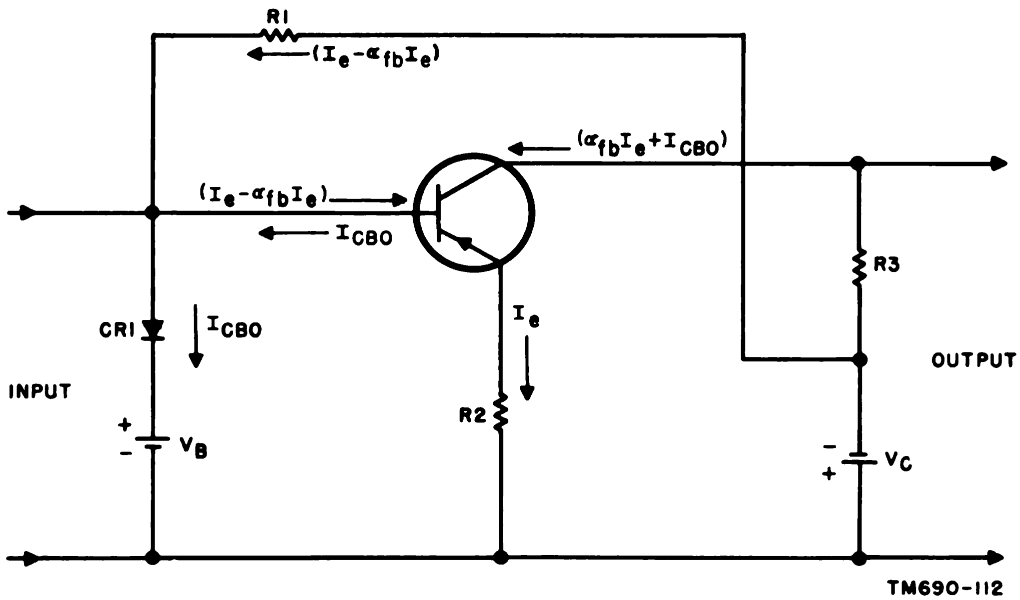


Figure 83. Transistor amplifier using single reverse-biased junction diode for compensation of temperature variations of saturation current.

Section V. TRANSISTOR STABILIZED CIRCUITS

89. General

Certain currents and voltages of a temperature-stabilized transistor may be used to temperature-stabilize another transistor or several transistors. Typical circuit arrangements employing this method are covered in paragraphs 90 through 92.

90. Common Emitter-Base Voltage

a. The emitter-base junction of a transistor has a negative temperature coefficient of resistance similar to that of a PN junction diode (par. 85). It is possible to use the variations of the emitter-base junction resistance of one transistor to control the emitter-base bias of a second transistor. A circuit employing this principle is shown in figure 84. The emitter-base voltage of transistor $Q1$ is used to bias the emitter-base junction of transistor $Q2$. Assuming zero dc resistance in the secondary winding of transformer $T2$, the base of transistor $Q2$ has a direct dc connection to the emitter of transistor $Q1$. Assuming zero dc resistance in the secondary of transformer $T1$, the emitter of transistor $Q2$ has a direct dc connection to the base of transistor $Q1$. Battery V_E provides forward bias for transistor $Q1$, an NPN type. Cross connection of the emitters and the bases of the two transistors provides forward bias for transistor Q2, because transistor $Q2$ is a PNP type

b. Transistor $Q1$ is temperature-stabilized by use of high-valued resistor $R1$ (swamping resistor) in the emitter lead, and low dc resistance in the base lead. This circuit is similar to the circuit shown in A, figure 76 and discussed in paragraph 79*a*. Swamping resistor $R1$ maintains a relatively constant emitter current in transistor $Q1$.

c. As the temperature increases, the base-emitter junction resistance of transistor $Q1$ decreases. Because the current through the junction remains constant, the voltage drop across the junction decreases. A decrease in this voltage represents a decrease in the forward bias of transistor $Q2$. The tendency of transistor $Q2$ collector current to increase with the temperature increase is offset.

d. If a curve of transistor $Q2$ collector current versus temperature were plotted, it would be similar to the double-diode stabilization curve (curve CC, fig. 82).

e. In the circuit of figure 84, transformer $T1$ couples the ac signal input to the base-emitter circuit of transistor $Q1$. Battery V_E supplies emitter-base bias voltage for transistor $Q1$ and collector voltage for transistor $Q2$. Resistor $R1$ is the emitter swamping resistor. Capacitor $C1$ bypasses the ac signal around resistor $R1$ and battery V_E. Battery V_C supplies transistor $Q1$ collector voltage. Transformer $T2$ couples the ac

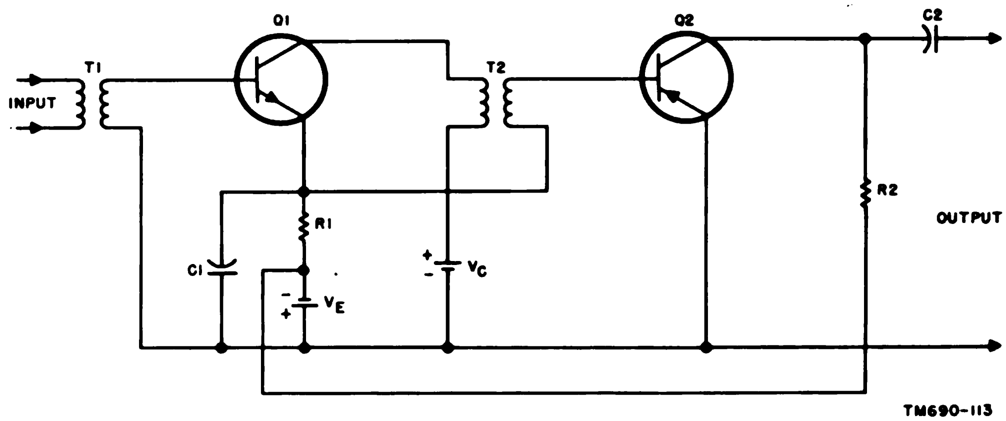


Figure 84. Circuit showing temperature stabilization of one transistor (Q2) *by the emitter-base voltage of another transistor* (Q1).

signal output of transistor $Q1$ to the input circuit of transistor $Q2$. Resistor $R2$ is the transistor $Q2$ collector load resistor and develops the output signal. Capacitor $C2$ blocks the dc voltage from and couples the ac signal to the following stage.

91. Common Emitter-Collector Current

a. A method of temperature-stabilizing the emitter-collector current of one transistor by using the stabilized emitter-collector current of another transistor is shown in figure 85. Transistor $Q1$ collector current is stabilized by use of swamping resistor $R2$. Resistor $R1$ is selected so that its value is low, thus limiting the accumulation of saturation current carriers in the base region (par. 74*a*).

b. The stabilized dc collector current of transistor $Q1$ is made to flow through the emitter-collector of transistor $Q2$ by connecting the collector of transistor $Q1$ directly to the emitter of transistor $Q2$. The current of this circuit is indicated by the arrows. Electrons that flow into the emitter lead of transistor $Q1$ flow through resistor $R2$, battery V_C, resistor $R4$, collector-emitter of transistor $Q2$, and into the collector of transistor $Q1$.

c. The advantage of this method of stabilizing the collector current of transistor $Q2$ is that it eliminates the need for a swamping resistor in the emitter lead of transistor $Q2$. The elimination of such a swamping resistor is especially important in power amplifiers that draw heavy emitter currents.

d. The functions of the circuit elements are as follows:

(1) Note that transistor $Q1$ is used in a *CC* amplifier and that transistor $Q2$ is used in a *CE* amplifier. Capacitor $C3$ places the collector of transistor $Q1$ and the emitter of transistor $Q2$ at ac zero potential.

(2) Capacitor $C1$ blocks the dc voltage from the previous stage and couples the ac signal to the base of transistor $Q1$. Resistor $R1$ provides a dc return path for base current. Battery V_{B1} provides base-emitter bias. Resistor $R2$ is the emitter load resistor and develops the output signal. Capacitor $C2$ blocks the emitter dc voltage from and couples the ac signal voltage to the base of transistor $Q2$.

(3) Resistor $R3$ provides a dc return path for base current. Battery V_{B2} provides base bias. Resistor $R4$ is the collector load resistor and develops the output signal. Battery V_C provides voltage for both transistors.

92. Bias Variation With Collector Current

a. Figure 86 shows two stages of a dc (direct-coupled) amplifier. A dc amplifier amplifies dc voltage and very low-frequency ac signals.

b. This circuit is arranged so that an increase in collector current caused by a temperature rise in transistor $Q1$ will reduce the forward bias in transistor $Q2$.

c. The circuit operates as follows:

(1) Transistor $Q1$ is used in a *CB* amplifier; its stability factor is ideal (par. 77*c*). However, some variation of its collector current does occur with temperature variation. Assume an increment in collector current (ΔI_C) occurs because of an increase in temperature. The increase in

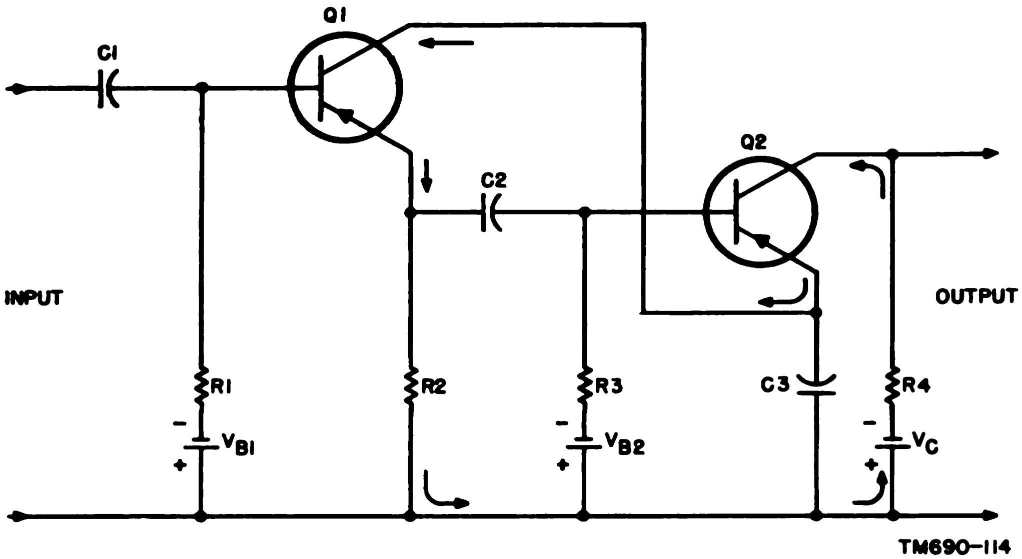


Figure 85. Circuit using common emitter-collector current for temperature stabilization, in two transistors.

collector current is in the direction of the arrow. A portion of the increment in collector current flows through resistor $R3$. This current is designated ΔI_{C1} and develops a voltage across resistor $R3$ with the polarity indicated. Another portion of the increment in collector current flows through resistor $R2$. This current is designated ΔI_{C2} and develops a voltage across resistor $R2$ with the polarity indicated. Note that the voltage polarities indicated are only for an *increment* in collector current, and are not necessarily the steady-state voltage polarities. The steady-state voltages are of no interest in this discussion and will not be treated.

(2) Consider the base-emitter bias circuit of transistor $Q2$. The base-emitter bias is the sum of the voltages across resistors $R3$, $R2$, and battery V_C. The indicated voltage across resistor $R3$ *aids* the forward bias; that across resistor $R2$ *opposes* the forward bias. By selecting the values of resistors $R2$ and $R3$ so that the voltage drop across resistor $R2$ is the larger, the resultant forward bias is decreased. This action limits the tendency of transistor $Q2$ collector current to increase with temperature.

d. In this circuit, resistor $R1$ is the emitter swamping resistor. Battery V_E supplies emitter bias voltage. Battery V_C supplies collector voltage for both transistors. Resistor $R4$ is the collector load resistor.

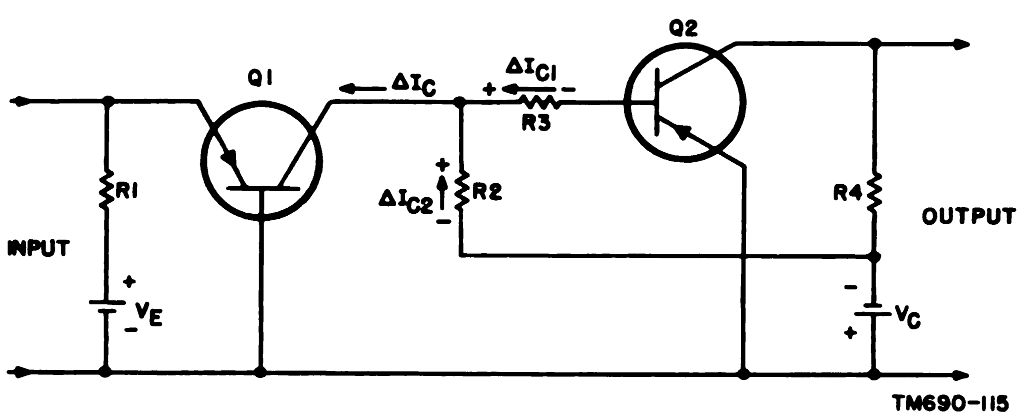


Figure 86. Two-stage temperature-stabilized dc amplifier.

93. General

a. In paragraph 80, the voltage stability factor (S_V) was explained and expressed by the following equation:

$$S_V = -[S_I R_E + R_C(1+\alpha_{fb} S_I)]$$

This equation shows that the voltage stability factor is directly proportional to the current stability factor (S_I). Therefore, the circuit techniques (pars. 77–92) for improving the current stability factor will also result in improvement of the voltage stability factor; that is, limiting the variations of collector voltage with temperature variations of saturation current.

b. Circuits used to stabilize the collector supply voltage are discussed in paragraphs 94 through 96.

94. Breakdown Diode Voltage Regulator

a. A graph of the current through and the voltage across a reverse-biased junction diode is shown in A, figure 87. Note that at a certain

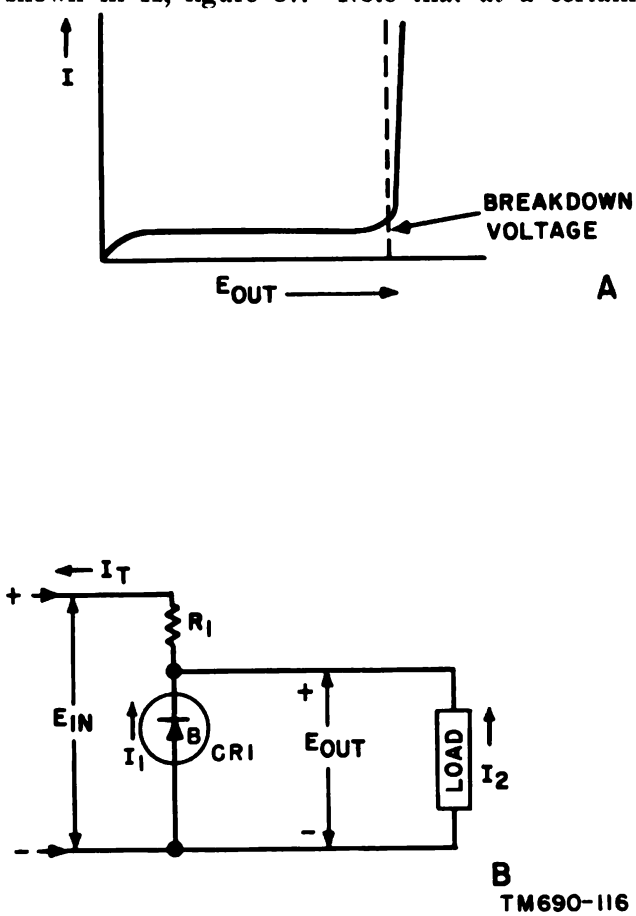


Figure 87. Breakdown diode current-voltage characteristics and voltage regulator.

value of reverse-bias voltage (E_{OUT}), the current increases rapidly while the voltage across the diode remains almost constant. The voltage at which this action occurs is called the breakdown, or Zener, voltage. The name Zener is that of the scientist who investigated this phenomena. When the reverse-biased diode is used to take advantage of this characteristic, it is called a *breakdown* or *Zener* diode. Note that this characteristic is similar to that of some gas-filled tubes, that start conducting at a particular voltage and continue to conduct varying amounts of current while the voltage across the tube remains almost fixed.

b. As in the case of the gas-filled tube, the breakdown diode can be used as a voltage regulator (B, fig. 87).

(1) When the current ($I2$) drawn by the load increases, the total current drawn from the source (E_{IN}) *does not increase.* The increased current for the load is diverted to the load from breakdown diode $CR1$; the voltage across breakdown diode $CR1$ and the load remain constant. A decrease in the current ($I2$) drawn by the load causes a corresponding increase in the current ($I1$) drawn by the breakdown diode. Under these conditions, the total current (I_T) drawn from the source (E_{IN}) remains constant, so that the voltage output (E_{OUT}) again remains constant.

(2) If the source voltage (E_{IN}) increases, total current (I_T) drawn from the source increases. The voltage drop across resistor $R1$ increases by the amount of increase in the source voltage, and current $I1$ increases by the amount of increase in current I_T. The load current ($I2$) and the load voltage (E_{OUT}) remain constant. A decrease in source voltage is compensated in the same manner by a decrease in voltage drop across resistor $R1$, and a decrease in current $I1$.

c. The voltage regulator discussed in *b* above, is capable of maintaining constant load voltage although variations occur in either or both source voltage and load current. Dependent on the characteristics of the particular breakdown diode, the breakdown voltage can be any value from 2 to 60 volts

95. Breakdown Diode Temperature Compensation

a. It was stated (par. 85*b*), that the reverse-biased junction diode has a negative temperature coefficient of resistance. This statement is true if the reverse-bias voltage does not equal or exceed the breakdown voltage (fig. 87). The breakdown diode has a *positive* temperature coefficient of resistance several times larger than the *negative* temperature coefficient of resistance of the forward- or reverse-biased junction diode.

b. The discussion in paragraph 94 concerning the voltage regulator (B, fig. 87) applies only if the temperature of the breakdown diode does not vary under operating conditions. One method of compensating for the increasing or decreasing breakdown diode resistance with increasing or decreasing temperature, respectively, is to place elements of negative temperature-coefficient in series with the breakdown diode. Figure 88 shows a circuit with two forward-biased diodes ($CR1$ and $CR2$) in series with the breakdown diode ($CR3$). The total resistance of the three diodes in series remains constant over a wide range of temperatures. The complete result is a constant voltage output although temperature, input voltage (E_{IN}), and load-current drain may vary.

c. Two diodes are used in this circuit because the temperature-coefficient of resistance of each is half that of the breakdown diode. Forward-biased diodes are used because of the very low voltage drop across them. Thermistors, or other temperature-sensitive elements, can also be used.

96. Voltage Stabilized Transistor Amplifier

a. The circuit shown in figure 89 uses a breakdown diode ($CR1$) to stabilize the collector voltage. Note that the current ($I2$) drawn from the battery divides into breakdown diode current ($I1$) and collector current (I_C). When the dc collector current increases because of the increase in temperature, current $I1$ decreases by the same amount, so that current $I2$ remains constant; the dc voltage drop across resistor $R2$ also remains constant. The voltage supplied to the collector (the battery voltage less the voltage drop across resistor $R2$) also remains constant.

b. The discussion in *a* above, neglects the variation in the resistance with temperature of breakdown diode $CR1$. A method for compensating for this change in resistance is discussed in paragraph 95.

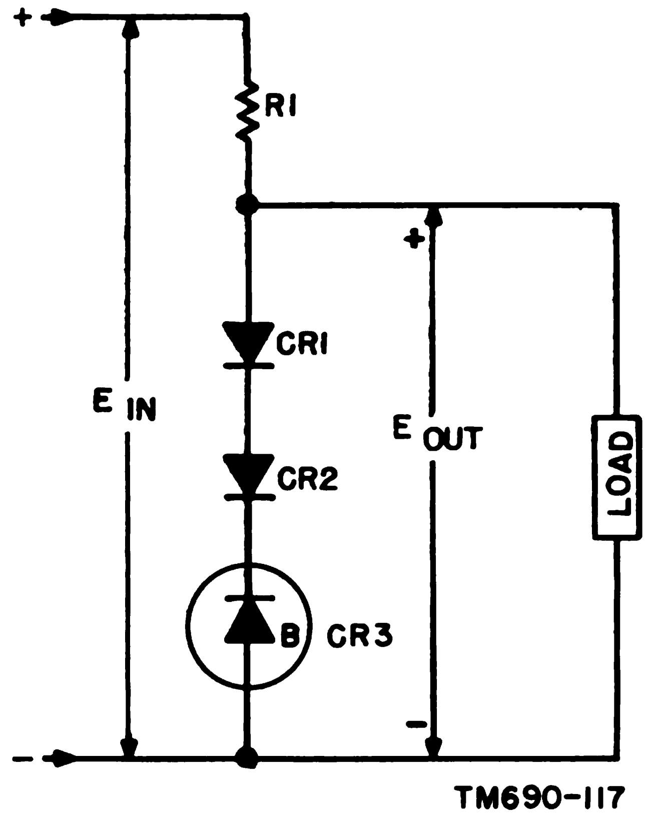


Figure 88. Breakdown diode temperature-compensated voltage regulator.

c. The ac resistance of a breakdown diode may vary from 5 ohms to 1,000 ohms depending on the particular diode. To avoid shunting collector load resistor $R2$ by the low ac resistance of diode $CR1$, a high impedance choke coil ($L1$) is placed in series with the diode.

d. Transformer $T1$ couples the ac signal to the base-emitter circuit. Resistor R1 is the emitter

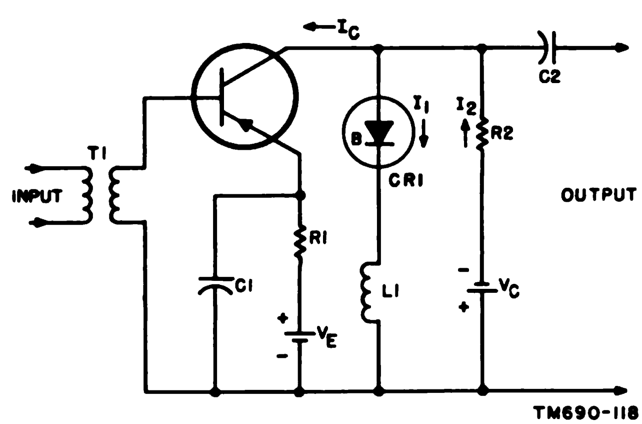


Figure 89. Breakdown diode used in an amplifier to obtain collector voltage stabilization.

swamping resistor. Battery V_E provides emitter-base bias voltage. Capacitor $C1$ bypasses the ac signal around resistor $R1$ and battery V_E. Capacitor $C2$ blocks the collector dc voltage from and couples the ac signal to the following stage.

97. Surge Protection by Junction Diode

a. If an excessive emitter-collector voltage occurs when the normally forward biased base-emitter circuit is *reverse* biased, internal oscillation may occur that can destory the transistor. This condition can occur in transistor amplifiers using transformers for input and output (fig. 90).

b. If the signal from the previous stage is suddenly terminated or if too strong noise signals drive the base-emitter circuit into a reverse bias condition, the collector current is rapidly cut off. The field surrounding transformer T2 collapses rapidly and produces a high emitter-collector voltage while the base-emitter circuit is reverse biased. As stated in *a* above, this condition causes strong oscillations which dissipate power in the transistor and can destroy it. To forestall the possible occurrence of this condition, a junction diode is connected between the base and the emitter to prevent the base-emitter circuit from reverse biasing. This action is referred to as shunt-limiting action.

c. The voltage divider consisting of resistors R1 and R2 introduces a voltage into the base circuit to forward-bias the base-emitter junction, and reverse bias junction diode CR1. Under normal operating conditions, junction diode CR1 can be considered an open circuit. A strong surge voltage of the polarity indicated can occur across transformer T1 secondary. If the surge voltage is greater than the voltage across resistor R1, junction diode CR1 will become forward-biased and conduct. When diode CR1 conducts, only a very small voltage appears across it and the voltage can be considered negligible. This action prevents the base-emitter junction from reverse biasing (*a* above).

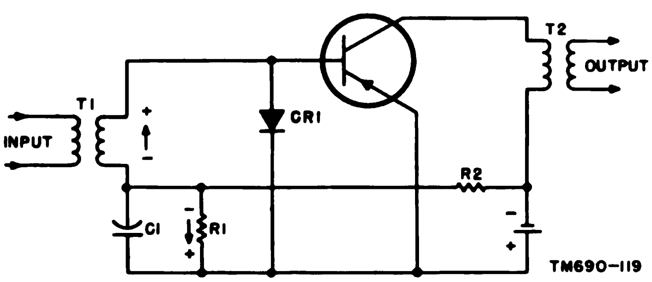


Figure 90. Use of junction diode to prevent base-emitter circuit from reverse-biasing.

d. Transformer T1 couples the ac signal to the amplifier. Capacitor C1 bypasses the ac signal around resistor R1. Transformer T2 couples the ac signal to the following stage.

98. Summary

a. Reverse-bias collector current I_{CBO} (fig. 72), also called saturation current, increases rapidly at high temperatures and causes increased emitter current (I_E).

b. Emitter-base junction resistance decreases with increasing temperature and causes increased emitter current, even when temperature rises from very low values (such as −25° C). See figure 73.

c. The current stability factor (S_I) is defined as the ratio of an increment in emitter current (ΔI_E) to an increment in saturation current (ΔI_{CBO}), and is expressed:

$$S_I = \frac{\Delta I_E}{\Delta I_{CBO}}$$

d. The current stability factor (fig. 74) in terms of external resistance is expressed:

$$S_I = \frac{1}{R_B} \Big/ \left(\frac{1}{R_B} + \frac{1}{R_F} + \frac{1-\alpha_{fb}}{R_E} \right)$$

e. The voltage stability factor (S_V) is defined as the ratio of an increment in collector voltage (ΔV_{CB}) to an increment in saturation current (ΔI_{CBO}) and is expressed:

$$S_V = \frac{\Delta_C V_B}{\Delta I_{CBO}}$$

f. The expression for voltage stability factor (fig. 74) is:

$$S_V = -[S_I R_E + R_C(1 + \alpha_{fb} S_I)]$$

g. An emitter swamping resistor minimizes variations in emitter current caused by variations in the emitter-base junction resistance.

h. Zero base resistance limits the accumulation of saturation current carriers in the base region and therefore limits the increase in emitter current due to this cause.

i. The basic CB amplifier (B, fig. 74) exhibits best temperature stability because it uses an emitter swamping resistor and zero base resistance.

j. The basic CE amplifier (C, fig. 74) exhibits poor temperature stability because it uses a base resistor and zero emitter resistance.

k. The temperature stability of the basic CC amplifier (D, fig. 74) depends upon the ratio of base resistance to emitter resistance.

l. CE amplifiers can be designed to use zero base resistors and an emitter-swamping resistor (A, fig. 76).

m. Dc negative feedback can be employed to minimize variations in emitter current with temperature (C, fig. 76).

n. Emitter current stability can be achieved by reducing emitter-base forward bias by 2.5 millivolts for each degree (centigrade) rise in temperature.

o. A forward-biased junction diode has a negative temperature coefficient of resistance.

p. A reverse-biased junction diode has a negative temperature coefficient of resistance provided the reverse-bias voltage does not equal or exceed the breakdown voltage.

q. Thermistors (figs. 77 and 78), forward-biased diodes (fig. 80), on reverse-biased diodes (fig. 83) can be used to reduce emitter-base forward bias (*n* above), as the temperature increases.

r. The currents and voltages developed in a temperature-stabilized transistor amplifier (figs. 84 and 85) can be used to temperature-stabilize other transistor amplifiers.

s. The breakdown (Zener) diode (figs. 87–90) can be used as a voltage regulator to stabilize collector voltage.

t. The breakdown diode has a positive temperature coefficient of resistance.

u. Temperature-sensitive elements such as forward- or reverse-biased junction diodes having a negative temperature coefficient of resistance can be used to temperature-stabilize breakdown diodes (fig. 88).

CHAPTER 6

TRANSISTOR ANALYSIS AND COMPARISON USING CHARACTERISTIC CURVES AND CHARTS

Section I. TRANSISTOR ANALYSIS

99. General

Paragraphs 100 through 103 cover the graphical analysis of the transistor.

a. The output static characteristic curves are used to calculate voltage, current, and power gain (par. 100).

b. The forward transfer static characteristic curves and the output static characteristic curves with its loadline are used to construct the dynamic transfer characteristic curve (par. 101). Paragraph 101 also discusses the use of the dynamic transfer characteristic curve to determine distortion under overload conditions.

c. Collector voltage and current readings at a number of points are used to construct a constant power dissipation line on the output static characteristic curve (par. 102) to determine operation of the transistor within its power dissipation ratio.

d. The effect of collector current and voltage on the capacitance existing between the collector and the base is covered graphically in paragraph 103.

100. Output Characteristic Curves, Calculation of Gain

a. General. Calculation of the current, voltage, and power gain of a common-emitter transistor amplifier (A, fig. 91) can be accomplished by using the common-emitter output static characteristic curves (B, fig. 91). The output characteristic curves plot the collector current against the collector voltage with the base current as the fixed value. The known information about the amplifier is as follows:

(1) Collector-supply voltage is 10 volts.

(2) Load resistor $R2$ is 1,500 ohms.

(3) The emitter-base input resistance (r_i) is 500 ohms.

(4) The peak-to-peak input current is 20 μa.

(5) The operating point (X) is 25 μa of base current and 4.8 volts on the collector.

b. Loadline. The first step in the procedure is to establish the loadline of load resistor $R2$ on the output characteristic curves. This is done by locating and connecting points Y and Z of the loadline.

(1) When the collector current is zero, the total collector supply voltage (10 volts) equals the collector voltage (V_{CE}). Point Z (one point of loadline) then is at the 10-volt mark on the horizontal axis.

(2) When the voltage on the collector is zero, the total collector-supply voltage (10 volts) is dropped across load resistor $R2$ (1,500 ohms). The total current (I_C) then is:

$$I_C = \frac{10\ v}{1{,}500 \text{ ohms}}$$
$$= 0.0066 \text{ ampere}$$
$$= 6.6 \text{ ma}$$

Point Y (second point of loadline) then is at the 6.6-ma mark on the vertical axis.

(3) Connecting points Y and Z with a straight line establishes the load line.

Note. The relationship of the loadline to the output characteristic curves for the transistor is equivalent to that for the electron tube. For details, refer to TM 11–662.

c. Operating Point and Waveforms. The operating point (a(5) above), is located at point X on the loadline. This point is the intersection of

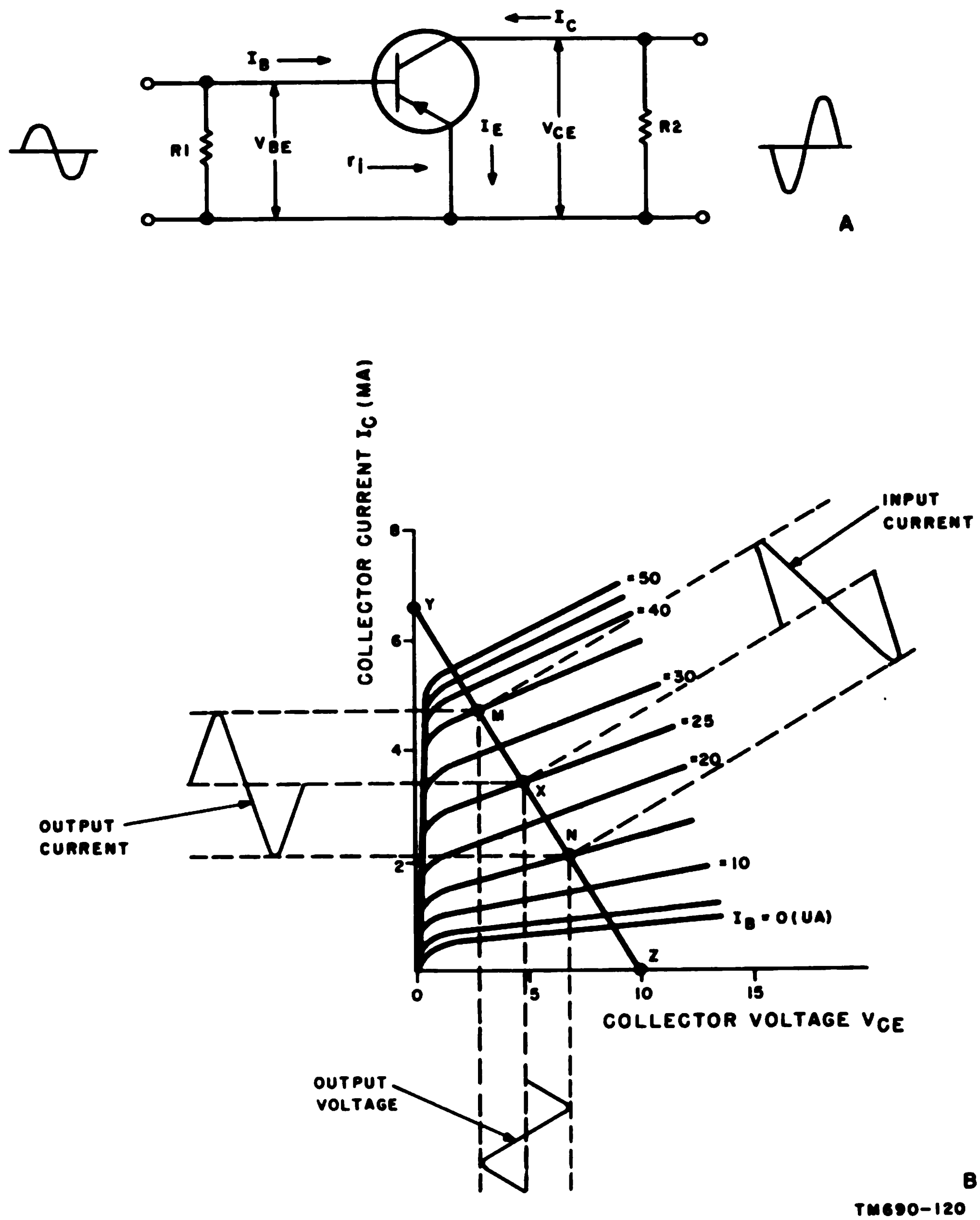


Figure 91. CE amplifier and its output characteristic curves with loadline.

a line drawn vertically from the 4.8-volt mark on the horizontal axis to the 25 μa curve of base current. Since the peak-to-peak input current is 20 μa (a(4) above), the deviation is 10μa above the operating point (point M) and 10 μa below the operating point (point N).

(1) Establish the waveform for the input current by extending a line (perpendicular to the loadline) from operating point X and each deviation point (M and N).

(2) Establish the waveform for the output current by extending a horizontal line through the vertical axis from the operating point and each of the deviation points (M, N, and X).

(3) Establish the waveform for the output voltage by extending a vertical line through the horizontal axis from the operating point and each of the deviation points (M, N, and X).

Note. The horizontal lines extended through the vertical axis provide the collector-current operating point and the deviation from the operating point. The vertical lines extended through the horizontal axis provide the collector voltage operating point and the deviations from the operating point.

d. Current Gain. Current gain in this configuration is the ratio of the change in collector current to the change in base current.

(1) Determine the current gain as follows:

$$A_i=\frac{\Delta I_C}{\Delta I_B}$$

$$=\frac{I_C(\max)-I_C(\min)}{I_B(\max)-I_B(\min)}$$

(2) Substitute known values in the formula:

$$A_i=\frac{4.7 \text{ ma}-2.1 \text{ ma}}{35\ \mu\text{a}-15\ \mu\text{a}}$$

$$=\frac{2.6 \text{ ma}}{20\ \mu\text{a}}$$

$$=\frac{2.6 \text{ ma}}{0.02 \text{ ma}}$$

$$=130$$

The current is amplified 130 times.

e. Voltage Gain. Voltage gain in this configuration is the ratio of the change in collector voltage to the change in base voltage.

(1) Determine the voltage gain as follows:

$$A_v=\frac{\Delta V_{CE}}{\Delta V_{BE}}$$

(2) Solve for ΔV_{BE}. The change in input voltage is the change in input current multiplied by the input impedance (500 ohms).

$$\Delta V_{BE}=\Delta I_B r_i$$

$$=20\ \mu a \times 500 \text{ ohms}$$

$$=0.00002 \text{ ampere} \times 500 \text{ ohms}$$

$$=0.01 \text{ volt}$$

(3) Substitute known values in the formula ((1) above).

$$A_v=\frac{6.7v-2.7v}{0.01}$$

$$=\frac{4v}{0.01v}$$

$$=400$$

The voltage is amplified 400 times.

f. Power Gain. The power gain is the voltage gain times the current gain.

(1) Determine the power gain as follows:

$$G=A_vA_i$$

(2) Substitute known values obtained in *d* and *e* above.

$$\text{G}=130\times 400$$

$$\text{G}=52{,}000$$

The power input is increased 52,000 times in going through the transistor.

(3) The power gain in db is:

$$G=10 \log 52{,}000$$

$$=10\times 4.7$$

$$=47 \text{ db}$$

101. Dynamic Transfer Characteristic Curve

a. Construction of Dynamic Transfer Characteristic Curve. The loadline on the output static characteristic curve tells a great deal (par. 100), but not as conveniently as does another type of characteristic curve. When the effects of the loadline are added to the forward transfer static characteristic curves, a resultant curve known as the *dynamic transfer characteristic curve* (fig. 92) is formed. It has become common practice, when studying the behavior of the collector current under the influence of a signal current applied to the base, to show the dynamic transfer characteristic and to plot the input signal and the resultant collector current along this characteristic (*b* below). In figure 92, the output characteristic with the 1,500-ohm loadline (B, fig. 91) is repeated. In addition, the transfer static characteristic is drawn to its left. To show the collector current-base current curve, which represents the collector current corresponding to certain base currents and the effect of the load in creating the effective collector voltages, the 1,500-ohm loadline will be projected on the transfer static characteristics. The two families of curves have three attributes in common—a common-collector current axis, like values of base current, and like values of collector-supply voltage, although the last two named are illustrated differently. To project the effect of the loadline on the transfer static characteristics, it is necessary to plot the collector current values for each value of base current (*P–Z*) shown on the output characteristic curve. This is done as follows:

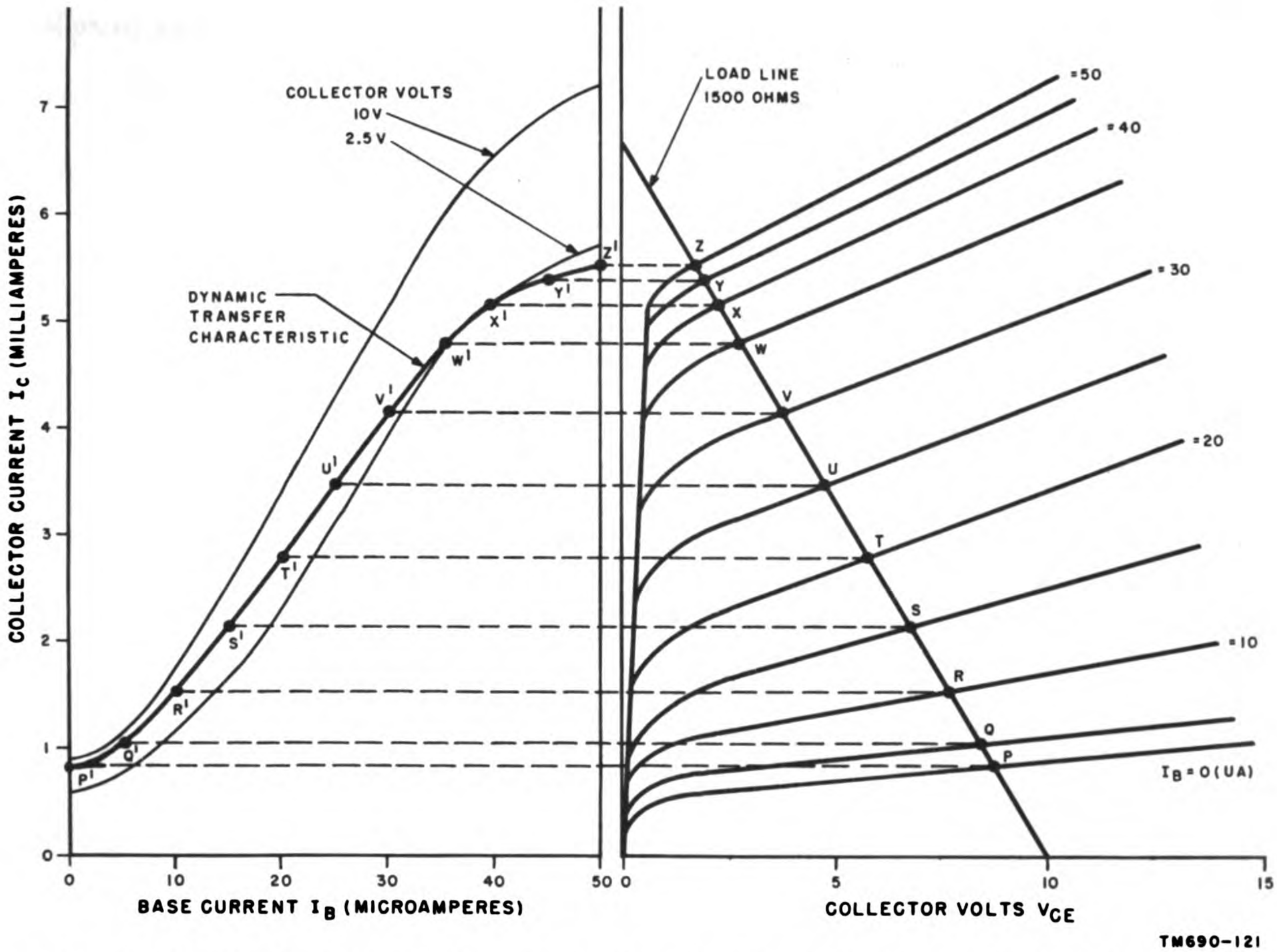


Figure 92. Construction of dynamic transfer characteristic curve from output static characteristic curves with loadline and forward transfer static characteristic curves.

(1) Extend a horizontal line from point P of the output characteristic curve to point P^1 ($I_B=0$ and $I_C=0.85$ ma) of the transfer static characteristic curves.

(2) Extend a horizontal line from each of the remaining points (Q through Z) to locate Q^1 through Z^1 of the transfer static characteristic curves.

(3) Connect points P^1 through Z^1 sequentially to establish the *dynamic* transfer characteristic curve.

(4) The two curves on the transfer characteristics marked 2.5V and 10V are obtained with collector voltage held constant and the output ac short circuited. This is a static condition. The collector voltage of the dynamic transfer characteristic curve is not shown as a constant value since the collector voltage now varies because of the presence of the load resistor.

Note. The dynamic transfer characteristic curve for the transistor is equivalent to that for the electron tube. For details, refer to TM 11–662.

b. Dynamic Transfer Characteristic Curve, Signal Analysis.

(1) *Linear operation* (fig. 93). When the proper operating point is established and if the change of base current is within the linear portion of the dynamic transfer characteristic curve, the transistor will operate linearly. When the transistor is operating linearly, the amplified output signal will be an exact reproduction of the input signal. Class A amplifiers are operated in this manner.

(2) *Overdriving* (fig. 94). When the proper operating voltage is established, and if the change of base current exceeds the linear portion of the dynamic transfer characteristic curve, the input signal is

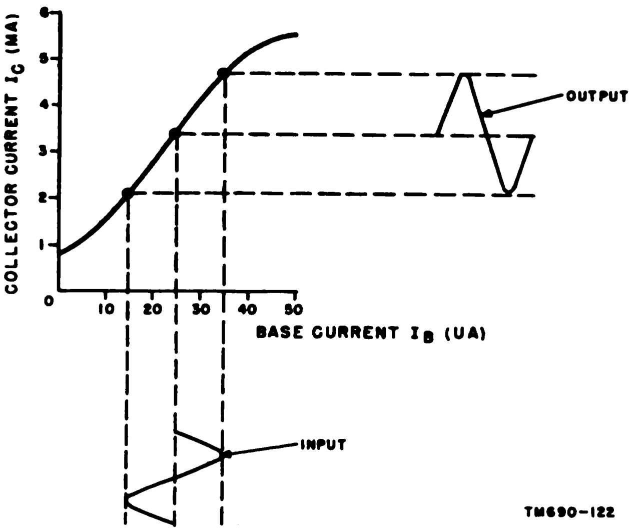


Figure 93. Dynamic transfer characteristic curve showing linear (class A) operation.

overdriving the transistor and the amplified output signal will be distorted.

(3) *Improper operating point* (fig. 95). When the improper operating point is established, the change of base current will automatically cause the linear portion of the dynamic transfer characteristic curve to be exceeded. The location of the operating point will determine whether the negative or positive change of base current will exceed the linear portion. When the transistor is functioning in this manner, the output signal will be distorted during either the negative or the positive change in collector current.

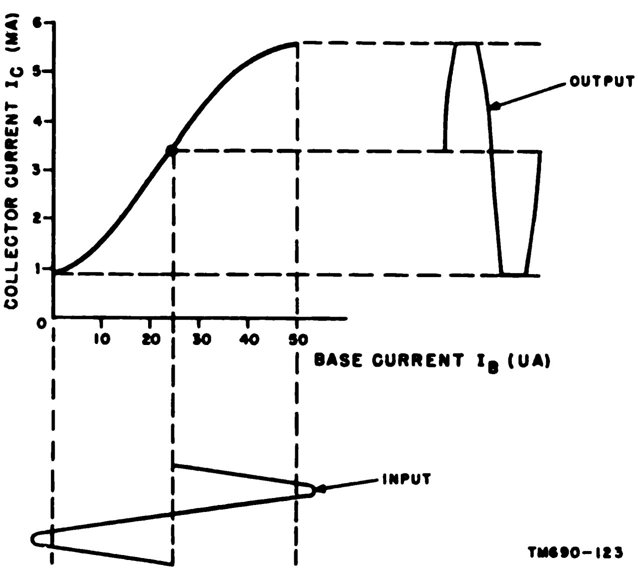


Figure 94. Dynamic transfer characteristic curve showing overdriving.

(4) *Class B operation.* Use of the dynamic transfer characteristic curve in analyzing distortion in class B operation of amplifiers is covered in chapter 7.

102. Constant Power Dissipation Line

a. Each transistor has a maximum collector power that it can safely dissipate without damaging the transistor. To insure that the maximum collector power dissipation rating is not exceeded, a *constant power dissipation line* (fig. 96) is drawn on the output static characteristic curves and the collector load resistor is selected so that its loadline falls in the area bounded by the vertical and horizontal axis and the constant power dissipation line.

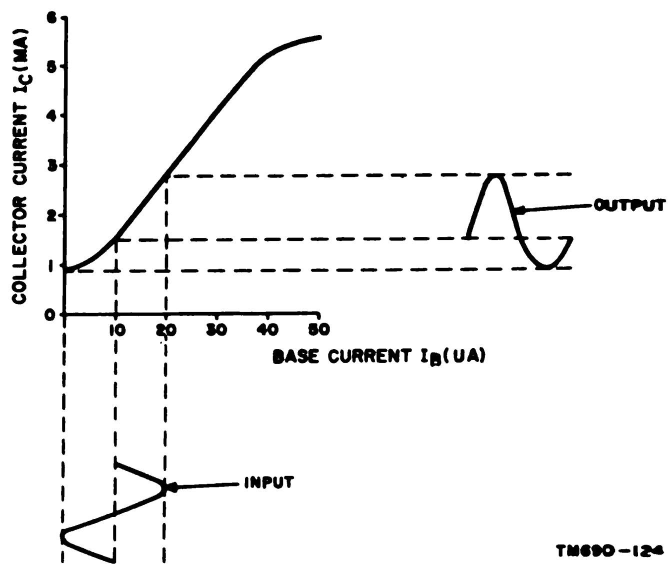


Figure 95. Dynamic transfer characteristic curve showing nonlinear operation.

b. The constant power dissipation line represents a number of points of collector voltage and collector current, the products of which (in power) are equal to the maximum collector power rating of the particular transistor. To draw the constant power dissipation line, the maximum collector power rating and a set of output static characteristic curves are required. Assume that, for a given transistor, the rating is 18 mw and the required curves are as shown in figure 96. To construct the constant power dissipation line, proceed as follows:

(1) Use the formula: *power equals voltage times current.*

$$P=EI$$

(2) Rearrange the formula:

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

(3) Substitute 18 mw for power in the formula, and select arbitrary values of voltage to find corresponding values of current. For instance, assume 15 volts is selected.

$$I=\frac{18\ mw}{15\ v}$$

$$=1.2 \text{ ma}$$

Continue this procedure until a sufficient number of points is obtained. Typical points are indicated in the chart below.

E (volts)	I (ma)	E (volts)	I (ma)
15	1. 2	4	4. 5
12	1. 5	3	6. 0
9	2. 0	2	9. 0
6	3. 0		

These voltage and current points are located and marked on the output static characteristic curves. When connected, the resultant curve represents the constant power dissipation line.

c. Any loadline selected so that it is tangent to the constant power dissipation line will insure maximum permissible power gain of the transistor while operating within the maximum collector power dissipation rating. This is important in the use of power amplifiers.

103. Interelement Capacitances

a. Figure 97 shows the interelement capacitances associated with the transistor. The capacitances are shown externally; however, the actual capacitance effects are produced by the PN junctions within the transistor. Because the width of the PN junctions vary in accordance with the voltages across them and the current flow through them, the capacitance values also vary. For example, the variation of collector-base capacitance (C_{cb}) with collector voltage and emitter current is shown in A and B, figure 98.

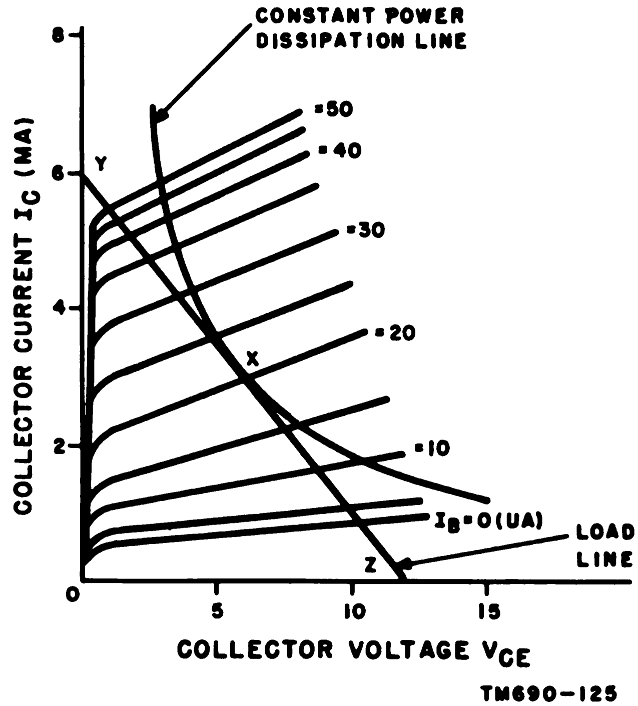


Figure 96. Output static characteristic curves with constant power dissipation line.

(1) The increase in the width of the PN junction between base and collector, as the reverse bias voltage (V_{CB}) is increased, is reflected in lower capacitance values. This phenomenon is equivalent to spreading apart the plates of a capacitor so that lower capacitance results.

(2) An increase in emitter current, most of which flows to the collector through the base-collector junction, increases the collector-base capacitance (C_{cb}). The increased current through the PN junction may be considered as effectively reducing the width of the PN junction. This phenomenon is equivalent to reducing the distance between the plates of a capacitor so that increased capacitance results.

(3) In high-frequency amplifier applications, the collector-base capacitance causes positive feedback that may result in oscillation. External circuits must be used to prevent oscillation (ch. 8).

(4) The average value of collector-base capacitance (C_{cb}) may vary from 2$\mu\mu$f for high-frequency transistors, to 50$\mu\mu$f for low-frequency (audio) transistors.

b. The collector-emitter capacitance (C_{ce}) of a transistor, also caused by the PN junction, normally is 5 to 10 times greater than that of the collector-base capacitance and also varies with emitter current and collector voltage (*a* above).

(1) The effect of the variation of interelement capacitance with current and voltage is utilized to frequency modulate oscillators (ch. 12).

(2) Because the collector-emitter capacitance is 5 to 10 times greater than the collector-base capacitance, common-base amplifiers have a better high-frequency (audio) responce than common-emitter amplifiers.

c. The emitter-base capacitance (C_{eb}), although very high because of normal forward bias and resultant small width of the PN junction, does not offer many problems in amplifier design because it is normally shunted by a low input resistance.

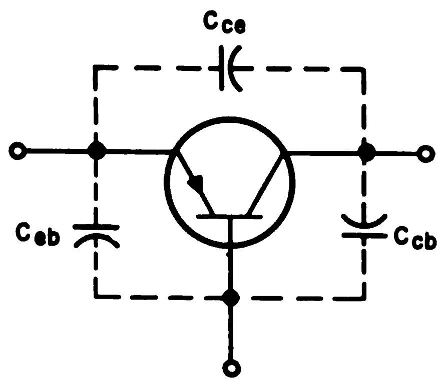


Figure 97. Effective interelement capacitances of the transistor.

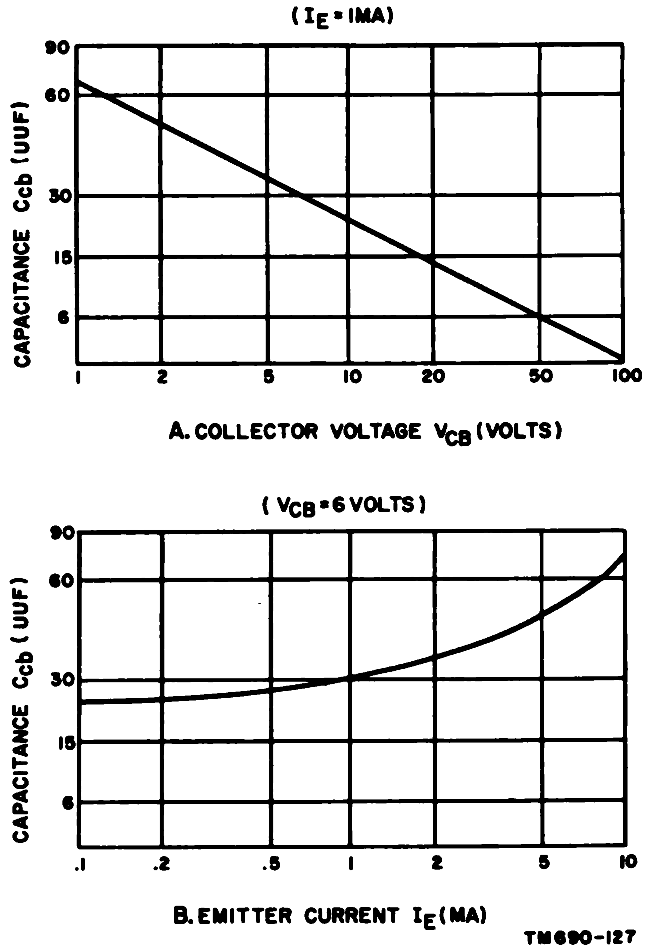


Figure 98. Variation of collector-base capacitance (C_{cb}) *with collector voltage and emitter current.*

Section II. COMPARISON OF TRANSISTOR CONFIGURATIONS

104. Comparison of Configurations for Resistances

a. Medium-Power Amplifier Stage. Figure 99 represents a typical medium-power amplifier stage. The block labeled *transistor* represents the common-collector, the common-emitter, or the common-base configuration. The input signal is represented by voltage source v_g and input resistance R_g of the voltage source. The transistor input resistance is represented by r_i and the output resistance is represented by r_o. The load resistance is represented by R_L.

b. Load Resistance and Input Resistance. Figure 100 shows the comparison of the load resistance to the input resistance for the common-collector, common-emitter, and common-base configurations. Input resistance r_i for the common-collector and the common-base configurations increases as load resistance R_L is increased. Input resistance r_i for the common-emitter configuration decreases as load resistance R_L is increased. Unlike the electron tube amplifier, the transistor amplifier input resistance depends upon the value of the load resistor.

c. Generator Resistance and Output Resistance. Figure 101 shows a comparison of generator resistance R_g to the output resistance r_o for the common-base, common-emitter, and common-collector configurations. Output resistance r_o increases as generator resistance R_g increases for the common-base and common-collector configurations. Output resistance r_o decreases for the common-emitter configuration as generator resistance R_g increases. Unlike the electron tube amplifier, the transistor

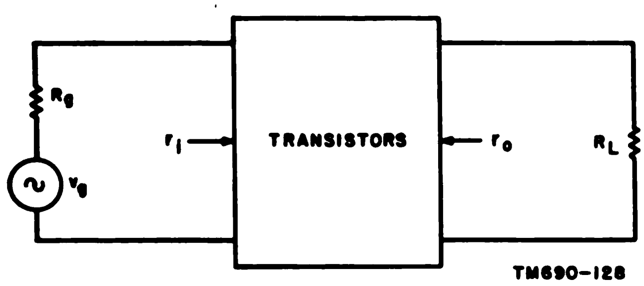


Figure 99. Medium power amplifier stage showing transistor as a block with a voltage source and a load resistance.

output resistance depends upon the generator resistance.

105. Comparison of Configurations for Gain

a. General. The three possible circuit configurations for the transistor were covered in chapter 3. Formulas were developed in chapter 4 calculating input resistance, output resistance, current gain, voltage gain, and power gain. Typical values of these quantities for medium-power transistors are given in *b* below. Charts that plot voltage, current, and power gain for each configuration of a particular transistor are discussed in *c*, *d*, and *e* below. This information will help the reader to

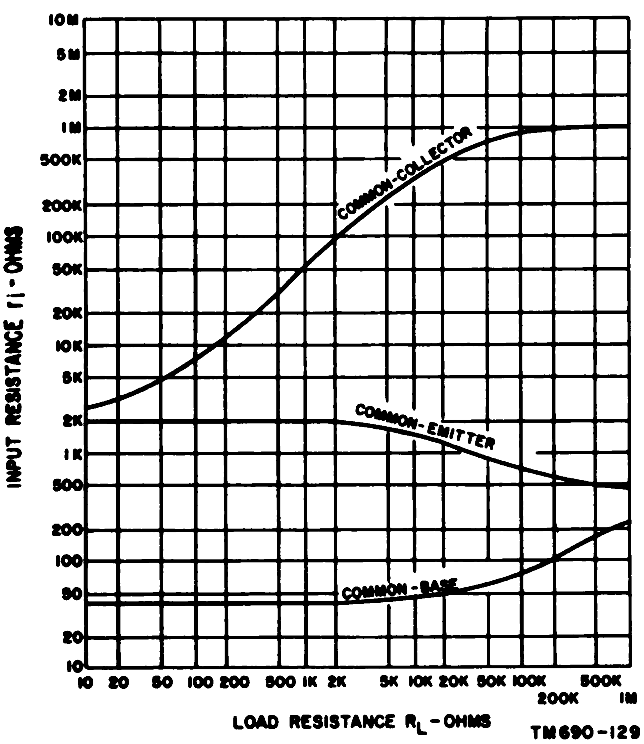


Figure 100. Variation of input resistance with load resistance for each configuration.

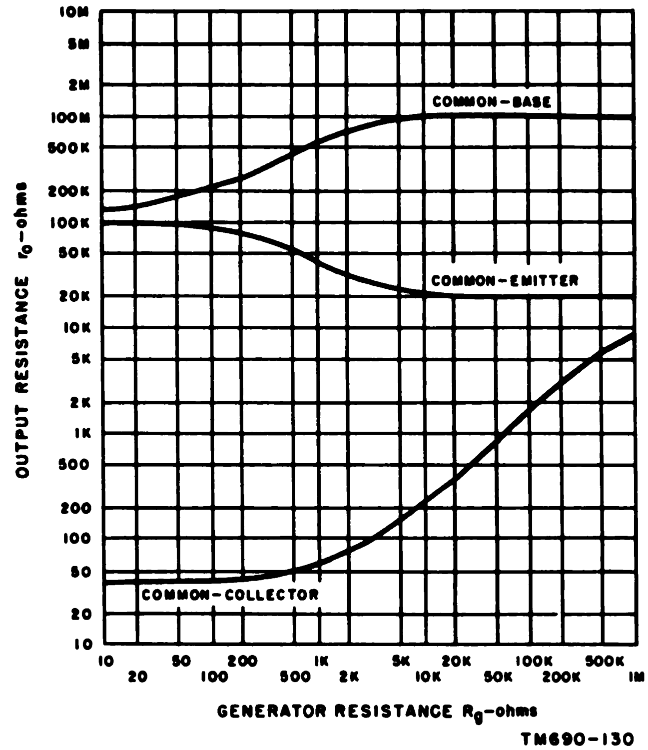


Figure 101. Variation of output resistance with generator resistance for each configuration.

appreciate the combinations of configurations used in audio amplifiers, and the combinations of configurations used to achieve a given result.

b. Typical Values. The table on the following page lists typical values of input resistance, output resistance, current gain, voltage gain, and power gain for the three configurations.

c. Common-Base Amplifier. Figure 102 shows the comparison of the various gains of a common-

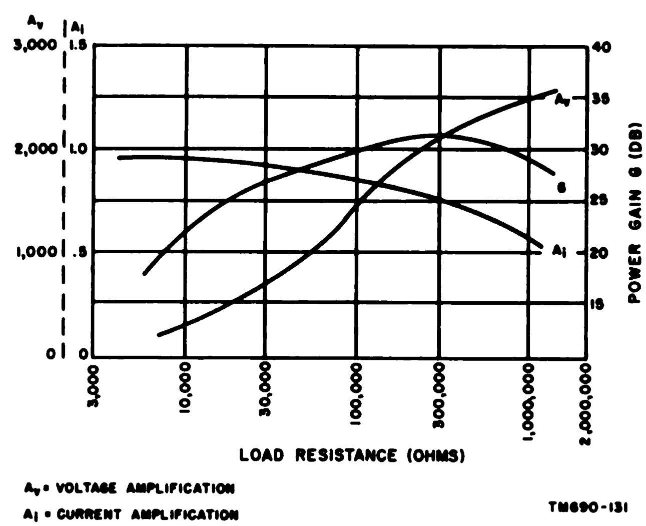


Figure 102. Common-base, voltage, current, and power gain characteristic curves.

Item	*CB* amplifier	*CE* amplifier	*CC* amplifier
Input resistance	30–150 ohms	500–1,500 ohms	20K–500K ohms.
Output resistance	300K–500K	30K–50K	50–1,000 ohms.
Voltage gain	500–1,500	300–1,000	Less than 1.
Current gain	Less than 1	25–50	25–50.
Power gain	20–30 db	25–40 db	10–20 db.

base amplifier. The voltage gain (A_v) is the ratio of the signal output voltage to the signal input voltage. The current gain (A_i) is the ratio of the signal output current to the signal input current. The power gain (G) is the ratio of the output signal power to the input signal power expressed in db. The current gain, the voltage gain, and the power gain are plotted against the load resistance. The current gain (A_i) is always less than one and decreases as the load resistance is increased. The voltage gain (A_v) increases as the load resistance is increased. The maximum power gain (G) for this particular transistor occurs when the load resistance is approximately 300,000 ohms.

d. Common-Emitter Amplifier. Figure 103 shows the comparison of the various gains of a common-emitter amplifier. The current gain (A_i) decreases as the load resistance is increased. The voltage gain (A_v) increases as the load re-

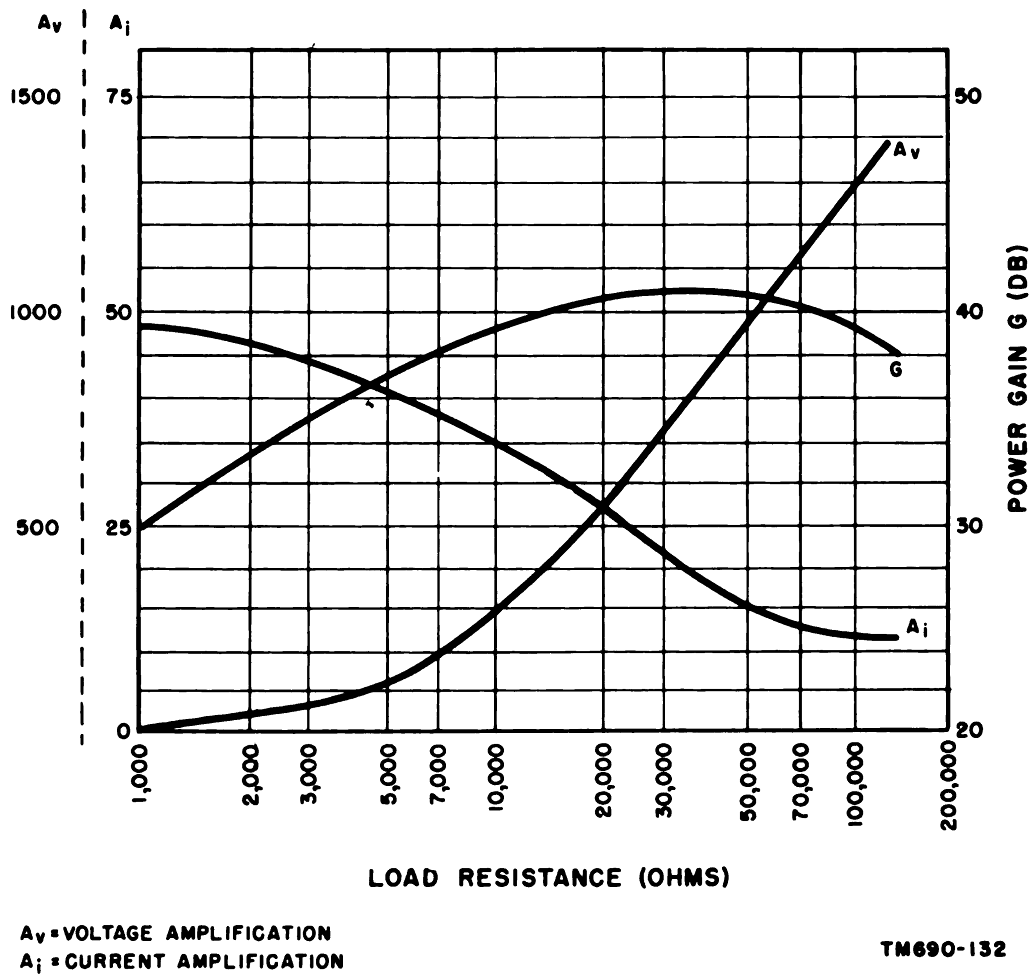


Figure 103. Common-emitter, voltage, current, and power gain characteristic curves.

sistance is increased. The maximum power gain (G) occurs when the load resistance is approximately 40,000 ohms. Note that current, voltage, and power gain for this configuration may exceed unity.

e. Common-Collector Amplifier. Figure 104 shows the comparison of the various gains of a common-collector amplifier. The current gain (A_i) decreases as the load resistance is increased. The voltage gain (A_v) increases as the load resistance is increased but it never exceeds unity. The maximum power gain (G) occurs when the load resistance is approximately 1,000 ohms.

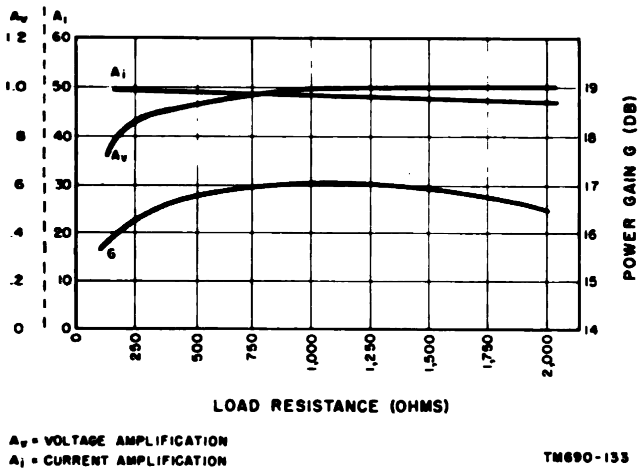


Figure 104. Common-collector, voltage, current, and power gain characteristic curves.

106. Summary

a. The current, voltage, and power gain of a transistor amplifier can be calculated from the output static characteristic curves upon which has been drawn the loadline.

b. The loadline displays the way in which the collector-supply voltage is divided between the load and the collector under various conditions of collector current

c. The dynamic transfer characteristic curve may be used to determine the linearity and nonlinearity of the output signal to the input signal for a specific operating point and a specific load resistance.

d. Nonlinearity in a transistor's behavior is a source of distortion and is undesirable except for specific types of operation.

e. The constant power dissipation line may be used to establish a loadline that will provide maximum gain without exceeding the maximum collector power dissipation rating of the transistor.

f. The interelement capacitances of the transistor are caused by the PN junctions in the transistor.

g. The curves for voltage, current, and power gain help in selecting a specific configuration to obtain a specific result or a combination of configurations to obtain a specific result.

CHAPTER 7

AUDIO AMPLIFIERS

Section I. INTRODUCTION

107. General

a. Audio amplifiers are used in equipment such as public address systems, sound recorders, sound reproducers and radio and television sets. The frequencies of the signals amplified are in the range of 10 to 20,000 cps.

b. The input circuit of a transistor amplifier may draw current from either the input device or the previous stage. In this respect, each transistor amplifier is considered as either a current or a power amplifier operating at a current or a power level higher than the previous stage and lower than the following stage. Preamplifiers (pars. 109–114) usually operate at power levels measured in micromicrowatts, or microwatts. Driver stages (pars. 126–129) usually operate at power levels measured in milliwatts. Power stages (pars. 130–133) usually operate at power levels measured in hundreds of milliwatts or in watts. These power levels are only approximate; the equipment in which these stages are used determine the power levels of the preamplifier, the driver, and the power stage.

108. Classification of Amplifiers

As in the case of electron tube amplifiers, transistor amplifiers can be operated class A, class B, class AB, or class C.

a. Class A. Class A amplifiers are operated on the linear portion of the collector characteristics (par. 99). The transistor amplifier is so biased that collector current flows continuously during the complete electrical cycle of the signal and even when no signal is present. Audio, amplifiers, operated class A, may be used in single-ended or in push-pull applications.

b. Class B. Class B amplifiers can be biased either for collector current cutoff or for zero collector voltage.

(1) When biased for collector current cutoff, collector current will flow only during that half cycle of the input signal voltage that *aids* the forward bias. This method of bias is most often used because it results in the best power efficiency.

(2) When biased for zero collector voltage, heavy collector current will flow when no signal is present. Almost all of the collector supply voltage will be applied across the collector load resistor. Power dissipation in the transistor will be zero since power is the *product of current and voltage.* The collector voltage will vary only during that half cycle of the input signal voltage that opposes the forward bias. This method of bias is seldom used because high power dissipation in the load resistor when no input signal is present results in low power efficiency. Also, when the collector current is heavy, the forward current transfer ratio (α_{fb}) is appreciably reduced.

(3) Class B audio amplifiers must be operated in push-pull to avoid severe signal distortion.

c. Class AB. Class AB amplifiers can be biased so that either the collector current or the collector voltage is zero for *less* than a half cycle of the input signal. The statements for the class B amplifier (*b* above) apply also to the class AB amplifier.

d. Class C. Class C amplifiers can be biased so that either the collector current or the collector voltage is zero for *more* than a half cycle of the input signal. The class C amplifier, either in single-ended or in push-pull application, should never be used as an audio amplifier because it will result in severe signal distortion.

109. General

a. Preamplifiers are low-level stages (par. 107*b*) that usually follow low-level output transducers such as microphones, hearing-aid pickup devices, and recorder-reproducer heads.

b. The most important characteristics of preamplifiers are the signal-to-noise level (noise factor) (par. 110), the input and output impedances (par. 111), and the required frequency response (pars. 112, 113, and 114).

110. Noise Factor

a. General. The quality of an amplifier with respect to noise is indicated by its noise factor (or fig.). The noise factor (F_o) is determined by measuring the ratio of signal-to-noise power (S_i/N_i) at the input of the amplifier and the ratio of signal-to-noise power (S_o/N_o) at the output of the amplifier. The noise factor is equal to the input signal-to-noise ratio divided by the output signal-to-noise ratio:

$$F_o = \frac{S_i/N_i}{S_o/N_o}$$

The smaller the value of this ratio (F_o) the better the noise quality of the amplifier. The noise factor of an amplifier with a given transistor is affected mainly by the operating point (*b* below), the resistance of the signal source (*c* below), and the frequency of the signal being amplified (*d* below).

Note. The logarithm of the ratio for noise factor gives the noise factor in db.

b. Operating Point. The operating point is established by the zero-signal base or emitter current and the collector voltage. The effect of several values of emitter current and collector voltage on the noise factor of a typical transistor amplifier is shown in figure 105. The curves indicate that for a given transistor, a low-noise factor can be achieved by operating the transistor at an emitter current of less than 1 ma, and a collector voltage of less than 2 volts. Note that increasing the collector voltage (at the higher values) increases the noise factor more rapidly than increasing the emitter current.

c. Signal Source Resistance. The resistance of the source feeding the transistor also affects the noise factor of an amplifier. The curve in the illustration (fig. 106) indicates that the noise factor for a typical transistor can be kept low by using a signal source resistance in the range of 100 to 3,000 ohms.

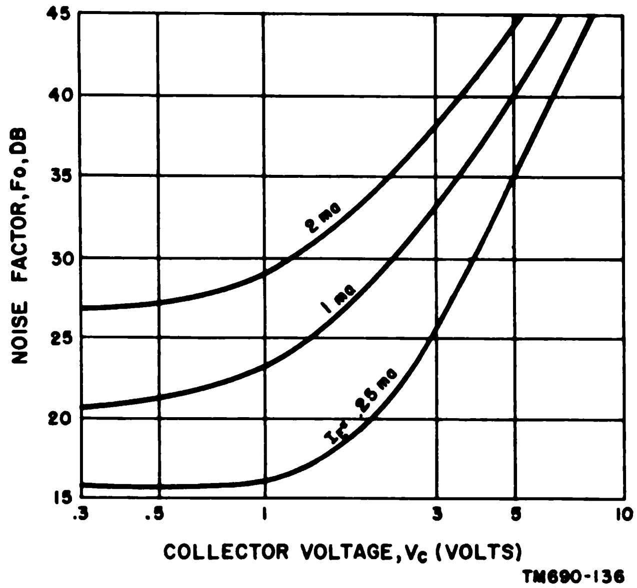


Figure 105. Chart showing variation of noise factor with collector voltage and emitter current.

d. Signal Frequency. Figure 107 shows the variation of the noise factor as the frequency is increased. At very low frequencies, the noise factor is high. As the frequency increases, the noise factor improves to approximately 50 kilocycles and then starts to increase slowly again. This curve indicates that low-noise dc amplifiers are difficult to design.

111. Input Resistance

a. Low Input Resistance. If a preamplifier is to be fed from a low-resistance signal source, either the CB configuration or the CE configuration may be used. The CB configuration has an input impedance which is normally between 30 and 150 ohms; the CE configuration has an input impedance which is normally between 500 and 1,500 ohms (par. 103).

b. High Input Resistance. Although it is undesirable to use a high-resistance signal source because of the large noise factor (par. 110*c*), it becomes necessary if a device such as a crystal pickup head is used. A high-input resistance preamplifier is also required. Three circuits may be used to obtain a high input resistance (r_{in})

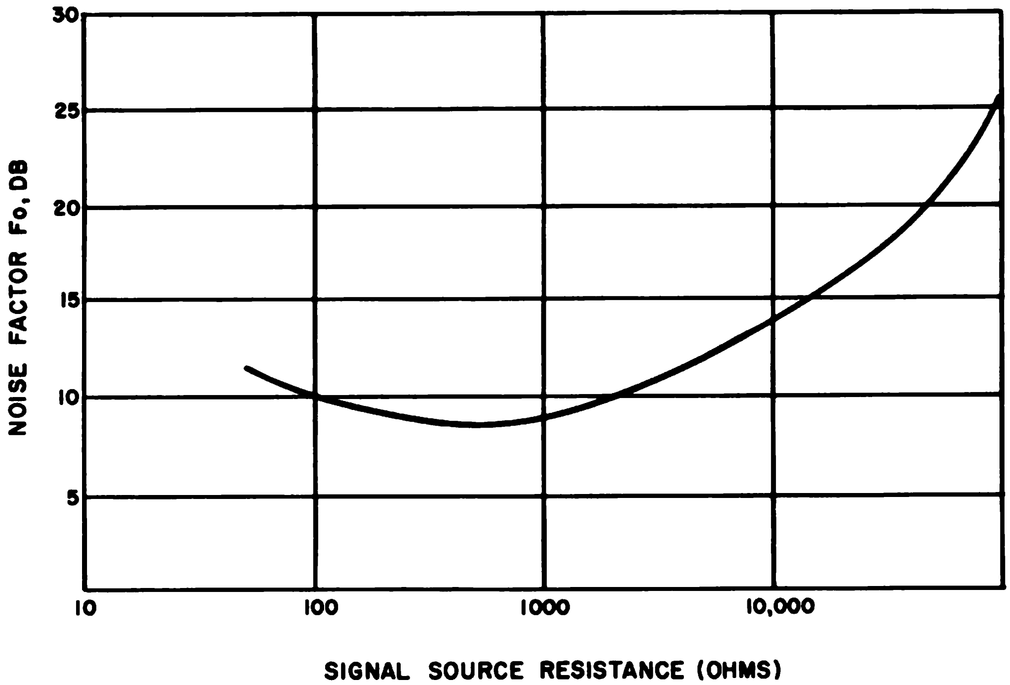


Figure 106. Chart showing variation of noise factor with signal source resistance.

without using a transformer. They are the *CC* configuration (A, fig. 108), the *CE* configuration with series input resistance (R_s) (B, fig. 108) and the degenerated (negative feedback) *CE* configuration (C, fig. 108).

(1) *CC configuration.* The input resistance to the *CC* configuration is high (fig. 100) because of the large negative voltage feedback in the base-emitter circuit (A, fig. 108). As the input voltage rises, the opposing voltage developed across resistor R_L substantially reduces the net voltage across the base-emitter junction. By this action, the current drawn from the signal source remains low. By Ohm's law it is known that a low current drawn by a relatively high voltage represents a high resistance. If a load resistance (R_L) of 500 ohms is used, the input resistance of a typical transistor will be over 20,000 ohms. The disadvantage of the *CC* configuration, however, is that small variations in the current drawn by the following stage cause large changes in the input resistance value.

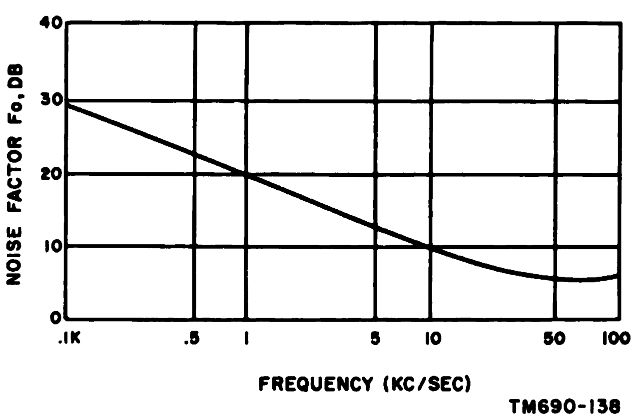


Figure 107. Chart showing variation of noise factor with frequency.

(2) *CE configuration with series resistors.* The base-emitter resistance (represented by r_i in B, fig. 108) for a typical transistor is about 2,000 ohms if a load resistance (R_L) of 500 ohms is used (fig. 100). To achieve a total input resistance (r_{in}) of 20,000 ohms, a resistor (R_s) of 18,000 ohms can be placed in series with the base-emitter resistance (r_i). This circuit has an advantage over the *CC* configuration ((1) above), in that the total input resistance, determined mainly by resistor R_s, remains relatively constant even with large variations in transistor parameters or current drain by the following stage. The disadvantage of this circuit, in addition to a small loss in current gain, is the large resistance in the base lead. This

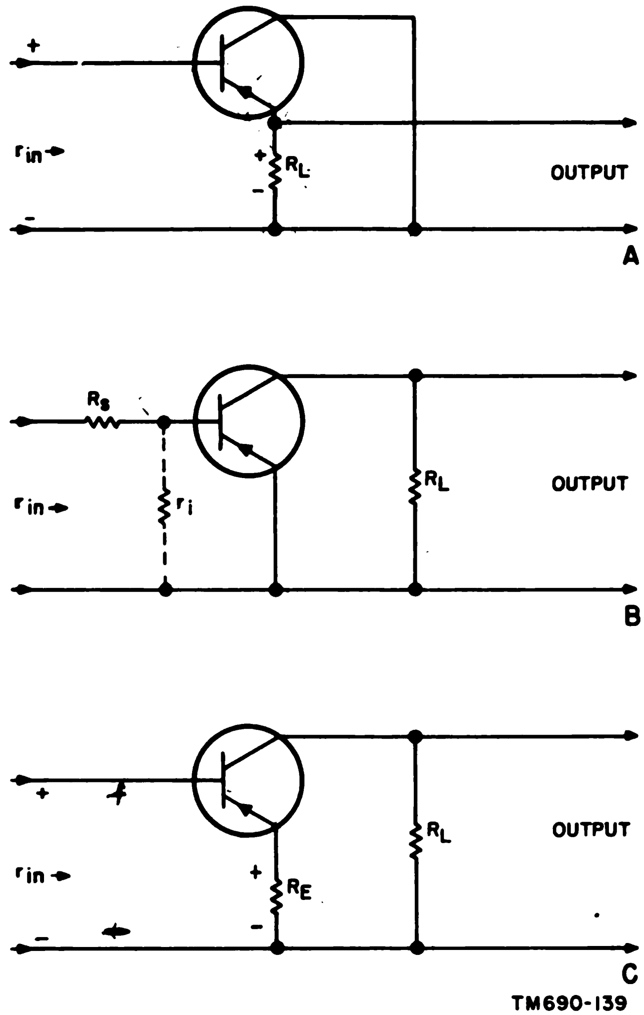


Figure 108. Three amplifier circuit arrangements for producing a high input resistance.

resistance leads to poor bias stability if the bias voltage is fed to the transistor through this resistor.

(3) *Degenerated CE configuration.* If an unbypassed resistor (R_E) is inserted in the emitter lead of a *CE* configuration (C, fig. 108), the signal voltage developed across this resistor opposes the input signal voltage. As in the case of the *CC* configuration ((1) above), this negative feedback voltage or degenerative voltage, causes an increase in the input resistance. Without the unbypassed resistor in the emitter lead the input resistance of the *CE* configuration would be 2,000 ohms ((2) above), if a load resistor (R_L) of 500 ohms were used. With an unbypassed resistor (R_E) of 500 ohms and a load resistor (R_L) of 500 ohms, the input resistance (r_{in}) of the degenerated *CE* configuration with a typical transistor is 20,000 ohms. The degenerated *CE* configuration is equal to the *CE* configuration with the series resistor ((2) above). However, the advantage of the degenerated *CE* configuration is that the unbypassed resistor (R_E) also acts as an emitter swamping resistor and helps to bias stabilize the transistor.

112. Two-Stage Direct-Coupled Preamplifier, Low-Frequency Compensation

a. Figure 109 shows a typical two-stage, direct-coupled preamplifier. The circuit is designed to compensate for the poor low-freqnency output of the transducer, microphone M1.

(1) By directly coupling the transistor Q1 output to the transistor Q2 input, use of a coupling capacitor is avoided. A coupling capacitor would attenuate the low frequencies.

(2) Capacitor C2 and resistor R3 form a low-pass filter. The high frequencies are partially shunted to ground (attenuated). Note that transistor Q2 is a degenerated CE stage (par. 111). The degenerated stage (using resistor R4 for negative feedback) has a very high input impedance (approximately 30,000 ohms). By using a high input impedance, a relatively low valued capacitor (C2, 0.2 μf) is sufficient to shunt the high frequencies to ground.

b. The collector voltage and the emitter current of transistor Q1 are kept very low to reduce the noise factor (par. 110*b*). With the signal amplified by the first stage, the emitter current and the collector voltage of the second stage can be increased as shown without increasing the noise factor.

c. Transformer T1 couples the output of microphone M1 to the base of transistor Q1. Transistor Q1 amplifies the signal. Resistor R1 is the emitter swamping resistor. Capacitor C1 bypasses the ac signal around resistor R1. Collector load resistor R2 develops the output signal. Capacitor C2 and resistor R3 constitute a low-pass filter. The signal is amplified by transistor Q2. Resistor R4 produces negative feedback. Resistor R5 is the emitter swamping resistor. Capacitor

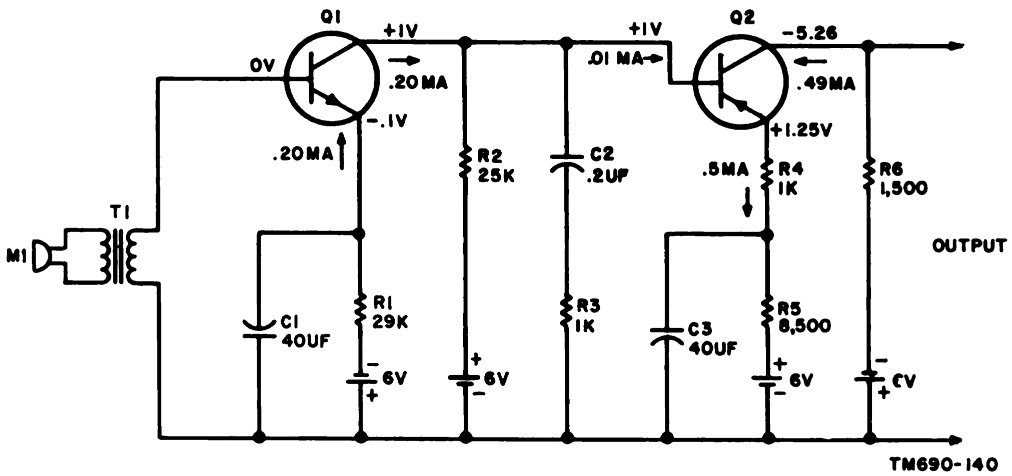


Figure 109. Typical two-stage direct-coupled preamplifier, providing low-frequency compensation for transducer.

C3 bypasses the ac signal around resistor R5. Collector load resistor R6 develops the output signal.

113. Three-Stage, RC-Coupled Preamplifier, Low-Frequency Compensation

a. General. The preamplifier shown in figure 110 is designed to compensate for the poor low-frequency output of a transducer. This is done by using two low-pass (high-frequency attenuating) networks. The first network (capacitor C3 and resistor R4) shunts the input of transistor Q2; the second network (capacitor C6 and resistor R8) shunts the input of transistor Q3. Transistors Q2 and Q3 are used in CE configurations; the input of each is approximately 1,000 ohms. Because of the low input impedance, each low-pass network must use a large capacitance (1–2 μf) and a low resistance (500–300 ohms) to partially shunt the high frequencies around the low input impedance of each stage. Compare the values used in these networks with those used in the comparable network (capacitor C2 and resistor R3, fig. 109). In the latter case, the high impedance of the CE degenerated stage permits the use of a smaller capacitance (C2, 0.2 μf) and a higher resistance (R3, 1,000 ohms).

b. Circuit Description.

(1) The first stage (transistor Q1) is a degenerated CE configuration. Negative feedback is provided by resistor R1. The

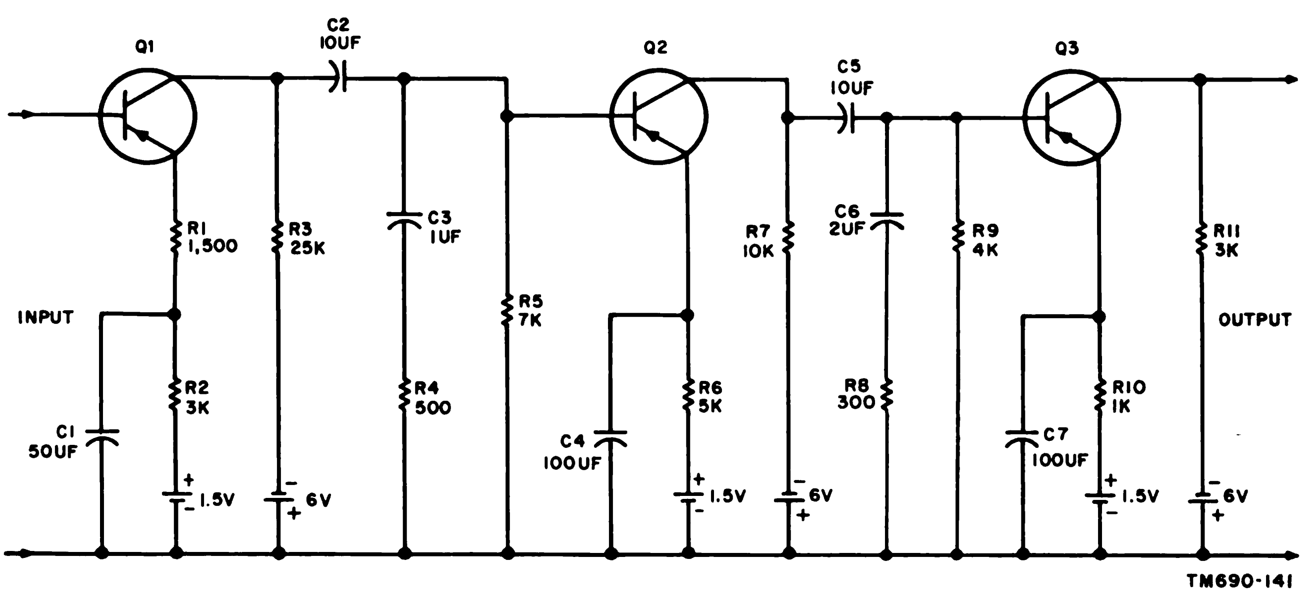


Figure 110. Typical three-stage, RC-coupled preamplifier with low-frequency compensation.

input impedance is approximately 55,000 ohms to accommodate a high-impedance transducer. Resistor R2 with resistor R1 swamps the emitter-base junction resistance. Collector load resistor R3 develops the output signal which is coupled to the following stage by dc blocking capacitor C2. Note the relatively large value of capacitor C2 (10 μf); this large value is required because of the low input impedance of the following stage and also to prevent the attenuation of the low frequencies.

(2) Capacitor C3 and resistor R4 form a high-frequency attenuator. Resistor R5 provides a base dc return path; its value is high to prevent shunting the signal current around the base-emitter circuit. Resistor R6 is the emitter swamping resistor. Capacitor C4 bypasses the ac signal around resistor R6. Collector load resistor R7 develops the output signal. Dc blocking capacitor C5 permits the ac signal to be coupled to the following stage.

(3) Capacitor C6 and resistor R8 form a high-frequency attenuator. Resistor R9 provides a base dc return path. Resistor R10 is the emitter swamping resistor. Capacitor C7 bypasses the ac signal around resistor R10. Collector load resistor R11 develops the output signal.

114. Two-Stage, RC-Coupled Preamplifier, High-Frequency Compensation

a. The typical two-stage preamplifier (fig. 111) is designed to compensate for poor high-frequency output of a transducer connected to its input. The equalizer network (in dashed lines) attenuates the low frequencies more than it does the high frequencies. Capacitor C1 is mainly a dc blocking capacitor, but it will attenuate the lower frequencies more readily than the higher frequencies. Because of its higher reactance at the lower frequencies, capacitor C2 bypasses the high frequencies around resistor R3, thus permitting maximum amplification of the higher frequencies. Capacitor C2 can be considered an open circuit at low frequencies; the low frequencies must be passed through resistor R3 and are therefore attenuated. The amount of attenuation experienced by the low frequencies, depends on the relative values of capacitor C2 and resistor R3.

b. Resistor R1 determines the base-emitter bias for transistor Q1. Collector load resistor R2 develops the output signal of transistor Q1. Capacitors C1 and C2 and resistor R3 form an equalizing network (*a* above). Resistor R4 determines the base-emitter bias for transistor Q2. Collector load resistor R5 develops the output signal.

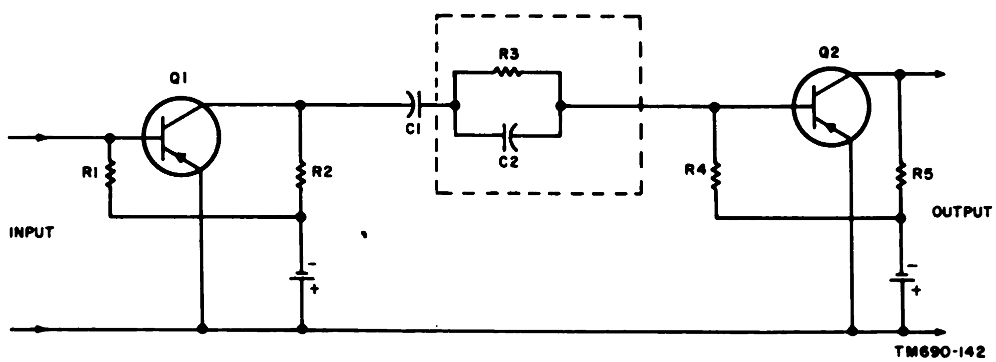


Figure 111. Typical two-stage preamplifier, providing high-frequency compensation for transducer.

Section III. COUPLING NETWORKS USING VARIABLE RESISTORS AS VOLUME AND TONE CONTROLS

115. General

a. Signal coupling between one transistor stage and another is accomplished with RC networks (A, fig. 112), transformers (B, fig. 112), impedance coupling networks (C, fig. 112), and direct coupling (D, fig. 112). The advantages, disadvantages, frequency response, and applications of these methods of interstage coupling are covered in paragraphs 116 through 119.

b. The current, voltage, and power gains of transistor stages are usually varied by using variable resistors in the coupling networks. The variable resistor is usually referred to as a *gain* control. At audio frequencies, the gain control is referred to as a *volume* control. Coupling networks that include volume controls are covered in paragraphs 120 through 124.

116. RC-Coupled Amplifiers

a. The RC network (in dashed lines, A, fig. 112) used between two transistor stages consist of a collector load resistor (R1) for the first stage, a dc blocking capacitor (C1), and a dc return resistor (R2) for the input element of the second stage.

(1) Because of the dissipation of dc power in the collector load resistor, the efficiency (ratio of ac power output to dc power delivered to stage) of the RC-coupled amplifier is low.

(2) The dc blocking capacitor prevents the dc voltage of the collector of the first stage from appearing on the input terminal of the second stage. To prevent a large signal voltage drop across the dc blocking capacitor, the reactance of the capacitor must be small compared to the input resistance of the following stage with which it is in series. Since the input resistance of the following stage is low (usually lower than 1,000 ohms), the capacitance value must be high. However, because of the low voltages used, the physical size of the capacitor can be kept small.

(3) The resistance of the dc return resistor is usually 7 to 15 times the input resistance of the second stage. This ratio is selected to prevent shunting the signal current around the input circuit of the second stage. The upper limit of the value of this resistor is dictated by dc bias temperature stabilization considerations (ch. 5).

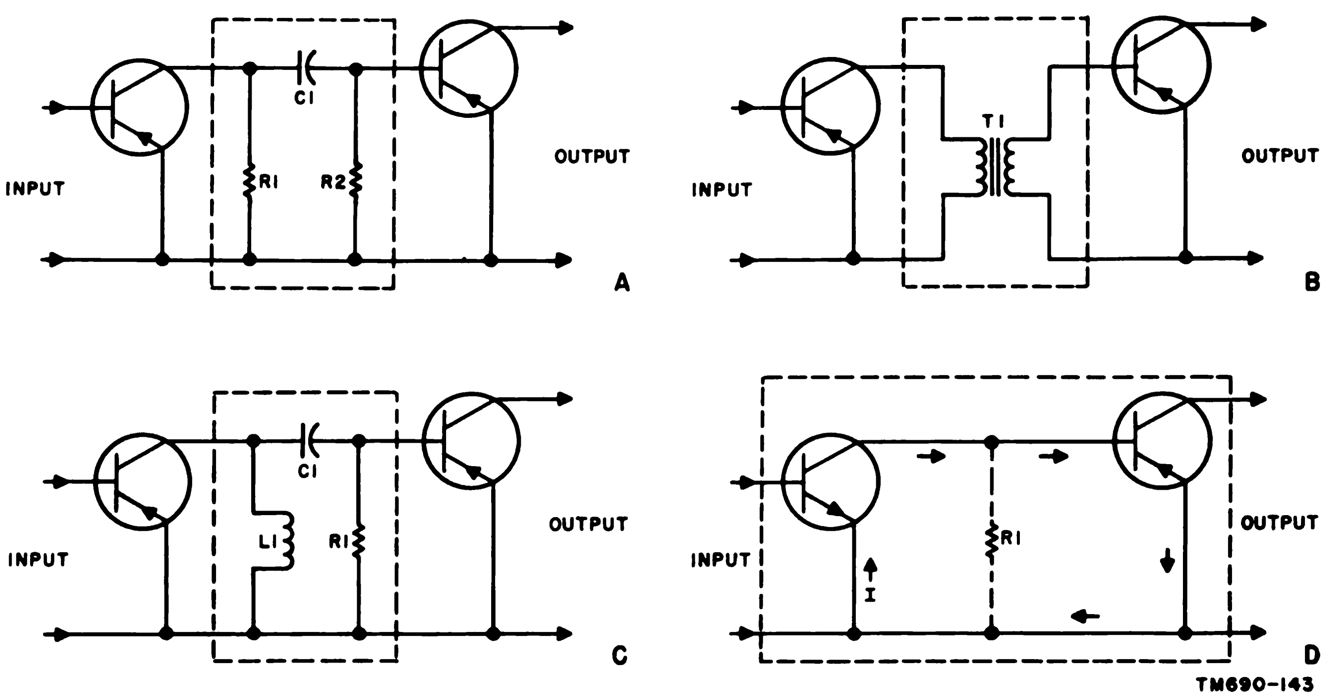


Figure 112. Interstage coupling networks.

b. The very low frequencies, *a*(2) above, are attenuated by the coupling capacitor whose reactance increases with low frequencies. The high-frequency response of the transistor is limited by the shunting effect of the collector-emitter capacitance of the first stage and the base-emitter capacitance of the second stage. Circuit techniques that can be used to extend the low-frequency and the high-frequency response of RC-coupled transistors are covered in chapter 9.

c. RC coupling is used extensively with junction transistors. High gain, economy of circuit parts, and small size can be achieved with RC coupling. By using emitter swamping resistors and self-bias (ch. 5), good temperature stability can be achieved. RC coupling is used extensively in audio amplifiers, such as low-level, low-noise preamplifiers up to high-level amplifiers for power stages. Use of RC coupling in battery operated equipment is usually limited to low power operation to limit battery drain.

117. Transformer-Coupled Amplifiers

a. Interstage coupling by means of a transformer is shown in B, figure 112. The primary winding of transformer T1 (including the ac reflected load from the secondary winding) is the collector load impedance of the first stage. The secondary winding of transformer T1 introduces the ac signal to the base and also acts as the base dc return path.

b. The very low resistance in the base path aids temperature stabilization of the dc operating point. With a swamping resistor in the emitter lead, the current stability factor is ideal (ch. 5).

c. Because there is no collector load resistor to dissipate power, the power efficiency of the transformer-coupled amplifier approaches the theoretical maximum of 50 percent. For this reason, the transformer-coupled amplifier is used extensively in portable equipment where battery power is used.

d. Transformers facilitate the matching of the load to the output of the transistor and the source to the input of the transistor to bring about maximum available power gain in a given stage.

e. The frequency response of a transformer-coupled stage is not as good as that of the RC-coupled stage. The shunt reactance of the primary winding at low frequencies causes the low-frequency response to fall off. At high frequencies the response is reduced by the collector capacitance, and the leakage reactance between primary and secondary windings.

f. In addition to the poor frequency response (*c* above), transformers are more expensive, heavier, and larger in size compared to resistors and capacitors required for coupling. Use of transformers therefore is normally confined to those applications requiring high power efficiency and high output power.

118. Impedance-Coupled Amplifier

a. An impedance-coupled amplifier is one in which one or both resistors of an RC-coupled amplifier (A, fig. 112) are replaced by inductors. A more frequently used arrangement is shown in C, figure 112. In this circuit, the collector load resistor is replaced by an inductor. This provides high power efficiency since the dc power loss is eliminated.

b. The low-frequency response is reduced by the shunt reactance of the inductor. The high-frequency response is reduced by the collector capacitance. Unlike the transformer-coupled amplifier, the impedance-coupled amplifier suffers no loss of high frequencies by leakage reactance. The frequency response of the impedance-coupled amplifier is better than that of the transformer-coupled amplifier, but not as good as that of the RC-coupled amplifier.

119. Direct-Coupled Amplifier

a. The direct-coupled amplifier (D, fig. 112) is used for amplification of dc signals and for amplification of very low frequencies.

b. In the circuit shown, an NPN transistor is connected directly to a PNP transistor. The direction of current flow is shown by the arrows. If the collector current of the first stage is larger than the base current of the second stage, then a collector load resistor (R1) must be connected as indicated.

c. Because so few circuit parts are required in the direct-coupled amplifier, maximum economy can be achieved. However, the number of stages that can be directly coupled is limited. Temperature variation of the bias current in one stage is amplified by all the stages, causing severe temperature instability.

120. Electron Tube and Transistor Volume Controls Compared

The designation of volume controls as voltage dividers or current dividers depends on the relative magnitude of the source and load impedances.

a. Voltage Divider. If the source impedance is low compared to the load impedance, the source is considered a constant-voltage source and the volume control is considered a voltage divider. In the case of an electron tube (A, fig. 113), the source (tube $V1$ plate) impedance is low compared to the load (tube $V2$ grid-cathode input). In this case the volume control ($R2$) is a voltage divider. In most cases, the grid-cathode circuit draws no current. The voltage (e_{in}) that appears between the movable arm (connected to grid) and the lower end of variable resistor $R2$ (connected to cathode) controls the signal current flow in the cathode-plate circuit of tube $V2$. By placing the movable arm at the upper or lower end of variable resistor R_2, the signal gain of tube $V2$ can be varied from maximum (tube $V1$ output voltage $e_o = e_{in}$) to zero ($e_{in}=0$) respectively.

b. Current Divider. If the source impedance is high compared to the load impedance, the source is considered a constant-current source and the volume control is considered a current divider. In the case of a transistor (B, fig. 113), the source (transistor $Q1$ collector) impedance is high compared to the load (transistor $Q2$ base-emitter) impedance. In this case the volume control ($R2$) is a current divider. Unlike the electron tube (*a* above), the input circuit of the transistor normally draws considerable current. Collector signal output current i_o divides at the movable arm to supply varying amounts of input current (i_{in}) to the transistor $Q2$ base-emitter circuit. Since the base-emitter junction resistance can be considered negligible compared to the normal resistance value of variable resistor $R2$, output current i_o is equal to input current i_{in} when the variable arm is in the upper position. This condition gives maximum gain. With the variable arm in the lower position, no signal input current ($i_o = o$) enters transistor $Q2$. This condition gives zero gain.

121. Volume Control Requirements

a. Noise. Volume controls should be so arranged in a given circuit that the volume control introduces no noise or a minimum amount of noise. This requirement can be achieved by

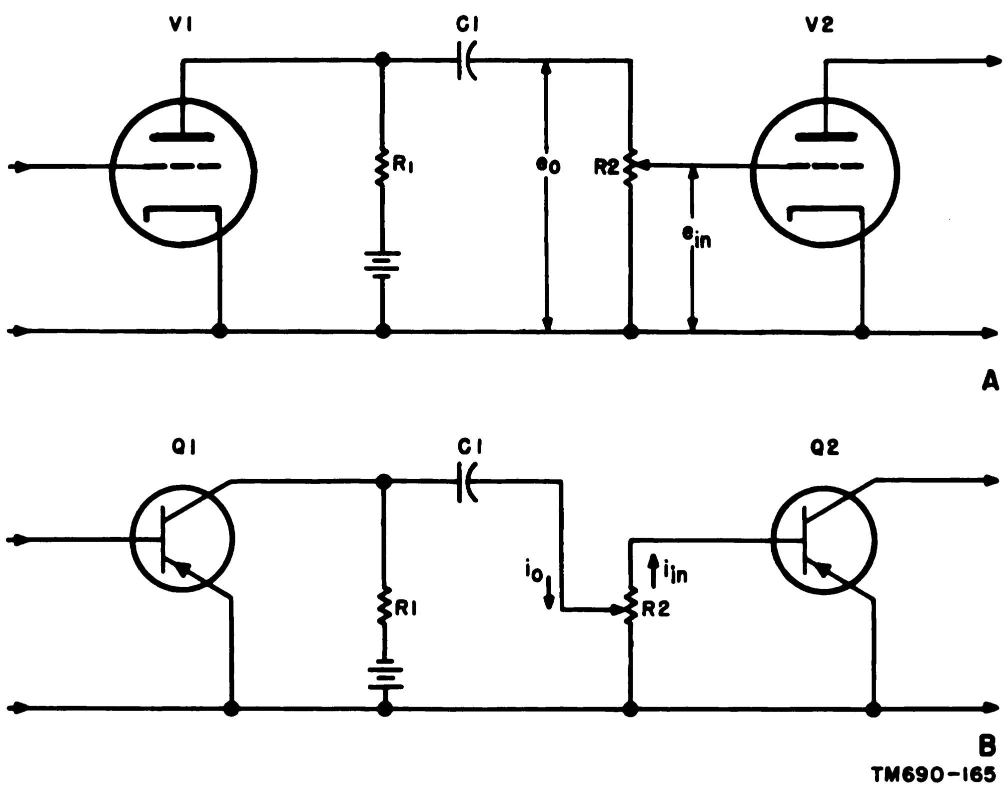


Figure 113. Volume controls used as voltage or current dividers.

avoiding the flow of heavy dc current through the volume control (par. 123).

b. Gain. The volume control and its associated circuit should permit the variation of gain from zero to maximum (par. 122).

c. Frequency. The volume control should be so arranged in a given circuit that all frequencies are attenuated equally for all positions of the variable arm of the control (par. 123).

122. Unsatisfactory Volume Control Circuits

Examples of several volume control circuits not meeting noise, gain, or frequency requirements (par. 121) are discussed below.

a. Use of variable resistor R1 (A, fig. 114) as the collector load and the volume control is unsatisfactory. The heavy dc collector current flowing through variable resistor R1 will cause excessive noise.

b. Use of variable resistor $R2$ (B, fig. 114) as a series volume control eliminates dc current through variable resistor $R2$, but the arrangement is unsatisfactory for two reasons—

(1) Unless the variable resistor is made infinite in value, the amplifier gain cannot be reduced to zero.

(2) The circuit increases the effective base response as more of variable resistor $R2$ is used in the circuit. With the variable arm at the extreme left (zero resistance), capacitor $C1$ used alone, will attenuate the lower frequencies more than the higher frequencies. As the variable arm moves to the right (increased resistance in circuit) the total coupling impedance will be determined mainly by the resistance. As a result, the relative attenuation of the high frequencies compared to the low frequencies will be greater. This action gives the effect of a base boost.

c. Use of a degenerative emitter volume control (variable resistor $R1$ in C, fig. 114) is unsatisfactory for the following reasons:

(1) Unless variable resistor $R1$ is made extremely large (20,000 to 50,000 ohms) the signal cannot be reduced to zero. A resistance of this value would require a very high value of battery (V_{EB}) voltage.

(2) The heavy emitter current flowing through variable resistor $R1$ causes excessive noise.

(3) The circuit increases the effective base response as more of variable resistor R1 is bypassed by capacitor C1. This action occurs because capacitor C1 becomes more effective in bypassing the low frequencies as the resistance it must bypass increases. This results in less degeneration of the low frequencies.

d. The two-stage amplifier (D, fig. 114) uses variable resistor R2 as a volume control. Collector load resistor R1 develops transistor Q1 output signal. Dc blocking capacitor C1 couples the signal to variable resistor R2 and the transistor Q2 input circuit. Resistors R4 and R3 form a voltage divider to establish the base bias voltage. Resistor R5 is the emitter swamping resistor and is ac bypassed by capacitor C2. Collector load resistor R6 develops the output signal. Placement of variable resistor R2 in the circuit as shown is unsatisfactory for the reasons stated below.

(1) The base bias voltage is that voltage developed across resistor R3. This voltage will vary from a maximum to zero depending on the position of the variable arm on variable resistor R2. The voltage developed across resistor R3 must cancel the opposing ac voltage developed across swamping resistor R5 and leave a net forward base-emitter bias. This condition will not be met with variable resistor R2 shunting resistor R3.

(2) The dc current flowing through variable resistor R3 will cause excessive noise.

(3) The gain of the low frequencies will vary according to the position of the variable arm on variable resistor R2. This condition can be understood by considering the base-emitter junction resistance of transistor Q2 as negligibly small. With this assumption, variable resistor R2 can be considered as shunting emitter swamping resistor R5. With all of the resistance of variable resistor R2 in the circuit, capacitor C2 effectively bypasses the low frequencies around the *parallel* combination of resistor R5 and variable resistor R2 to minimize degeneration of the low frequencies. As the variable arm moves to reduce the portion of variable resistor R2 in the circuit, capacitor C2 becomes less effective as a bypass

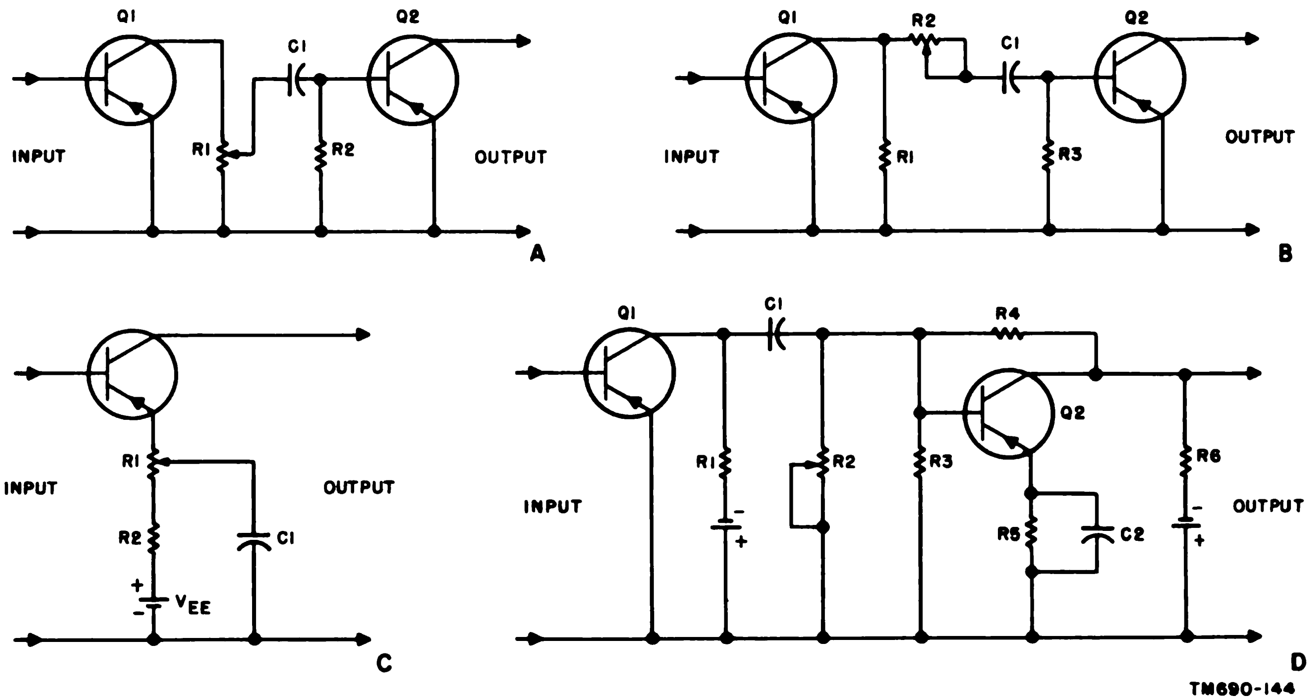


Figure 114. Unsatisfactory volume control circuits.

capacitor for the low frequencies, causing greater degeneration of the low frequencies compared to the high frequencies.

123. Satisfactory Volume Control Circuit

a. A large number of satisfactory volume control circuits can be designed. These circuits can use T, L, or H attenuating pads. A typical satisfactory volume control circuit is discussed in *b* below.

b. The circuit shown in figure 115 avoids the unsatisfactory conditions found in the circuit shown in D, figure 114. The functions of the parts of the two circuits bearing the same reference symbols are the same (par. 122*d*). Dc blocking capacitor C3 has been added to the circuit of figure 115.

(1) Capacitor C3 avoids the variation in base bias voltage by dc isolating resistor R2 from resistor R3 (par. 122*d* (1)). The noise factor of the amplifier is also improved by preventing dc current flow through the variable resistor (par. 122*d*(2)).

(2) By connecting the lower end of variable

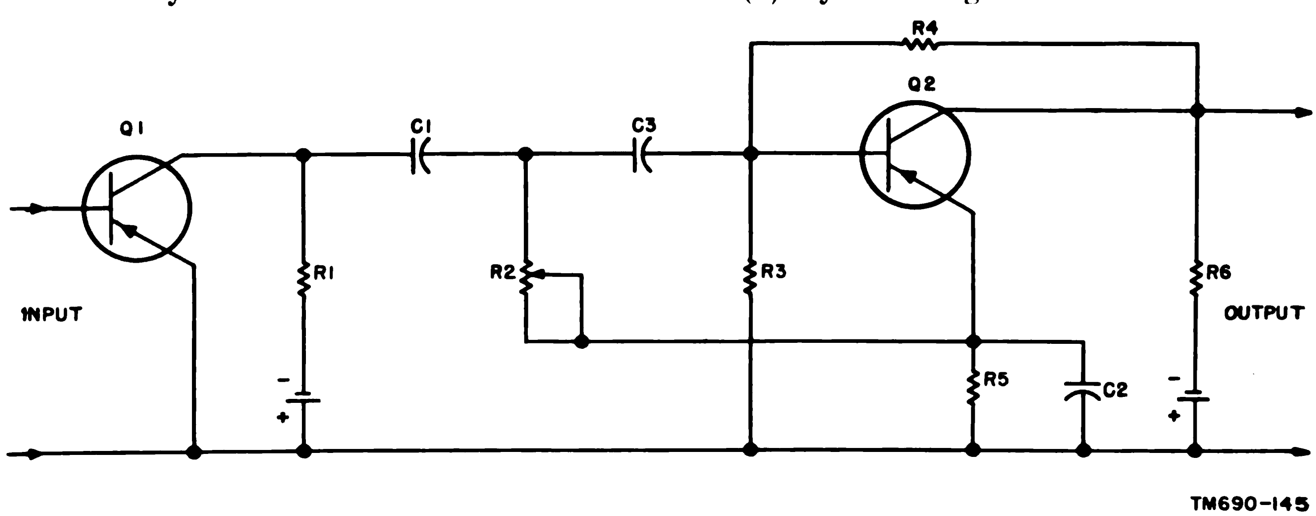


Figure 115. Satisfactory volume control circuit.

resistor R2 to the upper end of resistor R5 and capacitor C2, variable resistor R2 does not ac shunt resistor R5 and capacitor C2. This arrangement avoids the deterioration of the low-frequency response of the amplifier as the resistance of variable resistor R2 is reduced (par. 122*d*(3)).

124. Volume Controls in Transformer Coupling

a. Volume controls used in transformer-coupled amplifiers (fig. 116) must meet the requirements specified in paragraph 121. The volume control circuit should be designed so that a large change in reflected impedance is avoided. Such a large

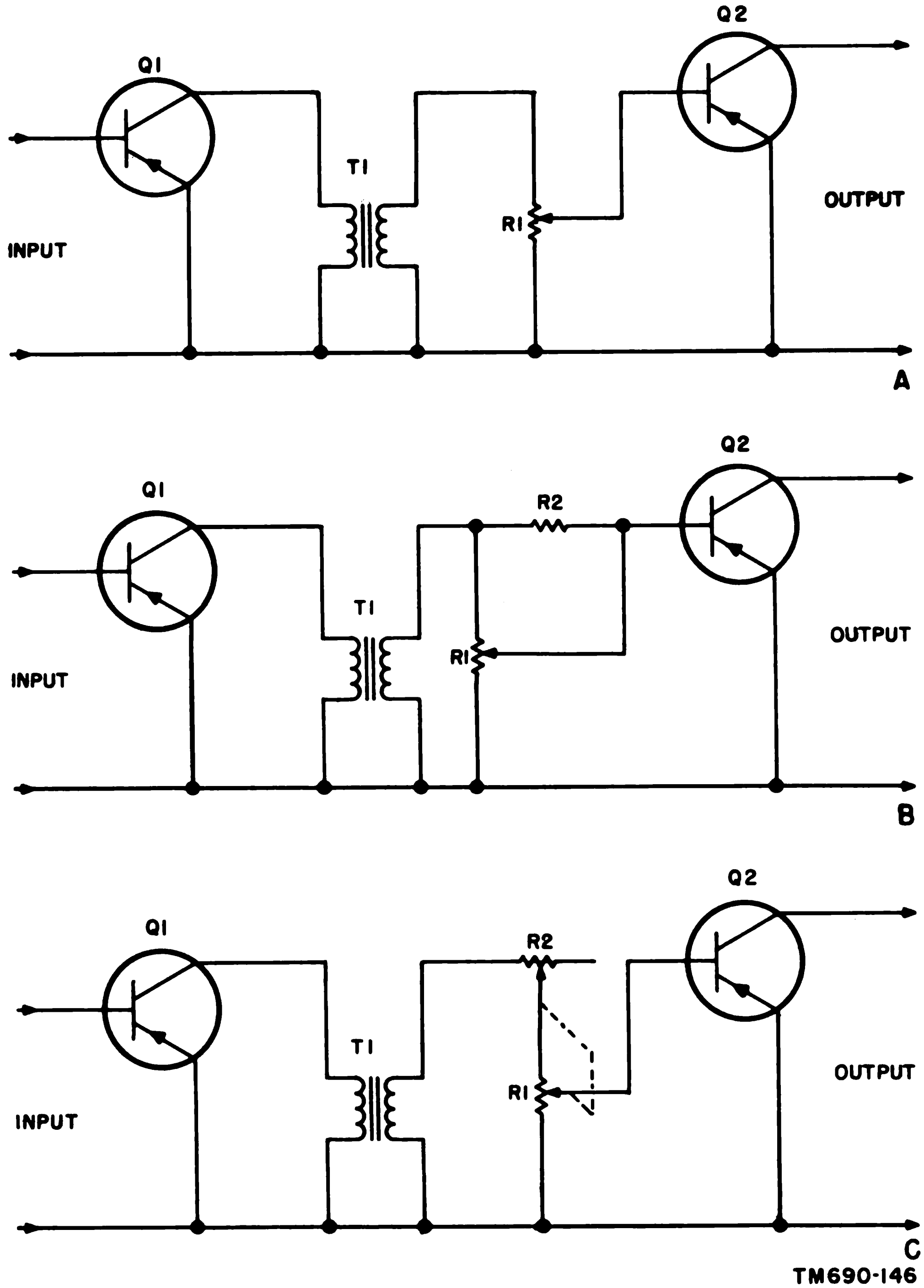


Figure 116. Volume control circuit arrangements used in transformer-coupled amplifiers.

change causes an unequal frequency attenuation. This condition is more important in transformer-coupled amplifiers than in RC coupled amplifiers because the impedance reflected to the primary circuit (usually the collector circuit) from the secondary circuit (usually the base-emitter circuit) is equal to the *secondary impedance multiplied by the square of the turns ratio of the transformer.* A small variation in the secondary circuit impedance can cause a very large variation in the primary circuit impedance.

b. An unsatisfactory volume control circuit is shown in A, figure 116. Variable resistor R1 is made low in value, usually 1,000 ohms, to minimize impedance variations. Transistor Q2 input is made less than 1,000 ohms. At zero volume, the secondary load impedance consists only of variable resistor R1, 1,000 ohms. At full volume, the secondary load impedance consists of variable resistor R1 in parallel with transistor Q2 input resistance; the total resistance is normally about 300 to 400 ohms. These values multiplied by the square of the turns ratio of transformer T1 cause extremely large variations in impedance in the primary circuit. In addition, the low value of variable resistor R1 shunts a large portion of the signal current at full volume.

c. An improvement of the circuit (*b* above) is shown in B, figure 116. Resistor R2 is placed in series with the transistor Q2 input to minimize impedance variations reflected to the primary winding. Usually the value of resistor R2 is made equal to the transistor Q2 input resistance. At maximum and zero volume, the secondary impedances are equal. At the midpoint of variable resistor R2, the secondary resistance rises approximately 25 percent of the value at maximum or zero volume.

d. An ideal arrangement is shown in C, figure 116. By using mechanically ganged dual variable resistors (R1 and R2), the secondary impedance can be maintained constant at all volume control settings. As a larger portion of variable resistor R1 is shunted by transistor Q2 input impedance to increase volume, a larger portion of resistor R2 is inserted in the circuit. This action is accomplished by having the variable arm of variable resistor R2 move to the right (in the diagram) as the variable arm of variable resistor R1 moves up.

125. Tone Controls

a. Tone control circuits as well as volume control circuits (par. 120), may be used as current dividers. The tone control permits manual adjustment of the frequency response of an amplifier.

b. Any fixed low-frequency, or high-frequency, compensating network can be changed to a tone control circuit by substituting a manually variable circuit element for one of the fixed circuit elements.

(1) In the circuit of figure 109, capacitor C2 and resistor R3 form a low-frequency compensating network (par. 112). By substituting a variable resistor for resistor R3, a low-frequency boost tone control circuit can be achieved.

(2) In the circuit of figure 111, capacitors C1 and C2 and resistor R3 form a high-frequency compensating network (par. 114). By substituting a variable resistor for resistor R3, a treble (high-frequency) boost tone control circuit can be achieved.

Section IV. PHASE INVERTERS USED AS DRIVER STAGES

126. General

a. In an equipment, driver stages are located immediately before the power output stage. Most power output stages are push-pull stages that require two input signals, each 180° out of phase with the other. The phase requirement for the push-pull stage can be accomplished by using a tapped-secondary transformer between a single-ended driver stage and the output stage. For economy purposes and for better frequency response, transformer coupling (par. 117) may not be desirable. RC coupling normally is more economical and gives better frequency response.

b. If RC coupling is used between the driver stage and a push-pull stage, the driver stage must provide phase inversion to produce two signals 180° out of phase. Phase inverters may use one stage (one transistor) (par. 127) or two stages (two transistors) (pars. 128 and 129).

127. One-Stage Phase Inverters

a. Figure 117 shows a split-load phase inverter (transistor Q1) feeding a push-pull output stage (transistors Q2 and Q3).

(1) Transistor Q1 output current flows through collector load resistor R3 and emitter load resistor R2. Resistors R2 and R3 are equal in value. Resistor R1 establishes the base bias voltage.

(2) When the input signal aids the forward bias (base becomes more negative), the output current (I_o) increases. The increased output current causes the top side of resistor $R3$ to become more *positive* with respect to ground, and the top side of resistor $R2$ to become more *negative* with respect to ground. When the input signal opposes the forward bias, the output current decreases and causes voltage polarities across resistors $R3$ and $R2$ opposite to those indicated. This action produces two output signals that are reversed 180° with respect to each other. The signal developed across resistor $R3$ is coupled to the input circuit of transistor $Q2$ through dc blocking capacitor $C1$. The signal developed across resistor $R2$ is coupled to the input circuit of transistor $Q3$ through dc blocking capacitor $C2$.

(3) In this circuit, equal voltage outputs are obtained by making resistor $R2$ equal in value to resistor $R3$. However, an unbalanced impedance results because the collector output impedance of transistor $Q1$ is higher than its emitter output impedance. Distortion can occur whenever strong signal current outputs occur (*b* below).

b. The circuit in figure 118 is similar to that shown in figure 117 (*a* above). However, the addition of resistor $R4$ (fig. 118) overcomes the unbalanced output impedance (*a*(3) above). The functions of the parts in this circuit are the same as the correspondingly referenced parts in the unbalanced circuit (*a* above).

(1) The values of resistors R2 and R4 are chosen so that the signal source impedance for transistor Q2 is equal to the signal source impedance for transistor Q3. This eliminates distortion at strong signal current values.

(2) The signal voltage loss across resistor R4 is compensated by making resistor R2 higher in value than resistor R3.

c. Because of the large negative feedback voltage developed across resistor R2, a large signal input is required to drive the one-stage phase inverter. This disadvantage can be overcome by using two-stage phase inverters (pars. 128 and 129). In addition, a two-stage phase inverter provides more power output than a one-stage phase inverter. This advantage is important if the driver stage must feed a large amount of power to a high-level push-pull power output stage.

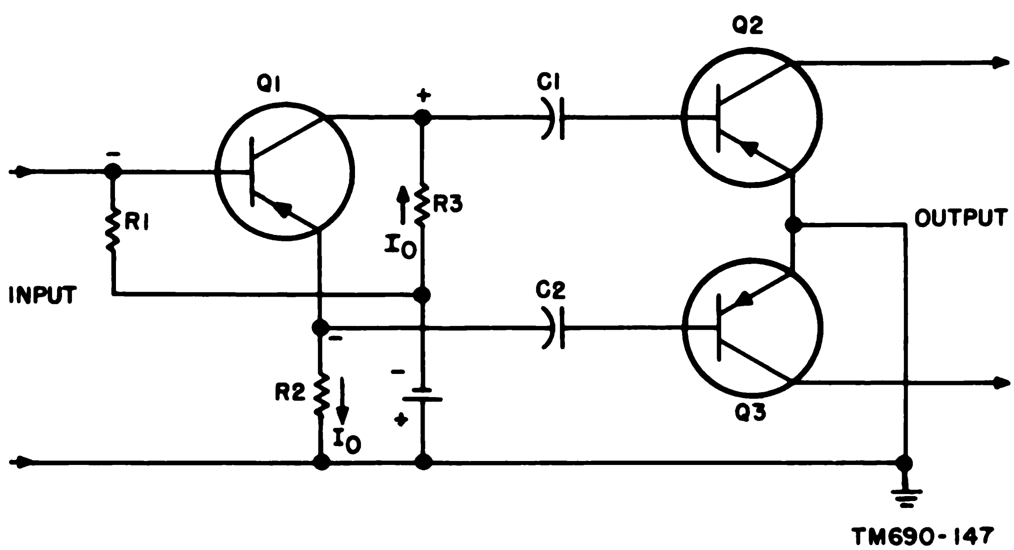


Figure 117. One-stage phase inverter.

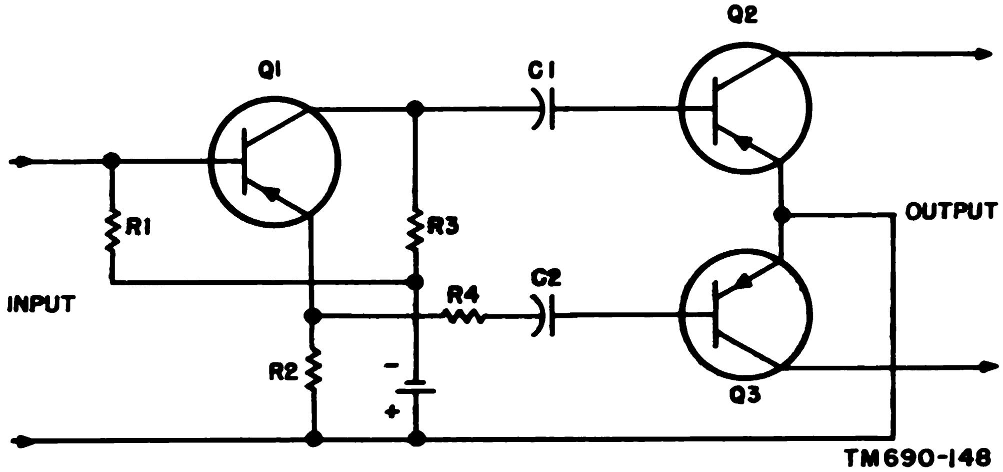


Figure 118. One-stage phase inverter with equalized output impedance.

128. Two-Stage Phase Inverter, CE and CB Configurations

Figure 119 shows a two-stage phase inverter using one CE configuration (transistor Q1) and one CB configuration (transistor Q2), feeding a push-pull output stage (transistors Q3 and Q4).

a. Resistor R1 establishes the base bias on transistor Q1 (par. 79*c*). Collector load resistor R4 develops the output signal of transistor Q1. Dc blocking capacitor C2 couples the output of transistor Q1 to the base of transistor Q3. Emitter load resistor R2 develops a small signal to feed transistor Q2. Capacitor C1 places the base of transistor Q2 at ac ground potential. Resistor R3 establishes the base bias voltage for transistor Q2. Collector load resistor R5 develops the output signal of transistor Q2. Dc blocking capacitor C3 couples the output of transistor Q2 to the base of transistor Q4.

b. Two signals reversed 180° in phase are obtained from the phase inverter. Assume that an input signal causes the base of transistor Q1 to go negative. Because of the 180° phase reversal of the signal in a CE configuration (ch. 3), the top side of collector load resistor R4 goes positive. The same signal current that causes the top side of resistor R4 to go positive, flows through resistor R2 and causes the top side of resistor R2 to go negative; that is, the emitter potential *follows* the base potential. Since there is no phase reversal of the signal through a CB configuration (ch. 3), the end of collector load resistor R5 that is connected to the collector goes negative with the emitter. Thus capacitor C2 couples a positive-going signal into the transistor Q3 base while capacitor C3 couples a negative-going signal into the transistor Q4 base to achieve the 180° phase reversal required. If the incoming signal at the transistor Q1 base were going positive, all polarities indicated would be reversed.

c. Although there is some negative feedback at the input of transistor Q1 caused by resistor R2, the input resistance to transistor Q1 remains low because resistor R2 is shunted by the low resistance of the transistor Q2 input circuit.

129. Two-Stage Phase Inverter, CE Configurations

a. Figure 120 shows a two-stage phase inverter consisting of two identical CE configurations. The application of this circuit is identical with that in paragraph 128.

(1) Two signals, 180° reversed in phase, are obtained from the phase inverter. Assume that an input signal drives the base of transistor Q1 negative. Because of the 180° phase reversal in the CE configuration, the transistor Q1 collector goes positive. One portion of this positive signal is coupled to the base of transistor Q2 through dc blocking capacitor C2 and attenuating resistor R4. The other portion of the positive signal is completed through dc blocking capacitor C4 to one input circuit of a push-pull output stage. The positive-going signal on the base of transistor Q2 causes a negative-going signal at the

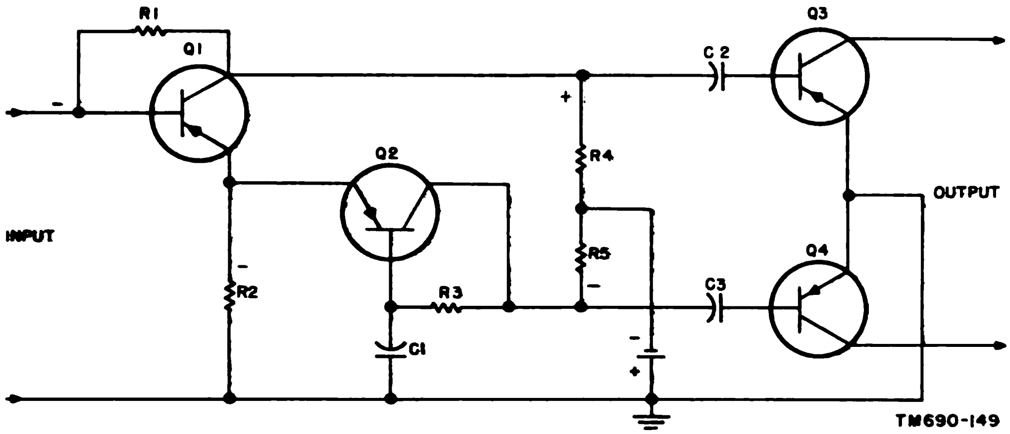


Figure 119. Two-stage phase inverter using a CE and a CB configuration.

collector of transistor Q2. This negative signal is coupled through dc blocking capacitor C5 to the other input circuit of a push-pull output stage.

(2) Resistor R1 provides base bias for transistor Q1. Collector load resistor R3 develops transistor Q1 output signal. Resistor R2 is the emitter swamping resistor and is ac bypassed by capacitor C1. Resistor R5 provides base bias for transistor Q2. Collector load resistor R6 develops transistor Q2 output signal. Resistor R7 is the emitter swamping resistor and is ac bypassed by capacitor C3.

b. Since two identical CE configurations are used, the source impedances are equal for the two input circuits of the push-pull output stage. In addition, the amount of power that can be delivered by the two-stage phase inverter is much greater than that of the split-load phase inverter (par. 127).

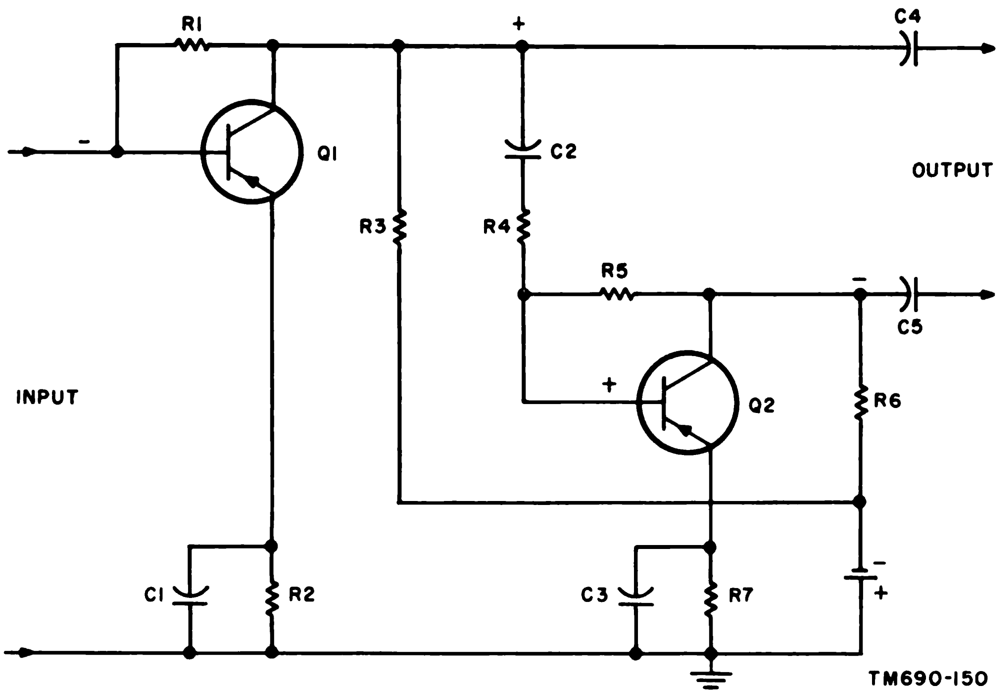


Figure 120. Two-stage phase inverter using two CE configurations.

130. General

a. Single ended and push-pull class A power amplifiers are used in equipments in which large power outputs are not required and high power efficiency is not a dominating factor. These amplifiers are used mainly in applications requiring minimum distortion.

b. Class B push-pull power amplifiers are used mainly in equipments requiring high power output and high power efficiency. The theory of operation of class B push-pull power amplifiers employing two similar type (two PNP or two NPN) transistors is discussed in paragraphs 131 through 133.

131. Class B, Push-Pull, Zero Bias Amplifier

a. Figure 121 shows a simplified circuit of a class B amplifier. The emitter-base junctions are zero biased. In this circuit each transistor conducts on alternate half cycles of the input signal. The output signal is combined in the secondary of the output transformer. Maximum efficiency is obtained even during idling (no input signal) periods, because neither transistor conducts during this period.

b. An indication of the output current waveform for a given signal current input can be obtained by considering the dynamic transfer characteristic (par. 101) for the amplifier. It is assumed that the two transistors have identical dynamic transfer characteristics. This characteristic, for one of the transistors, is shown in A,

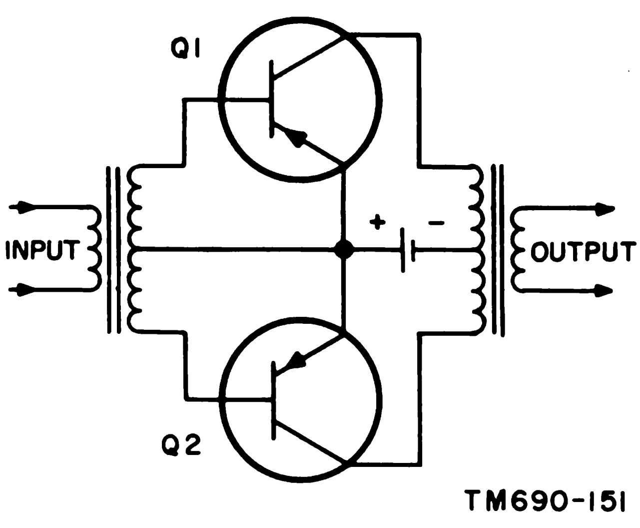


Figure 121. Class B, push-pull amplifier with zero input bias.

figure 122. The variations in output (collector) current is plotted against input (base) current under load conditions. Since two transistors are used, the overall dynamic transfer characteristic for the push-pull amplifier is obtained by placing two of the curves (A, fig. 122) back-to-back. The two curves are shown back-to-back *and combined* (B, fig. 122). Note that the zero line of each curve is lined up vertically to reflect the zero bias current. In C, figure 122, points on the input base current (a sine wave) are projected onto the dynamic transfer characteristic curve. The corresponding points are determined and projected as indicated to form the output collector current waveform. Note that severe distortion occurs at the crossover points; that is, at the points where the signal passes through zero value. This is called crossover distortion. This type of distortion becomes more severe with low signal input currents. Crossover distortion can be eliminated by using a small forward bias on both transistors of the push-pull amplifier (par. 132).

132. Class B, Push-Pull, Low Bias Amplifier

a. A class B push-pull amplifier with a small forward bias applied to the base-emitter junctions is shown in A, figure 124. A voltage divider is formed by resistors R2 and R1. The voltage developed across resistor R1 supplies the base-emitter bias for both transistors. This small forward bias eliminates crossover distortion (par. 131).

b. A study of the dynamic transfer characteristic curve of the amplifier demonstrates the elimination of crossover distortion. In A, figure 123, the dynamic transfer characteristic curve of each transistor is placed back-to-back for zero base current bias conditions. The two curves are back-to-back and not *combined.* The dashed lines indicate the base current values when forward bias is applied to obtain the overall dynamic characteristic curve of the amplifier. With forward bias applied, the separate curve of each transistor must be placed back-to-back and alined at the base bias current line (dashed line). The zero base current lines (solid lines) are offset (B, fig. 123). In C, figure 123, points on the input base current (a sine wave) are projected onto the dynamic transfer characteristic curve.

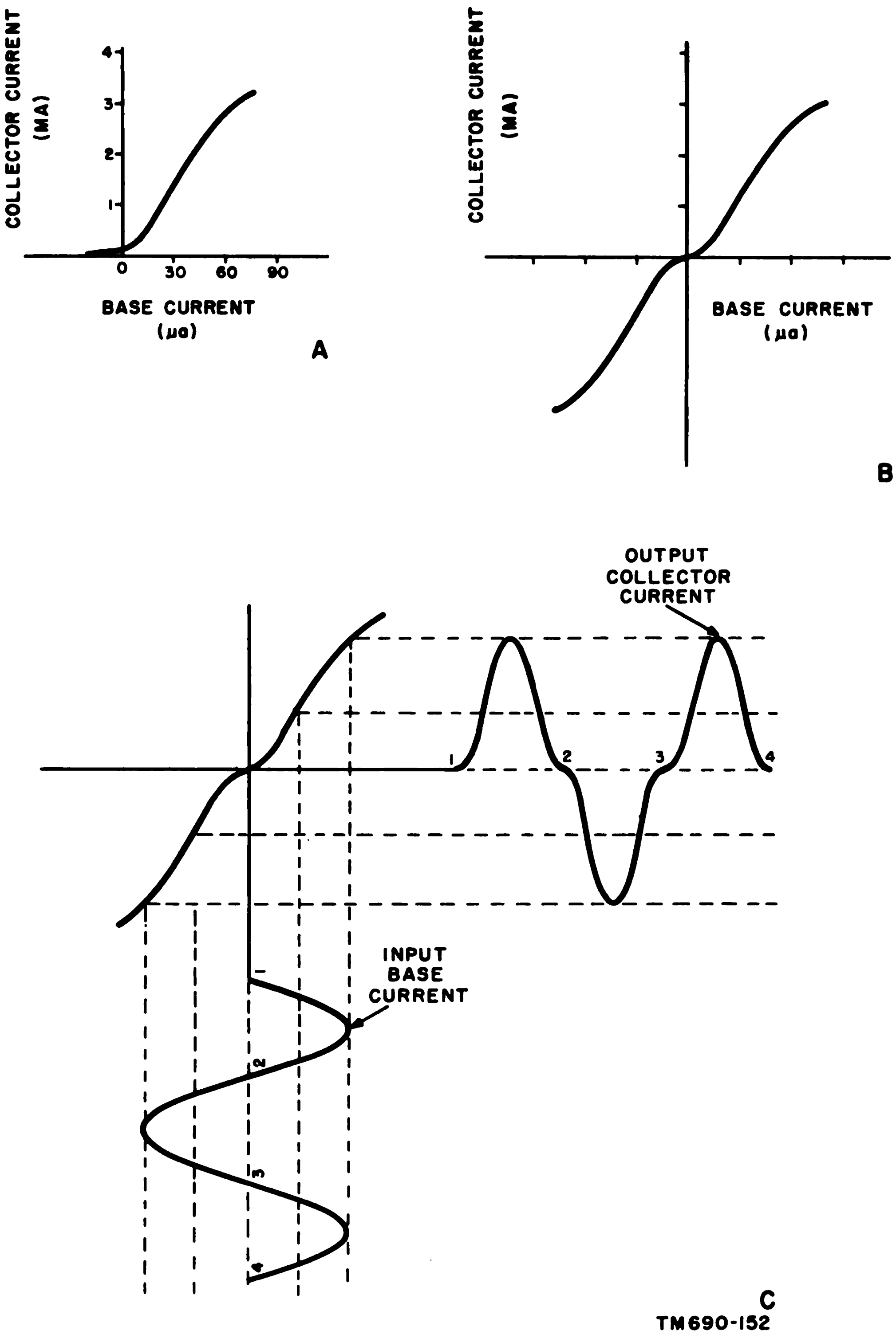


***Figure 122.** Dynamic transfer characteristic curves of class B, push-pull amplifier with zero bias, showing input and output current waveforms.*

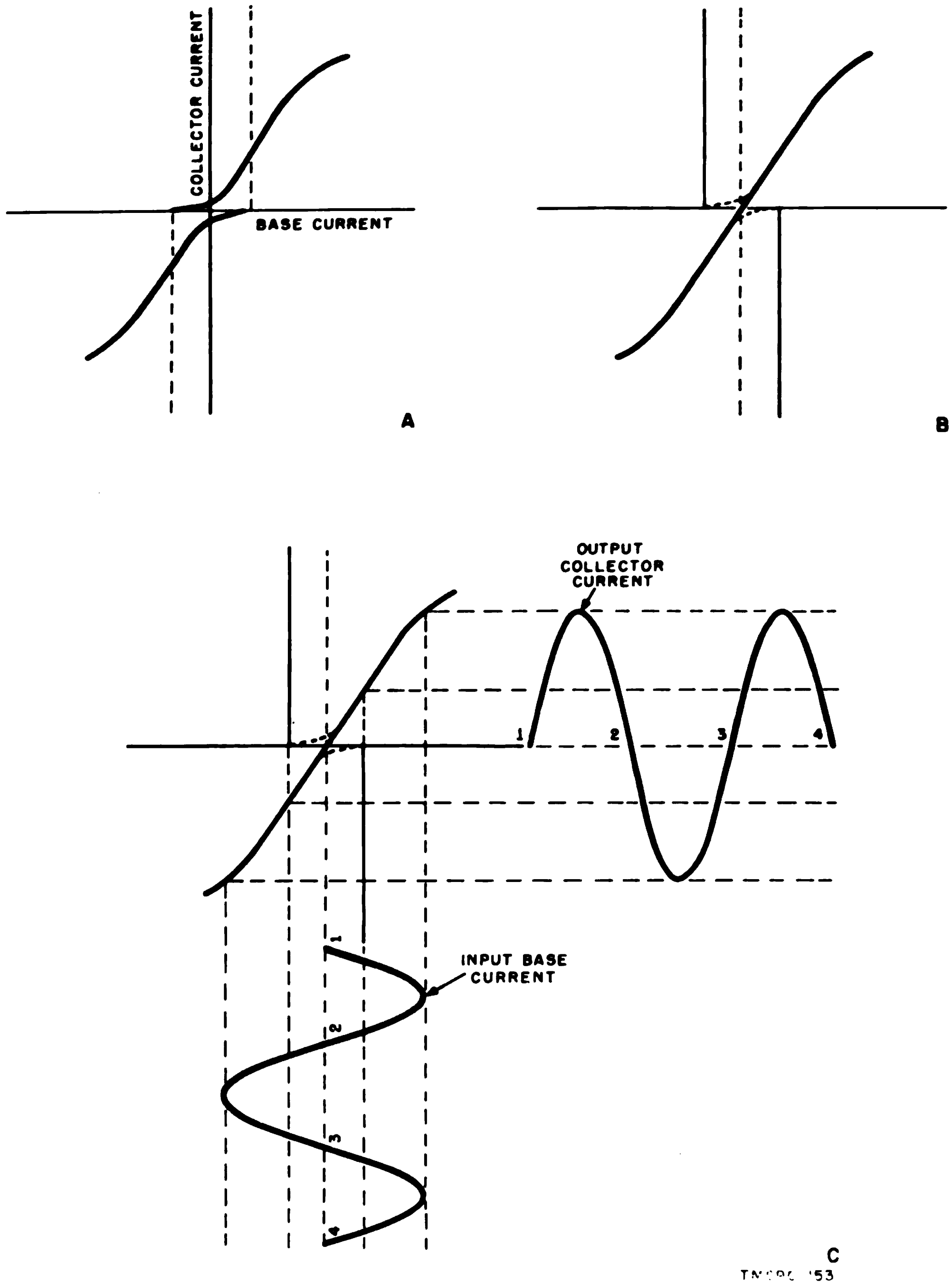


Figure 123. Dynamic transfer characteristic curves of class B, push-pull amplifier with small forward bias, showing input and output current waveforms.

The corresponding points are determined and projected as indicated to form the output collector current waveform. Compare this output current waveform with that shown in C, figure 122. Note that crossover distortion does not occur when a small forward bias is applied.

c. In the voltage divider (A, fig. 124), electron current flow from the battery is in the direction of the arrow. This current establishes the indicated polarity across resistor R1 to furnish the required small forward bias (*b* above). Note that no bypass capacitor is used across resistor

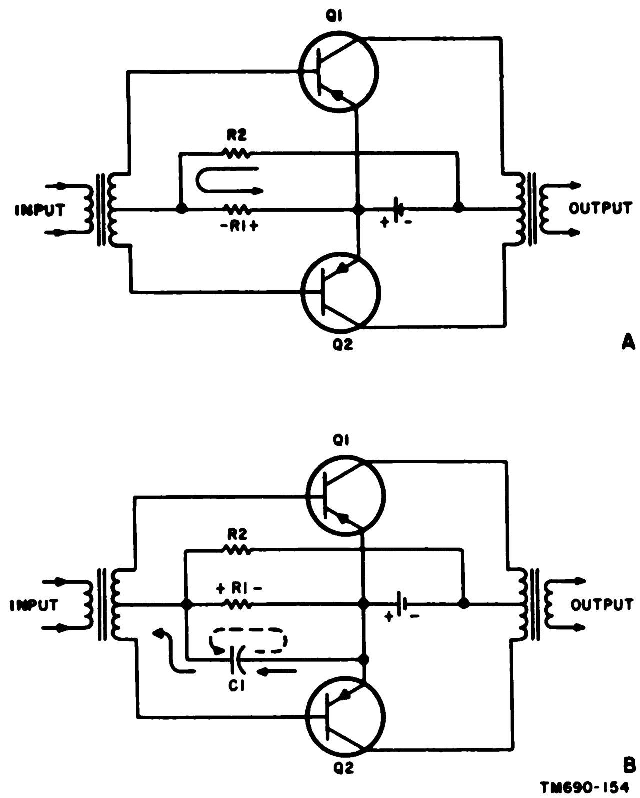


Figure 124. Class B, push-pull amplifiers with small bias voltage applied.

R1. If a bypass capacitor were used (B, fig. 124), the capacitor would charge (solid-line arrow) through the base-emitter junction of the conducting transistor (during the presence of a signal), and discharge (dashed-line arrow) through resistor R1. The discharge current through resistor R1 would develop a dc voltage with the polarity indicated. This is a reverse bias polarity that could drive the amplifier into class C operation with the resultant distortion even more severe than crossover distortion. The capacitor must not be used.

133. Class B Push-Pull Amplifier, Capacitance-Diode Coupling

a. RC coupling to the input of a class B push-pull amplifier causes rectification that can reverse bias the amplifier (A, fig. 125). In the class B push-pull amplifier, one transistor conducts for one-half cycle of the input signal while the other transistor remains nonconducting (except for a small forward bias that eliminates crossover distortion (par. 132)). Assume that the input signal causes transistor Q1 to conduct. Electrons (solid-line arrows) leave the right hand plate of coupling capacitor C1, enter the transistor Q1 base-emitter junction, flow to the emitter ground connection, through ground to the junction of resistor R1 and resistor R2, through resistor R1 to the left hand plate. Resistor R1 represents the output resistance of the previous amplifier. The charging of capacitor C1 occurs rapidly, because of the low-resistance path. Capacitor C1 cannot *discharge* through the transistor Q1 emitter-base junction; for practical purpose, the junction represents an open circuit to electron flow from emitter to base. Electrons on the left hand plate of capacitor C1 must flow through resistor R1, through ground to the ground connection at the emitter lead, through resistors R6 and R3 to the right hand plate. The discharge path is shown by the dashed-line arrows. Capacitor C1 discharges slowly because it must discharge through resistor R3. Normally resistor R3 is made large to avoid shunting signal current around the transistor Q1 base-emitter junction. The discharge current through resistor R3 develops a reverse-bias voltage with the polarity indicated. The reverse bias can cause class C operation with resultant severe distortion of the signal. The discharge current flowing through resistor R6 does not result in a reverse bias because the battery current flowing in the opposite direction maintains a forward bias.

b. The advantages of RC coupling (economy and better frequency response) can be retained by replacing resistor R3 with a diode (B, fig. 125). Diode CR1 is reverse biased and acts as a high-valued resistance when the input signal causes the transistor Q1 emitter-base junction to conduct. This prevents the shunting of signal current around the emitter-base junction. The charge path (solid-line arrows) for capacitor C1 is the same as that for the circuit shown in A, figure 125 (*a* above). The discharge path is also the same, except that resistor R3 has been replaced by diode CR1 which is forward biased during the discharge period. Diode CR1, when forward biased, exhibits negligible resistance. Capacitor C1 discharges rapidly, therefore, and reverse bias of the emitter-base junction is avoided.

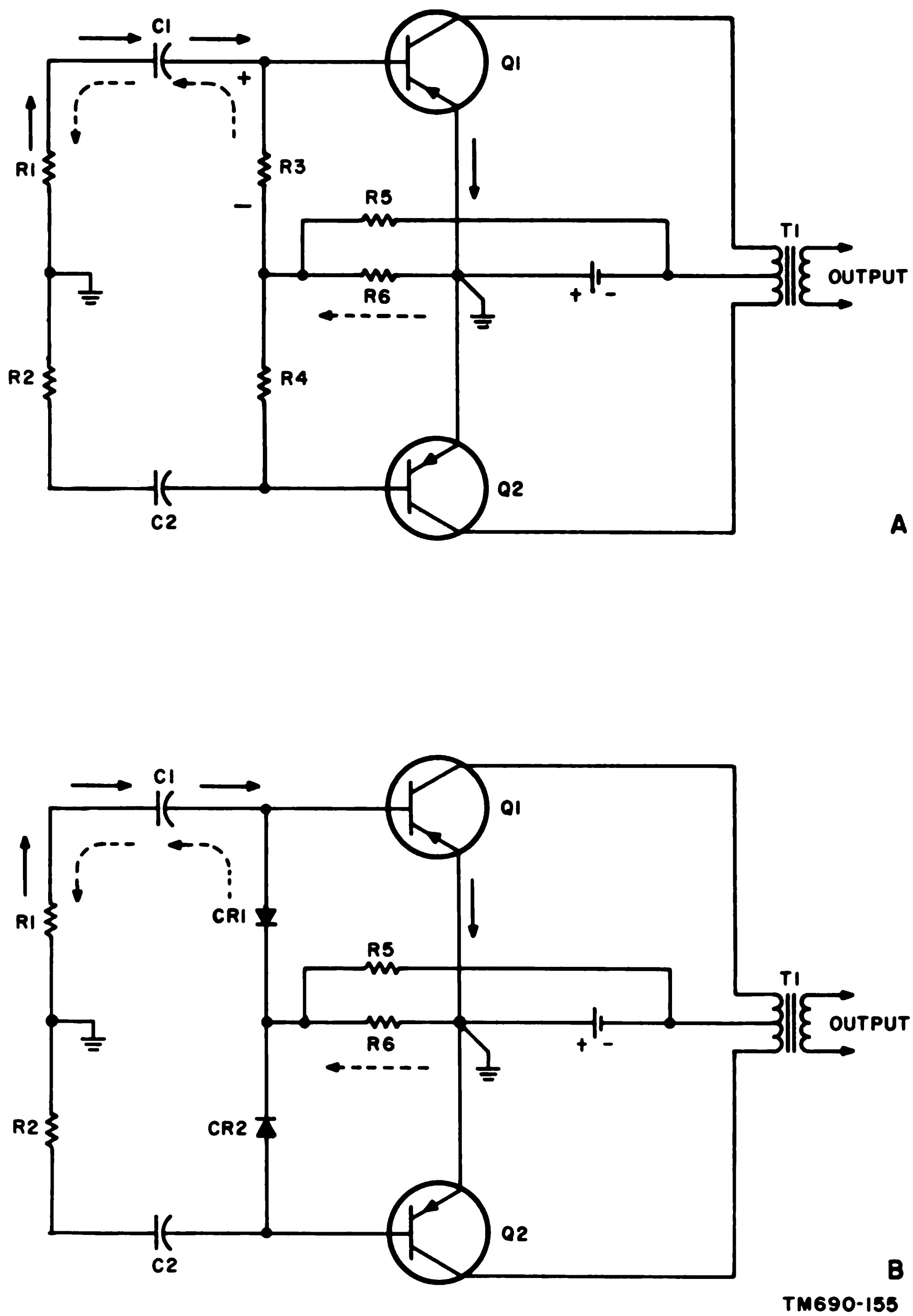


Figure 125. RC-coupled and capacitance-diode coupled class B, push-pull amplifier.

134. General

a. Junction transistors are available as PNP (A, fig. 126) and NPN (B, fig. 126) types. The direction of electron current flow in the terminal leads of the two types of transistors is indicated by the arrows. Note that the direction of current flow in one terminal lead of one transistor is opposite to that in the corresponding terminal of the other transistor. If the two types of transistors are connected in a single stage (fig. 127), the dc electron current path (indicated by arrows) in the output circuit is *completed* through the collector-emitter junctions of the transistors. When connected in this manner, the circuit is referred to as a *complementary symmetry* circuit. The theory of operation of this circuit is discussed in paragraph 135.

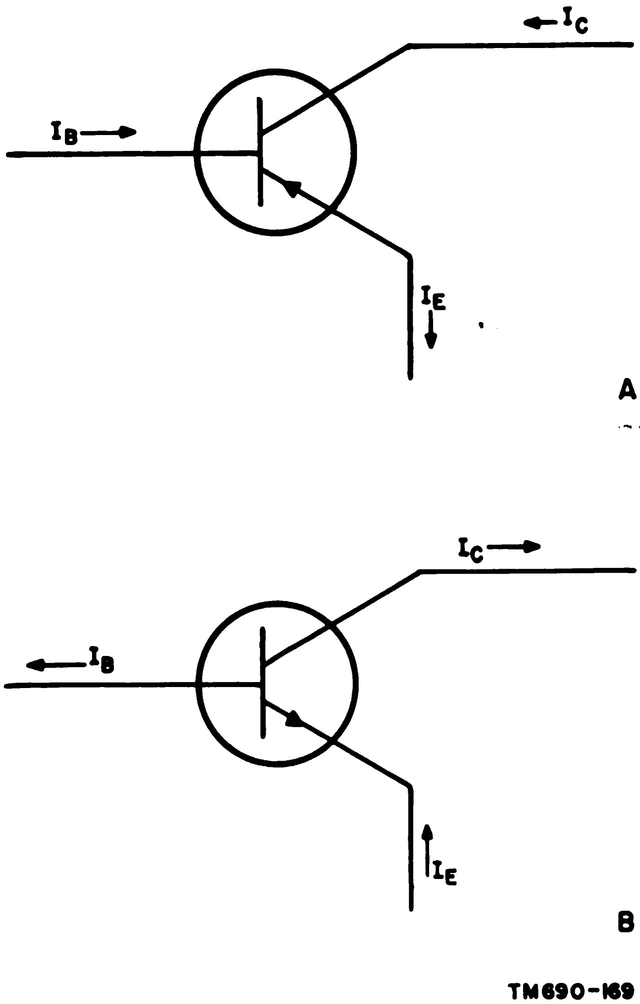


Figure 126. Dc current flow in PNP and NPN transistors.

b. The advantages of complementary symmetry circuits are as follows:

(1) The complementary symmetry circuit provides all the advantages of conventional push-pull amplifiers without the need for a phase-inverter driver stage, or for a center-tapped input transformer.

(2) The input coupling capacitor (fig. 127) charges through one transistor during the positive half cycle of the input signal and discharges through the other transistor during the negative half cycle of the input signal. This action eliminates the need for discharge diodes with capacitance coupling as is required in conventional class B push-pull amplifiers (par. 132).

(3) The parallel connection of the output circuit (fig. 127) with respect to the load eliminates the need for a tapped-primary transformer in the output circuit.

135. Complementary Symmetry Circuit

a. Figure 127 shows two transistors in a complementary symmetry connection. Transistor Q1 is a PNP transistor and transistor Q2 is an NPN transistor. A negative going input signal forward biases transistor Q1 and causes it to conduct. A positive going input signal forward biases transistor Q2 and causes it to conduct. As one transistor conducts, the other is nonconducting, because the signal that forward biases one transistor, reverse biases the other transistor.

b. The resultant action in the output circuit can be understood by considering the circuit of figure 128. This is a simplified version of the output circuit. The internal emitter-collector circuit of transistor Q1 is represented by variable resistor R1 and that of transistor Q2 by variable resistor R2.

(1) With no input signal and class B operation (zero emitter-base bias), the variable arms of the variable resistors can be considered to be in the OFF positions. No current flows through the transistors nor through load resistor R_L. As the incoming signal goes positive, transistor $Q2$ conducts (*a* above) and transistor $Q1$ remains nonconducting. Variable resistor $R1$ remains in the OFF position.

The variable arm of resistor $R2$ moves toward point 3 and current passes through the series circuit consisting of battery V_{CC2}, resistor R_L and variable resistor $R2$. The amount of current flow depends upon the magnitude of the incoming signal, the variable arm moving toward point 3 for increasing forward bias and toward point 4 for decreasing forward bias. The current flows in the direction of the dashed arrow, producing a voltage with the indicated polarity. When the input signal goes negative, transistor $Q1$ conducts and transistor $Q2$ becomes nonconducting. The same action is repeated with variable resistor $R1$. Current flows through battery V_{CC1}, variable resistor $R1$, and load resistor R_L in the direction shown by the solid-line arrow, and produces a voltage across resistor R_L with the polarity indicated.

(2) For class A operation, a forward bias is applied to the two transistors (fig. 127), so that collector current is not cut off at any time. In the simplified circuit (fig. 128), the variable resistors will not be in the OFF position at any time. Dc bias current in the output circuit flows out of the negative terminal of battery V_{CC2} into the positive terminal of battery V_{CC1}, through variable resistor $R1$, variable resistor $R2$, and into the positive terminal of battery V_{CC2}. No current flows through resistor R_L. Under these conditions, the output circuit can be considered a balanced bridge, the arms of the bridge consisting of resistors $R1$ and $R2$ and batteries V_{CC1} and V_{CC2}. When the input signal goes positive, transistor $Q2$ conducts more and transistor $Q1$ conducts less. In the simplified circuit, the variable arm of resistor $R1$ moves toward point 1 and that of resistor $R2$ moves toward point 3. This action results in an unbalanced bridge and electrons flow through resistor R_L in the direction of the dashed-line arrow, producing a voltage with the indicated polarity. When the input signal goes negative, transistor $Q1$ conducts more and transistor $Q2$ conducts less. In the simplified circuit, the variable arm of resistor $R1$ moves toward point 2 and that of resistor $R2$ moves toward point 4. Again the bridge is unbalanced, and electrons flow through re-

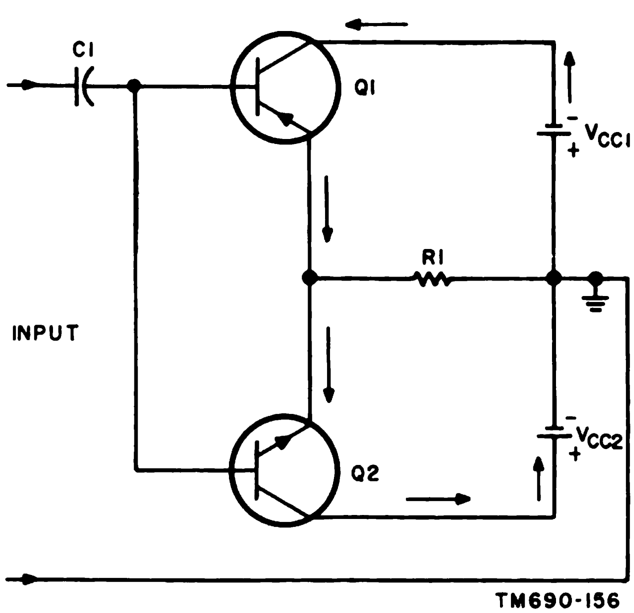


Figure 127. Zero bias complementary symmetry circuit.

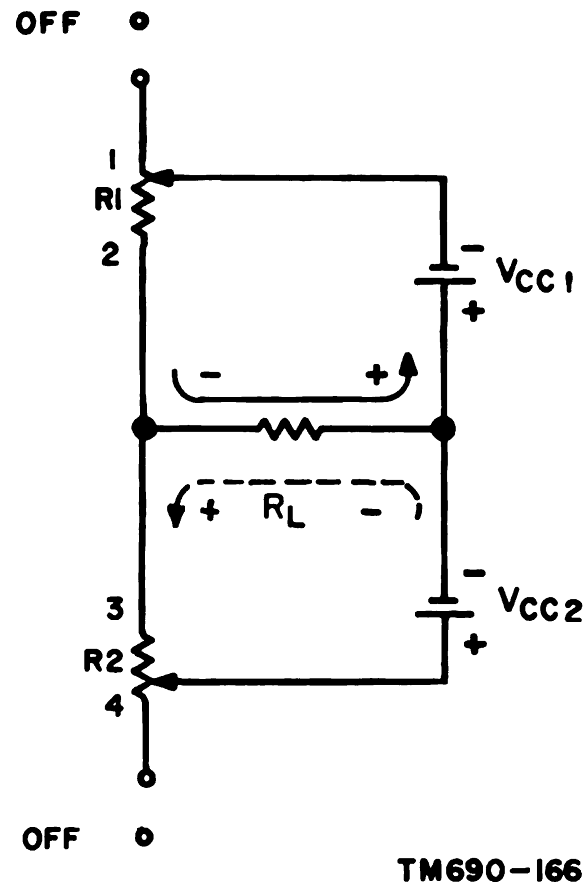


Figure 128. Simplified version of output circuit of complementary symmetry circuit.

sistor R_L in the direction of the solid-line arrow, producing a voltage with the indicated polarity.

c. In either class B (*b*(1) above) or class A (*b*(2) above), dc current does not flow through the load resistor. Advantage of this property can be taken by connecting the voice coil of a loudspeaker directly in place of resistor R_L. The voice coil will not be offset by dc current flow through it and thus distortion will not occur.

136. Additional Complementary Symmetry Circuits

a. Figure 129 shows a common collector complementary symmetry circuit to which a small forward bias has been applied. This circuit is similar to that shown in figure 127, except for the voltage divider consisting of resistors $R2$, $R1$, and $R3$ connected in series across batteries V_{CC1} and V_{CC2}. The voltage developed across resistor $R1$, with the indicated polarity, provides forward bias for each transistor. Resistor $R1$ is in series with the base-emitter junctions of the two transistors. Considering the dc resistance of the two junctions as being equal to each other, the voltage drop across each junction is half of that developed across resistor $R1$. Normally the value of resistor $R1$ is very small, so that its unbalancing effect on the input signal to transistor $Q2$ is negligible.

b. Figure 130 shows a common emitter complementary symmetry circuit with transformer input and transformer output. The voltage divider consisting of resistors $R3$ and $R1$, establishes the forward bias on transistor $Q1$. The voltage divider, consisting of resistors $R4$ and $R2$ establishes the forward bias on transistor $Q2$. Note that a tapped-secondary transformer is not required at the input. However, a split primary output transformer ($T2$) must be used; unlike conventional push-pull amplifiers, opposite polarity voltages are required for the collectors.

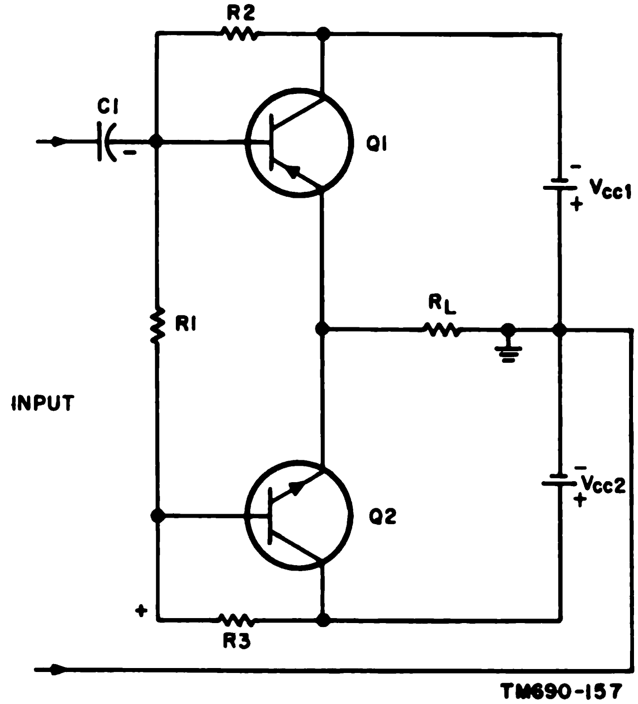


Figure 129. Forward biased common collector complementary symmetry circuit.

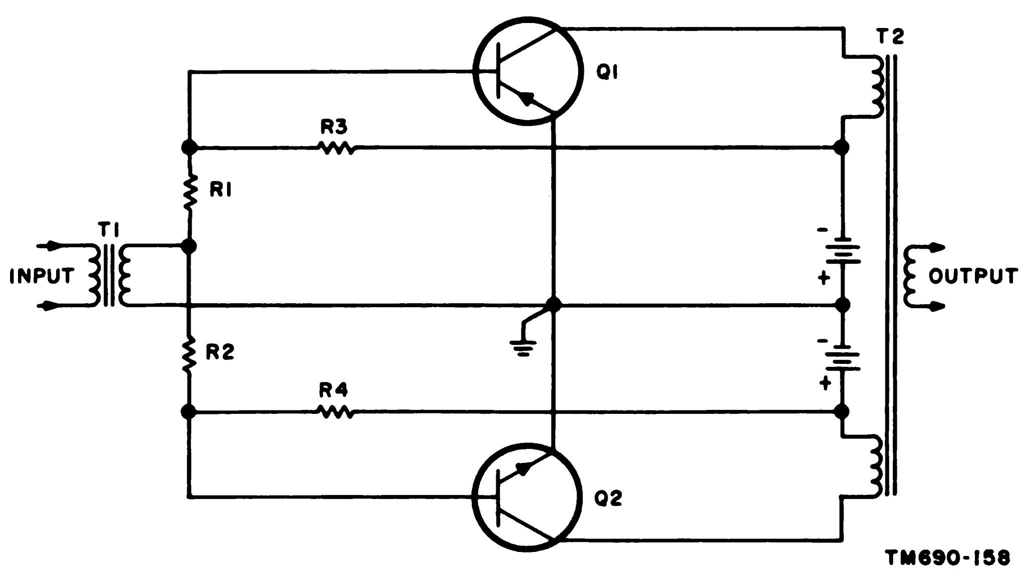


Figure 130. Complementary symmetry circuit with input and output transformer coupling.

c. A, figure 131, shows one common emitter complementary symmetry stage (transistors $Q3$ and $Q4$) being directly driven by another common emitter complementary symmetry stage (transistors $Q1$ and $Q2$). The input to transistors $Q1$ and $Q2$ is from a single-ended stage. When the input signal goes positive, transistor $Q1$ conducts and transistor $Q2$ remains nonconducting. Because of 180° phase reversal in a common emitter configuration, the collector of transistor $Q1$ goes

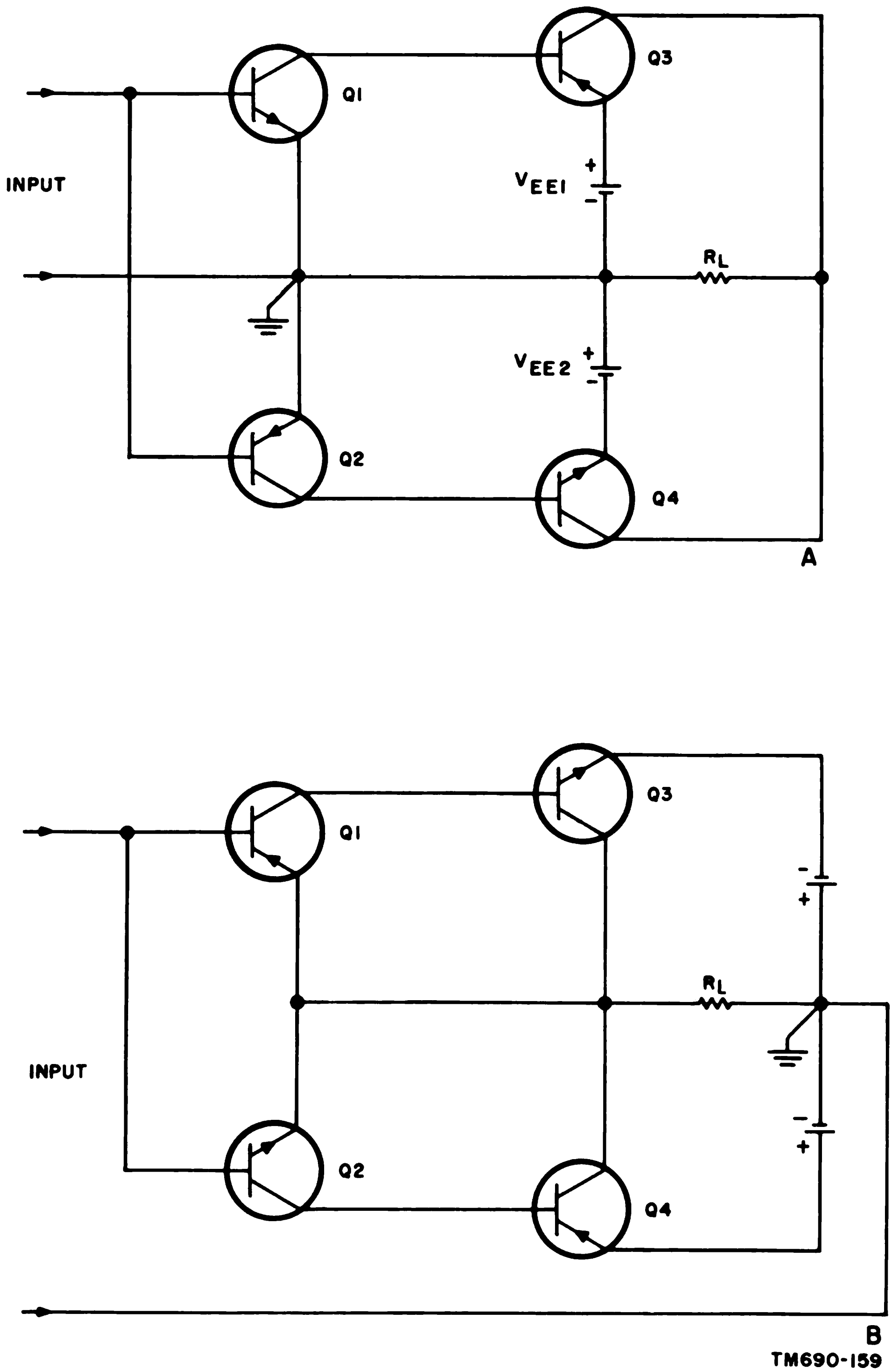


Figure 131. Direct-coupled complementary symmetry stages.

negative which causes transistor $Q3$ to conduct and the transistor $Q3$ collector goes positive. When the input signal goes negative, transistor $Q2$ conducts and its collector goes positive. This causes transistor $Q4$ to conduct and its collector goes negative. Transistors $Q1$ and $Q3$ are nonconducting during this period. Note that battery V_{BB1} supplies the required biasing voltages for transistors $Q1$ and $Q3$. Battery V_{BB2} supplies the required biasing voltages for transistors $Q2$ and $Q4$. The emitter-base junction of transistor $Q3$ is in series with the collector-emitter circuit of transistor $Q1$ and battery V_{BB1}. As a result the emitter of transistor $Q3$ is positive with respect to its base (forward bias), and the collector of transistor $Q1$ is positive with respect to its emitter as is required for electron flow through transistor $Q1$. A similar arrangement exists for transistors $Q2$ and $Q4$ and battery V_{BB2}.

d. A circuit similar in operation to that discussed in *c* above is shown in B, figure 131. However, the common collector configuration is used in this circuit. As in the case of single transistor amplifiers, maximum voltage, current, and power gain is obtained with the common emitter configuration; but its input resistance is low. The common collector provides higher input resistance with less current, voltage, and power gain.

Section VII. COMPOUND-CONNECTED TRANSISTORS

137. General

a. The current gain (par. 54), voltage gain (par. 55), and power gain (par. 56) of a transistor amplifier are directly proportional to the short circuit forward current amplification factor. This factor is the ratio of the output current to the input current measured with the output short circuited and is designated α_{fb} (common base), α_{fe} (common emitter), or α_{fc} (common collector). If α_{fb} of a given transistor is high, the α_{fe} and α_{fc} will be correspondingly high; therefore the following discussion is concerned only with α_{fb}.

b. To obtain maximum gain in a given amplifier, it is necessary to use a transistor having a high-valued short circuit forward current amplification factor (α_{fb}). Most transistors have an α_{fb} ranging from 0.940 to 0.985 with the average value of 0.96. Regardless of the average value of α_{fb}, this factor will decrease as emitter current increases. Figure 132 shows the variation of collector current (output) with emitter current (input) for a single transistor. The ratio of collector current to emitter current remains constant (straight line portion) to approximately 400 milliamperes; then the collector current increases less rapidly (current portion) as the emitter current increases. This indicates a reduction in the current amplification factor at high values of emitter current. This phenomenon is most pronounced in power amplifiers that draw heavy emitter current.

c. By compound connecting two transistors (fig. 133), the dropoff of collector current (or reduction in α_{fb}) at high emitter currents can be made negligible. The theory of operation of compound-connected transistors is covered in paragraph 138. The variation of total collector current with input emitter current of compound-connected transistors is shown in figure 132. Note that total collector current does not drop off as the emitter current increases.

138. Compound Connection

a. The dashed lines in figure 133 inclose two compound-connected transistors. Note that the base of one transistor ($Q1$) is connected to the

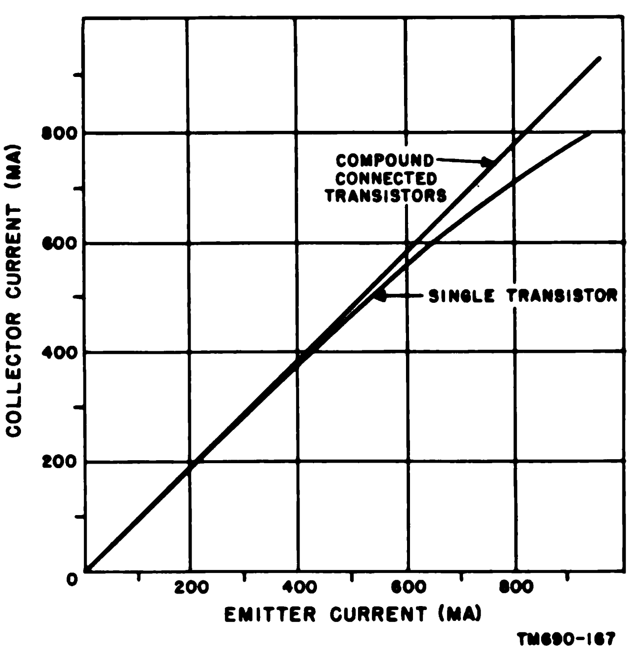


Figure 132. Variation of collector current with emitter current for a single transistor and for compound connected transistors.

emitter of the other transistor ($Q2$) and the two collectors are connected together. In the circuit the transistors are connected in the CB configuration. The computations (b below) will show that the α_{fb} for the compound-connected transistors (ratio of total collector current to input current (I_E)) is greater than that of a single transistor.

b. Assume that the short circuit forward current transfer ratio of each transistor is equal to 0.95; that is, α_{fb} (transistor $Q1$) or α_{fb} (transistor $Q2$) is equal to 0.95. The input current to transistor $Q1$ is designated I_E.

(1) Since in the CB configuration the collector current is equal to the emitter current multiplied by the short circuit forward current transfer amplification factor, the collector current (I_{c1}) of transistor $Q1$ is as follows:

$$I_{c1}=\alpha_{fb}I_E=0.95\ I_E$$

(2) The base current (I_{b1}) of transistor $Q1$ equals the emitter current less the collector current—

$$I_{b1}=I_E-0.95\ I_E=0.05\ I_E$$

(3) The base current of transistor $Q1$ is the emitter input current to transistor $Q2$. Transistor $Q2$ collector current then is as follows:

$$I_{c2}=(0.95)\ (0.05I_E)=0.0475\ I_E$$

Note. The base current of transistor $Q2$ equals the emitter current (0.05 I_E) less the collector current (0.0475 I_E). Its value is 0.0025 I_E. The value of base current of transistor $Q2$ is not required in the computation of α_{fb} in the compound connection of the transistors.

(4) The total output current (I_o) at point B is equal to the sum of the transistor $Q1$ collector current ((1) above) and the transistor $Q2$ collector current ((3) above) thus—

$$I_o=I_{c1}+I_{c2}$$

$$=0.95\ I_E+0.0475\ I_E$$

$$=0.9975\ I_E$$

c. The two compound-connected transistors can be considered as a single unit having an emitter, collector, and base at points A, B, and C respectively.

(1) The ratio of output current (0.9975 I_E) to input current (I_E) is 0.9975. This is the short circuit forward current amplification factor for the combination and represents a very high α_{fb}, actually an increase from a single unit having an α_{fb} of 0.95 to a compound unit having an α_{fb} of 0.9975.

(2) The effectiveness of the compound connection is better indicated by considering the short circuit forward current amplification factor in the CE configuration (α_{fe}). By using the conversion formulas listed in paragraph 66, it can be shown that a transistor having an α_{fb} of 0.95 has an α_{fe} of 19; a transistor (or compound connection) having an α_{fb} of 0.9975 ((1) above) has an α_{fe} of 399.

d. Compound-connected transistors in a circuit of any configuration can be considered, therefore, as a single unit with a high current amplification factor. To simplify the computations in the example (b and c above) each transistor selected has an α_{fb} of 0.95. Actually the advantages of the compound connection are not restricted to transis-

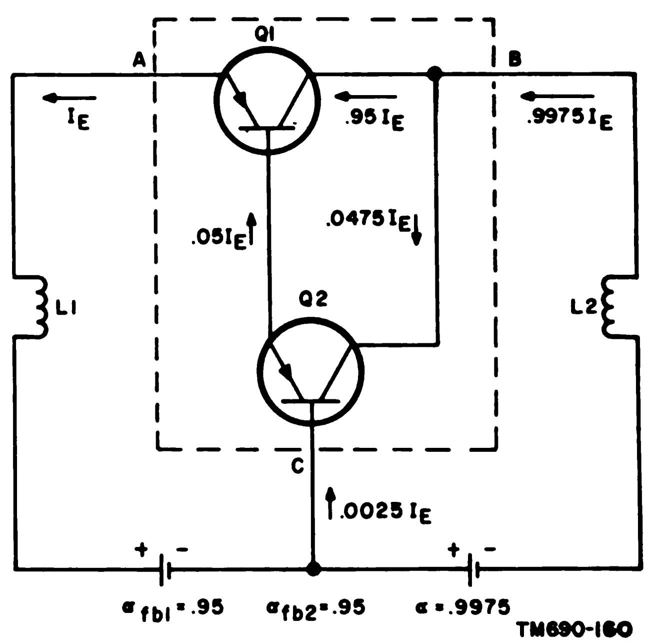


Figure 133. Current flow in compound-connected transistors.

tors having equal-valued current amplification factors.

139. Compound Connection in Complementary Symmetry

Compound-connected transistors (par. 138) can be used in single ended amplifiers, conventional push-pull amplifiers, or in complementary symmetry amplifiers. An example of the latter is shown in figure 134. This circuit is similar to the complementary symmetry circuit shown in figure 129 and discussed in paragraph 136. The junction of the correspondingly referenced parts is the same. Transistor Q1 (fig. 129) is replaced by the compound connection of transistors Q1A and Q1B (fig. 134); transistor Q2 (fig. 129) is replaced by the compound connection of transistors Q2A and Q2B (fig. 134). The gain of the compound-connected complementary symmetry circuit is greater than that of the complementary symmetry circuit.

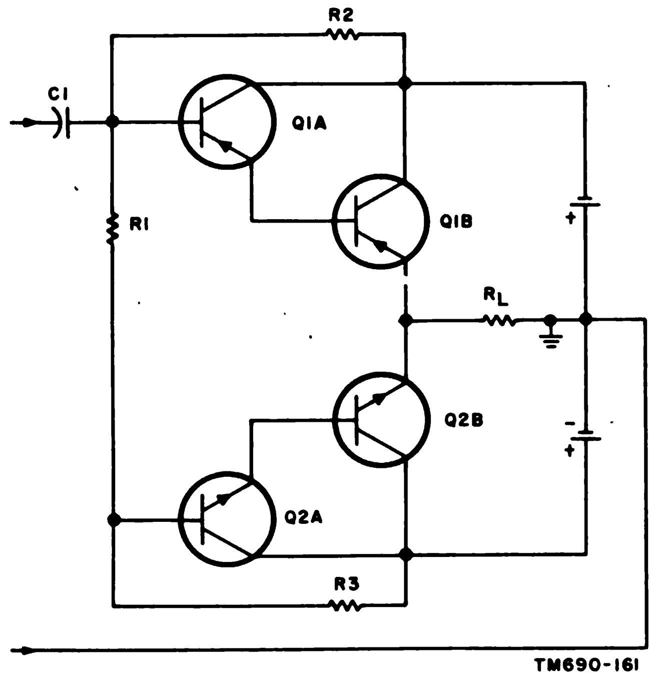


Figure 134. Compound-connected transistors in complementary symmetry circuits.

Section VIII. BRIDGE CONNECTIONS

140. Basic Bridge Circuit

a. A basic bridge circuit arrangement is shown in A, figure 135. The arms of the bridge, W, X, Y, and Z, can be circuit elements such as resistors, inductors, capacitors, transistors, or batteries. The bridge is said to be balanced when, regardless of the voltage applied at points 1 and 2, the voltage drops across the bridge arms are such that zero voltage occurs across points 3 and 4; that is, no current flows through resistor R_L.

b. The arms of the basic bridge circuit can be replaced by two transistors ($Q1$ and $Q2$) and two batteries (V_{CC1} and V_{CC2}) as shown in B, figure 35. If the two transistors are biased to draw equal or no emitter-collector current under quiescent (no signal) conditions, the bridge is balanced and no current flows through resistor R_L. If an input signal causes either transistor to conduct more than the other, the bridge becomes unbalanced and current flows through resistor R_L. Sine wave-input signals fed to points A–B and C–B 180° out of phase will produce an amplified sine wave signal across resistor R_L. If the transistors are operated class A or class B, no dc current will flow through resistor R_L.

c. The advantage of the bridge connection is that no dc current passes through resistor R_L. The resistor can be replaced by a speaker voice coil. Dc current through a speaker voice coil offsets the speaker coil and causes distortion. If a conventional push-pull amplifier were used, the speaker voice coil would have to be center-tapped and only half of the coil would be used for each half cycle of the output signal. This would lead to inefficiency in converting electrical energy into sound energy.

141. Complementary Symmetry Bridge Circuit Arrangement

Figure 136 shows a bridge circuit arrangement in which the arms of the bridge are transistors. Two transistors ($Q1$ and $Q3$) are PNP types, and two transistors ($Q2$ and $Q4$) are NPN types. The advantage of this circuit is that no part of the input circuit or the output circuit need be placed at ground potential. Furthermore this circuit can be driven by a single-ended driver; that is, two signals 180° out of phase are not required.

a. Assume that under quiescent conditions all the transistors are zero biased and draw no emitter-collector current (class B operation). If the transistors draw no current, there is no completed circuit across battery V_{CC} and no dc current flows through resistor R_L.

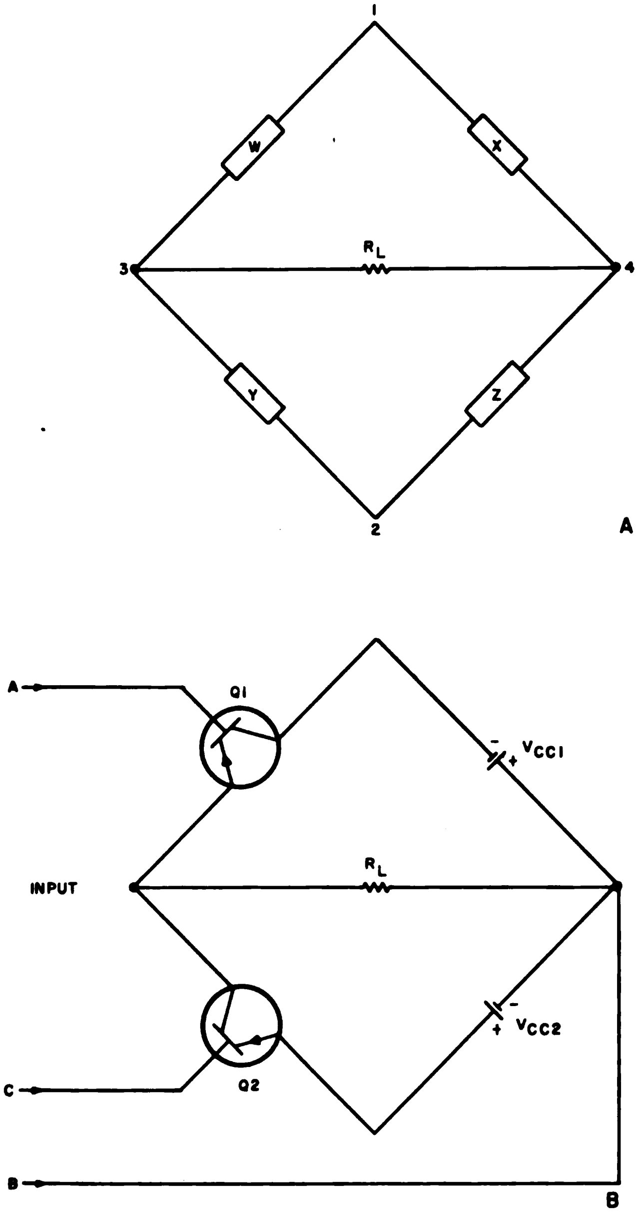


Figure 135. Basic bridge circuit arrangement, and bridge-connected transistors.

(1) Assume that an input signal causes point A to become negative with respect to point B. This will cause transistors $Q1$ and $Q4$ to become forward biased and transistors $Q3$ and $Q2$ to be reverse biased. Transistors $Q1$ and $Q4$ will conduct. The electrons will flow out of the negative terminal of the battery through the transistor $Q1$ collector-emitter, through resistor R_L, the transistor $Q4$ emitter-collector and back to the positive terminal of the battery. The path is indicated by the solid-line arrows. A voltage is developed across load resistor R_L, with the polarity indicated.

(2) If an input signal causes point A to become positive with respect to point B, then transistors $Q3$ and $Q2$ conduct while transistors $Q1$ and $Q4$ are nonconducting. The electron current path is indicated by the dashed-line arrows. A voltage is developed across load resistor R_L in the polarity indicated.

b. Assume that under quiescent conditions the transistors are biased to draw equal amounts of current (class A operation). (The circuits required to forward-bias the transistor are not pertinent to the discussion and are not shown.) Electrons emerge from the negative terminal of the battery. One-half of the electron current flows through the collector-emitter of transistor $Q1$ and through the emitter-collector of transistor $Q2$ into the positive terminal of the battery. The other half of the electron current flows through the collector-emitter of transistor $Q3$ and the emitter-collector of transistor $Q4$ back to the positive terminal of the battery. The bridge is therefore balanced and no current flows through resistor R_L.

(1) Assume that the input signal causes point A to become more negative with respect to point B. Transistors $Q1$ and $Q4$ become more forward biased and

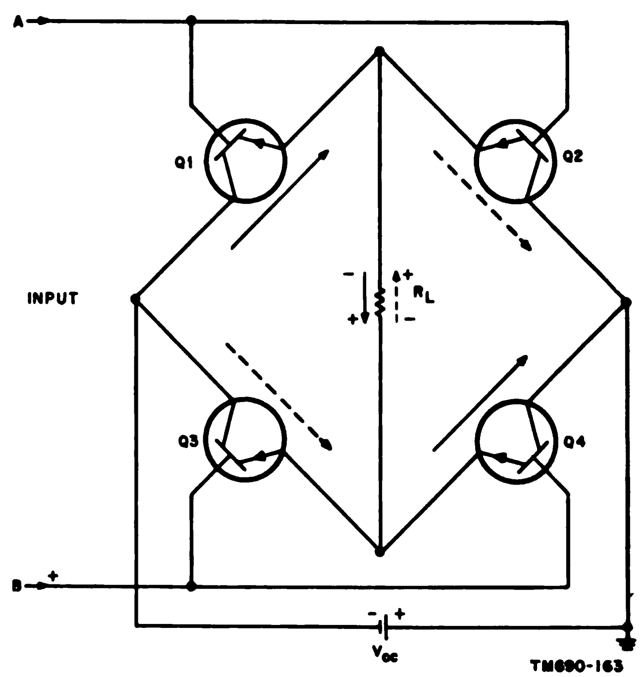


Figure 136. Complementary symmetry bridge circuit arrangement.

draw more collector current. Transistors $Q2$ and $Q3$ become less forward biased and draw less collector current. The bridge becomes unbalanced. The difference in current between transistors $Q1$ and $Q2$ flows through resistor R_L in the direction of the solid-line arrow and through transistor $Q4$ into the positive terminal of the battery. A voltage is developed across resistor R_L with the polarity indicated.

(2) If the input signal causes point A to become positive with respect to point B, transistors $Q2$ and $Q3$ become more forward biased and draw more collector current. Transistors $Q1$ and $Q4$ become less forward biased and draw less collector current. The difference in current between transistors $Q3$ and $Q4$ flows through resistor R_L in the direction of the dashed-line arrow and develops a voltage with the indicated polarity.

c. As in all cases of class A and class B amplifiers, bridge-connected class A amplifiers, though less efficient than bridge-connected class B amplifiers with respect to power consumption, afford less signal distortion.

142. Bridge Circuit Connection With Four PNP Transistors

a. Figure 137 shows a four transistor bridge circuit using four PNP transistors. The following discussion would apply if four NPN transistors were used, except that all voltage polarities indicated would have to be reversed.

b. Basically the operation of this transistor bridge circuit is similar to the operation of the complementary symmetry bridge-connected circuit (fig. 136) discussed in paragraph 141.

(1) Under quiescent conditions, and class B operation, none of the transistors conduct and no dc current flows through load resistor R_L (fig. 137). For class A operation, at the quiescent point all the transistors draw an equal amount of current, the bridge is balanced and no current flows through load resistor R_L.

(2) Assume that an incoming signal causes the bases of transistors $Q1$ and $Q4$ to go negative (forward bias) and the bases of transistors $Q3$ and $Q2$ to go positive (reverse bias). Transistors $Q1$ and $Q4$ conduct (class B) or conduct more heavily (class A) with electron current flow through R_L in the direction of the solid-line arrow. The polarity of the voltage developed across resistor R_L is in the direction indicated. Transistors $Q3$ and $Q2$ conduct less current (class A) or remain cut off (class B).

(3) If the incoming signal reverses in polarity ((2) above) then transistors $Q3$ and $Q2$ conduct (class B) or conduct more heavily (class A). The direction of current flow through resistor R_L is in the direction of the dashed-line arrow. The polarity of the voltage developed across resistor R_L is in the direction indicated. Transistors $Q1$ and $Q4$ conduct less current (class A) or remain cut off (class B).

(4) The main disadvantage of this circuit is the requirement for a transformer ($T1$) with five taps on the secondary winding. This is required because transistors $Q1$ and $Q3$ are *CC* configurations (output taken from emitter) with high input impedances requiring large voltage input signals. Transistors $Q2$ and $Q4$ are *CE* configurations (output taken from collector) with low input impedances requiring low input voltage signals for emitter-collector current to equal that of the *CC* configuration. This problem did not occur in the complementary symmetry bridge connected transistors (par. 141) because each transistor was operated in the *CC* configuration.

c. To overcome the need for a transformer with five taps on the secondary to feed a bridge connection of four similar transistors, the circuit of figure 138 can be used. This circuit is similar to that discussed in *b* above; only the method of driving the four transistors is different. The input signal is connected to the bases of transistors $Q1$ and $Q3$ only. A portion of the output of transistor $Q1$ is coupled to the base of transistor $Q4$ through current limiting resistor $R2$. A portion of the output of transistor $Q3$ is coupled to the base of transistor $Q2$ through current limiting resistor $R1$.

(1) The quiescent conditions for this circuit are the same as those for the circuit discussed in *b*(1) above.

(2) Assume that an incoming signal causes point A to go negative with respect to

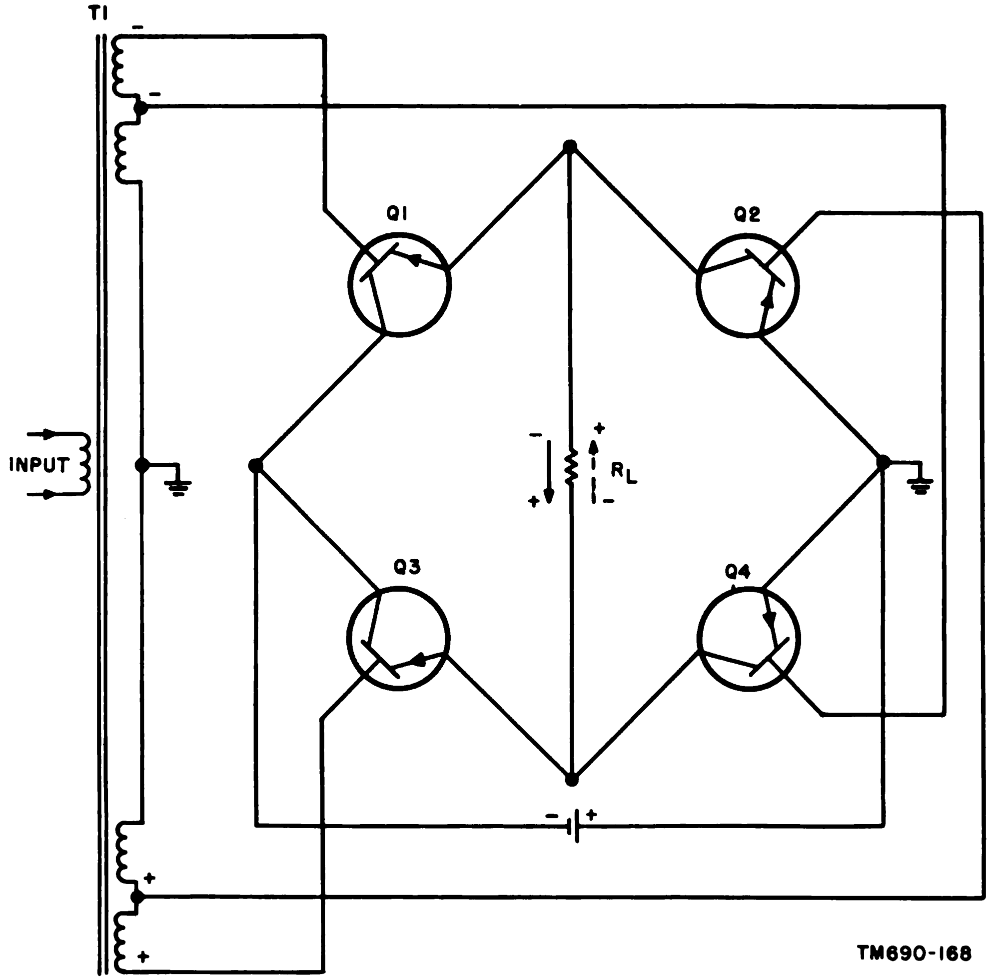


Figure 137. Bridge connection of four PNP transistors using separate input signals from transformer to each transistor.

ground and point B to go positive with respect to ground. Transistor $Q1$ conducts (class B) or conducts more heavily (class A). A portion of the output of transistor $Q1$ is fed to the base of transistor $Q4$ so that it conducts (class B) or conducts more heavily (class A). The resultant increase in current through transistors $Q1$ and $Q4$ flows through load resistor R_L in the direction of the solid-line arrow, producing the voltage polarity indicated. Transistors $Q3$ and $Q4$ remain nonconducting (class B), or conduct less.

(3) If the incoming signal causes point B to go negative with respect to ground and point A to go positive, transistors $Q3$ and $Q2$ then conduct (class B) or conduct more heavily (class A). The roles of conducting and nonconducting transistors are interchanged. Current flows through resistor R_L in the direction of the dashed-line arrow, developing a voltage with the indicated polarity.

143. Advantages of Four-Transistor Bridge Amplifier

The advantages of four-transistor bridge amplifiers (pars. 141 and 142) are as follows:

a. Dc current does not pass through the load. This is important when direct coupling to a speaker voice coil. Furthermore the speaker voice coil need not be center tapped.

b. The peak-to-peak output voltage is equal to *twice* the supply voltage less the small voltage drops in the conducting transistors. The maximum voltage experienced by each transistor is always less than the supply voltage. In transformer coupled output stages, this maximum voltage may be *twice* the supply voltage.

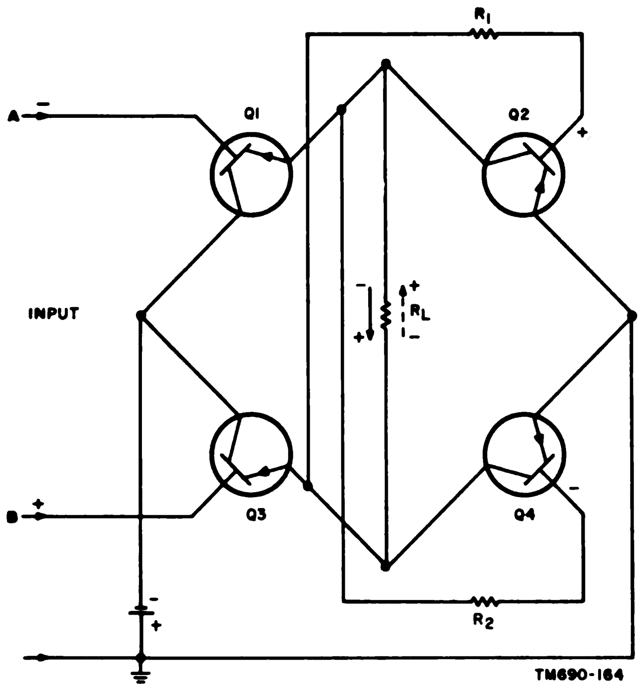


Figure 138. Bridge connection of four PNP transistors using portions of output of one transistor to feed its mating transistor.

c. A single battery source is required and it need not be tapped.

d. The power that must be dissipated by each transistor is half of that dissipated by each transistor in a conventional push-pull amplifier. This means that the four transistor bridge amplifiers can handle twice the power of a conventional push-pull amplifier without overheating the transistors.

144. Summary

a. The noise factor of an amplifier is the quotient of the signal-to-noise ratio at the output of the amplifier divided by the signal-to-noise ratio at the input of the amplifier.

b. The higher the collector voltage, the higher the noise factor of a transistor.

c. A degenerated CE amplifier (fig. 108) has a high input impedance.

d. Interstage coupling may use transformer, RC, impedance, or direct coupling (fig. 133).

e. Volume controls used in transistor amplifiers are considered current dividers (fig. 134).

f. The volume control should be so placed in a circuit that very low, or no dc current passes through the control. Variation of the control must not vary the frequency response of the amplifier.

g. Conventional push-pull amplifiers require a center-tapped secondary winding input transformer or a phase inverter as a driver.

h. A two-transistor phase inverter (figs. 119 and 120) compared to a one-transistor phase inverter (figs. 117 and 118) can deliver more power to the input circuit of the output stage.

i. Zero-biased, class B, push-pull amplifiers (fig. 121) produce crossover distortion (fig. 122). The distortion can be eliminated by placing a small forward bias on the input circuit (fig. 123).

j. If capacitance coupling to the input circuit of a class B amplifier is used, a discharge diode (fig. 125) must be placed in the input circuit to prevent reverse biasing of the amplifier.

k. Complementary symmetry (fig. 127) eliminates the need for a center-tapped secondary input transformer, a *phase-inverter* driver, and discharge diodes (*j* above). No dc current flows through the load circuit.

l. Compound-connected transistors (fig. 133) produce a high current amplification factor.

m. Bridge-connected transistors with the bridge using two or four transistors (figs. 135, 136, and 137) have no dc current flow through the load resistor.

n. The peak-to-peak output signal voltage of bridge amplifiers using four transistors approaches twice the supply voltage.

CHAPTER 8

TUNED AMPLIFIERS

Section I. COUPLING NETWORKS

145. General

a. The amplifiers treated in this chapter, are *narrow-band* amplifiers; that is, the bandwidth of frequencies amplified is a small percentage of the center frequency. *Wide-band* amplifiers usually amplify signals ranging from several cycles per second to several megacycles. These amplifiers are covered in chapter 9.

b. Tuned amplifiers amplify a selected bandwidth of frequencies and suppress unwanted frequencies. The selectivity of the amplifier is accomplished with parallel tuned interstage coupling networks. The properties of tuned amplifiers depend upon the characteristics of resonant circuits. Series resonant circuits are rarely used because they do not lend themselves to impedance transformation as readily as do parallel resonant circuits. The important properties of parallel resonant circuits are listed in paragraph 146. The detailed theory of resonant circuits is covered in TM 11–681.

c. Because the transistor amplifier basically is a *power* amplifier, the interstage coupling network must be designed to transfer maximum power from the output of one stage to the input of the following stage. This condition requires that minimum power loss be introduced by the coupling network, and that the coupling network match the output of one stage (usually high resistance) to the input of the following stage (usually low resistance).

146. Resonant Circuits

The important properties of parallel resonant circuits (fig. 139) are as follows:

a. The frequency of resonance (f_r) is:

$$f_r=\frac{1}{2\pi\sqrt{LC}}$$

b. For any given resonant frequency, the product of L and C is a constant (A, fig. 139). At low frequencies the product of L and C is large, and at high frequencies, the product of L and C is small.

c. For practical purposes the frequency at resonance may be considered independent of the resistance of the circuit.

d. The Q of a parallel resonant circuit is the ratio of the current in the tank (I_L or I_C) to the current in the line (I):

$$Q=\frac{I_C}{I} \text{ or,}$$

$$=\frac{X_L}{R} \text{ or,}$$

$$=\frac{Z}{X_c}$$

Z is total impedance of a parallel resonant circuit (tank).

e. Q is a measure of the selectivity of a circuit and varies inversely with the resistance in the circuit. The lower the resistance, the higher the Q and the greater the difference between voltages at the resonant frequency and the voltages at frequencies off resonance.

f. The effective limits on the band pass are points on the resonance curve that are 0.707 down from the maximum. Under these conditions, the band pass (A_f) is equal to the ratio of the center (resonant) frequency divided by the loaded (effective) Q of the circuit:

$$A_f=\frac{f_r}{Q}$$

g. The resonant frequency and the unloaded Q of a parallel tuned circuit (B, fig. 139) remain the same whether the signal is injected across points 1 and 3 or points 2 and 3. The input impedance

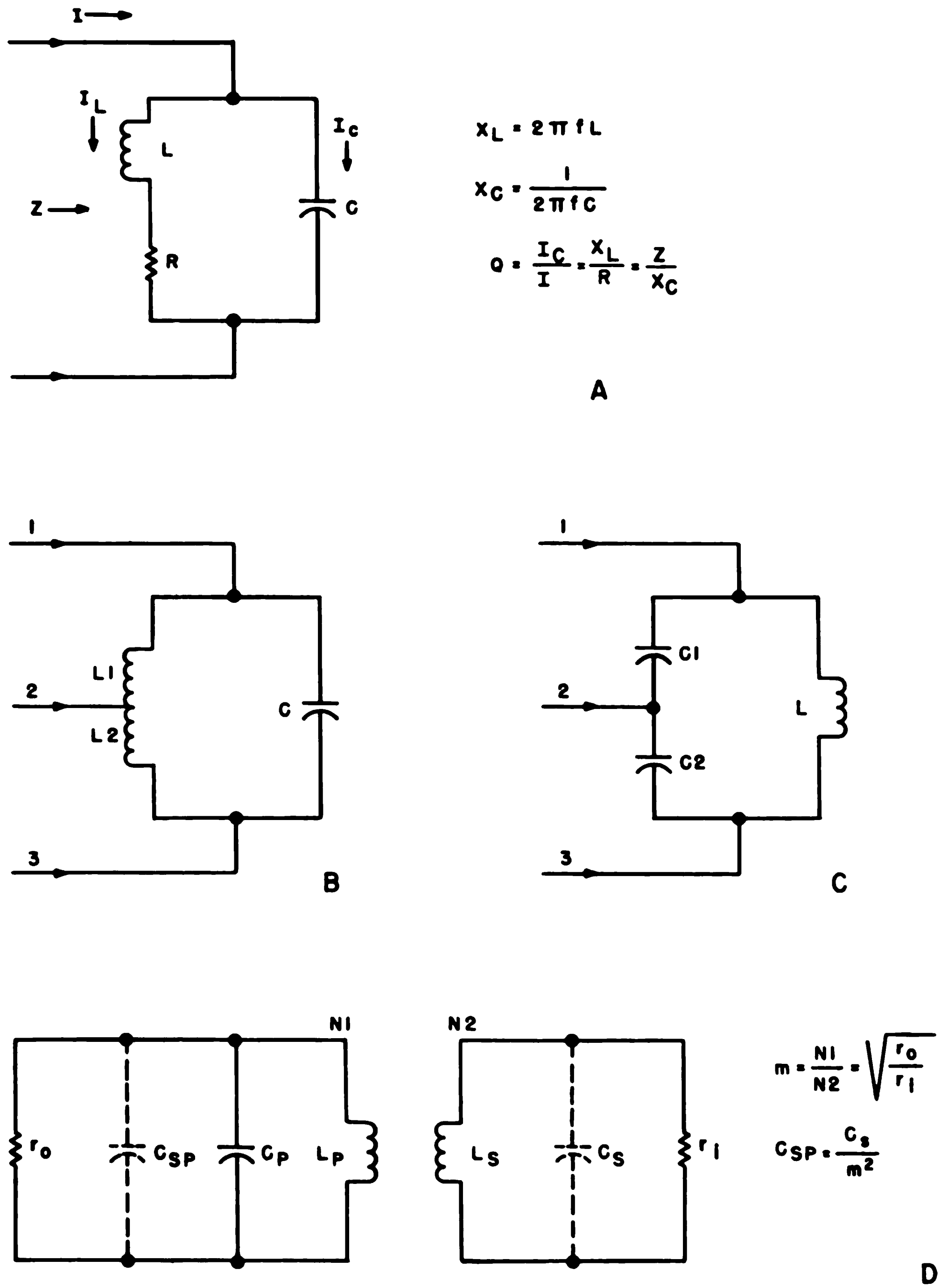


Figure 139. Parallel tuned circuits.

between points 2 and 3, however, is much smaller than the input impedance between points 1 and 3, depending on the square of the turns ratio between points 1 and 3, and points 2 and 3. For instance, if point 2 is at the center tap, the input impedance between points 1 and 3 is four times the input impedance between points 2 and 3.

h. The resonant frequency and the unloaded Q of a tuned circuit (C, fig. 139) remain the same whether the signal is injected across points 1 and 3 or points 2 and 3. The input impedance between points 2 and 3, however, is smaller than the input impedance between points 1 and 3.

i. In many applications, a tuned resonant circuit (capacitor C_p and inductance L_p) in the primary of a transformer (D, fig. 139) is coupled to the nonresonant secondary of the transformer. In this case, if N_1 represents the number of turns in the primary windings and N_2 represents the number of turns in the secondary winding, then the turns ratio (m) of primary to secondary under matched conditions is:

$$m=\frac{N_1}{N_2}=\sqrt{\frac{r_o}{r_i}}$$

If there is capacitance in the secondary circuit (C_s), it is reflected (or referred) to the primary circuit by transformer action as a capacitance (C_{sp}). Its value is:

$$C_{sp}=\frac{C_s}{m^2}$$

147. Transistor and Coupling Network Impedances

a. The output impedance of a transistor can be considered as consisting of a resistance (r_o) in parallel with a capacitance (C_o) (A, fig. 140). The input impedance of a transistor also can be considered as consisting of a resistance (r_i) in parallel with a capacitance (c_i).

b. Normally output capacitance C_o and input capacitance C_i are accounted for by considering them part of the coupling network (B, fig. 140). Assume that the required capacitance between terminals 1 and 2 of the coupling network is calculated to be 500 $\mu\mu f$. Assume also that capacitance C_o is 10 $\mu\mu f$. A capacitor of 490 $\mu\mu f$ then would be used between terminals 1 and 2 so that the total capacitance would be 500 $\mu\mu f$. The same technique is used to compensate for capacitance C_i at terminals 3 and 4.

c. For maximum transfer of power from transistor $Q1$ to transistor $Q2$, the input impedance (terminals 1 and 2) to the coupling network must equal resistance r_o; the output impedance of the coupling network (*looking* into terminals 3 and 4), must equal r_i.

148. Transformer Coupling, Tuned Primary

a. General. A single-tuned circuit used as a coupling network, and using inductive coupling is shown in A, figure 141.

(1) In this circuit, capacitance C_T includes transistor $Q1$ output capacitance (par.

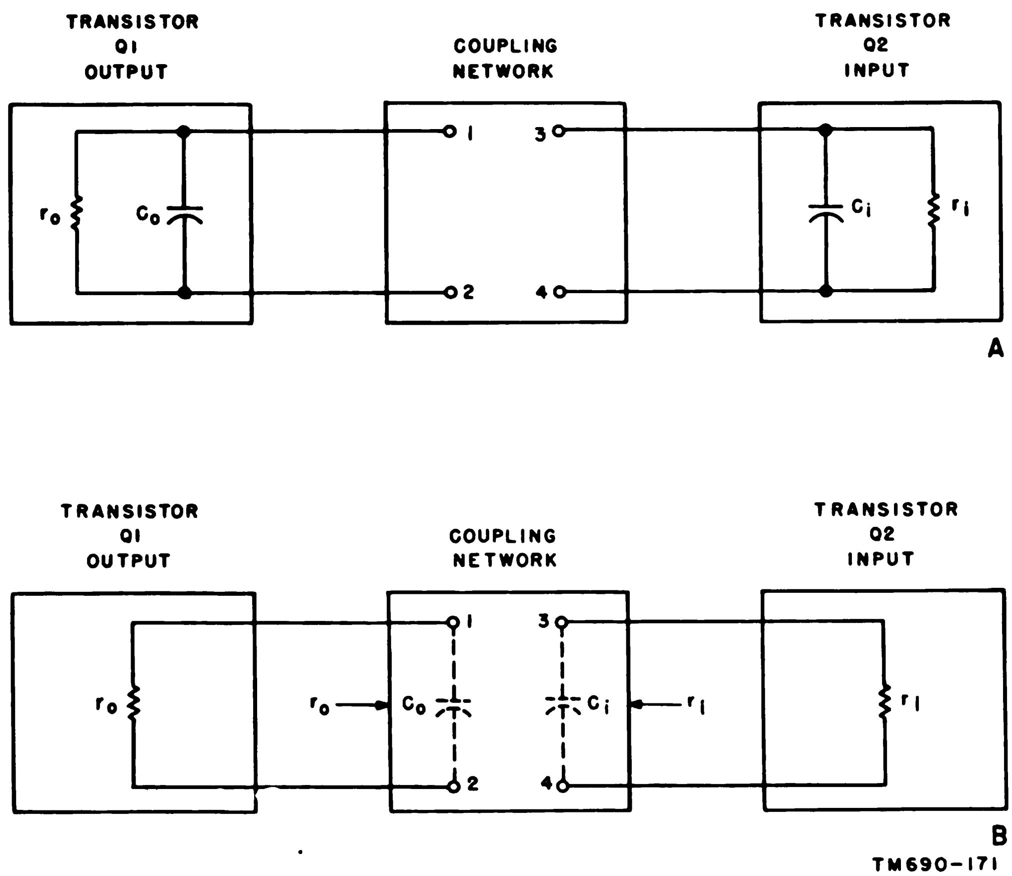


Figure 140. Equivalent output and input circuits of transistors connected by a coupling network.

147), and transistor $Q2$ input capacitance (referred to the primary of transformer $T1$). The output of transistor $Q1$ is matched to the input of transistor $Q2$ by selecting the proper turns ratio of primary and secondary windings for transformer $T1$.

(2) If L_P represents the inductive reactance between terminals 1 and 2 of transformer $T1$, then the resonant frequency (f_r) for this circuit is:

$$f_r = \frac{1}{2\pi\sqrt{L_P C_T}}$$

b. Application. Assume that it is desired to use the coupling network of A, figure 141, in an intermediate frequency amplifier using two common-emitter configurations.

(1) The following data is available:
Resonant frequency $f_r = 500$ kc.
Frequency bandwidth $\Delta f = 10$ kc.
Transistor $Q1$ output resistance $r_o =$ 12,000 ohms.
Transistor $Q1$ output capacitance $C_o =$ 10 $\mu\mu f$.
Transistor $Q2$ input resistance $r_i = 700$ ohms.
Transistor $Q2$ input capacitance $C_i = 170$ $\mu\mu f$.
Coupling network loss = 3 db.
Power response of frequencies at extremes of bandwidth $(f_r \pm \frac{1}{2} \Delta f) = \frac{1}{4}$ of response at resonant frequency (f_r).

(2) With the data and requirements listed in (1) above, calculations (beyond the scope of this manual) show that the unloaded Q of the tuned circuit (before connecting in amplifier) must be 288. The primary inductance (L_P) must be 7.7 μh. The secondary inductance (L_s) must be 0.45 μh. Capacitor C_T must be 13,100 $\mu\mu f$; this includes C_o of 10 $\mu\mu f$, and the effective C_i referred to the primary which is 10 $\mu\mu f$. To construct a transformer with such a high Q (288) and a very low inductance ($L_P = 7.7$ μh) is very difficult. In practice, the problem is solved by using a tapped primary winding (B, fig. 141). The inductance of the primary can be made many times the calculated inductance. For instance, the inductance ($L_{1\text{-}3}$) between terminals 1 and 3 (B, fig. 141) can be made 100 times the calculated inductance (L_P) for terminals 1 and 2 (A, fig. 141)—

$$L_{1\text{-}3} = 100\ L_P$$
$$= 100 \times 7.7$$
$$= 770\ \mu h$$

To maintain the same resonant frequency, the capacitance across terminals 1 and 3 ($C_{1\text{-}3}$) must be 100 times less than the calculated capacitance (C_T):

$$C_{1\text{-}3} = \frac{C_T}{100}$$
$$= \frac{13{,}100}{100}$$
$$= 131\ \mu\mu f$$

To maintain the impedance matching for maximum transfer of energy, the inductance between terminals 2 and 3 of transformer $T2$ must be equal to the calculated inductance (L_P) between terminals 1 and 2 of transformer $T1$ which in this case is 7.7 μh.

149. Autotransformer Coupling, Tuned Primary.

a. Inductive coupling from the output of one transistor to the input of a second transistor can

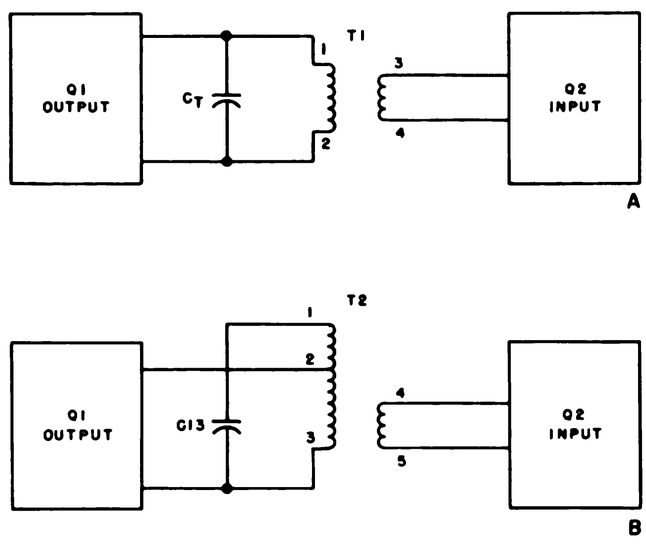


Figure 141. Interstage network using transformer coupling with tuned primary winding.

be accomplished with an autotransformer (A, fig. 142). The analysis of the operation of this circuit is the same as that for the circuit using a transformer with separate primary and secondary windings (par. 148). Capacitance C_T includes transistor $Q1$ output capacitance and transistor $Q2$ input capacitance referred to the primary. If L_{1-3} designates the inductance from terminal 1 to terminal 3 of transformer $T1$, then the resonant frequency (f_r) of the circuit is:

$$f_r=\frac{1}{2\pi\sqrt{L_{1-3}C_T}}$$

The tap at terminal 2 is positioned to provide impedance matching between the transistors.

b. If, in some applications, the inductance between terminals 1 and 3 is too small to achieve the required frequency selectivity (Q_o), the total primary inductance can be increased many times by using the arrangement shown in B, figure 142. To maintain the same resonant frequency, C_{1-4} must be reduced by the same factor so that the product of L_{1-3} and C_T (A, fig. 142), equals the product of L_{1-4} and C_{1-4} (B, fig. 142). Impedance matching is maintained provided that inductances L_{2-4} and L_{3-4} of transformer $T2$ equal inductances L_{1-3} and L_{2-3} respectively of transformer $T1$.

150. Capacitance Coupling

a. In many applications, transformer coupling (pars. 148 and 149) is not desirable because of the small number of turns required in the secondary winding. Difficulty in obtaining unity coupling between primary and secondary is often encountered. This problem becomes particularly severe in very-high frequency amplifiers operated in the common-base configuration. Such amplifiers have an input impedance below 75 ohms. In such cases, capacitance coupling (A, fig. 143) may be used.

(1) Impedance matching of transistor $Q1$ output to transistor $Q2$ input is achieved by selecting the proper ratio of $C1$ to $C2$. Capacitance $C2$ is normally much larger than capacitance $C1$. Their reactance values bear the opposite relationship. The total capacitance (C_T) across inductance $L1$ is as follows:

$$C_T=\frac{(C1)(C2)}{C1+C2}$$

The resonant frequency is:

$$f_r=\frac{1}{2\pi\sqrt{L1C_T}}$$

In many applications (particularly in common-base configurations), capacitor $C2$ is not used because the low input resistance effectively shunts it out of the circuit. Then the resonant frequency becomes:

$$f_r=\frac{1}{2\pi\sqrt{L1C1}}$$

(2) If inductance $L1$ is too small to achieve the required selectivity (Q_o), the circuit shown in B, figure 143, can be used. The inductance can be increased by any factor, and the total capacitance (C_T) reduced by the same factor to maintain the same resonant frequency. To maintain the matched condition, the inductance between terminals 2 and 3 must be made equal to inductance $L1$, and the ratio of $C3$ to $C4$ must equal the ratio of $C1$ to $C2$.

b. Another arrangement using capacitance coupling is shown in A, figure 144. The theory of operation of this circuit is similar to that of the circuit discussed in *a* above.

(1) The total capacitance (C_T) in the circuit is:

$$C_T=C1+\frac{(C2)(C_t)}{C2+C_t}$$

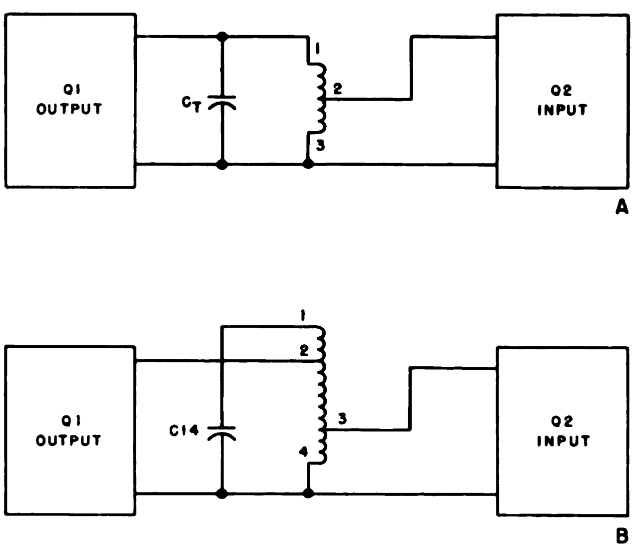


Figure 142. Interstage network using autotransformer coupling with tuned primary winding.

In very-high frequency common base amplifiers, C_i is effectively shunted by the low input resistance, and the total capacitance C_T is:

$$C_T = C1 + C2$$

In each case the resonant frequency is:

$$f_r = \frac{1}{2\pi\sqrt{L1C_T}}$$

(2) To achieve a high Q_o for selectivity, the total inductance can be increased (C_T decreased accordingly), so that the circuit arrangement appears as shown in B, figure 144.

151. Interstage Coupling, Double-Tuned Networks

a. General. The advantages of double-tuned, interstage coupling networks (fig. 145) over single-tuned networks (pars. 148, 149, and 150) are as follows:

(1) The frequency response is flatter within the pass band.

(2) The drop in response is sharper immediately adjacent to the ends of the pass band.

(3) Attenuation of frequencies not in the pass band is higher.

Note. For detailed theory on the response of double-tuned networks, refer to TM 11-681.

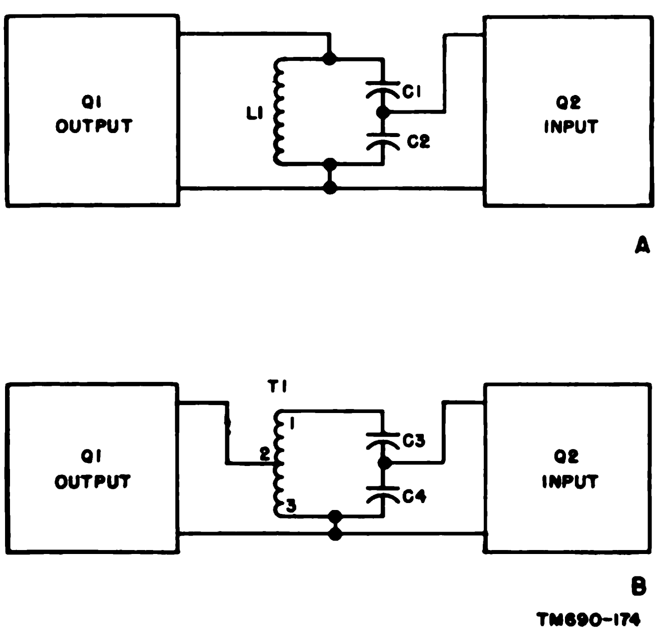


Figure 143. Capacitance coupling using a split capacitor.

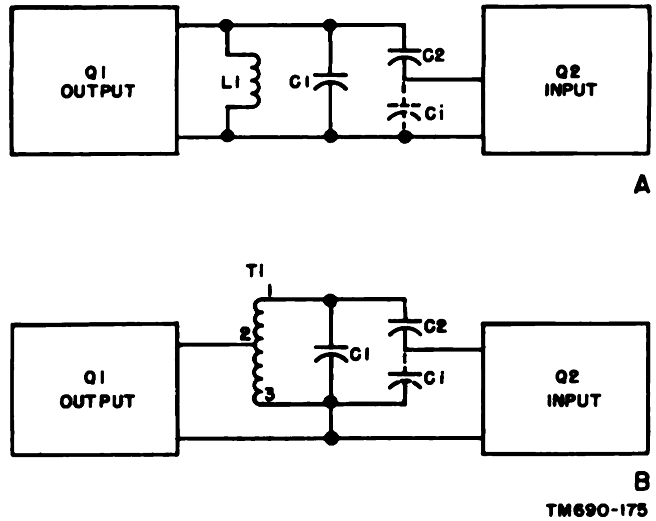


Figure 144. Capacitance coupling using two capacitors.

b. Inductive Coupling. The interstage coupling networks shown in A and B, figure 145, use two tuned circuits inductively coupled.

(1) In A, figure 145, capacitor $C1$ and the primary winding inductance L_p form a tuned circuit. Capacitor $C2$ and secondary winding inductance L_s also form a tuned circuit. Each circuit is tuned to the same resonant frequency (f_r) so that:

$$f_r = \frac{1}{2\pi\sqrt{L_p C1}} = \frac{1}{2\pi\sqrt{L_s\, C2}}$$

Impedance matching is obtained by properly selecting the turns ratio of the primary and the secondary windings.

(2) The circuit in B, figure 145, functions in the same manner as that discussed in (1) above. Tapped primary and secondary windings are used to facilitate the attainment of the desired selectivity (Q_o).

c. Capacitative Coupling. The interstage coupling networks shown in C and D, figure 145, use two capacitance-coupled tuned circuits.

(1) In C, figure 145, capacitor $C1$ and coil $L1$ form a resonant circuit. Capacitor $C3$ and coil $L2$ also form a resonant circuit. Each circuit is tuned to the same resonant frequency, so that:

$$f_r = \frac{1}{2\pi\sqrt{L1\,C1}} = \frac{1}{2\pi\sqrt{L2C3}}$$

Impedance matching is obtained by properly selecting the ratio of the reactance of capacitor $C2$ to the impedance

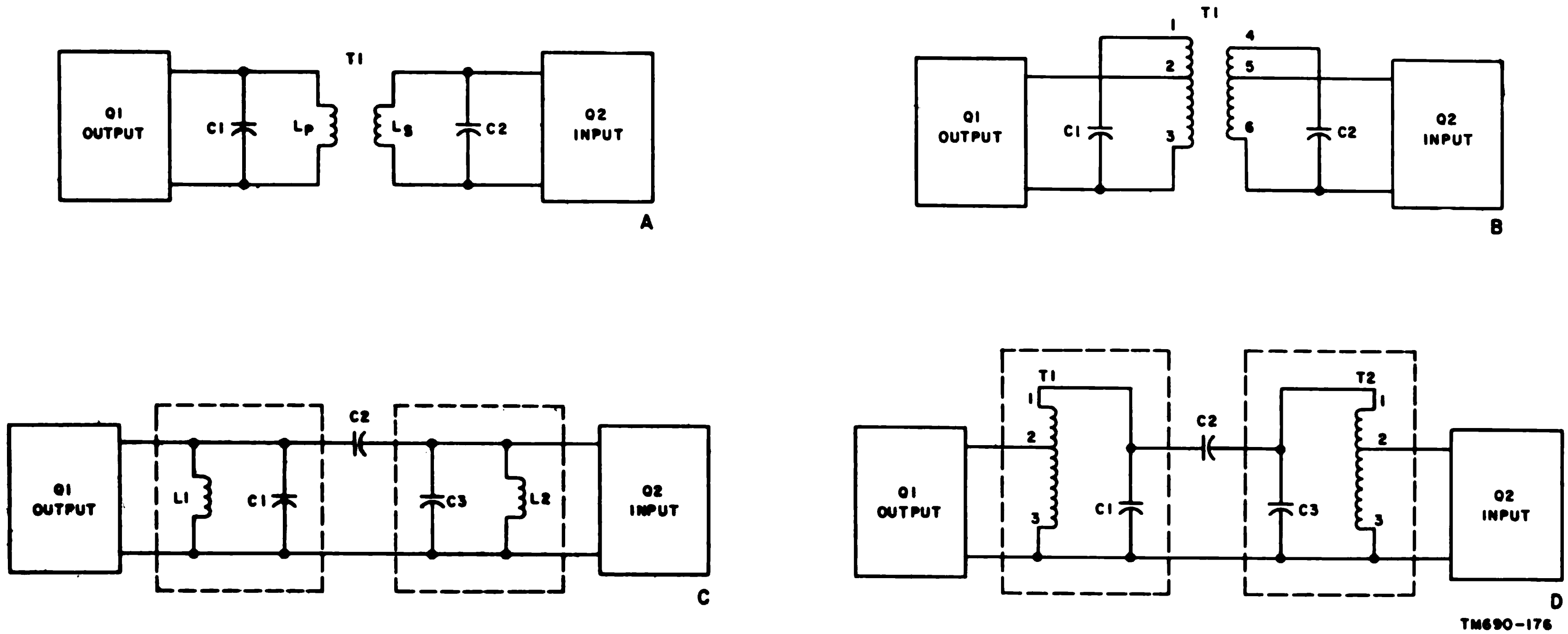


Figure 145. Double-tuned interstage coupling networks using inductive or capacitive coupling.

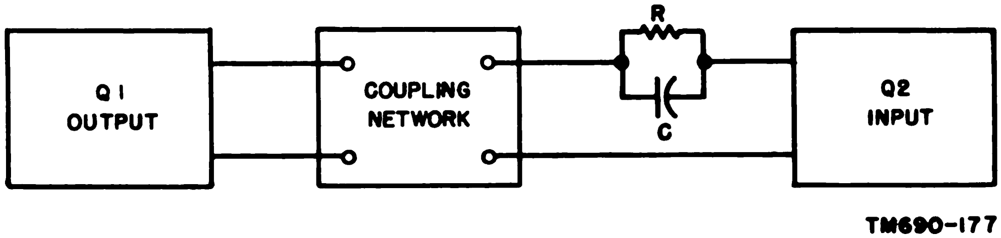


Figure 146. Coupled amplifiers with RC equalizer network.

of the input parallel circuit (capacitor $C3$ and coil $L2$).

(2) The circuit in D, figure 145, functions in the same manner as that discussed in (1) above. Tapped transformers (autotransformers) are used to facilitate the attainment of the desired selectivity (Q_o).

152. Gain Equalization

Many applications of tuned amplifiers require that the center frequency of the amplifier be varied over a wide range of frequencies. This is done by varying the capacitance or the inductance of the associated tuned circuits. Maintaining constant gain throughout a wide frequency range is difficult because the transistor current amplification factor (α_{fe}) drops in value at the higher frequencies. To compensate for the loss in gain at the higher frequencies, an equalizer network (fig. 146) resistor (R) and a capacitor (C) in parallel may be placed in series with the input of one of the amplifiers. This network attenuates the low frequencies more than it does the high frequencies. At the high frequencies, capacitor C effectively shunts resistor R. The low frequencies are attenuated by the resistor.

Section II. NEUTRALIZATION AND UNILATERALIZATION

153. General

a. A *unilateral* electrical device is one which transmits energy in one direction only. The transistor is not a unilateral device. In the derivation of the equivalent circuit of the transistor (ch. 4), it was shown that voltage variations in the output circuit cause voltage variations in the input circuit. Specifically, the internal voltage feedback to the input circuit from the output circuit was represented by a voltage generator placed in the equivalent circuit of the input circuit. In each transistor configuration, the feedback voltage aids the input voltage and is therefore a *positive* feedback. If the positive feedback voltage is large enough, the amplifier will oscillate. At low frequencies (audio frequencies), the voltage feedback is low; therefore special precautions to prevent oscillation are not usually required. High-frequency amplifiers with tuned coupling networks are susceptible to oscillations so that precautions must be taken (pars. 154–156).

b. The effect of positive or negative voltage feedback on the input circuit of an electrical device is to alter its input impedance. Normally, both the resistive and the reactive components of the input impedance are affected. The change in the input impedance of a transistor caused by the internal feedback, can be eliminated by using an *external* feedback circuit. If the external feedback circuit cancels both the resistive and the reactive changes in the input circuit, the transistor amplifier is considered to be *unilateralized*. If the external feedback circuit cancels only the reactive changes in the input circuit, the transistor amplifier is considered to be *neutralized*. Actually, neutralization is a special case of unilateralization. In either case, the external feedback circuit prevents oscillation in the amplifier.

154. Common-Base Amplifier, Unilateralized

a. A, figure 147, shows a tuned common-base amplifier. The dc biasing circuits are not shown. Transformer T1 couples the input signal to the amplifier. Capacitor C1 and the transformer T1 secondary winding form a parallel resonant circuit. Transformer T2 couples the output of the amplifier to the following stage. Capacitor C2 and the transformer T2 primary winding form a parallel resonant circuit.

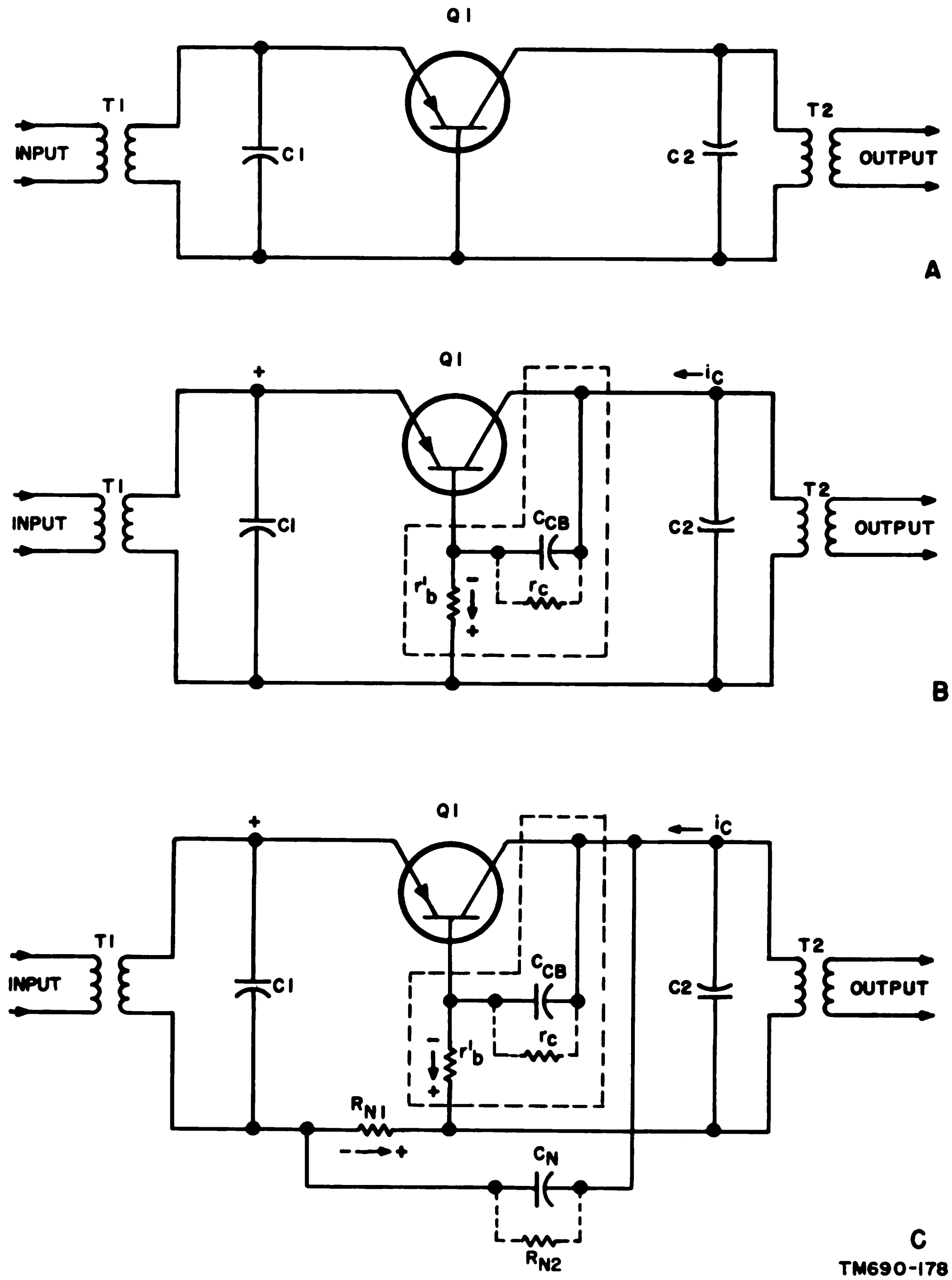


Figure 147. Common-base amplifier showing internal feedback elements and external unilateralizing circuit.

b. The *internal* elements of the transistor that may cause sufficient feedback and oscillation are shown in dashed lines (B, fig. 147).

(1) Resistor r'_b represents the resistance of the bulk material of the base and is referred to as the *base-spreading resistance.* Capacitor C_{CB} represents the capacitance of the base-collector junction. Resistor r_c represents the resistance of the base-collector junction. This resistance is very high in value because of the reverse bias on the base-collector junction. At very-high frequencies, capacitor C_{CB} effectively shunts resistor r_c.

(2) Assume that the incoming signal aids the forward bias (causes the emitter to go

more positive with respect to the base). Collector current i_c increases in the direction shown. A portion of the collector current passes through capacitor C_{CB} through resistor r'_b in the direction shown, producing a voltage with the indicated polarity. The voltage across resistor r'_b *aids* the incoming signal and therefore represents a positive feedback that may cause oscillation.

c. To overcome the possibility of oscillation, the external circuit (consisting of resistors R_{N1} and R_{N2} and capacitor C_N) can be added to the circuit (C, fig. 147).

(1) Resistors R_{N1} and R_{N2} and capacitor C_N correspond to resistors r'_b and r_c and capacitor C_{CB} respectively. The need for resistor R_{N2} depends on the frequency being amplified; the higher the frequency, the less the need for resistor R_{N2}.

(2) When the incoming signal aids the forward bias (*b*(2) above), the collector current increases. A portion of the collector current passes through capacitor C_{CB} through resistor r'_b, in the direction shown, producing a voltage with the indicated polarity. A portion of the collector current also passes through capacitor C_N through resistor R_{N1}, producing a voltage with the indicated polarity. The voltages produced across resistors r'_b and R_{N1} are opposing voltages. If the voltages are equal, no positive or negative feedback from output circuit to input circuit occurs. The amplifier is considered to be unilateralized.

155. Common-Emitter Amplifier, Partial Emitter Degeneration

a. A common-emitter amplifier using partial emitter degeneration to unilateralize the amplifier is shown in figure 148. Capacitor C_N and resistors R_{N1} and R_{N2} form a unilateralizing network. Transformer *T*1 couples the input signal to the base-emitter circuit. Resistor *R*1 forward biases the base-emitter circuit. Capacitor *C*1 prevents shorting of base bias voltage by the transformer *T*1 secondary. Transformer *T*2 couples the output signal to the following stage. Capacitor *C*2 and the transformer *T*2 primary winding form a parallel resonant circuit. Capacitor *C*3 blocks battery dc voltage from resistor R_{N1} and couples a portion of the collector current (i_{c2}) to the emitter. Inductor *L*1, an rf choke, prevents ac shorting to ground of resistor R_{N1} through capacitor *C*3.

b. When the input signal aids the forward bias, collector current i_c increases in the direction shown. Internally, a portion of the collector current is coupled to the base-spreading resistance through the collector-base junction capacitance (par. 154*b*(1)). The voltage developed across the base-spreading resistance aids the incoming signal and constitutes a positive feedback. To offset the positive feedback, a portion of the collector current (i_{c1}) is directed through resistor R_{N1} and the parallel combination of capacitor C_N and resistor R_{N2}. The voltage developed across resistor R_{N1} is a degenerating voltage equal and

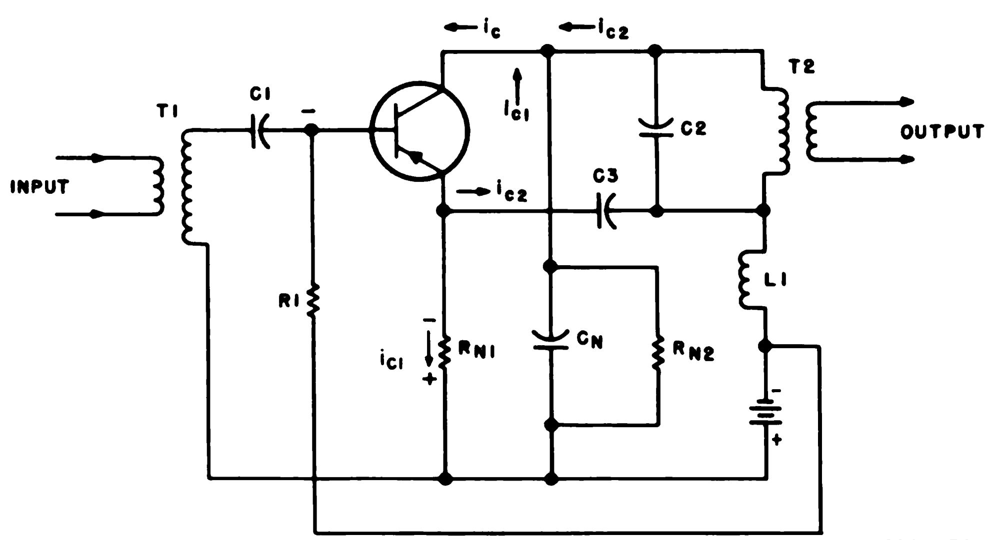


Figure 148. Common-emitter amplifier with partial emitter degeneration.

opposite to that developed across base-spreading resistance. The net voltage feedback to the input circuit is zero, and the amplifier is thus unilateralized. The values of capacitor C_N and resistors R_{N1} and R_{N2} depend on the internal values of the collector-base junction capacitance, the base-spreading resistance, and the collector resistance, respectively.

156. Common-Emitter Amplifier, Bridge Unilateralization

a. A, figure 149, shows a common-emitter amplifier that is unilateralized by using transformer $T2$ winding 2–3 and the network consisting of resistor R_N and capacitor C_N. Transformer $T1$ couples the input signal to the base-emitter circuit. Transformer $T2$ winding 1–2 couples the output signal to the following stage. Resistor $R1$ forward biases the transistor. Capacitor $C1$ prevents shorting of the base bias voltage by the transformer $T1$ secondary. Capacitor $C2$ bypasses the collector battery and places terminal 2 of transformer $T2$ at ac ground potential. Capacitor $C3$ tunes the primary of transformer $T2$.

b. The operation of this circuit can be demonstrated by showing the bridge arrangement formed by the transistor internal feedback elements and the external unilateralizing network (B, fig. 149).

(1) This bridge does not involve transformer $T1$, capacitors $C1$, $C2$, and $C3$, resistor

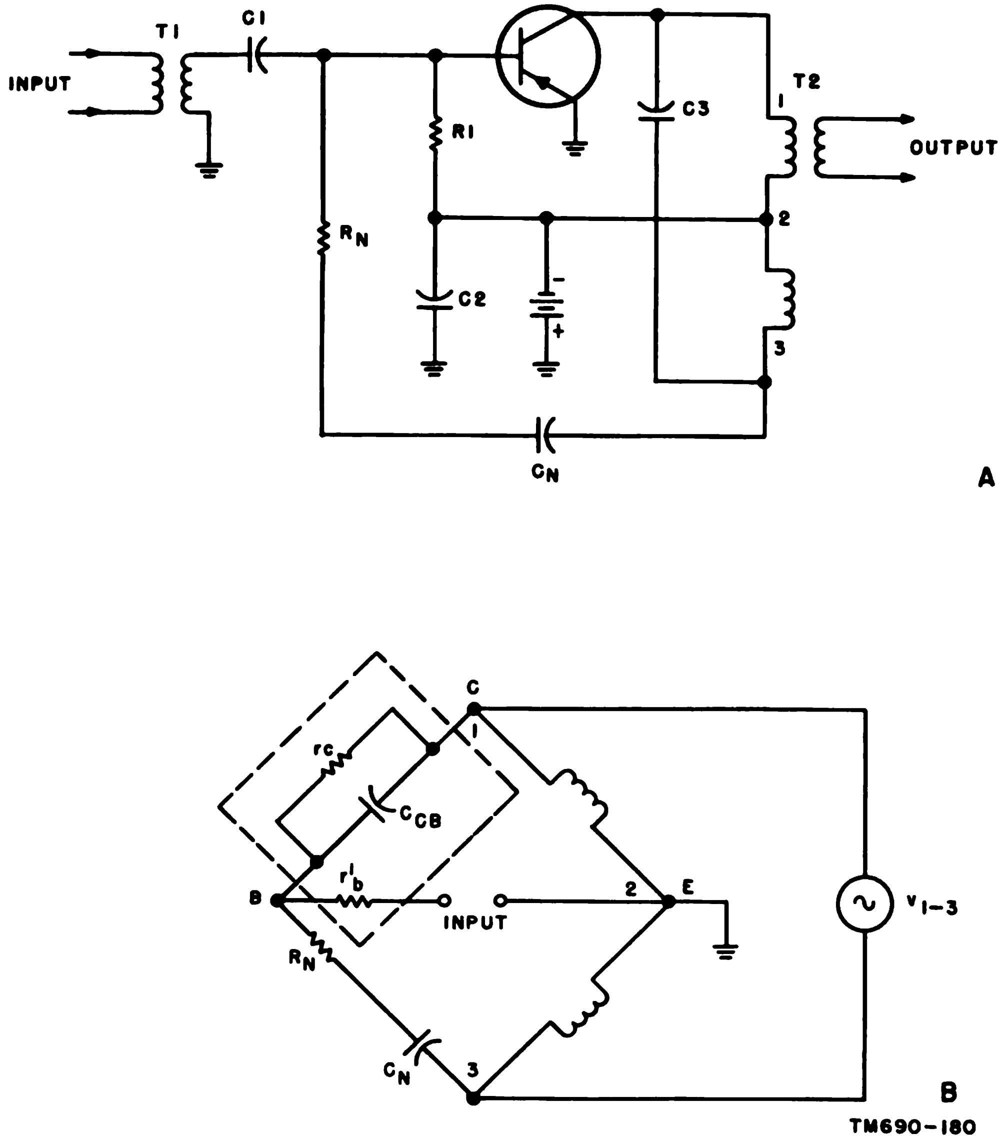


Figure 149. Common-emitter amplifier, using bridge arrangement to prevent oscillation.

$R1$, the transformer $T2$ secondary, and the collector battery.

(2) Points B, C, and E on the bridge represent the base, collector, and emitter terminals, respectively, of the transistor. The parts shown in dashed lines are the transistor internal feedback elements (par. 154). Points marked 1, 2, and 3 correspond to the terminals of the transformer $T2$ primary. The voltage developed across terminals 1 and 3 of transformer $T2$ is represented by a voltage generator with output voltage v_{1-3}.

(3) When the bridge is balanced, no part of voltage v_{1-3} appears between points B and E. The amplifier is considered to be unilateralized (par. 153). The bridge is balanced when the ratio of the voltages between points B and C and points B and 3 equals the ratio of the voltages between points C and E and points E and 3. In addition, the phase shift introduced by the network between points B and C must equal the phase shift introduced by the network between points B and 3.

(4) In many applications, use of capacitor C_N is sufficient to prevent oscillations in the amplifier; then resistor R_N is not used. Under these conditions, the amplifier is neutralized.

Section III. AUTOMATIC GAIN CONTROL

157. General

a. Tuned amplifiers are used extensively in radio and television receivers as rf and if amplifiers. In these applications, it is often desired to automatically vary the gain of the amplifier in accordance with the strength of the rf signal carrier received at the antenna. That is, lower gain is required for strong signals and higher gain is required for weak signals. One component of the output of a second detector of a receiver is a dc current and voltage that is directly proportional to the strength of the rf signal carrier received. It is this dc component that can be used to vary the gain of the tuned amplifier.

Note. For detailed theory on the operation of second detectors, refer to TM 11–662. Transistor detector circuits are discussed in chapter 12.

b. One method of varying the gain of a transistor amplifier is to vary the dc emitter current. The variation of power gain with dc emitter current of a common-base amplifier and a common-emitter amplifier is shown in A, figure 150. The curves shown are those of 455-kc amplifiers operated with a fixed collector voltage of 4 volts. Another method of varying the gain of a transistor amplifier is to vary the dc collector voltage. The variation of power gain with dc collector voltage of a common-base amplifier and a common-emitter amplifier is shown in B, figure 150. The curves shown are those of 455-kc amplifiers operated with a fixed emitter current of 1 ma. Circuits using these methods of automatic gain control (agc) are discussed in paragraphs 158 and 159.

158. Automatic Control of Emitter Current or Collector Voltage

a. Dc Emitter Current Control. The power gain of the common-emitter amplifier (A, fig. 151) is controlled by feeding the agc voltage to the base of the amplifier to vary the dc emitter current.

(1) In this case, the tuned amplifier is also operating as a dc amplifier to increase the dc current output of the second detector. If the dc current output of the second detector is sufficiently large, the dc emitter current of the tuned amplifier may be varied directly.

(2) In the circuit shown, resistors R1 and R4 form a voltage divider and establish the no-signal negative (forward) bias on the base. The agc voltage input from the second detector is *positive* with respect to ground and is fed to the base through dropping resistor R2. When the dc output of the second detector increases (due to a high carrier signal input to the detector), the positive dc voltage fed to the base of transistor Q1 through dropping resistor R2 *reduces* the net negative (forward) bias on the base and decreases the emitter current. The gain of the amplifier is also decreased (par. 157*b*). When the dc output of

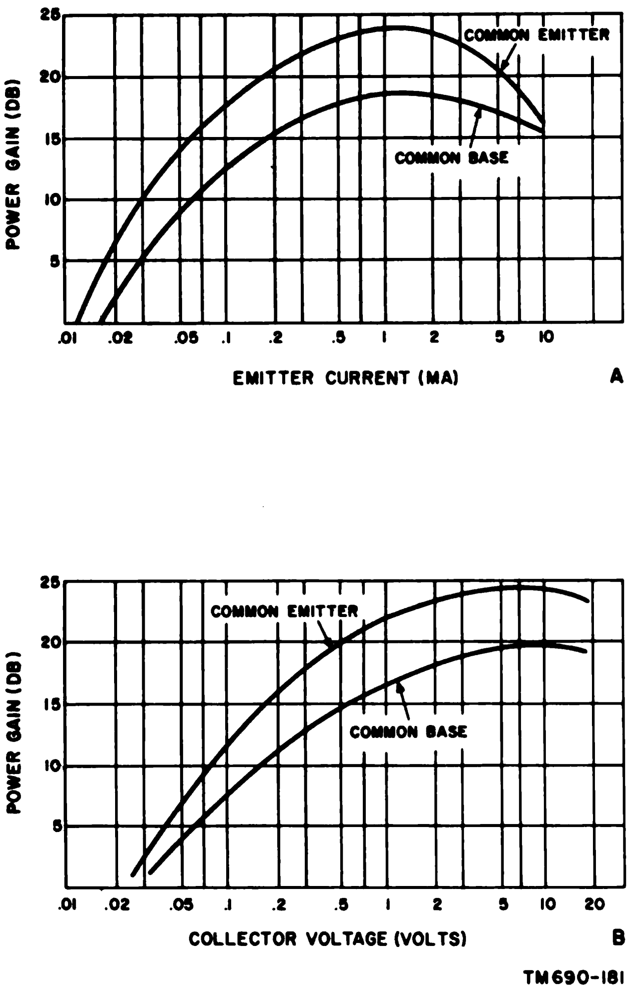


Figure 150. Variation of power gain with dc emitter current and dc collector voltage.

the second detector decreases, the net forward bias of transistor Q1 increases and increases the emitter current. The gain of the amplifier increases.

(3) Resistor *R*3, ac bypassed by capacitor *C*1, is the emitter swamping resistor. Capacitor *C*2 and the transformer *T*1 primary form a tuned parallel circuit to develop the collector output signal. Transformer *T*1 matches the collector output of transistor *Q*1 to the input of the following stage. The transformer *T*1 primary is tapped to obtain the desired amplifier selectivity (Q_o).

b. Dc Collector Voltage Control. The power gain of the common-emitter amplifier (B, fig. 151) is controlled by feeding the agc voltage to the base of the amplifier to vary the dc emitter current which in turn varies the dc collector voltage. Variation of the dc collector voltage is accomplished by passing the dc collector current through resistor *R*5. To be effective, resistor *R*5 must be 10,000 ohms or higher in value.

(1) In this case, the tuned amplifier is also operating as a dc amplifier to increase the dc voltage output of the second detector. If the dc voltage output of the second detector is sufficiently large, the dc collector voltage of the tuned amplifier may be varied directly.

(2) In the circuit shown, resistors R1 and R4 form a voltage divider and establish the no-signal negative (forward) bias on the base. The agc voltage from the second detector is *negative* with respect to ground and is fed to the base through dropping resistor R2. When the dc output of the second detector increases (due to a high carrier signal input to the detector), the negative dc voltage fed to the base of transistor Q1 through dropping resistor R2 *increases* the net nega-

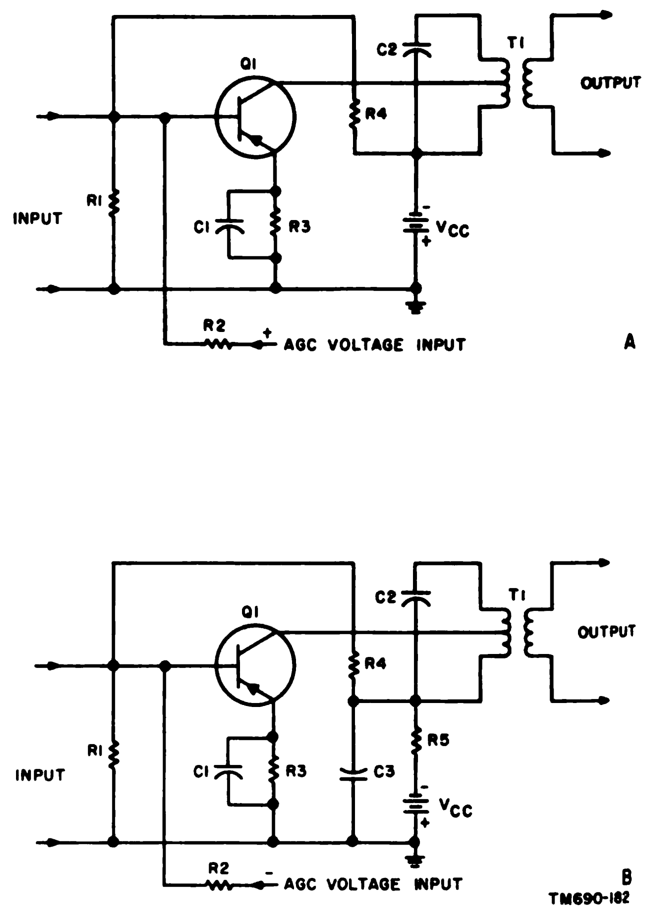


Figure 151. Common-emitter amplifier with automatic gain control of emitter current and collector voltage.

tive (forward) bias on the base and increases the emitter current and therefore the collector current. The flow of increased collector current through resistor R5 reduces the collector voltage. The gain of the amplifier is also reduced (par. 157*b*). When the dc output of the second detector decreases, the net forward bias of transistor Q1 decreases. Emitter and collector current decreases; collector voltage increases and the gain of the amplifier increases.

(3) Capacitor C3 ac bypasses resistor R5 to ground. Correspondingly referenced circuit elements (A and B, fig. 151) perform the same circuit function (*a*(3) above).

159. Typical Tuned Amplifier With Agc

a. Figure 152 shows a 455-kc amplifier (transistor *Q*1) and a second detector (crystal diode *CR*1). The input signal to the amplifier is an if carrier amplitude modulated by an audio signal. The output of the crystal diode is an audio signal. The amplifier is neutralized to prevent oscillation. The gain of the amplifier is controlled by an agc circuit.

b. Transformer *T*1 matches the output impedance of the previous stage to the input impedance of transistor *Q*1, and couples the signal to the transistor *Q*1 base. Capacitor *C*1 and the transformer *T*1 secondary form a parallel resonant (455-kc) circuit. The transformer *T*1 secondary is tapped to obtain the desired selectivity (Q_o). Transistor *Q*1 amplifies the signal which is developed across the parallel resonant (455-kc) circuit formed by capacitor *C*2 and the transformer *T*2 primary. Transformer *T*2 matches the output impedance of transistor *Q*1 to the input impedance of crystal diode *CR*1 and couples the transistor *Q*1 output signal to crystal diode *CR*1. The rectified output of crystal diode *CR*1 is filtered by capacitor *C*7 and resistor *R*4. That is, the if carrier is filtered out by the capacitor and the resistor so that the voltage across capacitor *C*7 consists of an audio signal and a dc voltage (positive with respect to ground) that is directly proportioned to the magnitude of the if carrier. This dc voltage is fed to the base of transistor *Q*1 through dropping resistor *R*1 to provide agc (par. 158*a*). In addition, resistor *R*1 and capacitor *C*2 form an audio decoupling network to prevent audio feedback to the transistor *Q*1 base.

c. Capacitor C5 provides negative feedback to the transistor Q1 base to neutralize the internal positive feedback of the transistor (pars. 153 and 156).

d. Resistor R3 provides fixed forward base bias. Capacitor C3 provides a low-impedance path for the 455-kc signal from the lower end of the input tuned circuit to the emitter. Resistor R2, bypassed for 455 kc by capacitor C4, is the emitter swamping resistor.

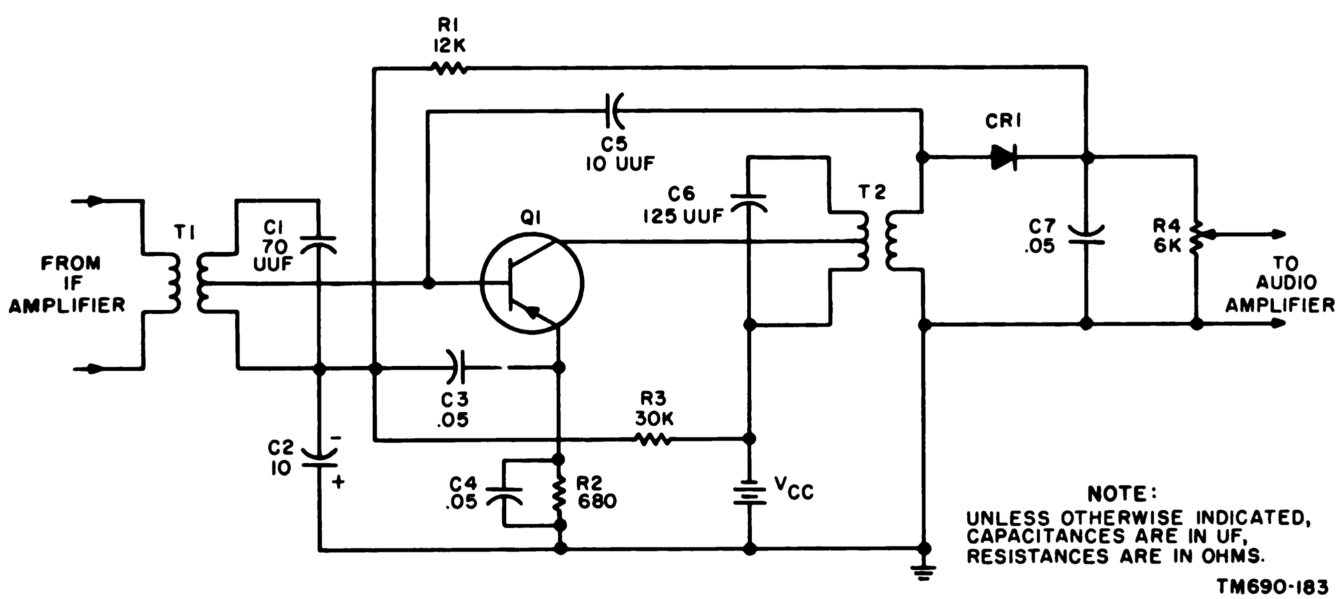


Figure 152. Typical if amplifier (and second detector) with agc and neutralizing circuits.

160. Summary

a. The selectivity of tuned narrow-band amplifiers is achieved by using resonant circuits in the interstage coupling networks.

b. Because the transistor is essentially a power amplifier (even at high frequencies), impedance matching between the output of one stage and the input of the following stage is important for maximum gain.

c. Impedance matching is achieved by selecting the proper turns ratio of the primary and secondary winding in the transformer coupling (A, fig. 141) or using split capacitors (A, fig. 143).

d. Because of low impedances associated with transistors, small inductance values are required in the tuned circuits. With small inductance values, it is difficult in practice to obtain a high unloaded Q_o (selectivity) for the tuned circuit. Use of tapped inductances (B, fig. 141) permits use of higher inductance values to obtain a high Q_o.

e. Double-tuned circuits (fig. 145) improve the uniformity of response in the pass band and more sharply suppress frequencies outside the pass band.

f. Positive feedback occurs in a transistor amplifier through the collector-base junction capacitance and the base-spreading resistance (fig. 147). At low (audio) frequencies, this feedback is negligible; at high (radio) frequencies the amount of feedback may be sufficient to sustain oscillations.

g. To eliminate the possibility of oscillation in high-frequency tuned amplifiers, unilateralizing or neutralizing circuits are used (par. 153).

h. Automatic gain control of a transistor amplifier is achieved by controlling the emitter dc current or the collector dc voltage (par. 157).

CHAPTER 9

WIDE-BAND AMPLIFIERS

161. General

a. A very important factor in certain applications of an amplifying device is its ability to amplify a nonsinusoidal signal. Nonsinusoidal signals, such as sawtooth signals in oscilloscope sweep circuits, consist of a fundamental component and a large number of harmonics. To provide an output signal that is an exact reproduction of the input signal, the amplifying device must be capable of amplifying with uniform gain all of the components (harmonics) of the sawtooth signal. This type of amplifying device is referred to as a *wide-band* or *video amplifier*.

b. Television and radar circuits which amplify rectangular waveforms (pulses) also require wide-band amplifiers. Experiment has shown that amplifiers for such pulses must have a uniform gain from about one-tenth of the lowest pulse-repetition frequency to about 10 times the highest pulse-repetition frequency. The frequency of pulses occurring one-tenth second apart is 10 cps; and that of pulses 1 microsecond apart is 1 megacycle. The amplifying device must provide a uniform gain for 1 cycle to 10 megacycles. When the wide-band amplifier is used in circuits for amplifying pulses, it may also be called a *pulse amplifier*.

c. Figure 153 shows a comparison between a narrow-band amplifier (A) and a wide-band amplifier (B). The narrow-band amplifier (A) is not capable of amplifying all of the harmonics and produces an output wave form that is highly distorted when compared with the input wave form. The wide-band amplifier (B) can amplify

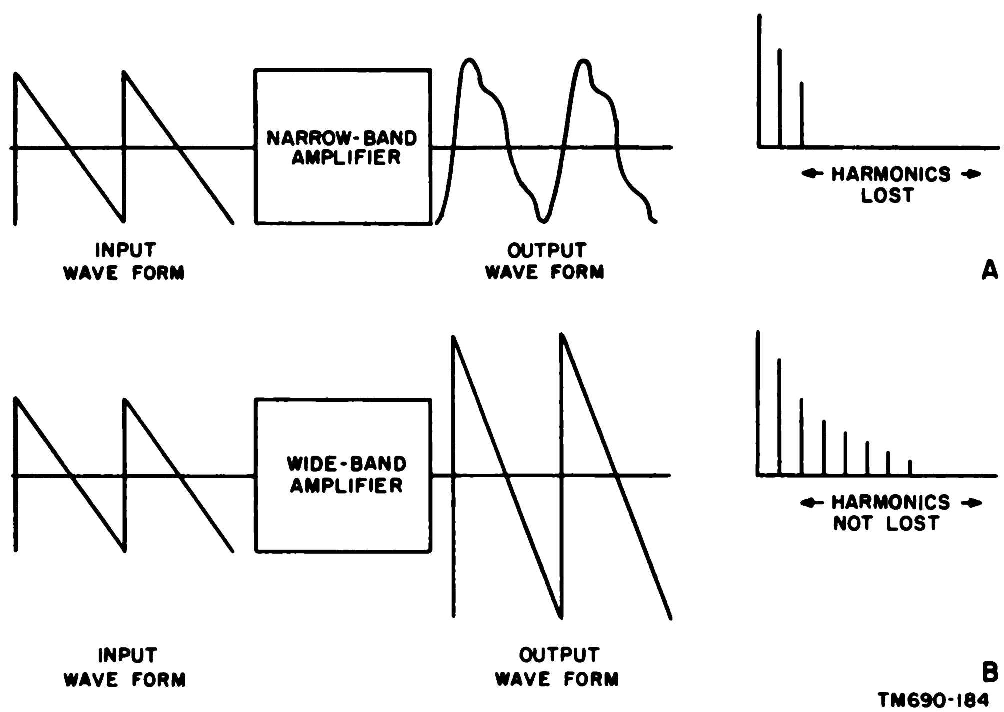


Figure 153. Distortion and reproduction of sawtooth wave form by narrow-band and wide-band amplifiers.

all of the harmonics and thus can produce an output wave form that is an exact reproduction of the input wave form.

162. Coupling Circuits

a. Although transformer coupling is successfully used in audio amplifiers, the frequency response curve (A, fig. 154) shows that the gain is not uniform for the range of audio frequencies. A more uniform frequency response curve (B, fig. 154) is obtained when resistance-capacitance (RC) coupling is employed. The frequency response curve shows a flatness over a wider range of frequencies, although the voltage gain is slightly less. Wide-band amplifiers normally employ a modified RC coupling (pars. 165 and 166). The modified RC coupling of wide-band amplifiers provides flatness over a still wider range of frequencies.

b. Figure 157 shows three types of compensation circuits used in wide-band amplifiers. The shunt compensation (A), the series compensation (B), or the shunt-series (combination) compensation (C), increase the gain at high frequencies.

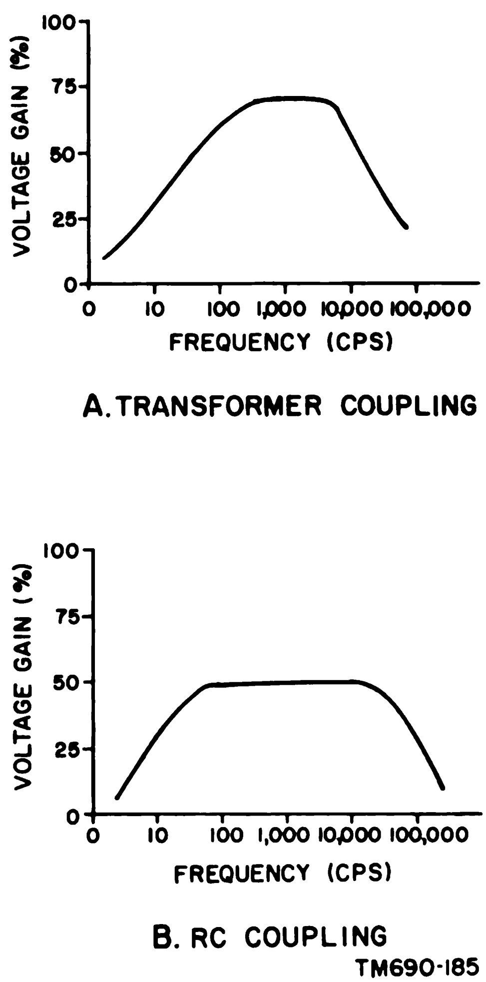


Figure 154. Frequency response curves for transformer or RC coupling.

c. Figure 158 shows a compensation circuit employed in wide-band amplifiers that compensates for reduced gain at low frequencies. This filter coupling circuit is discussed in paragraph 166.

d. Wide-band amplifier frequency response is usually obtained by using both high- and low-frequency compensation. These circuits are treated in detail in paragraphs 165 through 168.

163. Frequency Limiting Elements

Figure 155 shows an *RC* coupled amplifier. Two frequency response limitations must be overcome before the circuit can be employed for wide-band amplification.

a. The high-frequency response is limited by C_o and C_i. The stray capacitance and the capacitance of the output impedance are represented by C_o. The capacitive effect of the input impedance is represented by C_i. Since the reactances of C_o and C_i decrease as the frequency increases, the gain falls off as the frequency increases (par. 165).

b. The low-frequency response is limited by the time constant of capacitor C_c and resistor R_g. This time constant must be long in comparison to the lowest frequency to be amplified (par. 166).

164. Low-Frequency Phase Distortion

A wide-band amplifier must have exceptionally good low-frequency response to reproduce square waves of long duration. Radar reflections received from a long object such as a cloud, and the square waves used in amplifier testing have long durations. The wide-band amplifier must be capable of amplifying all of the harmonics of this type of wave form for an exact reproduction.

a. Low-Frequency Response Distortion.

(1) The fundamental frequency and the low-frequency harmonics of a square wave signal have the greatest amplitudes. The higher-frequency harmonics grow progressively smaller in amplitude as the frequency increases. When compared to the fundamental frequency, the amplitude of the third, the fifth, and the seventh harmonics of a square wave are respectively decreased by one-third, one-fifth, and one-seventh. Since the fundamental frequency and the low-frequency

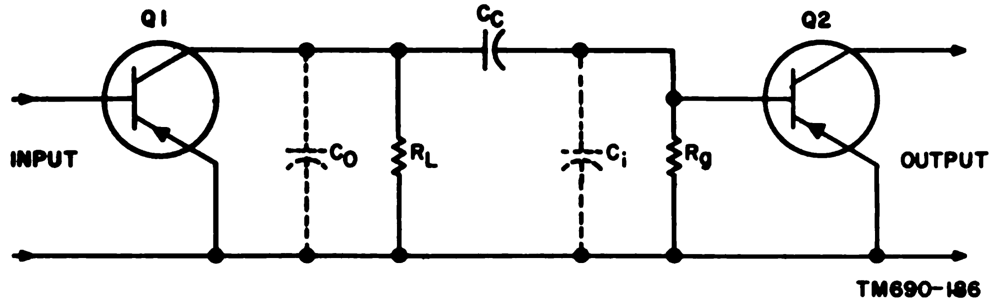


Figure 155. RC-coupled amplifier showing capacitive effect at high frequencies.

harmonics have the greatest amplitudes, any small variation in the phase of the low-frequency response is extremely noticeable. The effect on the higher-frequency harmonics would be negligible since their amplitudes are progressively decreased.

(2) A variation in the phase of the low-frequency response is more noticeable than a variation in the gain. The gain of the low-frequency response will drop to only 99.4 percent of its midfrequency value when the phase shift varies 2° from the phase at midfrequency. In many electronic applications, a phase shift of 2° is the maximum tolerable shift.

b. Phase Distortion Correction. Phase shift distortion can be minimized by allowing the phase shift to vary directly with the frequency. Figure 156 shows a wide-band amplifier of this type. The fundamental frequency and its second harmonic are shown in the input. The fundamental frequency and the second harmonic are shown in the output but shifted in phase. The amount of phase shift is exaggerated for clarification. A delay time of 1/800 of a second causes a 45° phase shift in the fundamental frequency. Since the delay time is constant and the phase shift varies directly with the frequency, the second harmonic is also delayed 1/800 of a second. The resultant phase shift on the second harmonic is 90°. As the frequency doubles, the phase shift doubles and the fundamental frequency and the second harmonic remain in the same relative phase relationship. When the phase shift varies directly with the frequency, there is no phase distortion.

165. High Frequency Compensation

a. Shunt Compensation. For shunt compensation (A, fig. 157), inductor L_1 is added in series with load resistor R_L. This compensates for the shunting effect of output impedance capacitance C_o and impedance capacitance C_i. The capacitive reactance of output impedance capacitance C_O and input impedance capacitance C_i shunt inductor L_1 and load resistor R_L. Capacitor C_C is practically a short circuit at high frequencies. Since inductor L_1 (in series with resistor R_L and capaci-

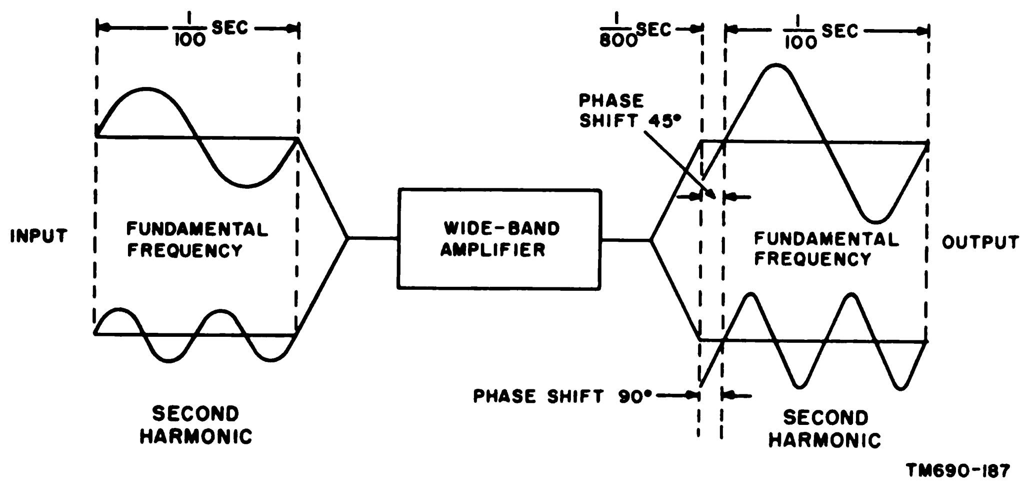


Figure 156. Wide-band amplifier, phase relationship comparison.

495909 O—59——11

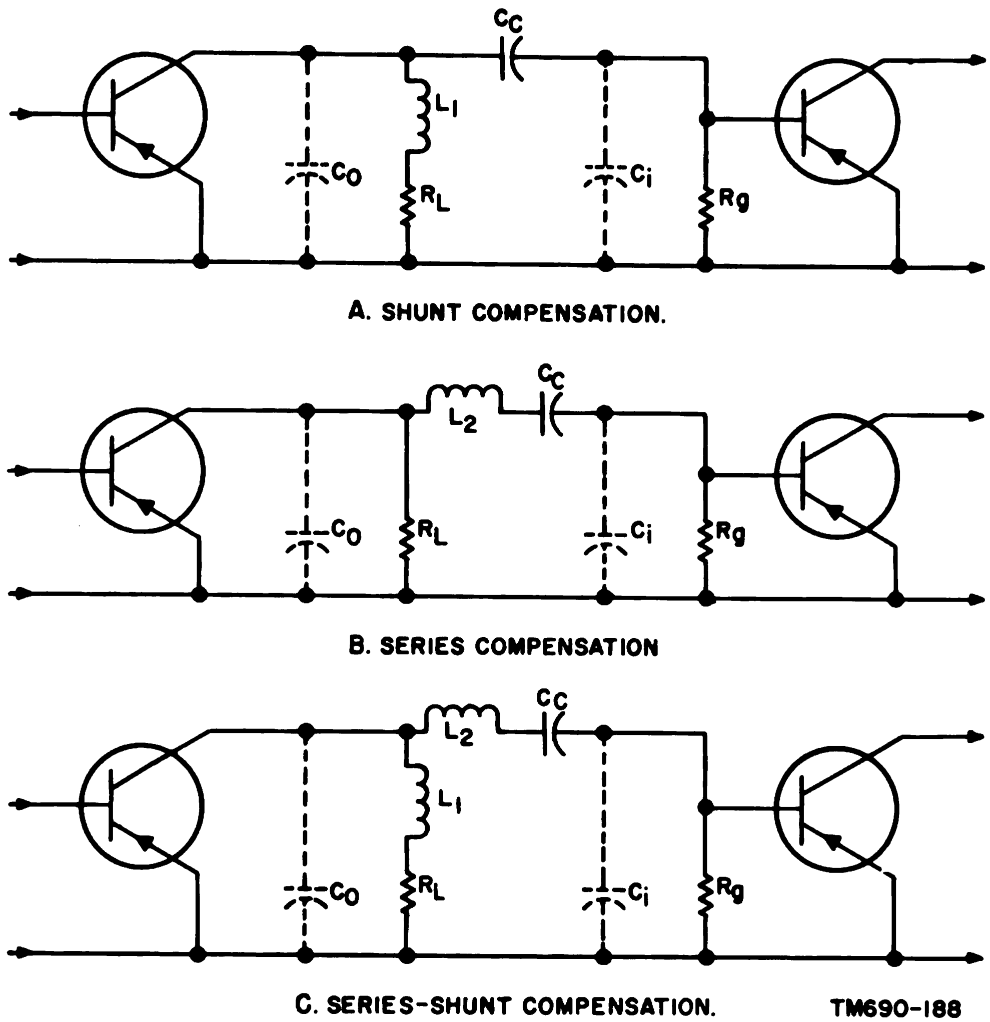


Figure 157. Wide-band amplifier, high-frequency compensation coupling.

tors C_i and C_O form a parallel resonant circuit having a very broad response, this type of compensation may also be called shunt peaking. The resonant peak of this parallel combination maintains a practically uniform gain in the high-frequency range. In the uncompensated circuit (fig. 155), the high-frequency gain was reduced by the capacitive reactances of capacitors C_O plus C_i. The impedance of the parallel resonant circuit has a value that is approximately the same as load resistor R_L. When the frequency increases, the decrease in the capacitive reactances of C_O plus C_i is exactly compensated by the increase in the inductive reactance of inductor L_1. The frequency response is increased, extending the flatness of the response curve over a much higher range of frequencies.

b. Series Compensation. For series compensation (B, fig. 157), inductor L_2 is added in series with capacitor C_c. Considering capacitor C_c as a short circuit at high frequencies, inductor L_2 and capacitor C_i form a series resonant circuit at very high frequencies. As very high frequencies are approached, the series circuit approaches resonance. Thus, current flow through capacitor C_i increases as the frequency increases. The capacitive reactance of capacitor C_O decreases the voltage across load resistor R_L as the frequency increases. Since the current flow through capacitor C_i increases with frequency it compensates for the decrease in voltage across load resistor R_L. The frequency response is approximately the same as for the shunt peaking coupling circuit. The high frequency gain of a wide-band amplifier using series peaking is about 50 percent greater than that using shunt peaking.

c. Series-Shunt Compensation. For series-shunt compensation (C, fig. 157), the series-compensation and the shunt-compensation coupling circuits are combined. This type of coupling is called *combination compensation* or *combination peaking*. The qualitative analysis discussed in *a* and *b* above apply to this circuit. The frequency response is approximately the same as that of the shunt peaking or the series peaking coupling circuits. The high frequency gain of the combination peaking coupling circuit is approximately 80 percent greater than the series peaking coupling circuit.

166. Low-Frequency Compensation

On the low-frequency end of the frequency response range, the input and the output capacitances of the transistors have no effect on the frequency response. The low-frequency response is limited by capacitor C_C (fig. 158) and resistor R_g. The time constant (R_gC_C) must be large to prevent the low-frequency response from falling off and to prevent phase distortion. The loss of gain at low frequencies is minimized by adding a compensating filter in series with load resistor RL. The compensating filter consists of resistor R_F and capacitor C_F. The filter increases the collector load impedance at low frequencies and compensates for the phase shift produced by capacitor C_C and resistor R_g. For high frequencies, capacitor C_F is practically a short circuit. Thus, at high frequencies, the collector load impedance consists only of resistor RL. As the frequency decreases, the reactance of capacitor C_F *increases*. For very low frequencies capacitor C_F is practically an open circuit. At very low frequencies the collector load impedance consists of resistor RL and resistor R_F. This combination extends the frequency response curve over a much lower range of frequencies. The gain and phase response becomes more uniform over the low frequency range of the frequency response curve.

167. Square Wave Characteristics

Figure 159 shows a wide-band amplifier with the input and the output characteristics for a square wave under various conditions. The effects

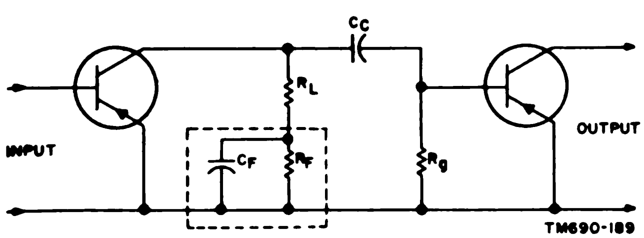


Figure 158. Wide-band amplifier, low-frequency compensation coupling.

of capacitor C_C (fig. 158) and resistor R_g are compensated for by the compensating filter (R_F and C_F). The value of resistor R_F and capacitor C_F determines the best low-frequency response and most uniform time delay. The square wave input (fig. 159) consists of the fundamental frequency and the odd harmonics. With the value of resistor RL (fig. 158) too small, low-frequency attenuation results (A, fig. 159). With the value of resistor RL (fig. 158) too great, too much low-frequency gain (B, fig. 159) results. When the time constant of the compensating filter (R_FC_F, fig. 158) does not compensate for the phase shift across capacitor C_C and resistor R_g, phase distortion is the result (C and D, fig. 159). With the correct values of resistor R_F (fig. 158) and the time constant of the compensating filter (R_FC_F) a uniform gain and time delay (E, fig. 159) is obtained.

168. Simultaneous High- and Low-Frequency Compensation

Figure 160 shows a typical wide-band amplifier (30 cps to 4 mc) using both high- and low-frequency compensation. The high- and the low-compensating circuits operate independently and do not interfere with each other. At low frequencies the series reactance of inductor $L1$ is very small and has no effect on the collector load impedance. The series reactance of inductor $L2$ is also very small and has no effect on the input circuit. The reactances of output impedance capacitance C_O (not shown) and input impedance capacitance C_i (not shown) are so large that they have no effect at low frequencies. This means that the combination peaking coupling has no effect at low frequencies. Similarly, at high frequencies, the reactance of capacitor C_F is very small and is practically a short circuit. Resistor R_F has no effect on the collector load impedance at high frequencies. Thus, compensating filter R_FC_F has no effect at high frequencies.

169. Summary

a. A wide-band amplifier must be capable of producing an undistorted amplified output of a nonsinusoidal signal input.

b. Wide-band amplifiers find application in television, radar, and many other electronic applications.

c. Wide-band amplifiers employ RC coupling circuits with high- and low-frequency compensating networks.

d. The high-frequency response is approximately uniform for all three types of high-frequency peaking circuits.

e. The high-frequency response gain is lowest with shunt peaking.

f. High-frequency response gain with series peaking is greater than shunt peaking.

g. High-frequency response gain with combination peaking is greater than shunt or series peaking.

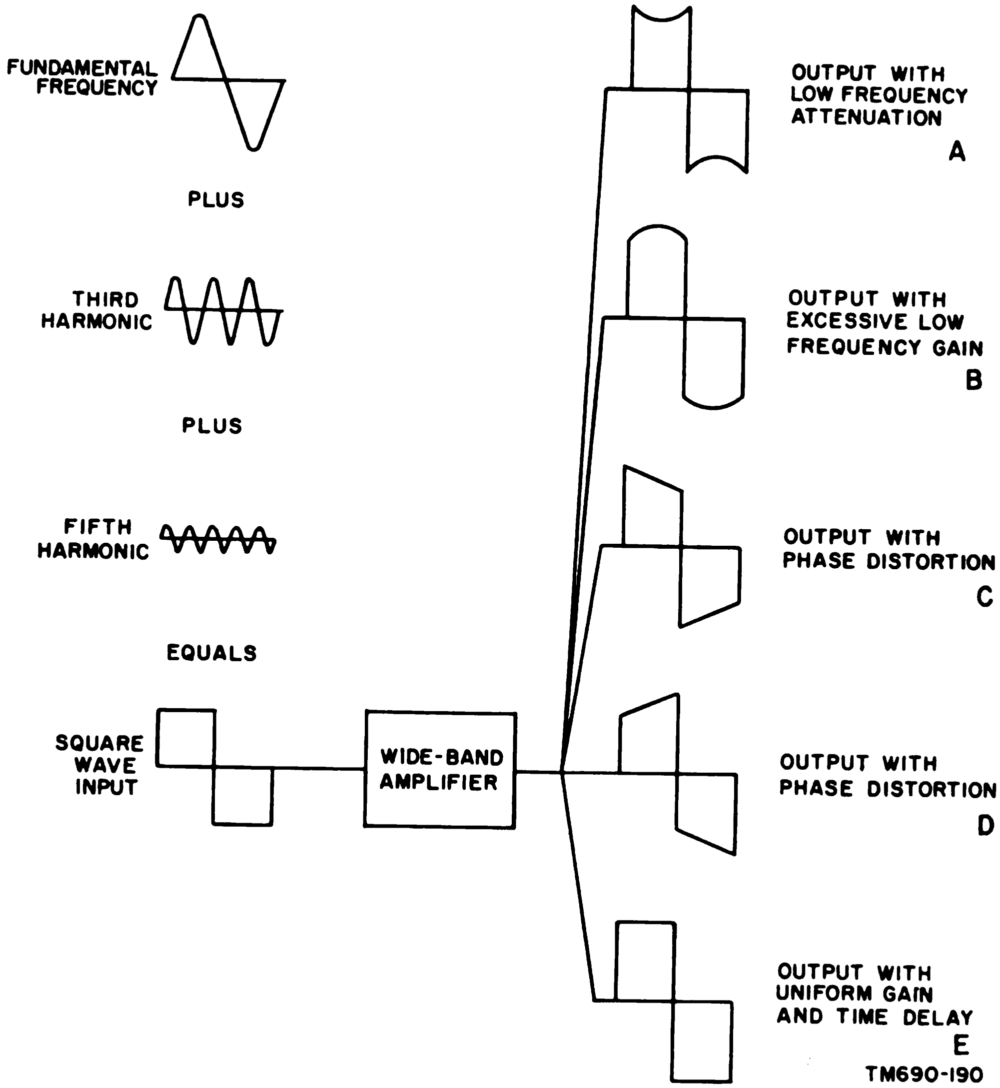


Figure 159. Wide-band amplifier showing square wave input and output characteristics.

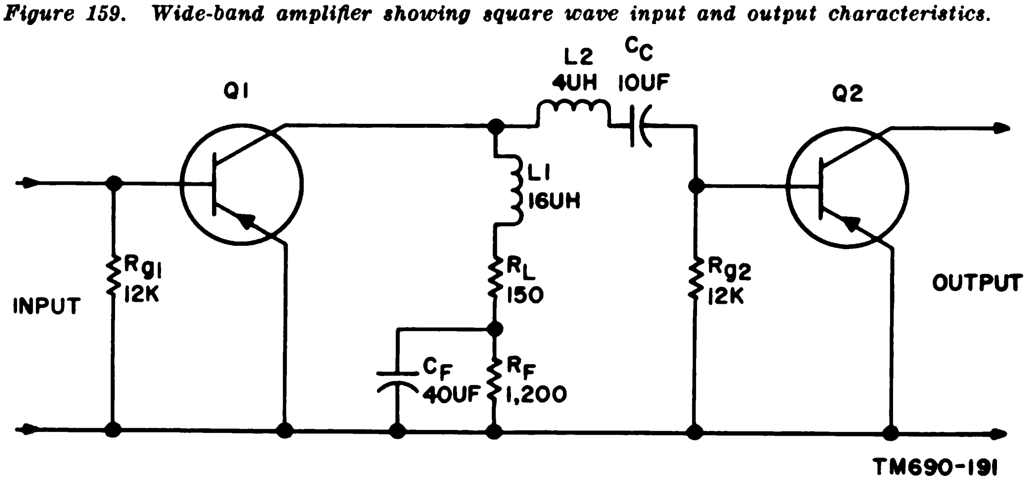


Figure 160. Wide-band (30 cps to 4 mc) amplifier showing typical values of components.

CHAPTER 10

OSCILLATORS

Section I. INTRODUCTION

170. General

a. Regenerative Feedback. To generate ac power with a transistor amplifier (fig. 161) a portion ($p_{feedback}$) of the output power (p_{out}), must be returned to the input in phase (regenerative or positive feedback) with the starting power (par. 171). The power delivered to the load (p_{load}), will be the output power less the feedback power.

b. Oscillating Frequency and Dc Power. Besides the requirement for regenerative feedback, frequency determining elements ((1) below) and the necessary dc bias voltages ((2) below) must be incorporated in the transistor oscillator circuit.

(1) The frequency determining circuit elements may consist of an inductance-capacitance (*LC*) network, a crystal, or a resistance-capacitance (*RC*) network. Transistor oscillator circuits that use these networks are described in paragraphs 173 through 183.

(2) Bias voltage requirements for the transistor oscillator are similar to those for the transistor amplifier (pars. 38 and 39). Stabilization of operating point (pars. 74–98) also is an important factor for consideration in the transistor oscillator circuit, since instability of the operating point affects the output amplitude, waveform, and frequency stability (par. 172).

c. General Comparison of Input and Output Differences Between Electron Tubes and Transistors. The input and output impedances of an electron tube circuit are high; the feedback signal suffers little loss, because of the mismatch in the feedback network. In the common-base configuration of the transistor, the input impedance is low and the output impedance is high. Coupling the feedback signal from the output to the input requires a feedback network to match the unequal impedances; sometimes the loss due to mismatch may be compensated for by providing more feedback energy. Use of the other transistor configurations involve similar problems (*d* below).

d. Selection of Transistor Configuration. The choice of a transistor configuration for an oscillator is determined by the oscillator requirements and the advantages of a particular transistor amplifier configuration. The following factors concerning each configuration are important in making the choice.

(1) The common-base configuration has the lowest input impedance and the highest output impedance. Compensation for the loss in the feedback circuit must be made unless matching occurs (*c* above). Current gain for the junction transistor is less than one. Voltage and power gains are greater than one. No phase reversal exists between input and output.

(2) The common-emitter configuration has moderate input and output impedances, reducing the requirement for matching with the feedback circuit. This configuration closely resembles the grounded cathode electron tube circuit in that a phase reversal occurs between the input and output circuits. Current, voltage, and power gains are all greater than one. Highest power gains are obtained with this configuration making it generally desirable for use in transistor oscillator circuits.

(3) The common-collector configuration has a high input impedance and a moderate output impedance. The requirement for

matching the input and output impedances with the feedback circuit also exists with this configuration ((1) above). Voltage gain is less than one. Current and power gains are greater than one. No phase reversal exists between input and output.

e. Frequency Limitations.

(1) In general, the data sheet of a transistor includes the cutoff frequency of the transistor in its high-frequency characteristics. This frequency is given with relation to the current gain of the transistor when used in either the common-base or the common-emitter configuration and is designated f_{α_b} or f_{α_e}, respectively. The cutoff frequency is defined as the point at which the current gain drops 3 db or 0.707 times its low-frequency (normally 1,000 cps) value.

(2) Oscillators can operate efficiently above the cutoff frequency and could operate up to but not including the f_{max} of the transistor. The maximum frequency of the transistor is defined as the frequency at which the power gain of the transistor amplifier is unity. That is, an input signal of a specific level, would appear in the output of the transistor amplifier at the same level (no loss or gain). Since power gain is required to overcome losses in the feedback circuit, operation of the transistor oscillator at f_{max} is not possible. The operating frequency is therefore chosen at some value, below f_{max}, at which sufficient power gain for feedback and output is obtained.

171. Initial and Sustained Oscillation

a. Initial Oscillation. Oscillation may be initially caused in a transistor circuit by using the same technique as a vacuum tube oscillator circuit (TM 11–662). The circuit may be externally excited (triggered) or self-excited. Triggered circuits are discussed in paragraphs 185 through 193. In the self-excited circuit, at the moment that dc power is applied, the energy level does not instantly reach maximum but, instead, gradually approaches it. Oscillations build up to a point limited by the normal operation of the amplifier, the feedback energy, and the nonlinear condition of the circuit.

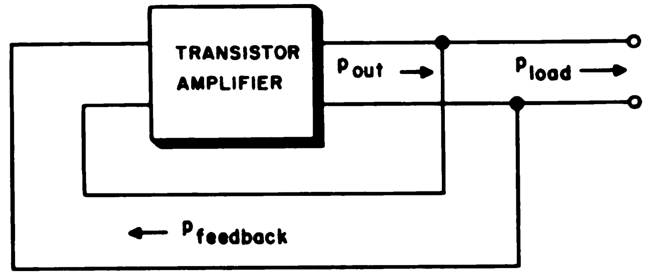
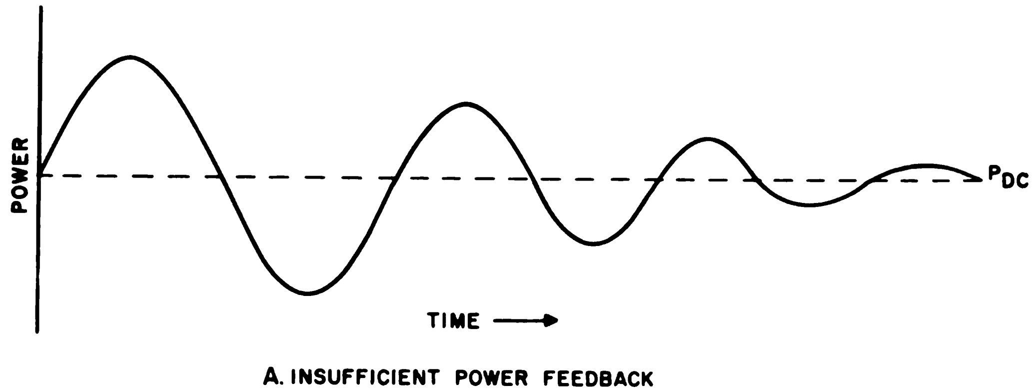


Figure 161. Transistor oscillator showing application of feedback, block diagram.

b. Feedback. The output power, p_{out} (fig. 161), is equal to the power gain of the amplifier multiplied by the input power.

(1) When the amplifier power gain is less than one, a damped oscillation is obtained (A, fig. 162); that is, the oscillations become smaller with time until they no longer exist. For example, if the power gain of the transistor amplifier is equal to 0.9 and the starting signal is 1 dbm, the output power after the first pulse is 0.9 dbm. If all of this power is fed back to the input and amplified, the output power on the second cycle becomes 0.81 dbm. Feeding this power back to the input results in a further reduction of the output power. This process continues until oscillation ceases. This process is similar to the action of oscillation in a charged parallel-tuned circuit, where the resistance present dampens the oscillation. Thus, to sustain oscillation (B, fig. 162), the power gain factor of an amplifier must be equal to or greater than one.

(2) In the practical transistor oscillator circuit (fig. 163), the power output is divided. A portion (p_{load}) is supplied to the load with the remainder ($p_{feedback}$) going to the feedback network. Power losses occur in the feedback network. The feedback power is equal to the input power (p_{in}) plus the loss incurred in the feedback network. For example, if the power gain of a transistor amplifier is 20 and the input power (p_{in}) level is 1 dbm, the output power (p_{out}) is 20 dbm. If the loss incurred in the feedback network is 10 dbm, the feedback power ($p_{feedback}$) must be 11 dbm to sustain oscillation.



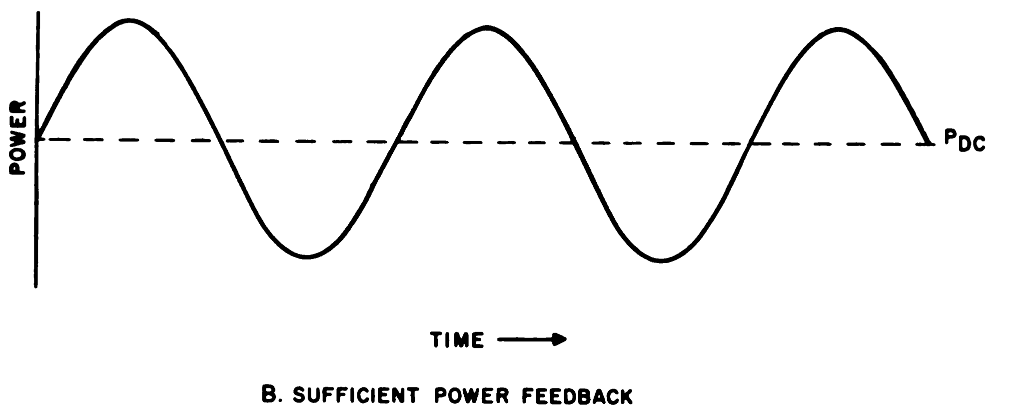


Figure 162. Waveforms showing effects of sufficient and insufficient power feedback.

That is, $p_{feedback}$ less the loss in the network equals the p_{in} and is 1 dbm. Thus, all values remain constant and oscillation is sustained.

(3) The power delivered to the load (p_{load}) is the output power less the feedback power.

172. Frequency Stability

Frequency instability caused by transistor characteristics rather than external circuit elements is discussed in *a* and *b* below.

a. Operating Point.

(1) The dc operating point is usually chosen so that the operation of the transistor circuit is over the linear portion of the transistor characteristic. When the operation of the circuit falls into a nonlinear portion of the transistor characteristic because of variations in bias voltages, the transistor parameters vary (pars. 51 and 52). Since these parameters are basic to the transistor, and therefore affect frequency of oscillation, frequency variation will occur with changes in bias voltages. A constant supply voltage is thus a prime requirement for frequency stability.

(2) The collector-to-emitter capacitance parameter of the transistor affects frequency stability more than the other parameters. This reactive element (sometimes referred to as the barrier capacitance) varies with changes in collector or emitter voltages and with temperature (par. 103). In high-frequency oscillators, when the collector-to-emitter capacitance becomes an important factor, the effect of changes in the capacitance may be minimized by inserting a relatively large

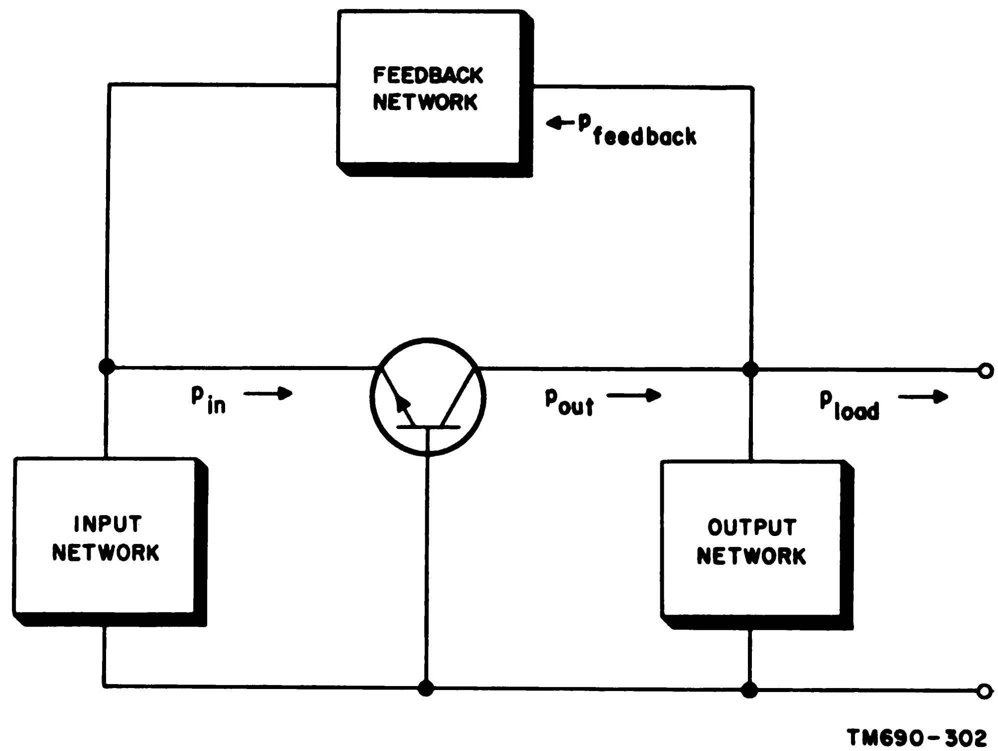


Figure 163. Transistor oscillator showing important networks in block form.

swamping capacitor across the collector-to-emitter electrodes. The total capacitance of the two in parallel results in a circuit which is less sensitive to variations in voltage. The added capacitor may be part of a tuned circuit (par. 175).

(3) The use of a common bias source for both collector and emitter electrodes maintains a relatively constant ratio of the two voltages. In effect, a change in one voltage is somewhat counteracted by the change in the other, since an increased collector voltage causes an increase in the oscillating frequency and an increased emitter voltage causes a decrease in the oscillating frequency. However, complete compensation is not obtained since the effects on the circuit parameters of each bias voltage differ.

b. Temperature. Changes in operating point with changes in temperature are encountered in the transistor oscillator. The effects of changes in temperature on the transistor amplifier are given in paragraph 74. Since the transistor oscillator is an amplifier with additional circuitry (par. 170), the means of temperature stabilization are the same for both oscillators and amplifiers.

173. Basic Transistor Oscillator Circuits

The circuits in figure 164 illustrate the basic component arrangements and polarity applications in transistor oscillators. Each circuit provides amplification, regenerative feedback, and inductance-capacitance tuning. Exact bias and stabilization arrangements are not illustrated. However, the relative transistor dc potentials required for normal functioning of the oscillator are shown.

Note. In all circuits illustrated, a PNP-type transistor is used. NPN transistors may be substituted in any of the circuits, as long as the dc polarities are reversed.

a. The circuits in A and B, figure 164, are similar. In both cases, feedback is coupled from the collector to the base by a transformer (tickler coil (par. 174)). The circuit in B of the figure is a shunt fed version of the circuit in A. Since the feedback signal is from collector to base in each case, the necessary feedback signal phase inversion is accomplished by the transformer which, when properly connected, provides 180° phase shift.

b. The circuits in C and D, figure 164, are also tickler-coil oscillators. In C, regenerative feedback with zero phase shift is obtained in the tuned collector-to-emitter circuit by the proper

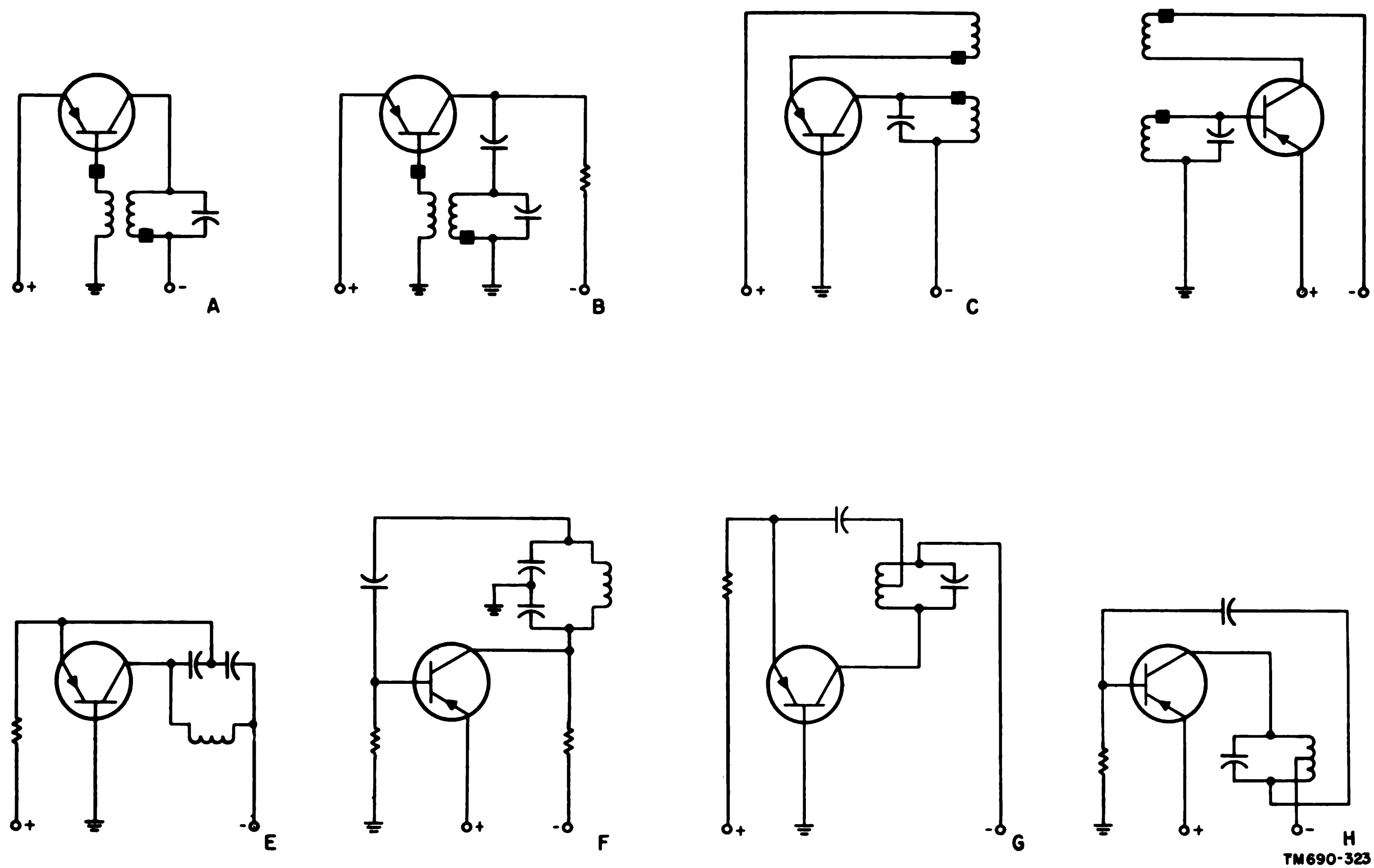


Figure 164. Basic transistor oscillator circuits.

connection of the transformer. In D, since the feedback is from collector to base, 180° phase shift is required. The signal in the untuned collector winding is coupled and inverted in phase to the tuned-base winding.

c. The circuits in E and F of the figure are transistorized versions of the Colpitts-type electron tube oscillator (par. 175). In E, the signal from the tuned collector is coupled to the emitter in phase. In F, the signal from the tuned collector is coupled to the base 180° out of phase.

d. The circuits in G and H of the figure are similar to those in E and F, except that a split inductance is used to provide the necessary feedback in place of the split capacitance. These are transistorized versions of the Hartley-type electron tube oscillator. In each of these circuits, the collector is tuned. Since each coil functions as an autotransformer, feedback in proper phase is accomplished by induction. In G, the feedback signal is coupled from collector to emitter with no phase shift. In H, the feedback signal is coupled from collector to base with 180° phase shift.

Note. **Resistances and capacitances (external to tuned circuits) are added in figure 164 to indicate ac and dc blocking components respectively. More detailed descriptions are given in succeeding paragraphs.**

Section II. LC RESONANT FEEDBACK OSCILLATORS

174. Tickler Coil Oscillators

a. Basic Oscillator Analysis. A transistor oscillator using a tickler coil for inductive feedback is shown in A, figure 165. Dc bias circuits have been omitted for simplicity (the emitter-base diode is forward-biased and the collector-base diode is reverse-biased). The nonlinear conditions for amplitude limitation (par. 171*a*) in this circuit are that the emitter-base diode becomes reverse-biased (transistor is cut off) or the collector-base diode becomes forward-biased (transistor is saturated). The currents and voltages indicated on the figure are ac and are instantaneous. The waveforms of emitter and collector currents (i_e and i_c, respectively) are shown in B, figure 165.

(1) Oscillations begin in the circuit when the bias conditions of the transistor are normal and power is applied. The amplitude of current flow in the circuit will increase steadily (between points X and Y in B, figure 165) due to the regenerative feedback coupled from the collector circuit to the emitter circuit by the transformer windings (winding 3–4 to winding 1–2).

(2) A point (Y) is reached at which time the collector current can no longer increase. The transistor is saturated (par. 207*b* (3)), and since no further change in collector current exists, feedback ceases and emitter current i_e begins to fall. This decrease in emitter current results in a decrease in collector current i_c.

(3) The decrease in collector current causes the induced feedback voltage to reverse and the emitter current decreases. The emitter current decreases steadily (between points Y and Z) because of the regenerative feedback until another point (Z) is reached at which time the emitter current is cut off. The emitter-base diode is reverse-biased at this point. Collector current ceases to flow causing the feedback current to cease flowing. Once the feedback current, which had driven the transistor to cutoff, ceases, the bias conditions begin to revert to their original state ((1) above) and the process is repeated. Basically, the transistor is driven to saturation, then to cutoff, then back to saturation, etc. The time for change from saturation to cutoff is primarily determined by the tank circuit, which, in turn, determines the frequency of oscillation.

b. Tuned-Base Oscillator (fig. 166). The tuned-base transistor oscillator is similar to the electron tube tuned-grid oscillator. In this application, one battery is used to provide bias for the common-emitter configuration. Resistors R_B, R_C, and R_F provide the necessary bias conditions. Resistor R_E is the emitter swamping resistor. The components within the dotted lines comprise the transistor amplifier. The collector shunt-fed arrangement prevents dc current flow through tickler winding 1–2 of transformer $T1$. Feedback is accomplished by the mutual inductance between the windings of the transformer.

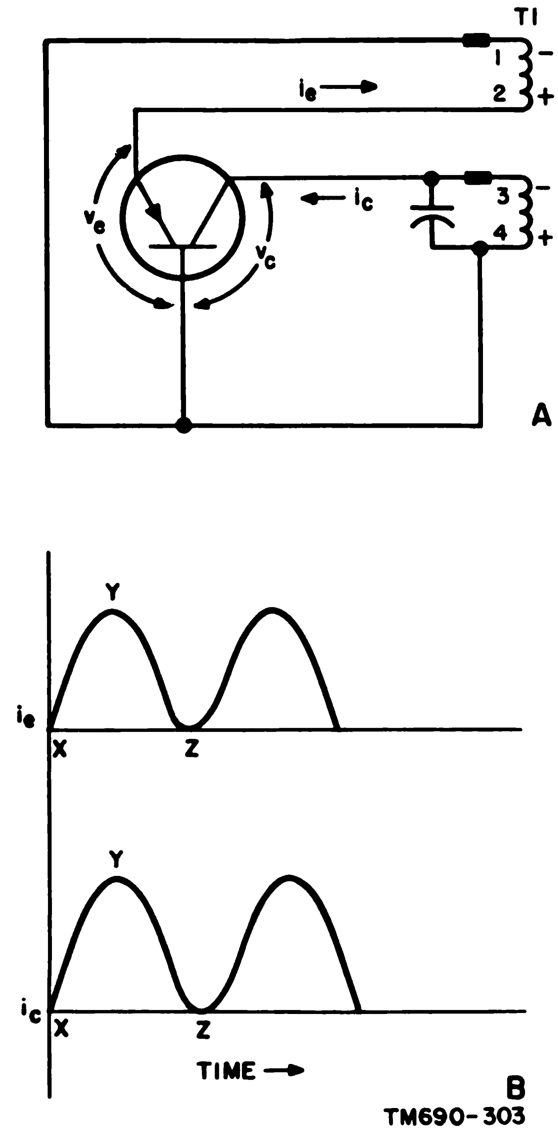


Figure 165. Typical transistor oscillator and current waveforms, bias circuits omitted.

(1) The tank circuit consists of transformer winding 3–4 and variable capacitor $C1$ and constitutes the frequency determining components of the oscillator. Variable capacitor $C1$ permits tuning the oscillator through a range of frequencies. Capacitor C_C couples the oscillatory signal to the base of the transistor and also blocks the dc. Without capacitor C_C, the base bias condition would be determined primarily by the low dc resistance of transformer winding 3–4. Capacitor C_B bypasses the ac signal around emitter swamping resistor R_B and prevents degeneration.

(2) Oscillations in this circuit are started the moment the dc power (battery V_{CC}) is connected in the circuit. At that moment, a surge of current flows through the transistor, and the base circuit goes into oscillation. The oscillations, amplified by the transistor, appear across transformer winding 1–2, and are coupled to transformer winding 3–4. The feedback signal is regenerative and of sufficient magnitude to sustain oscillations.

(3) The output signal is coupled from the collector through coupling capacitor C_O to provide a fixed output to the load.

c. Tuned-Collector Oscillator (fig. 167). The tuned-collector transistor oscillator is similar to the electron tube tuned-plate oscillator. Resistors R_F and R_B establish the base bias. Resistor R_E is the emitter swamping resistor. Capacitors C_B and C_E bypass ac around resistors R_B and R_E, respectively. Although a series-fed arrangement is illustrated, a shunt-fed arrangement of the tuned-collector oscillator is also possible with slight circuit modification. This arrangement would result in a circuit which would be almost identical with the tuned-base oscillator, except for the placement of the tank circuit ((1) below).

(1) The tuned circuit consists of transformer winding 3–4 and variable capacitor C1. Capacitor C1 permits tuning the oscillator through a range of frequencies. Oscillations in this circuit are started the moment the dc power is applied (*b*(2) above). Regeneration is accomplished by coupling the feedback signal from transformer winding 3–4 to transformer winding 1–2.

(2) Transformer winding 5–6 couples the signal output to the load.

175. Colpitts and Clapp Oscillators

a. Colpitts Oscillator. Figure 168 illustrates a transistor Colpitts oscillator circuit. Regenerative feedback is obtained from the tank circuit and applied to the emitter of the transistor. Base bias is provided by resistors R_B and R_F. Resistor R_C is the collector load resistor. Resistor R_E develops the emitter input signal and also acts as the emitter swamping resistor. The tuned circuit consists of capacitors $C1$ and $C2$ in paral-

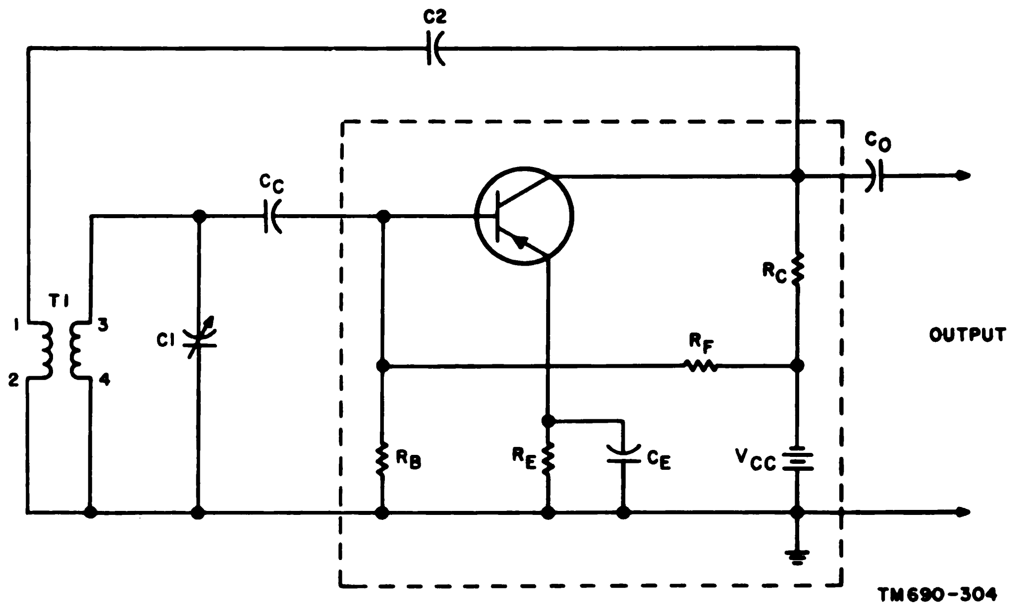


Figure 166. Tuned-base oscillator.

lel with transformer winding 1–2. Capacitors $C1$ and $C2$ form a voltage divider. The voltage developed across $C2$ is the feedback voltage. Either or both capacitors may be adjusted to control the frequency and the amount of feedback voltage. For minimum feedback loss, the ratio of the capacitive reactance of $C1$ to $C2$ should be approximately equal to the ratio of the output impedance to the input impedance of the transistor.

b. Clapp Oscillator. Modification of the Colpitts oscillator by including a capacitor in series with winding 1–2 of the transformer, results in the Clapp oscillator (fig. 169). This added capacitance improves the frequency stability of the oscillator. When the series capacitance of capacitors C1 and C2 is large compared with the added capacitance, the oscillator frequency is effectively determined by transformer winding 1–2 and capacitance C series combination. This capacitance may be made variable to provide os-

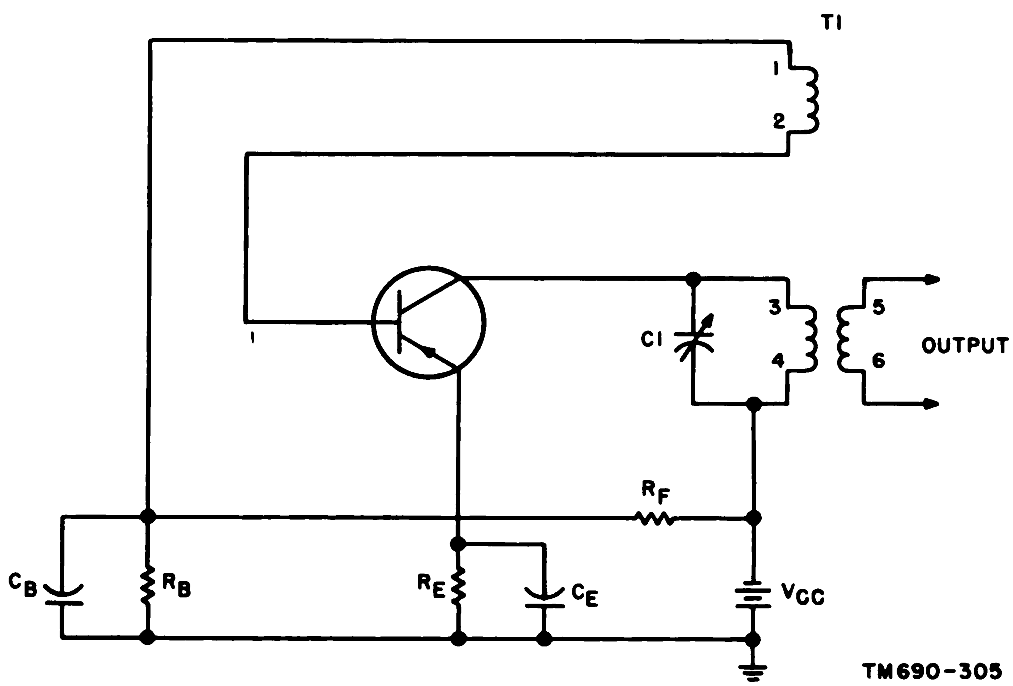


Figure 167. Tuned-collector oscillator.

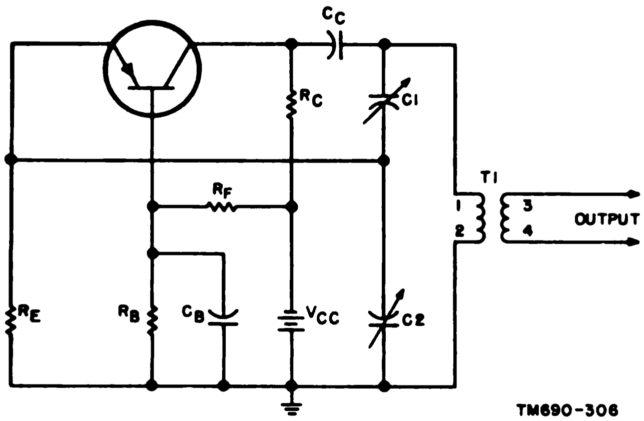


Figure 168. Transistor Colpitts Oscillator.

cillations over a range of frequencies. At the resonant frequency, the shunting impedance of the LC series combination is at a minimum, thereby making the oscillating frequency comparatively independent of the transistor parameter variations.

c. Simplified Feedback Networks. The common-base configuration is used in the Colpitts oscillator of figure 168. No phase shift exists between the signal on the collector and the signal on the emitter. Therefore, the feedback signal is not phase shifted ((1) below). The common-emitter configuration is used in the Clapp oscillator of figure 169. A 180° phase shift exists between the signal on the collector and the signal on the base. In this instance, the feedback signal is inverted to provide an inphase relationship at the input ((2) below).

Note. Only ac conditions of the feedback signal are discussed in (1) and (2) below and illustrated on figure 170. Thus, the polarities indicated at the transistor and capacitor are instantaneous.

(1) When the common-base configuration is used, the feedback signal is returned between emitter and ground (A, fig. 170). As the emitter goes negative, the collector also goes negative, developing the potential polarities across capacitors C1 and C2, as shown. The feedback voltage developed across capacitor C2, which is fed back between the emitter and ground, also goes negative. The inphase relationship at the emitter is maintained.

(2) When the common-emitter configuration is used, the feedback signal is returned between base and ground (B, fig. 170). As the base goes positive, the collector goes negative, developing the potential polarities across capacitors C1 and C2 as shown. The voltage developed across capacitor C2 is fed back between base and ground. Since the negative going side of capacitor C2 is grounded, the voltage fed back is inverted, and the inphase relationship at the base is maintained.

176. Hartley Oscillator

The Hartley oscillator (figs. 171 and 172) is similar to the Colpitts oscillator except for the use of a split inductance instead of a split capacitance to obtain feedback. The shunt-fed (*a* below), and series-fed (*b* below) oscillators are operationally similar. They differ primarily in the method of obtaining collector bias. A modified Hartley-type oscillator which provides greater power output, is described in *c* below.

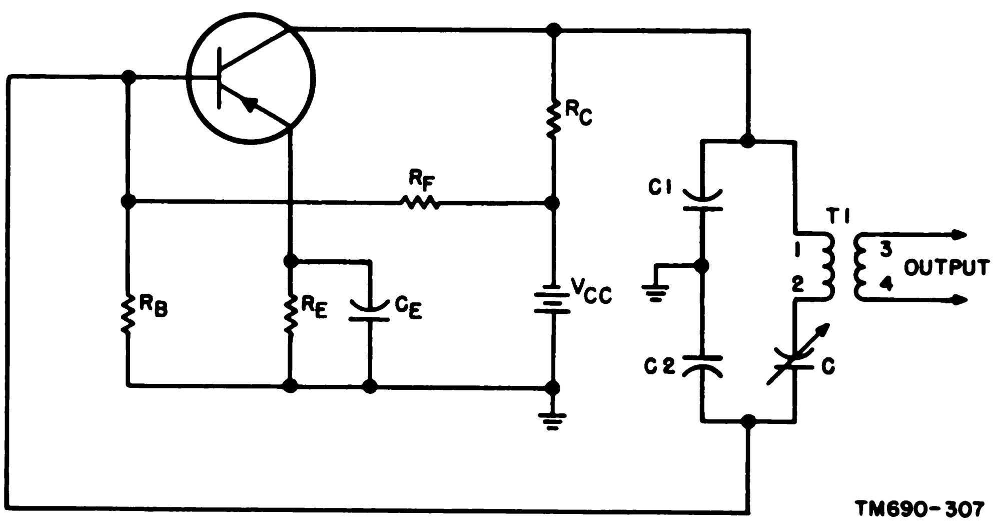


Figure 169. Transistor Clapp oscillator.

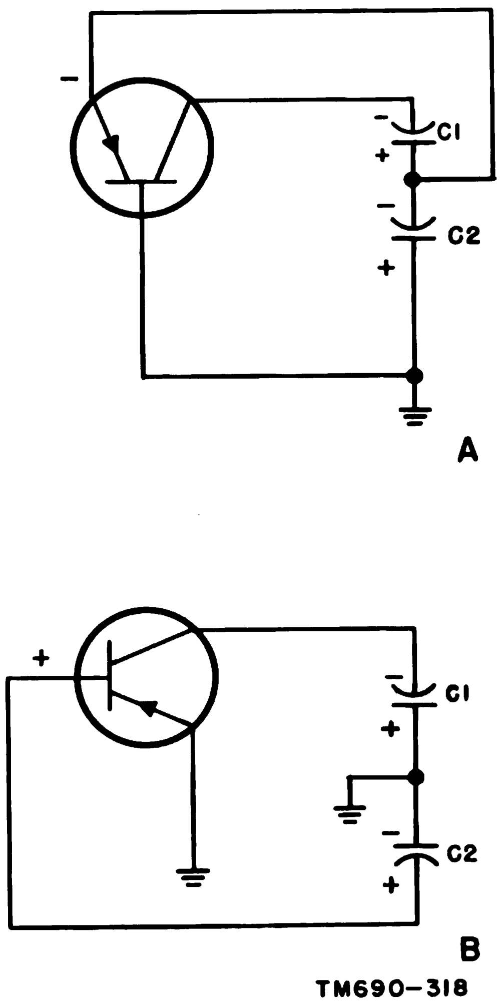


Figure 170. Simplified feedback networks in Colpitts and Clapp oscillators.

a. Shunt-Fed Hartley Oscillator (fig. 171).

(1) Resistors R_B, R_C, and R_F provide the necessary bias conditions for the circuit. The frequency determining network consists of the series combination of transformer windings 1–2 and 2–3 in parallel with capacitor $C1$. Since capacitor $C1$ is variable, the circuit may be tuned through a range of frequencies. Capacitor $C2$ is a dc blocking capacitor. Capacitor C_E provides an ac bypass around emitter swamping resistor R_E.

(2) The inductance functions as an autotransformer to provide the regenerative feedback signal. The feedback is obtained from the induced voltage in transformer winding 2–3, coupled through capacitor C_C to the base of the transistor. By shunt feeding the collector through resistor R_C, direct-current flow through transformer $T1$ primary is avoided.

b. Series-Fed Hartley Oscillator (fig. 172).

(1) Resistors R_B and R_F provide the necessary bias for the base-emitter circuit. Collector bias is obtained through transformer winding 1–2. Capacitor C_E provides an ac bypass around emitter swamping resistor R_E.

(2) The feedback is obtained from the induced voltage in transformer winding 2–3 coupled through capacitor C_C to the base of the transistor. Capacitor C2

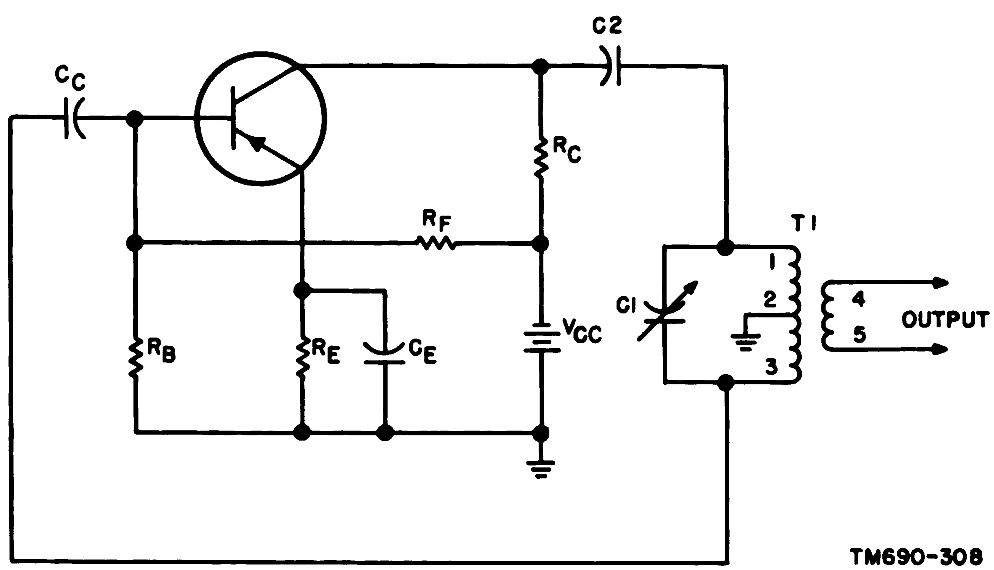


Figure 171. Transistor Hartley oscillator, shunt-fed.

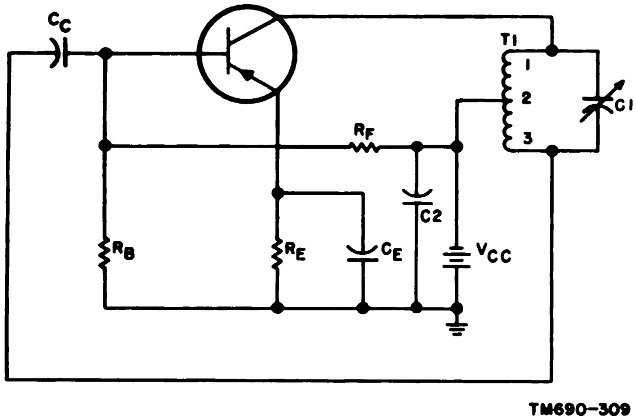


Figure 172. Transistor Hartley oscillator, series-fed.

places terminal 2 of transformer $T1$ at ac ground potential.

c. Push-Pull Oscillator (fig. 173). The oscillator circuit illustrated in figure 173 is similar to the Hartley-type oscillator circuit of figure 171, which has been modified for push-pull operation. Push-pull operation of transistors Q_1 and Q_2 provides greater output than the single-ended transistor oscillator.

(1) Base bias for this circuit is established by resistors R_B and R_F. Capacitor C_B provides an ac bypass around emitter swamping resistor R_E.

(2) Frequency of oscillation is essentially determined by the tank circuit consisting of transformer winding 4–6 and capacitor $C2$. The regenerative signal is applied between base and emitter of each transistor by means of the induced voltages in transformer windings 1–3 and 4–6. Circuit operation, after the feedback signal is applied to transformer winding 1–3, is similar to that of the push-pull amplifier (par. 132). Capacitor $C1$ places terminal 2 of transformer $T1$ at ac ground potential through capacitor C_E.

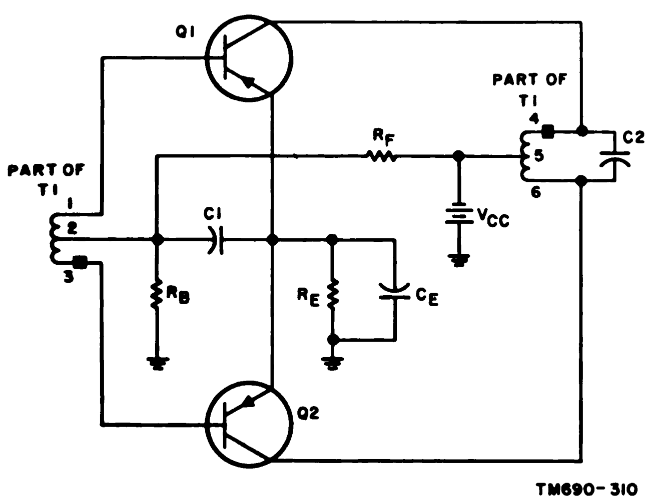


Figure 173. Transistor push-pull oscillator.

177. Crystal Oscillators

Three transistor crystal oscillator circuits are described in *b* and *c* below. A quartz crystal (*a* below), is normally used in an oscillator circuit because of its extremely high Q (narrow bandwidth) and good frequency stability over a given temperature range. In practical applications when the range of operating temperature or frequency tolerance is extremely critical, a constant temperature chamber is used to house the crystal.

a. Crystal Characteristics. The quartz crystal and its equivalent circuit are shown in figure 174.

(1) Resistance R, inductance L, and capacitance C_S in series (B, fig. 174) represent the electrical equivalent of the mechanical vibrating characteristic of the crystal. At series resonance of capacitance C_S and inductance L, the impedance of the crystal is low and the resonance frequency of the oscillator circuit is determined only by the mechanical vibrating characteristics of the crystal.

(2) Parallel capacitance C_P represents the electrostatic capacitance between the crystal electrodes (A, fig. 174). Above the frequency of series resonance ((1) above), the inductive reactance of inductance L is greater than the capacitive reactance of capacitance C_S. The combination (L and C_S) then appears as a net inductance. This net inductance forms a parallel resonant (antiresonant) circuit with capacitance C_P and any circuit capacitance appearing across the crystal. In this instance, the impedance of the crystal is high and the resonant frequency is determined by the crystal and by externally connected circuit elements.

(3) Figure 175 illustrates the impedance versus frequency characteristics of a typical quartz crystal. The lowest impedance occurs at the series resonant frequency (f_1). The highest impedance occurs at the parallel resonant frequency (f_2). The slopes of both curves are steep, indicating a high Q. For most crystals, the difference in frequency be-

tween f_1 and f_2 is very small compared to the series resonant frequency of the crystal.

(4) Either the series or parallel mode of oscillation of the crystal may be used in an oscillator circuit. The mode of operation is primarily determined by the impedance of the circuit to which the crystal is connected.

b. Tickler-Coil Feedback Oscillator (fig. 176). This circuit uses the series mode of operation of the crystal and functions similarly to the tuned-collector circuit of figure 167. Increased frequency stability is obtained with the insertion of

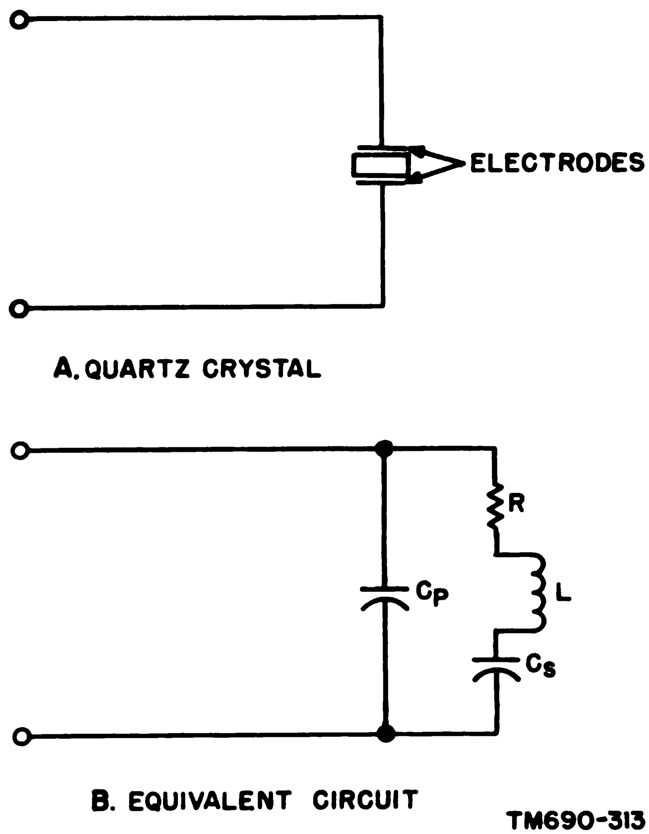


Figure 174. Quartz crystal and its equivalent circuit.

a crystal in the feedback path. However, the frequency is essentially fixed by the crystal. To obtain a variable frequency oscillator, different crystals may be inserted in the circuit while making $C1$ variable over a range of frequencies.

(1) Regenerative feedback from collector to base is accomplished through the mutual inductance between transformer windings 1–2 and 3–4. This provides the necessary 180° phase shift for the feedback signal.

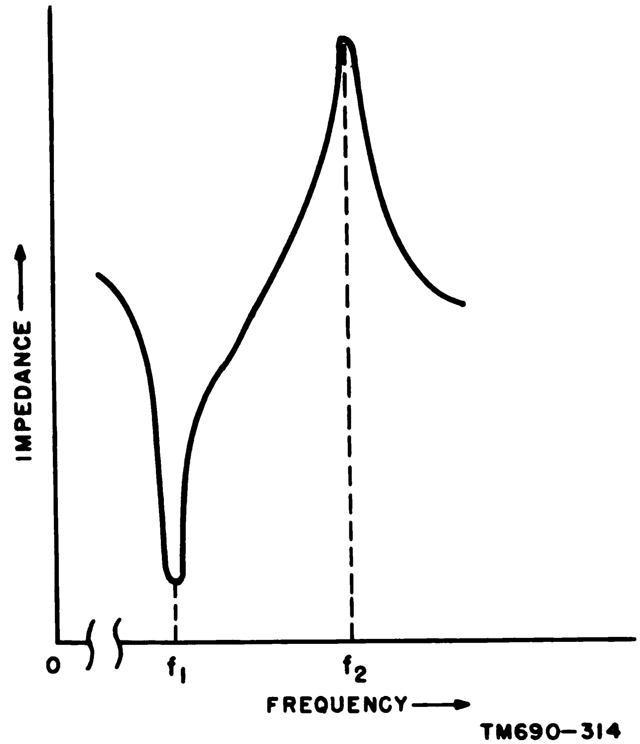


Figure 175. Quartz crystal impedance characteristic.

Resistors R_B, R_F, and R_C, provide the base and collector bias. Emitter swamping resistor R_E is ac bypassed by capacitor C_E.

(2) At frequencies above and below the series resonant frequency of the crystal, the impedance of the crystal increases and reduces the amount of feedback. This in turn prevents oscillation at frequencies other than the series resonant frequency.

c. Colpitts-Type Crystal Oscillators. The common-base (fig. 177) and the common-emitter (fig.

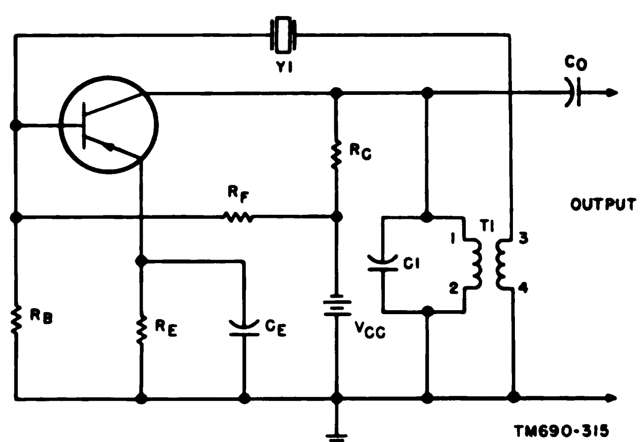


Figure 176. Crystal-controlled tickler-coil feedback oscillator.

178) Pierce oscillators use the parallel mode of resonance of the crystal. If the crystal were to be replaced by its equivalent circuit (B, fig. 174) the functioning of each circuit would become analogous to that of the Colpitts oscillator (fig. 168).

(1) The circuit of figure 177 shows the common-base configuration with the feedback supplied from collector to emitter through capacitor $C1$. Resistors R_B, R_C, and R_F provide the proper bias and conditions for the circuit. Resistor R_B is the emitter swamping resistor. Capacitors $C1$ and C_B form a voltage divider connected across the output. Capacitor $C2$ is an ac bypass around base biasing resistor R_B. Since no phase shift occurs in this configuration, the feedback signal must be connected so that the voltage across capacitor C_B will be returned to the emitter with no phase shift occurring (A, fig. 170). The oscillating frequency of this circuit is determined not only by the crystal but by the parallel capacitance offered by capacitors $C1$ and C_B. These are normally made large to swamp both the input and output capacitances of the transistor and make the oscillations comparatively independent of changes in transistor parameters. Since the parallel capacitance of $C1$ and C_B affects the oscillator frequency, the operation of the crystal is in the inductive region of the impedance versus frequency characteristic between the series and parallel resonant frequencies (fig. 175).

(2) The circuit of figure 178 shows the common-emitter configuration with the feedback supplied from collector to base. The resistances illustrated provide the proper bias and stabilizing conditions for the circuit. Capacitors $C1$ and $C2$ form the voltage divider for this circuit. Capacitor C_B is the emitter bypass capacitor. In contrast to the circuit described in (1) above, a 180° phase shift exists between input and output signals. This is accomplished through the arrangement of the voltage divider network (B, fig. 170). Note that the connection between the capacitors is grounded. The voltage developed across capacitor $C1$ is applied between base and ground providing the 180° phase reversal necessary. The oscillating frequency of this circuit is determined as in (1) above, by the crystal and the capacitors connected in parallel with it.

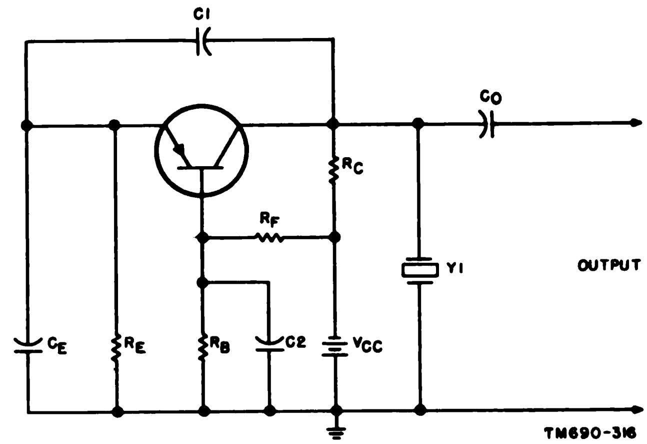


Figure 177. Colpitts-type crystal oscillator, collector-emitter regeneration.

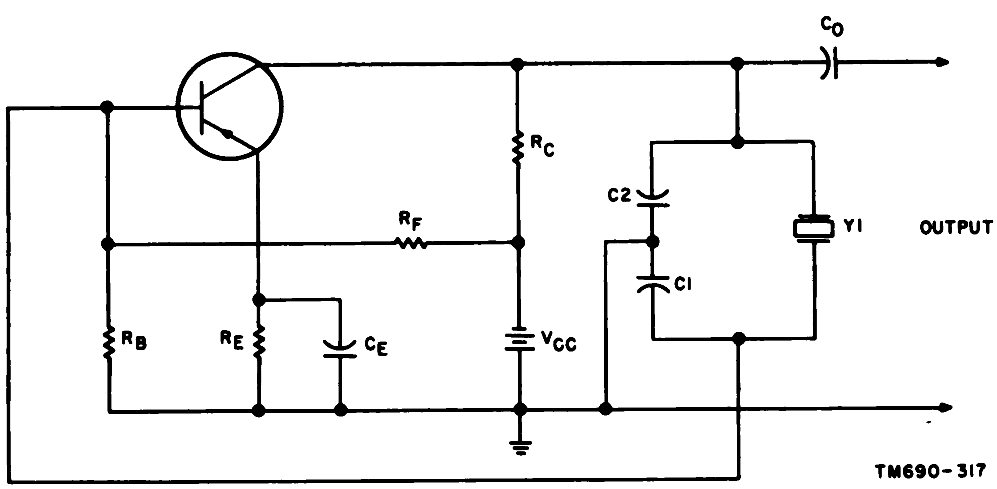


Figure 178. Colpitts-type crystal oscillator, collector-base regeneration.

178. Phase-Shift Oscillators

A sinewave output may be obtained from an oscillator using an RC network in place of the inductance-capacitance network. The RC network determines the frequency of oscillation and provides regenerative feedback between output and input (fig. 179). Since, in the common-emitter configuration, the signal between base and collector is reversed 180° inphase, an additional 180° phase shift is necessary to make the feedback signal positive when returned from output to input. This is accomplished by an RC network consisting of three sections, each contributing a 60° phase shift at the frequency of oscillation. In figure 179, the three RC sections are comprised of C1R1, C2R2, and C3RB.

a. RC Section Analysis.

(1) Current flow in a circuit, consisting of resistance and capacitance in series, is determined by the applied voltage divided by the series impedance of the components (Ohm's law). Since the impedance of an RC circuit is capacitive, current leads the applied voltage by a specific phase angle. This phase angle is determined by the numerical relationship of resistance and capacitance. The voltage drop across the resistor, determined by the current flow through it, therefore leads the applied voltage by the afore-mentioned phase angle.

(2) If the capacitance is fixed, a variation in the resistance value will cause a variation in the phase angle. When the resistance becomes equal to zero, a maximum phase angle of 90° will exist between applied voltage and current. This, however, is of no value since no voltage is developed across a zero resistance, and the only voltage across the capacitor will be the applied voltage. With a minimum of resistance in the section, the phase angle will be slightly less than 90°. Thus, at least three *RC* sections are required to provide the required 180° phase shift necessary for positive feedback. The value of resistance is generally chosen so that each section will provide a 60° phase shift.

Note. A more detailed analysis of the RC section is given in TM 11-670.

b. Circuit Operation. Resistors R_B, R_F, and R_C provide base and collector bias for the oscillator. Capacitor C_E ac bypasses emitter swamping resistor R_E. Capacitors $C1$, $C2$, and $C3$ and resistors $R1$, $R2$, and R_B constitute the feedback and phase-shifting networks.

(1) Oscillations are started by any random noise, or transistor noise, when power is applied. A change in the flow of base current results in an amplified change in collector current, phase shifted 180°. The signal returned to the base is inverted 180° by the action of the phase shifter (*b* above), making it regenerative.

(2) The output waveform is very nearly sinusoidal; the output frequency is fixed. That is, with fixed resistance and capacitance in the phase-shifting network, the 180° phase shift occurs at only one frequency. At other frequencies, the capactive reactance increases or decreases, causing a variation in phase relationship, and the feedback is degenerative (not in phase), preventing sustained oscillations. The phase-shift oscillator is normally made variable over particular ranges by providing ganged variable capacitors or resistors in the phase shifter.

(3) A high gain transistor must be used with the three-section phase-shifting network because the losses incurred in the network are high. Increasing the number of phase-shifting *sections* causes a reduction in the loss of the overall network. This is so, since additional sections reduce the phase shift necessary for each section and the loss for each section is lowered as the phase shift is reduced. It is therefore not uncommon to encounter a phase-shift oscillator which employs more than three sections in the phase-shifting network.

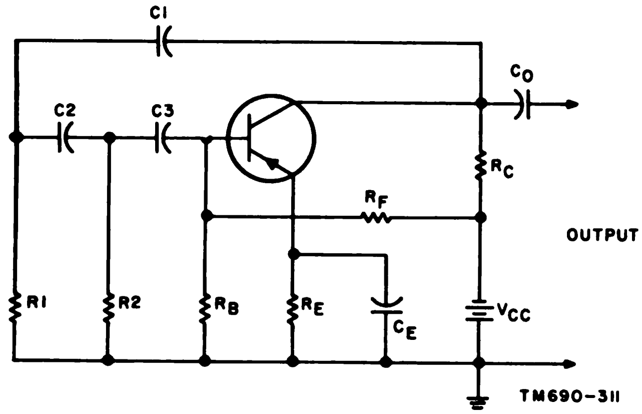


Figure 179. Transistor RC phase-shift oscillator.

179. Wien-Bridge-Type Oscillator

(fig. 180)

Another circuit which uses a resistance-capacitance network for the development of a sinusoidal output is the Wien-bridge-type oscillator. The circuit illustrated, using two transistors connected in the common-emitter configuration, is directly analogous to the type employing electron tubes. The second stage functions as an amplifier and phase inverter from which the feedback signal is taken in proper phase. As in the case of the electron tube equivalent circuit, regeneration is provided for oscillation and degeneration is provided to obtain frequency stability and distortionless output.

a. Bridge Analysis.

(1) The feedback voltage developed between the collector of transistor $Q2$ and ground is impressed across the entire bridge network (B, fig. 180). The voltage developed across capacitor $C2$ is regenerative and is applied to the input circuit of transistor $Q1$. These voltages are in phase at only one frequency. This is due to the phase relationships between the voltages developed across the series combination of resistor $R1$ and capacitor $C1$ and the parallel combination of resistor R_{B1} and capacitor $C2$. These vary in opposite directions as the frequency changes. At the resonant frequency, the phase shift is zero. At frequencies above and below the resonant frequency, a phase shift exists and the positive feedback voltage is reduced in value.

(2) Besides leaving the emitter resistors of both stages unbypassed, providing negative feedback, additional negative feedback is provided through resistor R2 and applied to the emitter of transistor Q1. The amount of negative feedback is a fixed quantity since the feedback is developed across resistors. This quantity is determined primarily by resistor R2 (which may be made variable to control output amplitude). Normally, its value provides greater negative feedback voltage at other than the resonant frequency. Therefore, at other frequencies, the negative feedback exceeds the positive feedback and a highly stable oscillator results.

(3) The frequency of oscillation is determined by capacitors $C1$ and $C2$ and resistors $R1$ and R_{B1}. The resonant frequency (f_r) is obtained from the following expression:

$$f_r = \frac{1}{2\pi\sqrt{R1C1R_{B1}C2}}$$

However, if $R1$ is equal to R_{B1} and $C1$ is equal to $C2$, the expression is reduced to

$$f_r = \frac{1}{2\pi R1C1}$$

When a variable frequency oscillator is desired, either capacitors $C1$ and $C2$ may be variable or resistors $R1$ and R_{B1} may be made variable in steps.

(4) The emitter resistance of transistor Q1 is a thermistor, with a positive temperature coefficient of resistance. In some circuits, a lamp is substituted for the thermistor. In either case, its function is to stabilize the output amplitude of the oscillator. When the output increases for any reason, an increased feedback voltage results. As a result of the increased feedback voltage, the current through the thermistor increases, increasing the resistance of the thermistor which in turn increases the negative feedback voltage developed across it. This reduces the gain of the amplifier and effectively returns the output voltage to what it had been. Thus, additional amplitude stabilization is provided. In cases where this stability is not a requirement, an ordinary resistor is used in series with the emitter electrode.

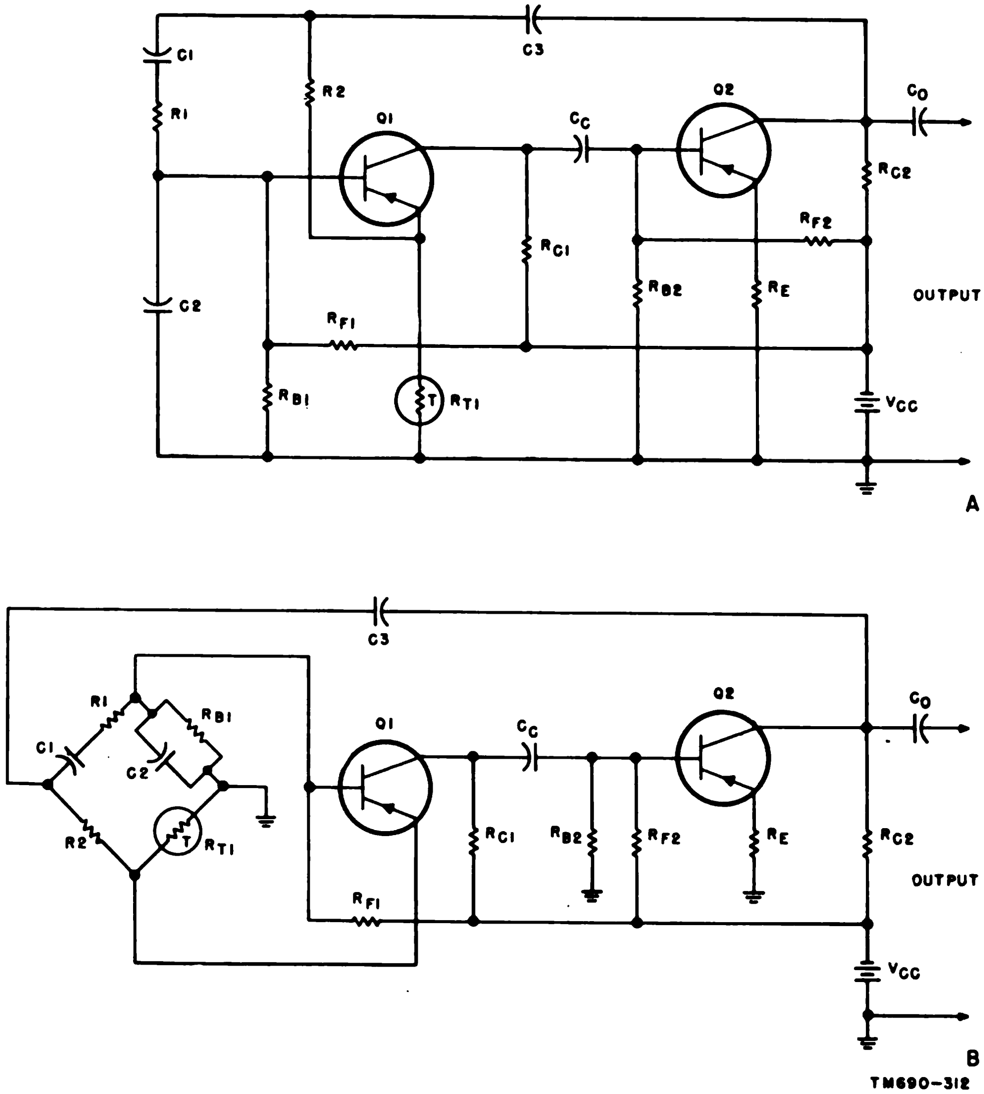


Figure 180. Transistor Wien-bridge oscillator.

b. Circuit Description. The circuits of A and B, figure 180, are identical. A, figure 180, is arranged so that the feedback paths are easily identified, while B of the figure is arranged so that the frequency selective bridge components are easily identified.

(1) Both transistors use the same methods of biasing and stabilization. Resistors R_{B1}, R_{C1}, and R_{F1}, and thermistor $RT1$ establish the bias in the amplifier circuit of transistor $Q1$. Thermistor $RT1$, also serves as the emitter swamping resistor. Resistors R_{B2}, R_{C2}, R_E, and R_{F2} provide the bias in the amplifier circuit of transistor $Q2$. Resistor R_B is left unbypassed, providing degeneration and increased output waveform stability. Dc power for both stages is provided by battery V_{CC}.

(2) The output of transistor $Q1$ is coupled to the input of transistor $Q2$ through capacitor C_C. Capacitor $C3$ couples a portion of the output of amplifier stage $Q2$ to the bridge network to provide the necessary feedback (both positive and negative (a above)). The output to the load is coupled through capacitor C_O.

Section IV. FREE-RUNNING NONSINUSOIDAL OSCILLATORS

180. Relaxation Oscillators, General

a. An oscillator circuit in which the output waveform is nonsinusoidal is generally classified as a relaxation oscillator. The relaxation oscillator uses a regenerative circuit in conjunction with RC or resistance-inductance (RL) components to provide a switching action. The charge and discharge times of the reactive elements are used to produce sawtooth, square, or pulse output waveforms.

b. Multivibrator and blocking oscillators are examples of relaxation oscillators. Multivibrator (par. 181) and blocking oscillator (par. 183) circuits generally use an RC time constant for determination of output waveform and frequency. A form of multivibrator circuit, used primarily to convert low voltage dc to high voltage ac for rectification in dc to dc power supplies, is described in paragraph 182.

c. Multivibrators and blocking oscillators are further classified as either free-running or driven (triggered) oscillators. The free-running oscillator is one in which the oscillations begin once power is applied to the circuit. The oscillations then become continuous while the circuit is in operation. The triggered oscillator (pars. 185–193) is controlled by a synchronizing or triggering external signal.

d. The following properties of a transistor amplifier circuit should be considered in the analysis of multivibrator and blocking oscillator circuits. These properties are applied to the common-emitter configuration which is used because a 180° phase shift occurs, as in an electron tube amplifier, between the output and input signals.

(1) An increase in base current causes an increase in collector current through the transistor; conversely, a decrease in base current causes a decrease in collector current.

(2) An increase in collector current causes the collector voltage to decrease; a decrease in collector current causes the collector voltage to increase towards the value of the source voltage. With PNP-type transistors, since the collector is reverse-biased, an increase in collector current causes the collector voltage to become less negative (or more positive). With NPN-type transistors, the collector voltage becomes less positive (or more negative).

(3) For normal functioning of a transistor amplifier, the base-emitter diode is forward-biased and the collector-base diode is reversed-biased. The polarity is determined by type of transistor used (PNP or NPN).

(4) A transistor is saturated when a further increase in base current causes no further increase in collector current.

(5) A transistor is cut off when either the base voltage polarity causes reverse bias or the collector voltage polarity causes forward bias.

(6) Capacitors and inductors require a definite amount of time to charge or discharge through a resistor. The measure of this time, called the time constant, is determined by multiplying the resistance by the capacitance ($R \times C$) or by dividing the inductance by the resistance ($L \div R$).

181. Multivibrators

The free-running (astable) multivibrator (par. 190*a*) is essentially a nonsinusoidal two-stage oscillator in which one stage conducts while the other is cut off until a point is reached at which the stages reverse their conditions. That is, the stage which had been conducting cuts off, and the stage that had been cut off conducts. This oscillating process is normally used to produce a square wave output. Most transistor multivibrator circuits are counterparts of those using electron tubes. For example, the emitter-coupled transistor multivibrator is analogous to the cathode-coupled electron tube multivibrator circuit; the collector-coupled transistor multivibrator (fig. 181) is analogous to the common plate-coupled electron tube multivibrator. Since most multivibrator circuits function similarly, only the collector-coupled transistor multivibrator is discussed in *a* and *b* below.

a. General. The basic collector-coupled transistor multivibrator of figure 181, is a two-stage resistance-capacitance coupled common-emitter amplifier with the output of the first stage coupled to the input of the second stage and the output of the second stage coupled to the input of the first stage. Since the signal in the collector circuit of a common-emitter amplifier is reversed in phase with respect to the input of that stage, a portion of the output of each stage is fed to the other stage in phase with the signal on the base electrode. This regenerative feedback with amplification is required for oscillation. Bias and stabilization are established identically for both transistors.

b. Circuit Description (figs. 181 and 182).

(1) Because of the variation in tolerances of the components, one transistor will conduct before the other or will conduct more heavily than the other.

(2) Assuming transistor $Q1$ is conducting more heavily than transistor $Q2$, more current i_{b1} will flow in the base circuit of transistor $Q1$ than in the base circuit of $Q2$. Collector current i_{c1} in transistor $Q1$ increases rapidly, causing collector voltage V_{c1} (fig. 182), and voltage at the junction of resistors R_{C1} and R_{F1} to decrease (become more positive). This increasing positive voltage is applied through capacitor C_{F1} to the base of transistor $Q2$.

(3) As base voltage v_{b2} of transistor $Q2$ becomes more positive, the forward bias decreases, resulting in a rapid decrease in base current i_{b2} and collector current i_{c2} in transistor $Q2$. Collector voltage V_{c2}, and thus the voltage at the junction of resistors R_{C2} and R_{F2} becomes more negative. This negatively increasing voltage is fed back through capacitor C_{F2} to the base of transistor $Q1$, increasing the forward bias.

(4) The process in (1) and (2) above continues until a point is reached where base voltage v_{b2} of transistor $Q2$ is made so positive with respect to the emitter that transistor $Q2$ is cut off (reverse bias is applied) and transistor $Q1$ is saturated (total dc voltage V_{CC} appears across resistor R_{C1}). That is, the current through transistor $Q1$ increases steadily as the current through transistor $Q2$ decreases steadily until transistor $Q2$ is cut off. Point A, figure 182, represents this action. This entire action happens so quickly that capacitor C_{F1} does not get a chance to discharge and the increased positive voltage at the collector of transistor $Q1$ appears entirely across resistor R_{B2}.

(5) During the period from A to B, figure 182, collector current i_{c1} and collector voltage v_{c1} remain constant and capacitor C_{F1} discharges through resistor R_{F1}. As capacitor C_{F1} discharges, more of the previously increased positive voltage at the collector of transistor $Q1$ appears across capacitor C_{F1} and less across resistor R_{B2}. This decreases the reverse bias on the base of transistor $Q2$. This action continues until the time at point B, figure 182, is reached, and forward bias is reestablished across the base-emitter diode of transistor $Q2$.

(6) Transistor $Q2$ conducts. As collector current i_{c2} in transistor $Q2$ increases, the collector voltage V_{c2} becomes less negative or more positive. This voltage, coupled through capacitor C_{F2} to the base of transistor $Q1$, drives it more positive and causes a decrease in current flow through transistor $Q1$. The resulting increased negative voltage at the collector of transistor $Q1$ is coupled through capacitor C_{F1} and appears across resistor R_{B2}. The collector current of transistor $Q2$ therefore increases. This process continues rapidly until transistor $Q1$ is cut off. Transistor $Q1$ remains cut off (and transistor $Q2$ conducts) until capacitor C_{F2} discharges through resistor R_{F2} enough to decrease the reverse bias on the base of transistor $Q1$ (C, fig. 182). The cycle is repeated ((1) above).

(7) The oscillating frequency of the multivibrator is usually determined by the values of resistance and capacitance in the circuit. In the collector-coupled multivibrator of figure 181, collector loads are provided by resistors R_{C1} and R_{C2}. Base bias for transistor $Q1$ is established through voltage divider resistors R_{B1} and R_{F2}. Base bias for transistor $Q2$ is established through voltage divider resistors R_{F1} and R_{B2}. Stabilization is obtained with emitter swamping resistor R_{E1} for

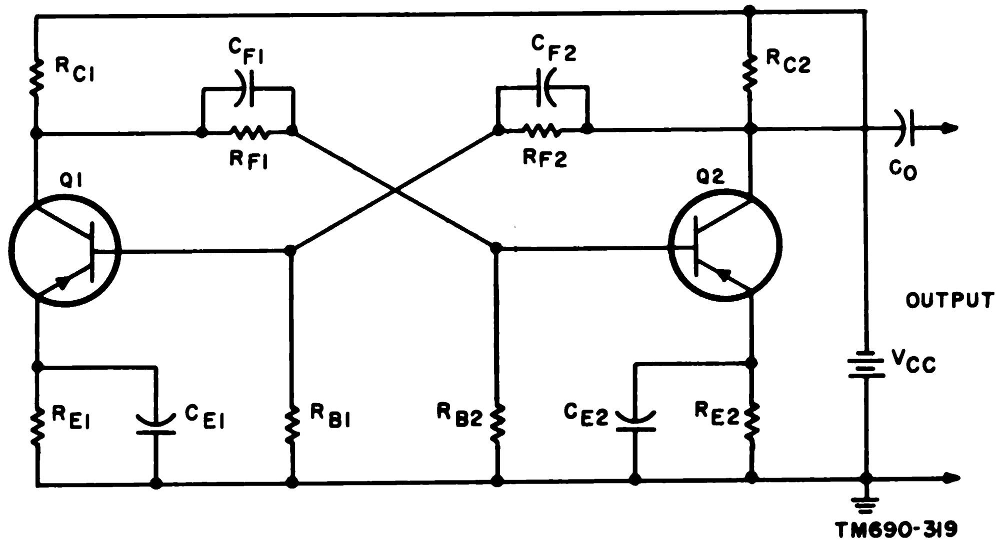


Figure 181. Transistor multivibrator.

transistor $Q1$, and resistor R_{E2} for transistor $Q2$. Emitter capacitors C_{E1} and C_{E2} are ac bypass capacitors.

(8) The output signal is coupled through capacitor C_O to the load. This output waveform, which is essentially square, may be obtained from either collector. To have a sawtooth output, a capacitor is usually connected from collector to ground for development of the output voltage.

Note. The multivibrator may be modified to produce a sinusoidal output wave. This is accomplished through the connection of a parallel-tuned circuit between the base electrodes of each transistor.

182. Saturable-Core Square Wave Oscillator

Transistors, functioning as high-speed switching elements, used in conjunction with transformers, the core materials of which exhibit rectangular hysteresis loops, provide a form of multivibrator circuit, designed to produce a square wave output. This type of circuit (fig. 184) operates with greater efficiency and at higher frequencies than the common mechanical vibrator (*a* below) normally found in portable power supplies. The transistors, functioning as switches or relays, operate from cutoff to saturation (*b* below). The square wave output of this type of circuit may be stepped up or down and rectified, providing a dc voltage higher or lower than the available source. When the circuit is used with a full-wave rectifier, almost no filtering is required to eliminate ripple, because of the square wave output.

a. General. Fundamental switching action such as occurs with a mechanical vibrator, is shown in A, figure 183. The vibrator contacts are simulated by ganged switches $S101$ and $S102$. The action is such that when one switch is opened, the other switch is closed. With switch $S101$ closed, battery voltage V_{CC} is applied across primary winding 1–2 of transformer $T1$. The polarities of the voltages induced in primary winding 3–4 and secondary winding 5–6 are as indicated (B, fig. 183). When the switches are reversed, battery voltage V_{CC} is applied across primary winding 3–4 with polarity reversed. Similarly, the voltage developed across secondary winding 5–6 is reversed. When switching is accomplished quickly, the voltage waveform (V_{RL}) across load resistor R_L is essentially square.

b. Circuit Description (fig. 184). Operation of the inverter circuit basically depends on the switching of the transistors. That is, when transistor $Q1$ conducts heavily, it may be compared to closing switch $S101$ (A, fig. 183). Driving current is applied to the base of the transistor. Simultaneously, transistor $Q2$, is at cutoff, being comparable to opening switch $S102$. No driving current is applied. The transition from cutoff to heavy conduction and vice versa is extremely rapid to prevent transistor burnout and to provide a square wave output. At cutoff, the voltage across the transistor is high but current flow is low (effectively zero). During heavy conduction, cur-

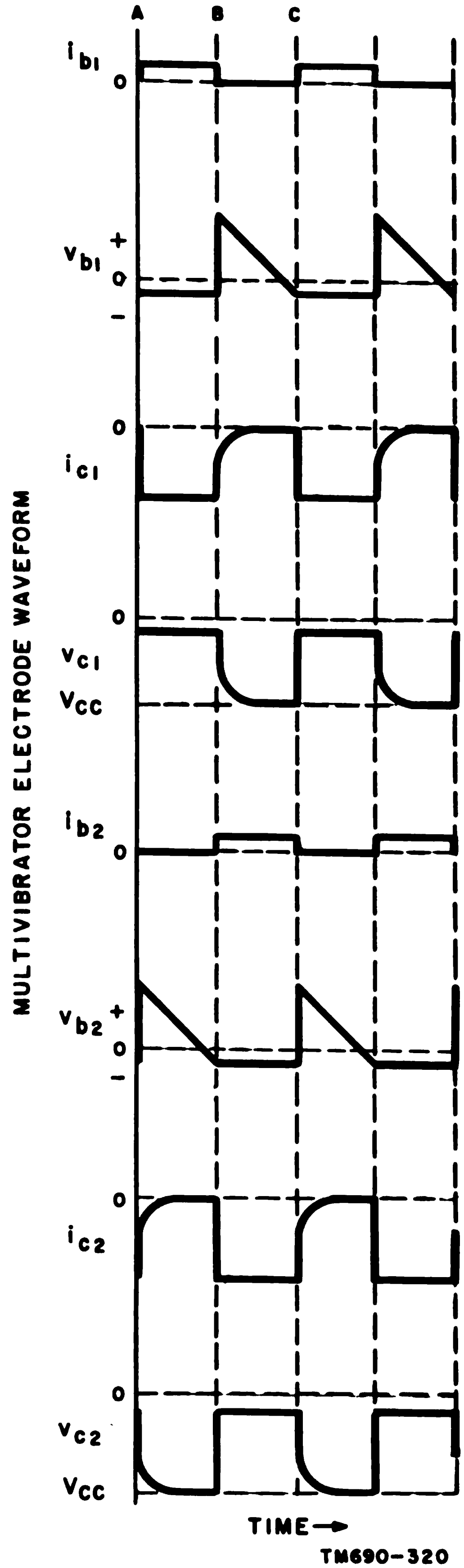


Figure 182. Multivibrator waveforms.

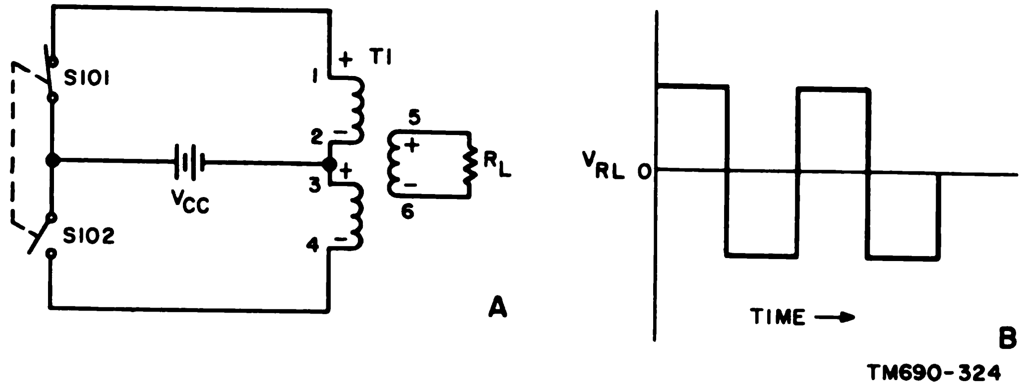


Figure 183. Mechanical switching circuit.

rent flow is high but the voltage across the transistor is low. The switching time between these points is low, maintaining a low average power dissipation.

Note. **Resistors R_s and R_f establish the initial starting bias required for the bases of transistors Q1 and Q2.**

(1) Starting oscillation of the circuit depends on the unbalance existing between the apparent identical circuits of transistors $Q1$ and $Q2$. This is due to both external circuitry and transistor unbalance. Because of this unbalance, more current will flow in one of the primary windings of transformer $T1$ than in the other.

(2) Assume more current flows through primary transformer winding 3–4 and the starting point on the hysteresis curve (B, fig. 184) is at point F (negative saturation region of flux in the transformer core). Transistor $Q1$ conducts, and the core flux moves from point F toward point B (positive saturation region) inducing a positive voltage in transformer windings 1–2, 4–5, and 6–7. Since current flow in the collector circuit is from the negative terminal of source voltage V_{CC}, the induced voltages will be positive at the top of each winding as indicated by the polarity markings of transformer $T1$. Thus, the base voltage of transistor $Q1$ is made more negative, driving it to heavy conduction while the base voltage of transistor $Q2$ is made more positive, driving it to cutoff.

(3) The flux in the core changes at a relatively constant rate until positive saturation is reached (point B). At core saturation, further increase in current does not increase the flux in the core. Since the flux remains comparatively constant, no voltage is induced in windings 1–2, 4–5, and 6–7. This condition corresponds to point C on the hysteresis curve. The induced voltages quickly fall to zero, removing the base driving current from transistor $Q1$ which makes it nonducting. Current flow in the collector circuit and through winding 3–4 stops.

(4) With the removal of current flow in winding 3–4, the flux in the core falls from the value at point C towards the value at point A. Due to this small decrease of flux, voltages of opposite polarity to that which previously existed ((2) above) are induced in all transformer windings. Transistor Q1 is driven further into cutoff by the positive potential on the base from feedback winding 1–2; transistor Q2 is driven to conduction by the negative potential on its base from feedback winding 6–7. The flux in the core, therefore is being driven from point A toward negative saturation at point E. During this period, transistor Q1 is maintained at cutoff while transistor Q2 conducts heavily. Once negative saturation of the core is reached, the switching of transistor Q1 from cutoff to conduction and transistor Q2 from conduction to cutoff occurs. The cycle is ready for repetition.

(5) The output voltage developed across winding 8–9 is essentially a square wave. This is so because the core flux changes at a relatively constant rate from points F to B and from A to E. The output

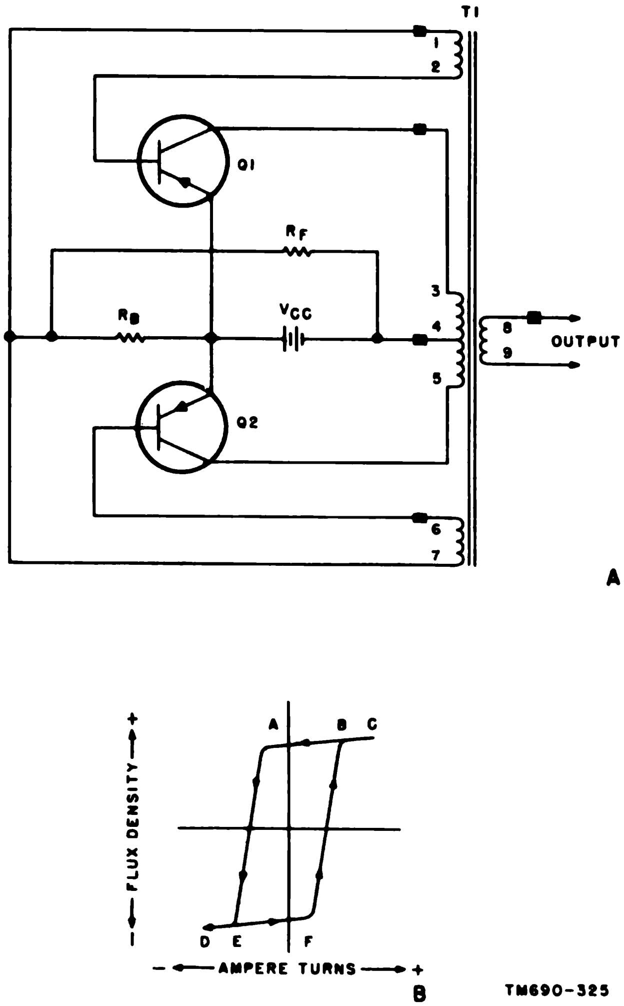


Figure 184. Saturable-core square wave oscillator.

frequency and voltage are determined by the turns ratio of the primary and secondary windings of transformer T1 and the saturation flux of the core.

183. Blocking Oscillators

a. A blocking oscillator conducts for a short period of time and is cut off (blocked) for a much longer period. Blocking oscillators may be either *free-running* or *triggered* (par. 190). Only the free-running type is discussed in this paragraph.

b. A basic circuit for the blocking oscillator is shown in figure 185. The similarity between this circuit and that of the tickler-coil oscillator of figure 166 should be noted.

(1) When the circuit is energized, current rises rapidly in the base due to the forward bias established between base and emitter by dc power supply V_{CC}. Current flow in the collector circuit increases in accordance with this base driving current. Due to the increasing collector current, a voltage which becomes more negative is induced in transformer winding 1–2. This voltage charges capacitor C_F through the small forward resistance of the base-emitter diode, and appears across this resistance increasing the forward bias. Regeneration continues rapidly until the transistor is saturated.

(2) At saturation, the collector current becomes constant. Therefore, there is no longer any induced voltage in winding 1–2 and a charging potential is no longer applied across capacitor C_F. Capacitor C_F begins discharging through resistor R_F. The field around winding 1–2 collapses and induces a voltage in the winding in the reverse direction. That is, the base is driven positive (reverse bias). Base and collector current fall to zero. The transistor is held at cutoff until capacitor C_F, discharging through resistor R_F, reaches the point at which the transistor is forward-biased and conduction begins again. The cycle is then repeated.

(3) The output waveform (not shown) is a pulse, the width of which is primarily determined by winding 1–2. The time between pulses (resting or blocking time) is determined by the time constant of resistor R_F and capacitor C_F. The output is coupled through transformer winding 3–4 to the load by transformer winding 5–6.

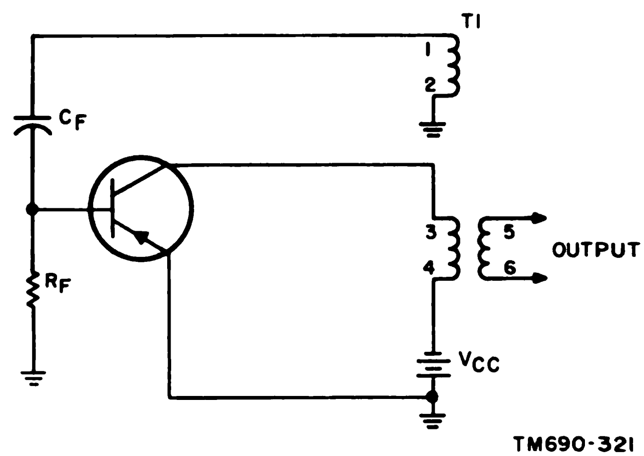


Figure 185. Blocking oscillator.

184. Summary

a. An oscillator is an amplifier with feedback of proper phase and amplitude. Transistors may be used in circuits similar to electron tube oscillator circuits.

b. Power losses occur in the feedback and the frequency determining networks. The regenerative feedback must overcome these losses to permit sustained oscillation in the circuit. The amplitude of the oscillations is limited by the normal operation of the amplifier, the amount of feedback energy, and the nonlinear conditions of the circuit.

c. Most transistor circuit parameters are changed if the operating point is changed, thus affecting the frequency of oscillation. Temperature stabilization is also a prime factor in the stabilization of operating point and frequency.

d. Bias and stabilization techniques for the transistor oscillator circuit are essentially the same as those for the transistor amplifier.

e. The regenerative feedback signal must undergo a 180° phase shift when feedback is accomplished from the collector to base. No phase shift is required when the feedback is from collector to emitter or from emitter to base.

f. When feedback is obtained by simple transformer coupling, the oscillator is called a tickler-coil oscillator. Tickler-coil oscillators may take the form of a *tuned-collector oscillator* or a *tuned-base oscillator*.

g. As in its electron tube counterpart the Colpitts transistor oscillator uses a split capacitor in the tuned circuit. A modification of the Colpitts transistor oscillator which provides increased frequency stability is called the *Clapp* oscillator.

h. The Hartley oscillator is similar to the Colpitts oscillator except that a tapped coil instead of a split capacitor is used.

i. A quartz crystal may be operated as either a series or parallel resonant circuit. Either mode may be used in an oscillator circuit.

j. Phase-shift oscillators use an RC network to provide the necessary regeneration. The output waveform is very nearly sinusoidal.

k. The Wien-bridge-type oscillator is a form of phase-shift oscillator with increased amplitude and frequency stability.

l. Oscillators, such as the multivibrator and the blocking oscillator which produce nonsinusoidal output waveforms, are called relaxation oscillators.

m. The collector-coupled transistor multivibrator is similar to the plate-coupled electron tube multivibrator.

n. The saturable-core square wave oscillator has higher efficiency and operates at higher frequencies than mechanical vibrators found in most portable equipments. The near square wave output when used in a full-wave rectifier circuit requires minimum filtering.

CHAPTER 11

PULSE AND SWITCHING CIRCUITS

Section I. TRANSISTOR SWITCHING CHARACTERISTICS

185. General

a. Circuit Applications. Pulse and switching circuits are used in the radar, television, telemetering, pulse-code communication, and computing fields. The circuits function as generators, amplifiers, inverters, frequency dividers, and wave shapers providing limiting, triggering, and gating and signal routing functions. Some typical circuits performing these functions are described in succeeding paragraphs.

b. Nonlinear Operation. Pulse and switching circuits are normally characterized by large-signal, or *nonlinear*, operation of the transistor (par. 186). Operation of these circuits is normally governed by the application of a pulse; a discontinuous change in the level of voltage or current is usually experienced by the circuit (*c* below). The input signals (trigger pulses) produce output signals having large and sudden changes in voltage or current. Such nonlinear operation results in an output wave form which may differ considerably from the input wave form (par. 188).

c. Unit Step Voltage.

(1) Pulse wave forms widely encountered in large-signal operation (par. 186) of the transistor are illustrated in figure 186. The changes in voltage levels with time are considered *instantaneous* and represent the theoretical (ideal) pulse. The effect of the transistor on the ideal pulse is given in paragraph 187.

(2) A voltage which undergoes an instantaneous change in amplitude from one constant level to another is called a unit step voltage. In pulse and switching circuit applications when the unit step voltage is the *applied signal*, it is usually of sufficient magnitude to cause the circuit to change from a state of conduction to one of cutoff or vice versa (par. 186). When the unit step voltage is the *result* of an applied signal to a switching circuit, the circuit will have changed its state.

(3) A, figure 186, illustrates a *positive* unit step voltage. At time *t*1, the voltage level is *increased* (positively) by the magnitude *V*. The voltage level does not have to necessarily increase from zero to a positive voltage. If the initial voltage level was at a negative potential and then changed to zero, a positive unit step voltage would be involved. B, figure 186 illustrates a *negative* unit step voltage. At time *t*2, the voltage level is *decreased* (negatively) by the magnitude *V*. In this instance, the change in level could be from a high positive potential to one that is less positive. C, figure 186 illustrates a square or rectangular pulse. At time *t*1, the voltage level is increased by the magnitude *V*. Between time *t*1 and *t*2, a new constant voltage level is established. At time *t*2, the voltage level is decreased by the magnitude *V*. For instance, a square pulse (C, fig. 186) may be referred to as two unit step voltages, one positive and one negative.

Note. The explanation of a *unit step current* is identical with that of unit step voltage. Current levels, rather than voltage levels, undergo instantaneous changes, either positively or negatively.

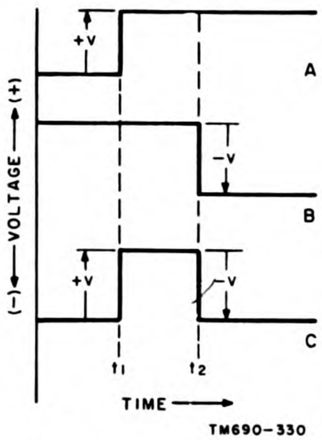


Figure 186. Unit step voltage waveforms, showing formation of a pulse.

186. Large-Signal Operation

In large-signal (nonlinear) operation, the transistor acts as an overdriven amplifier with resultant changes in the conduction state and the dc characteristics of the transistor (*a* below). The overdriven amplifier may be compared to a switch or relay (*b* and *c* below). Advantages and disadvantages of using the different transistor configuration as switches or relays are given in *d* below.

a. Output Characteristics.

(1) The *CE* output characteristics of a typical PNP transistor are shown in A, figure 187. The characteristics are arranged in three regions; *cutoff*, *active*, and *saturation.* An arbitrarily chosen loadline and the maximum permissible power dissipation curve (par. 102) are also shown. The cutoff and saturation regions are considered the *stable* or quiescent regions of operation. A transistor is considered in the *off* (nonconducting) or *on* (conducting) state when it is operated in the cutoff or saturation regions respectively. The third region of operation, referred to as the active region, is considered the *unstable* (transient) region through which operation of the transistor passes while changing from the *off* to *on* state.

(2) The effect of base bias voltage V_{BE} on collector current I_C (*b*(1) below), in the cutoff and active regions is illustrated in B, figure 187.

b. Transistor Switching Circuit. A typical transistor switching circuit, is illustrated in A, figure 188. Switch *S*1 controls the polarity and amount of base current from battery V_{B1} or V_{B2}.

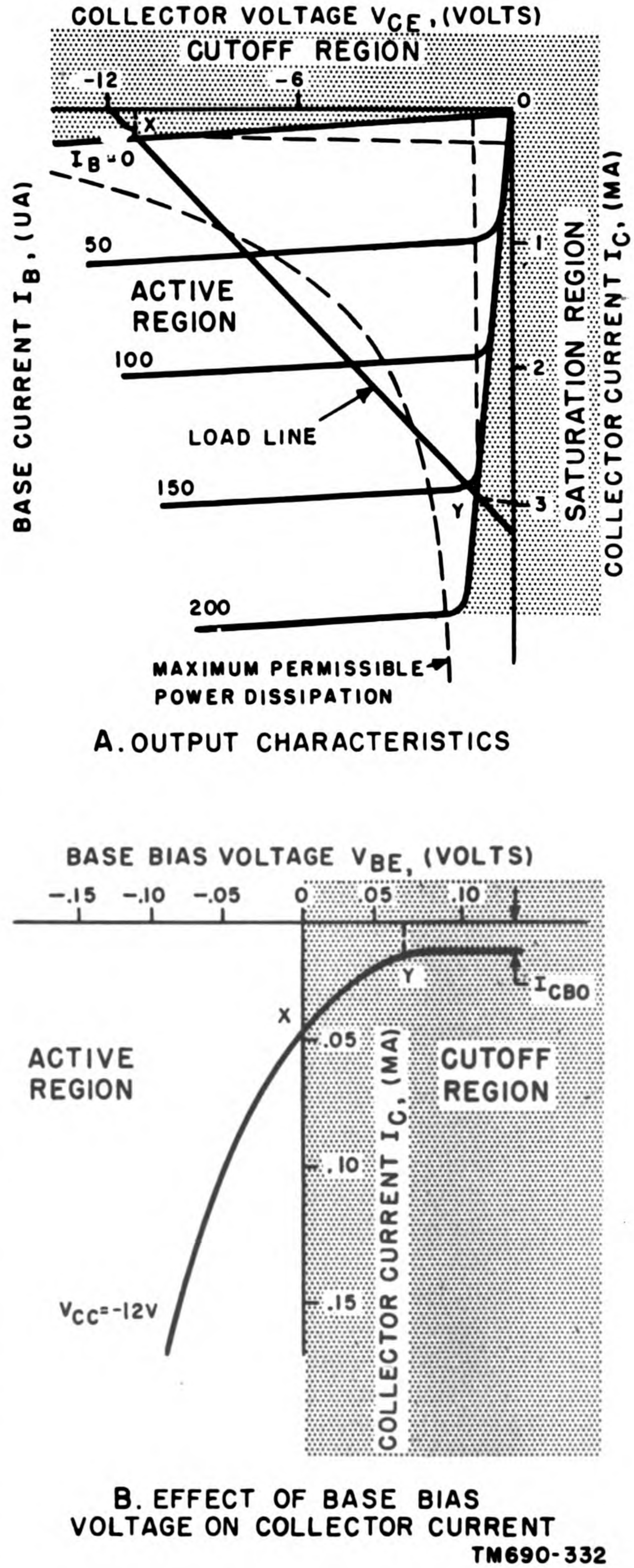


Figure 187. Transistor characteristic curves, switching application.

Resistors R_{B1} and R_{B2} are current-limiting resistors. The emitter-base and collector-base diode and switch equivalent circuits representing the *off* and *on* (dc) conditions of the transistor switching circuit are shown in B and C, figure 188. Each region of operation (a(1) above) is discussed in (1) through (3) below.

(1) *Cutoff region.*

(a) The cutoff region (A, fig. 187) includes the area above the zero base current curve ($I_B=0$). Ideally, with no initial base current, there would be zero collector current; the collector potential would equal battery voltage V_{CC} (A, fig. 188). However, at point X on the loadline (A, fig. 187), a small amount of collector current is measured. (See horizontal projection to collector current axis.) This is more clearly indicated in B, figure 187, where at zero base bias voltage (point X), collector current I_C equals approximately 0.05 ma. This is the reverse bias collector current (par. 43) for the CE configuration. Note that the application of a small reverse base bias voltage V_{BB} (approximately 0.075 volt, positive for a PNP transistor), reduces the value of reverse bias collector current to the value of I_{CBO} (par. 188). The collector voltage V_{CE} is indicated by the vertical projection from point Y (A, fig. 187) to the collector voltage axis. This value is equal to the difference in magnitude between the battery voltage (12 volts in this instance) and the voltage drop produced by reverse bias collector current flow through load resistor R_L (A, fig. 188). Normal quiescent conditions for a transistor switch in this region require that both emitter-base junction and collector-base junction must be reverse biased ((b) below).

(b) With switch $S1$ (A, fig. 188) in the OFF position, the emitter-base junction is reverse biased by battery V_{B2} through resistor R_{B2}. This is comparable to the application of a positive unit step voltage (par. 185c). The collector-base junction is reverse biased by battery V_{CC} through load resistor R_L; the transistor is in the *off* (cutoff) state. In the diode equivalent circuit of the unit step voltage (par. 185c). The col- transistor (B, fig. 188), diodes CR_B and CR_C represent the emitter-base and collector-base junctions, respectively. Diode CR_E is reverse biased by voltage V_{BE}; diode CR_C is reverse biased by voltage V_{CB}. Ideally there is no current flow through load resistor R_L and collector-emitter voltage V_{CE} (output voltage) equals collector battery voltage V_{CC}. The circuit, as a switch, is *open* (C, fig. 188). The initially applied bias causes switches S_1 and S_2 (which represent the diode equivalents (*1*) above) to open the output circuit.

(2) *Active region.*

(a) The active linear region (A, fig. 187) is the only region providing normal amplifier gain. In the linear region, the collector-base junction is reverse biased and the emitter-base junction is forward biased. Transient response of the output signal (par. 187) is essentially determined by transistor characteristics (ch. 4) in this region. In switching circuits, this region represents the transition region.

(b) Operation of switch $S1$ (A, fig. 188) to the ON position is comparable to the application of a negative unit step voltage. Forward bias is established by battery V_{B1}, through resistor R_{B1}, on the emitter-base junction, and the base current I_B and collector current I_C, become transitory in nature, moving from point X (A, fig. 187) on the loadline to point Y; here, collector current reaches saturation ((3) below). The signal passes through this region rapidly. In switching circuits, this region is of importance only for design considerations.

Note. The loadline (A, fig. 187) is shown passing through the area beyond the maximum permissible power dissipation curve. For *normal amplifier operation* of a transistor, this is undesirable. In switching circuits, however, the excursion of the collector current through this area is very rapid, and the *average power dissipated* within the transistor falls within the acceptable maximum limitation (par. 102).

(3) *Saturation region.*

(*a*) In the saturation region (A, fig. 187), an increase in base current *does not* cause an appreciable increase in collector current I_C. At point Y, on the loadline, the transistor is in the saturation region. Collector current I_C (measured by a horizontal projection from point Y) is at a maximum, and collector voltage V_{CE} (measured by a vertical projection from point Y) is at a minimum. This value of collector voltage is referred to as the *saturation voltage* (V_{SAT}), and is an important characteristic of the transistor. Deep saturation is generally avoided because of its effect on the transient response of the transistor (par. 187*c*).

Note. The saturation region is also referred to as the *bottomed* region.

(b) When collector current I_C reaches its limited value (battery voltage V_{CC} (A, fig. 188) divided by the value of load resistance R_L), the transistor saturates and is in the *on*, or conducting state. In the diode equivalent circuit of the transistor in saturation (B, fig. 188), diodes CR_E and CR_C are forward biased. Diode CR_B is forward biased by input voltage V_{BB} ((2) above).

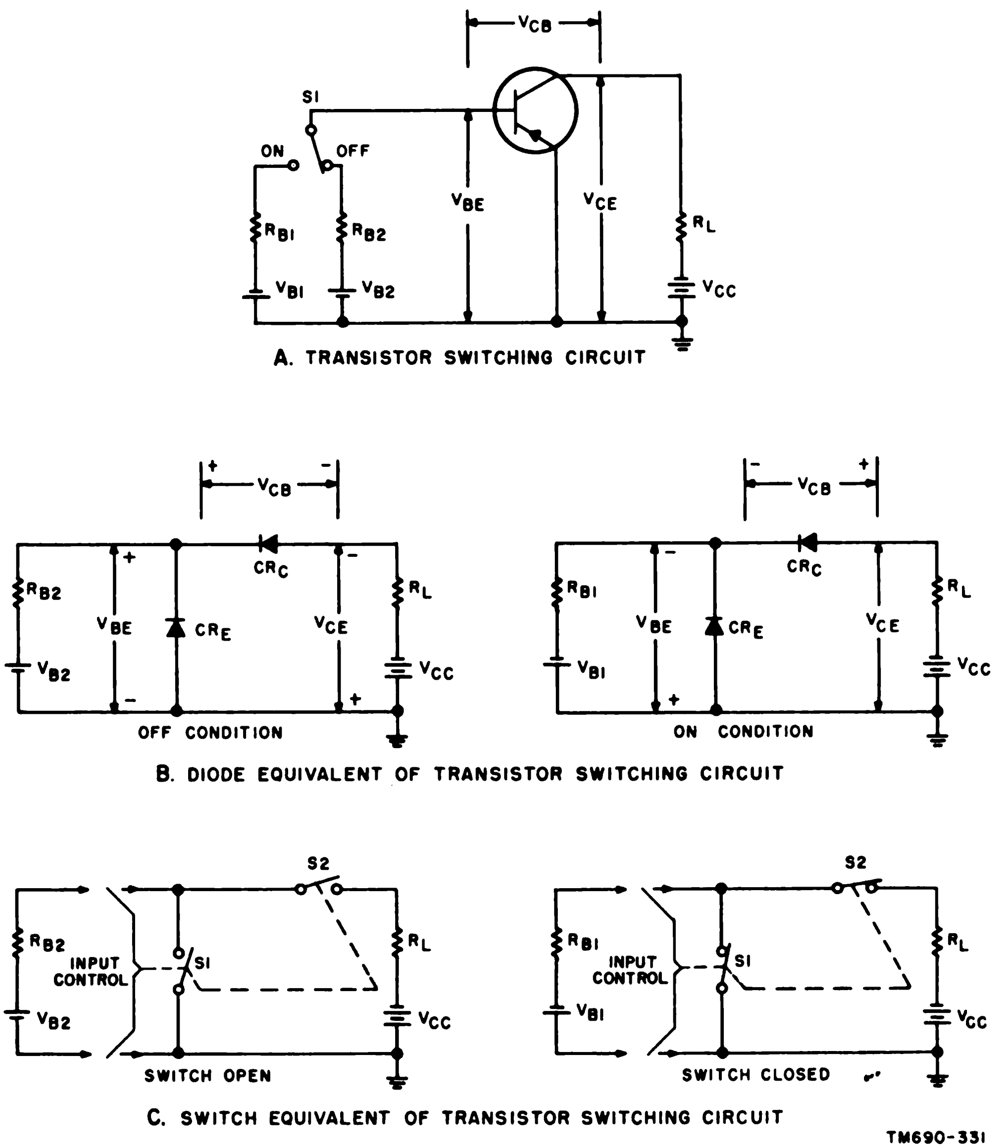


Figure 188. Transistor switching circuit.

Saturation (output) voltage V_{CB}, drops to a smaller negative value than input voltage V_{BB}; the difference in potential of these voltages causes the forward bias (V_{CB}) on diode CR_C. The equivalent switch circuit (C, fig. 188) is *closed*. Switches *S*1 and *S*2 close the circuit for battery V_{CC} through load resistor R_L.

Note. The circuit may also be switched from the *on* state to the *off* state in a similar manner. Small input voltage or current pulses may be used to control large output voltage or current pulses. For example, an input voltage swing of approximately .2 volt causes an output voltage swing of approximately 11.1 volts for a particular transistor. Correspondingly, an input current swing of 180 microamperes results in an output current swing of approximately 2,800 microamperes. An average current gain α_{FB} of approximately 15.5 is obtained.

c. PNP Transistor Switching Conditions in Operating Regions. The following chart lists the transistor items of interest (first column) relative to biases, resistances, gains, currents, and voltages and gives their condition or relative value in the three regions of operation. The chart covers a PNP transistor in the CE configuration (A, fig. 188). Voltage polarities are given with respect to ground.

Item	Region		
	Cutoff	Active (on, transitory)	Saturation (on, fully)
Emitter-base junction	Reverse biased	Forward biased	Forward biased.
Collector-base junction	Reverse biased	Reverse biased	Forward biased.
Input resistance	High (infinite)	Low (zero)	Low (zero).
Output resistance	High (infinite)	High	Low (zero).
Current gain	Zero	Normal	Zero.
Base current	Zero	Transitory	High.
Base voltage	Positive (note)	Negative (transitory)	Negative.
Collector current	Zero	Transitory	High (maximum).
Collector voltage	Negative (high)	Negative (transitory)	Zero.

Note. For NPN transistors, conditions are as in the above chart except that polarities are reversed.

d. Transistor Configurations as Switches. Regions of operation (*a* above) are similar for all transistor configurations used as switches. When both junctions of a transistor are reverse biased (cutoff condition), the output current is very small, and the output voltage is high. When both junctions are forward biased (saturated condition), the output current is high, and the output voltage is small. For most practical purposes, the small output current in the cutoff condition of the transistor and the small output voltage in the saturated condition may be neglected.

(1) *Common base* (A, fig. 40). For large-signal operation, the CB configuration acts as a series switch. The output (collector) current is essentially a large percentage of the input (emitter) current. For practical purposes, these currents may be considered equal. Transient response (par. 187) of this configuration is better than that of the other configurations ((2) and (3) below). The CB configuration is best for use in high-speed switching circuits. In addition, its low input resistance and high output resistance make it effective for impedance matching purposes.

(2) *Common emitter* (C, fig. 40). When used in pulse and switching circuits, this configuration acts as a shunt switch. That is, the input (base) current acts only to regulate or control the output (collector) current. The transient response of the CE configuration to an input pulse is poorer than that of the CB configuration. This is caused by the large drop in current gain at high frequencies. This configuration is used at frequencies lower than that of the CB configuration (assuming the same transistor is involved). Despite its lower frequency response, however, the CE configuration is preferable because of its

high power gain. In addition, only this configuration provides phase inversion.

Note. The cutoff frequency (f_{α}) of a transistor is the frequency at which the current gain drops 3 db from the maximum gain. Because of the direct relationship between the current gain of the *CB* configuration and the *CE* configuration, a small change in $f_{\alpha b}$ results in a large change in $f_{\alpha e}$. Therefore, when $f_{\alpha b}$ drops 3 db, $f_{\alpha e}$ drops considerably more. Thus, the cutoff frequency of the *CE* configuration is lower than that of the *CB* configuration.

(3) *Common collector* (E, fig. 40). The CC configuration (*emitter* follower) like the CE configuration functions as a shunt switch ((2) above). The voltage gain is essentially unity. The properties of a high input impedance and low output impedance of this configuration in its active region make it generally applicable as a buffer and impedance matching circuit. Circuits using this configuration are normally not driven to saturation because of the loss in input and output impedance properties.

187. Large-Signal Pulse Characteristics

When switch $S1$ (A, fig. 188) is operated in sequence from OFF to ON and then back to OFF, the resultant input current pulse I_B is similar to that shown in A, figure 189. In large-signal operation, rectangular input signal I_B drives the transistor from cutoff to saturation and back to cutoff. The distorted output current pulse I_C (B, fig. 189) results because the transistor cannot respond instantaneously to a change in signal level. The response of the transistor during rise and fall time is called the transient response of the circuit. The output pulse characteristics (*a* through *d* below), are governed primarily by the ac characteristics of the transistor.

Note. The transient response of a transistor basically determines the maximum repetition rate (speed of switching) at which the transistor may be driven.

a. Rise Time (t_r). The rise time (also referred to as buildup time or turn-on time) is time required for the leading edge of the pulse to increase in amplitude from 10 to 90 percent of its maximum value. Nonlinear characteristics of the transistor, the external circuit, and energy storage effects (*c* below) contribute to the time. Carriers moving from emitter to collector suffer collision and dispersion and do not reach the collector at the same time. Overdriving the transistor results in decreased rise time. However, the overdrive is normally held to a small value, since the turnoff time (storage time plus fall time (*c* and *d* below)), is affected.

b. Pulse Time (t_p). The pulse time, or duration time, is the length of time that the pulse remains at, or near, its maximum value. The pulse time duration is measured from the point on the leading edge where the amplitude of the pulse has reached 90 percent of its maximum value to the point on the trailing edge where the amplitude has fallen to 90 percent of its maximum value.

c. Storage Time (*Minority Carrier Storage*) (t_s). When the input current, I_B, is cut off, the output current, I_C, does not immediately fall to zero, but remains almost at its maximum value for a length of time before falling to zero (*d* below). This period is called the storage time, or saturation delay time. Storage time results from injected minority carriers being in the base region of the transistor at the moment when the input current is cut off. These carriers require a definite length of time to be collected. The length of storage time is essentially governed by the degree of saturation into which the transistor is driven and the time spent in saturation. The base current (I_B) reversal that occurs between points X and Y, at the end of the input pulse, is the result of the stored carriers contributed by the current gain at the transistor multiplied by the initial input current, I_B. When this current value decays to a value equal to the value of maximum current at saturation, the collector-base diode becomes reverse biased, and both I_B and I_C decay exponentially to zero. For high speed switching, storage time is an undesirable condition. ***Minority carrier storage*** may be avoided by switching a transistor from its *off* state to a point in the active region. Collector clamping (par. 189) prevents operation of the transistor in the saturation region.

d. Fall Time (t_f). In the *fall* or *decay* time of the pulse, the amplitude falls from 90 to 10 percent of its maximum value. The fall time of the pulse is essentially determined by the same factors which determine its rise time (*a* above). Fall time may be slightly reduced through the application of a reverse current at the end of the imput pulse.

Note. The *turnoff time* of a transistor switch is generally determined by the sum of the storage time and fall time. Decreasing either the storage time or fall time or both results in decreased turnoff time. The pulse repetition rate of the circuit is thereby increased.

495909 O—59——13

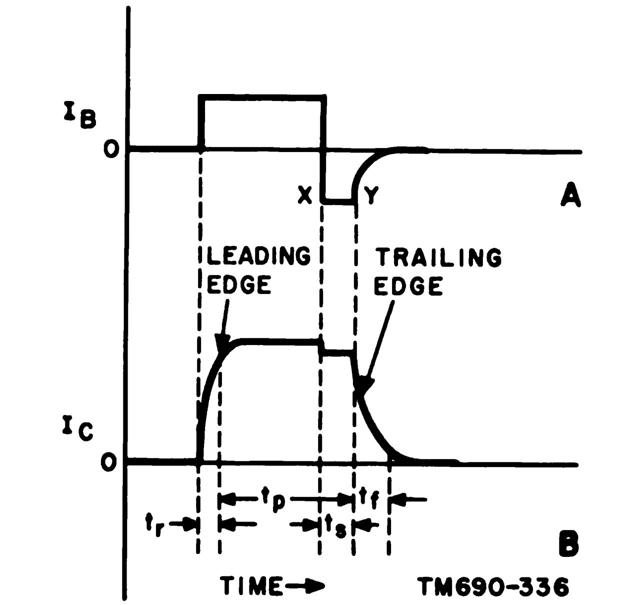


Figure 189. Current pulse characteristics, switching circuit.

188. Leakage Currents

a. CE *and* CB *Configurations.* For a transistor to act as a switch (par. 186), the output current should be zero when the transistor is in the cutoff stage (switch open). This ideal condition, however, cannot be achieved in practice.

(1) In the *CB* configuration (A, fig. 190) with the emitter current equal to zero ($I_E=0$), a reverse bias collector current flows. This current is caused by minority carriers in the collector and the base regions. A detailed discussion of the internal conduction mechanism that causes this current (I_{CBO}) is covered in paragraph 43. In the *CB* configuration, the reverse bias current, also called leakage current, is normally small (measured in microamperes) and can be tolerated in most applications. The effect of temperature on leakage current I_{CBO} is discussed in chapter 5.

(2) In the *CE* configuration (B, fig. 190), leakage current I_{CEO} from emitter to collector is much larger than that of the *CB* configuration ((1) above). Leakage current I_{CBO} is measured from collector to emitter with the base terminal open ($I_B=0$). Leakage current I_{CEO} is caused by leakage current I_{CBO}. Reserve bias minority carriers (electrons, solid line arrow) in the collector region enter the base region and combine with holes (dashed-line arrow) from the emitter region. Before combining, however, the electrons cause a heavy hole current to flow from the emitter to the collector. That is, the electrons from the collector act as a base current bias to cause an amplified collector current. The magnitude of this current is the product of the *CE* forward current amplification factor (α_{fe}) and the base current (I_{CBO}). The total leakage current then is:

$$I_{CEO}=I_{CBO}+\alpha_{fe}I_{CBO}$$
$$=I_{CBO}\ (1+\alpha_{fe})$$

Note that the factor $(1+\alpha_{fe})$ ranges from 25 to 50 depending on the particular transistor. Method for reducing I_{ECO} are discussed in *b* below.

b. Reduction of I_{CEO}. Leakage current I_{CEO} can be reduced by several circuit arrangements. Regardless of the method used, leakage current I_{CEO} cannot be made less than leakage current I_{CBO}.

(1) If the dc base resistance is reduced to zero (A, fig. 191), the minority carriers (I_{CBO}) that enter the base region from the collector easily move from the base region through the external circuit to the collector. This low resistance is provided by inductor *L*1. In this case, with no input ($I_B=0$), the collector leakage current (I_{CEO}) equals I_{CBO}.

(2) In cases where an inductor or dc short cannot be used, a low-valued resistor (R_1) between base and emitter may be used (B, fig. 191). The flow of current I_{CBO} through resistor *R*1 causes a small forward bias which results in the flow of base-emitter current I_B. The latter, emplified by a factor, α_{fe}, appears in the collector as $\alpha f_e I_B$. The smaller *R*1, the lower the collector current I_{CEO}. In this case:

$$I_{CEO}=I_{CBO}+\alpha f_e I_B.$$

(3) If a high input resistance is required with a low leakage current (I_{CEO}), a reverse bias battery (V_{BB}) in series with a high-valued resistance can be used (C, fig. 191). Note that the reverse bias battery voltage attracts the minority current (electron)

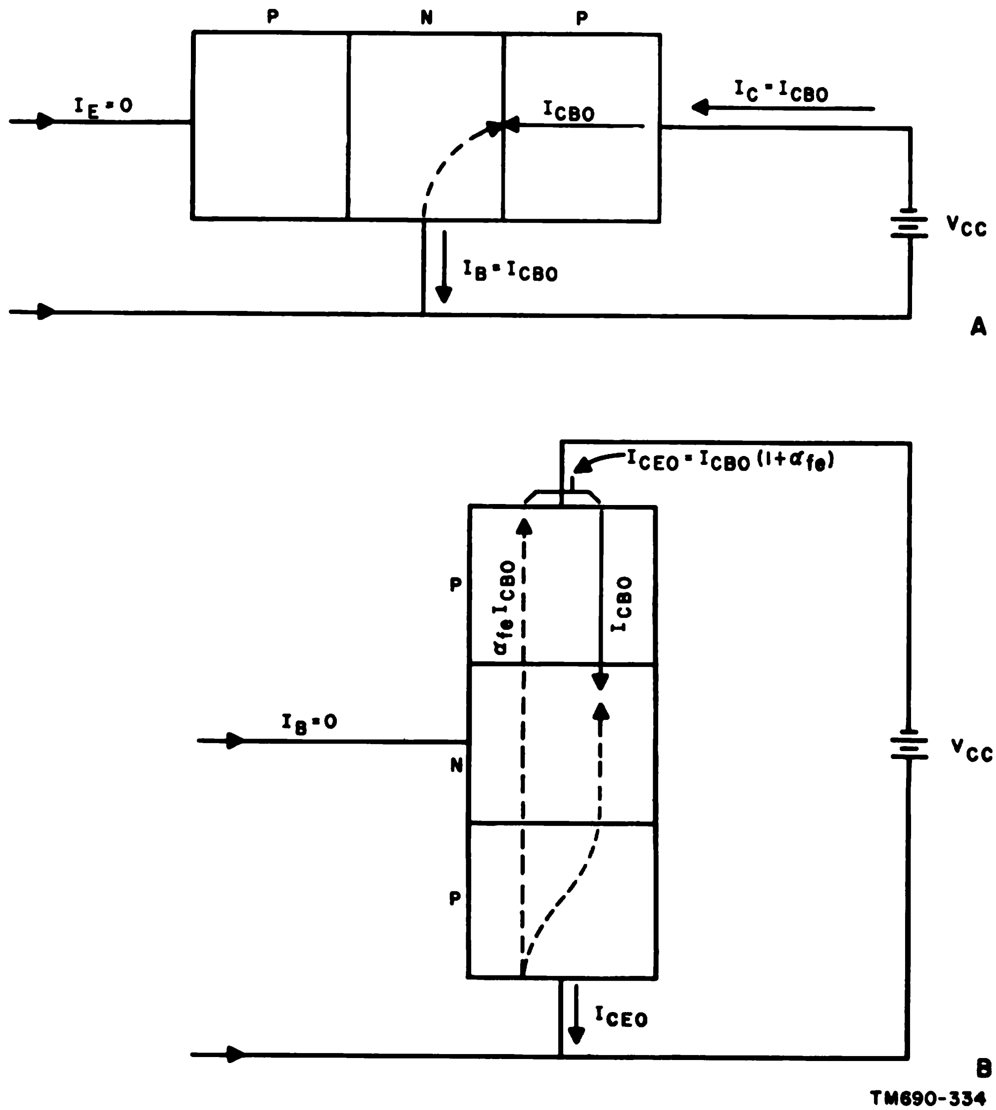


Figure 190. Leakage currents I_{CBO} *and* I_{CEO}.

carriers (I_{CBO}) from the base region. In addition, emitter-base current (I_B) is recurrent flowing is leakage current I_{CBO} (which is leakage current I_{CEO}).

189. Cutoff and Saturation Clamping

When a transistor is driven to saturation and cutoff, it may introduce undesirable effects on the output wave form. Variations in collector potential occur when either the temperature-dependent cutoff current or the load changes. This change in collector potential may cause erratic operation of the succeeding stages. When a transistor is driven to saturation, minority carrier storage delay is introduced (par. 187*c*) with the effects of pulse widening and reduction in the maximum repetition rate at which the circuit may be driven. The use of junction diodes to avoid transistor operation at cutoff or at saturation is given in *a* below. Alternate methods of diode clamping (*b* below) have been devised to avoid saturation of the transistor.

a. Collector-Emitter Clamping. In the circuit shown in A, figure 192, a PNP transistor is used in the *CE* configuration as a simple switching circuit. Current I_B is assumed to be of sufficient magnitude to drive the transistor from cutoff to saturation under normal conditions (that is, without clamping diodes *CR*1 and *CR*2, and their respective bias batteries V_{CO} and V_{CS}). The output characteristics with load line of resistor R_L are shown in B, figure 192. In this example, collector bias battery voltage V_{CC} is 12 volts. Diode *CR*1 and bias bat-

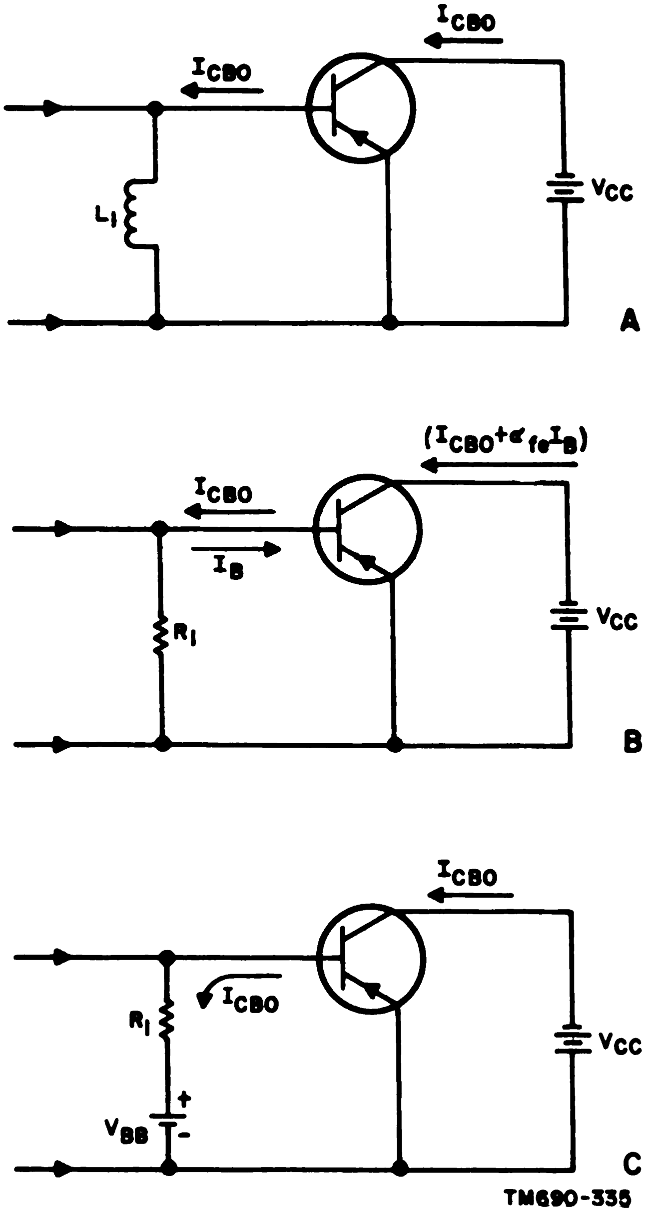


Figure 191. Circuits for reduction of leakage current I_{CBO}.

tery V_{CO} are used for cutoff clamping ((1) below). Junction diode $CR2$ and its bias battery, V_{CS}, are used for saturation clamping ((2) below). Clamping the upper and lower levels of the output permits the substitution of one transistor for another in this circuit.

Note. Since operation with clamping diodes generally results in operation over the linear portion of the output characteristics, the low current at cutoff and the low voltage at saturation are eliminated. Average power dissipation is increased and the load line must be chosen so that operation of the circuit falls within the maximum permissible power dissipation curve (fig. 187) to avoid damage to the transistor.

(1) *Cutoff clamping.* Assume that input current I_B has begun to fall from 150 ua at point Y (B, fig. 192). The collector potential increases negatively from 2 volts at this point. As indicated, the bias potential applied to diode $CR1$ by battery V_{CO} is 8 volts. As the collector potential increases from 2 volts toward 8 volts, diode $CR1$ remains reverse biased (nonconducting). When the collector potential reaches point X (I_B=50μa) on the load line (8 volts), diode $CR1$ becomes forward biased and begins to conduct. Further decrease in input current I_B to zero has no effect on the collector potential, which remains fixed at 8 volts even though collector current I_C decreases (V_{CO} holds V_{CE} at 8 volts). Current flow through load resistor R_L, however, is maintained at the value at point X (about 1 ma) on the collector current axis and consists of collector current I_C) and the current through forward-biased diode $CR1$ and battery V_{CO}. The voltage drop across resistor R_L is equal to the difference (approximately 4 volts) between the voltages of battery V_{CC} and battery V_{CO}. Any change in collector current I_C, which does not cause it to exceed the value at point X, is compensated for in the amount of current drawn from battery V_{CC}. Equilibrium is maintained and the output potential is fixed at 8 volts.

Note. Throughout the foregoing action, diode CR2 remains reverse biased and is effectively an open circuit.

(2) *Saturation clamping.* Disregarding diode $CR1$ and its bias battery V_{CO}, which are involved only in establishing the cutoff collector potential ((1) above), assume that input current I_B is increased to 150 ua at point Y (B, fig. 192). The collector voltage, indicated by the vertical projection to the collector voltage axis, has fallen from a high negative value to 2 volts. The voltage drop produced by the increasing collector current I_C flowing through load resistor R_L causes the voltage on the collector to fall. The bias voltage provided by battery V_{CS} is equal to 2 volts. A slight further increase in

collector current I_C, due to an increase in driving current I_B, further decreases the negative potential on the collector, forward-biasing diode $CR2$, which conducts. When diode $CR2$ conducts, its resistance is negligible and bias battery V_{CS} is effectively placed between collector and ground. This voltage cannot change and, since it is in parallel with collector dc supply V_{CC} and load resistor R_L, the voltage across this branch cannot change. The current flow through load resistor R_L is fixed to provide the necessary voltage drop so that the collector to ground potential is equal to that of bias battery V_{CS}. In this instance, the load current is approximately 3 ma. Although the collector current may further increase, because of an increase in base current, the potential at the collector remains fixed at 2 volts. The additional collector current is drawn from battery V_{CS} through conducting diode $CR2$. The vertical extension of the loadline from point Y represents the zero resistance loadline of forward-biased diode $CR2$. The *normal loadline* extending from point Y to the collector current axis (dashed line) can be seen to enter the saturation region, which falls beyond the knee of the curve of I_B=200 ua. The collector-base diode would be forward-biased in this region (par. 186*b*). When the driving current is reduced to zero, the collector current, I_C, falls rapidly from its maximum value to point Y and then follows the normal loadline toward point X. However, any minority carrier storage introduced by the clamping diode would affect the width of the output pulse. To the left of point Y on the loadline, the collector potential is more negative than 2 volts. Diode $CR2$ is reverse biased and is therefore in a nonconducting state.

(3) *Combination clamping.* The effect of cutoff and saturation clamping ((1) and (2) above) is to keep the collector potential in the range of –2 and –8 volts, thereby preventing cutoff or saturation and the resultant waveform distortion.

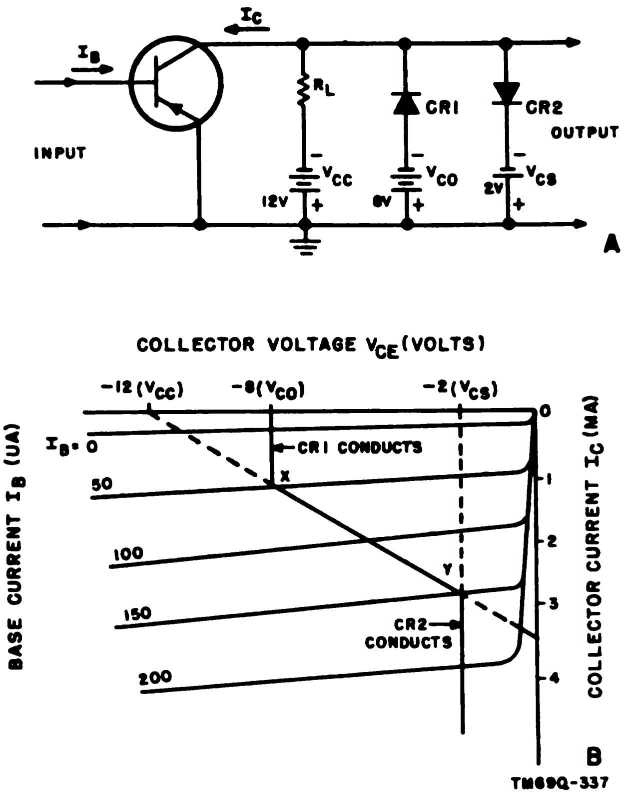


Figure 192. Cutoff and saturation clamping.

b. Collector-Base Saturation Clamping (fig. 193). More efficient switching action is obtained through the use of the single-diode ((1) below) and double-diode ((2) below) collector-to-base clamp.

(1) *Single-diode clamping* (A, fig. 193).

(*a*) In the cutoff condition of the transistor, diode $CR1$ and the emitter-base diode are reverse-biased by the voltage divider network consisting of resistors $R1$, $R2$, and R_B in series with collector and base bias batteries V_{CC} and V_{BB}. When batteries V_{CC} and V_{BB} are equal, the sum of the resistances of resistors $R1$ and $R2$ is made slightly greater than the resistance of resistor R_B. This provides the required initial reverse bias for the emitter-base junction. Resistor $R1$ is very much larger than resistor $R2$, the ratio of the two resistors being determined by the desired clamping voltage when the transistor is conducting.

(*b*) Assume that an applied signal (I_{IN}) forward-biases the emitter-base junction, and drives the circuit to saturation. The potential at the junction of resistors $R1$ and $R2$ is negative and effectively fixed near the very low saturation voltage of the transistor used. The collector potential falls from its high negative value (V_{CC}) to the value of the potential at the junction of resistors $R1$ and $R2$. During this period, diode $CR1$ is nonconducting and base current I_B is equal to the input current, I_{IN}. Load current I_L is equal to collector current I_C and is the amplified driving current, I_{IN}. When the collector potential falls just below the potential at the junction of resistors $R1$ and $R2$, diode $CR1$ conducts. As input current I_{IN}, increases further, base current I_B remains essentially constant and the excess current is shunted through resistor $R2$ and diode $CR1$. Collector current I_C increases by approximately the small value of the excess input current rather than by the amplified input current prior to the clamping effect. Since relatively small currents are passed through conducting diode $CR1$, the minority carrier storage delay contributed by the clamping diode is smaller than that contributed by the clamping diode in the circuit of figure 192 (par. 189*a*(2)).

(2) *Double-diode clamping.* Battery power dissipated in resistors $R1$ and $R2$ (A, fig. 193) is avoided by using double-diode clamping. Diode $CR1$ (B, fig. 193) functions as a clamping diode ((1) above). Forward-biased diode $CR2$ is substituted for resistor $R2$.

(*a*) The emitter-base junction is reverse biased by battery V_{BB} through resistor R_B. Collector reverse bias is provided by battery V_{CC} through load resistor R_L. Diode $CR2$ remains forward-biased throughout the functioning of circuit. (A negative signal must be applied to the base to drive the transistor into conduction.) Generally, a germanium diode, with a low forward voltage drop, is used as the clamping diode ($CR1$) and a silicon diode, with a slightly higher forward voltage drop, is used as the biasing diode ($CR2$). For example, in a typical application, diode $CR1$ provides a forward voltage drop of 1 volt and diode $CR2$ provides a forward voltage drop of 1.2 volts. This essentially maintains a reverse bias of 0.2 volt between collector and base after clamping action takes place ((*b*) below).

(*b*) Circuit functioning is similar to that of the single-diode clamping circuit ((1)(*b*) above). All of input current I_{IN} is applied to the base while diode $CR1$ remains reverse-biased. This condition exists until load current I_L drops the collector potential just below the potential at the junction of diodes $CR1$ and $CR2$. (The base potential is more positive than the potential at this point by the value of the voltage drop across diode $CR2$.) Diode $CR1$ becomes forward biased and shunts the excessive in-

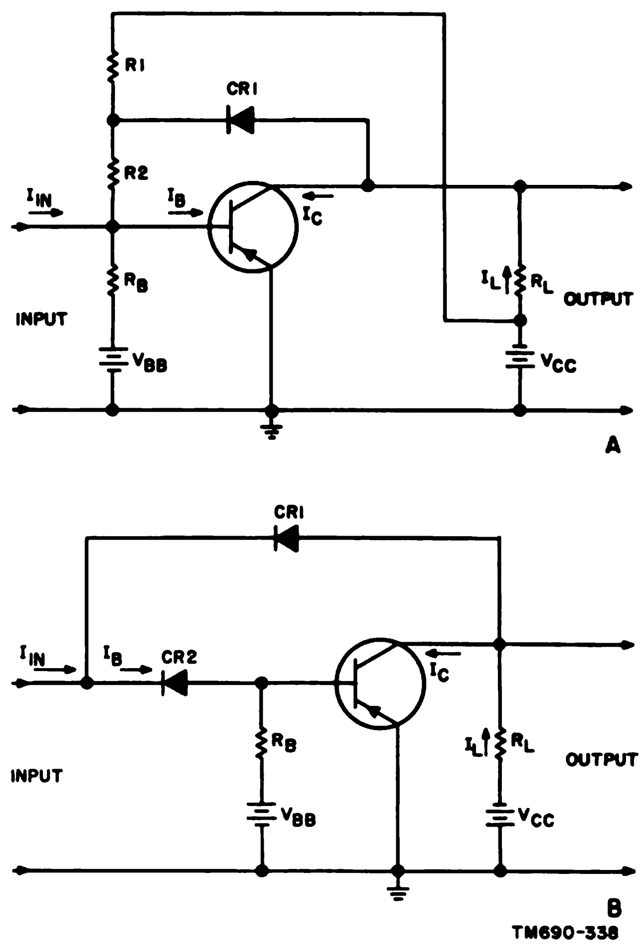


Figure 193. Single-diode and double-diode (collector-base) saturation clamping.

put current to the collector. The smaller voltage drop across diode $CR1$ maintains the reverse collector-base bias at the difference between the forward voltage drops across diodes $CR1$ and $CR2$ ((*a*) above). During the clamped condition, load current I_L is equal to $\alpha_{fe}I_B$ and collector current I_C is equal to the sum of load current I_L and the shunted portion of input current I_{IN}. The latter is again small and not an amplified current ((1)(*b*) above). Thus, it contributes little to output pulse widening. The maximum pulse repetition rate of the switching circuit is thereby increased.

Section II. TRIGGERED CIRCUITS

190. Triggered Circuits, General

A triggered circuit is one in which an externally applied signal causes an instantaneous change in the operating state of a circuit. Once the applied signal (trigger pulse) has initiated the change, the circuit uses its own power to complete the operation (trigger action). Triggered circuits operate in the astable (*a* below), monostable (*b* below), or bistable (*c* below) mode of operation. Each mode of operation passes through stable regions (cutoff and saturation) and unstable regions (active or transition).

a. Astable Operation. The astable mode of operation results in relaxation-type oscillations. The operating point for this mode of operation is in the active region (fig. 187). Astable circuits are free-running circuits. The free-running multivibrator circuit is a typical astable circuit (par. 181). Trigger pulses are used for synchronizing purposes.

b. Monostable Operation. The quiescent (operating) point of the monostable mode of operation is in one of the stable regions. When the circuit is triggered by an external pulse, its operating point is moved from the initial stable region to the other stable region. The time constant of the circuit elements holds the operating point in the new stable region for a period of time. The operating point then moves back to the original stable region. An example is the monostable multivibrator (par. 191), also referred to as a one-shot, single-shot, or single-swing multivibrator. The operation is also referred to as *flip-flip* action; a trigger pulse is required to *flip* the state of the circuit, which of its own accord *flips* back to its original state.

c. Bistable Operation. The quiescent operating point of the bistable mode of operation is also in either of the stable regions (*b* above). When the circuit is triggered by an external pulse, its operating point moves from one stable region to the other. The circuit *remains* in this second state *until a second trigger pulse* is applied; the circuit then returns to its original state. The Eccles-Jordan multivibrator (par. 193*a*) operates in the bistable mode. Operation of this type of circuit is also referred to as *flip-flop* action. That is, two trigger pulses are required; one to *flip* the state of the circuit and the second to *flop* the circuit back to its original state.

191. Monostable Multivibrator

A basic (monostable) multivibrator is shown in figure 194. Circuit voltage wave forms are shown in figure 195. Bias arrangements (*a* below) and transistor regeneration hold transistor $Q2$ in saturation and $Q1$ at cutoff during the quiescent or steady state period. A pulse of short duration (par. 187*b*) is applied and the multivibrator circuit functions until 1 full cycle is completed (*b* below).

a. Quiescent Condition. Battery V_{CC} provides the necessary collector bias voltages for both transistors and forward bias for transistor $Q2$. Transistor $Q2$ is in saturation during the quiescent period. Negative potential v_{c2}, at the collector of transistor $Q2$, is effectively zero or ground. The reverse bias provided by battery V_{BB} maintains transistor $Q1$ at cutoff. Collector potential v_{c1} is negative and equal to the value of battery voltage V_{CC}. Capacitor C_{F1} provides rapid application of the regenerative signal from the collector of transistor $Q1$ to the base of transistor $Q2$, and it is charged to battery voltage V_{CC} through resistor R_{L1} and the essentially shorted base-emitter junction of forward-biased transistor $Q2$.

b. Circuit Operation. The circuit is described with a negative pulse applied at the input ((1), (2) and (3) below), but other means of triggering may be used.

(1) A negative pulse is applied through coupling capacitor C_C to the base of tran-

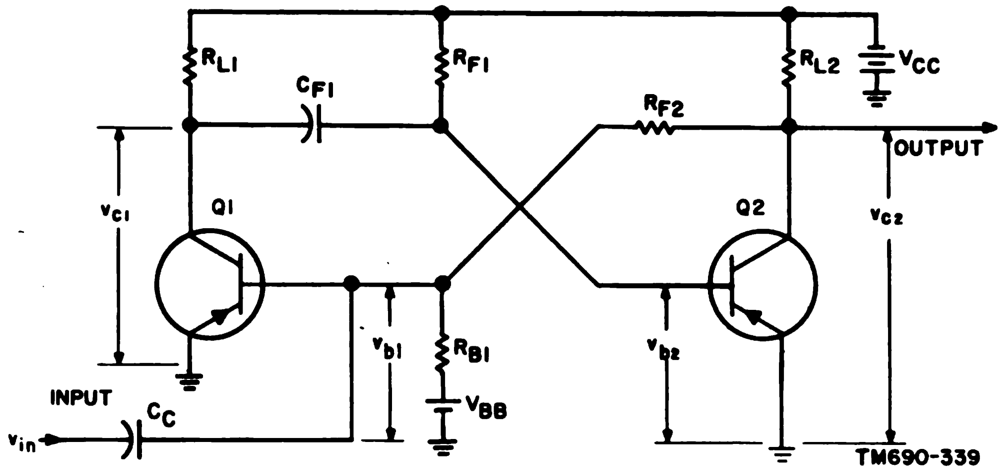


Figure 194. Monostable multivibrator.

sistor $Q1$, and transistor $Q1$ begins to conduct. The high negative voltage at the collector of transistor $Q1$ begins to

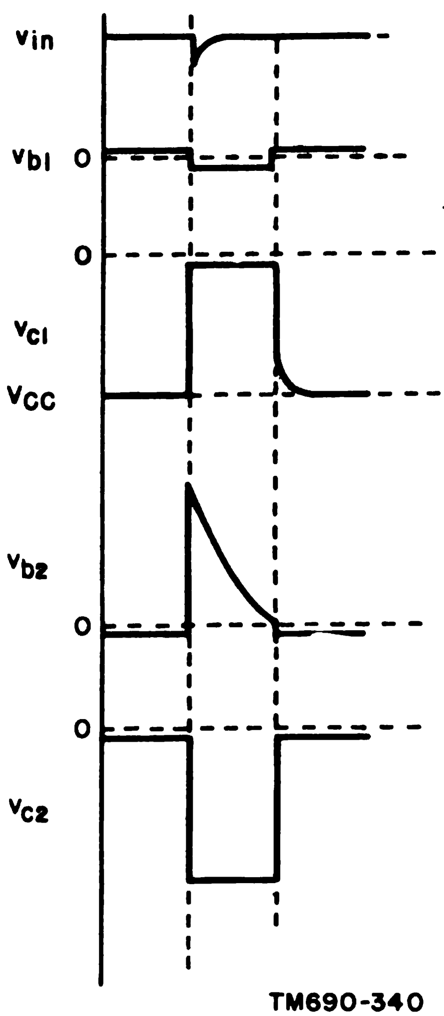


Figure 195. Monostable multivibrator waveforms.

fall (becomes more positive). This positive-going voltage is coupled to the base of transistor $Q2$, and the forward-bias is decreased. The base current and collector current of transistor $Q2$ begin to decrease. The collector voltage of transistor $Q2$ increases negatively. A portion of this voltage is coupled through resistor R_{F_2} to the base of transistor $Q1$, increasing its negative potential. This regeneration results in a rapid change of both transistors; it drives transistor $Q1$ into saturation and transistor $Q2$ into cutoff. Since capacitor C_{F_1} was initially charged to a potential almost equal to battery voltage V_{CC} (*a* above), the base of transistor $Q2$ is at a positive potential almost equal to the magnitude of battery voltage V_{CC}.

(2) Capacitor C_{F_1} discharges through resistor R_{F_1} and the low saturation resistance of transistor $Q1$. Base potential v_{b_2} of transistor $Q2$ becomes less positive. When the base potential of transistor $Q2$ becomes slightly negative, transistor $Q2$ again conducts. The collector potential of transistor $Q2$ increases positively and is coupled to the base of transistor $Q1$, driving it into cutoff. Transistor $Q1$ is again at cutoff and transistor $Q2$ is in saturation with its collector voltage almost at zero. This stable condition is maintained until another pulse triggers the circuit.

(3) The output is taken from the collector of transistor $Q2$. The wave form at this

point is essentially square. Time duration of the output pulse is primarily determined by the time constant of resistor R_{F1} and capacitor C_{F1} during discharge ((2) above).

192. Triggered Blocking Oscillator

The major difference between the triggered (monostable) blocking oscillator (A, fig. 196) and the astable or free running blocking oscillator (par. 183) is due to the bias arrangement. In the astable mode of operation, the base-emitter junction is forward-biased and the circuit functions as soon as dc power is applied. In the monostable mode of operation, the base-emitter junction is reversed-biased (*a* below) and the transistor is held at cutoff. The output obtained is a single cycle for each pulse applied at the input (B, fig. 196). The blocking oscillator exhibits minority carrier storage (par. 188*c*) and excessive backswing which are undesirable. The modified triggered blocking oscillator (*b* below), overcomes these characteristics.

a. Typical Blocking Oscillator (fig. 196). In the quiescent state, the transistor is held at cutoff

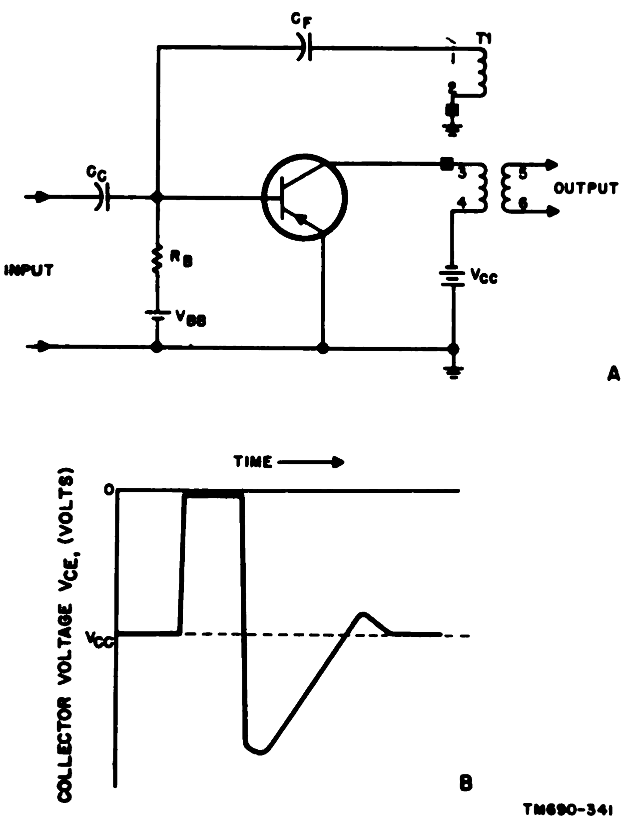


Figure 196. Triggered blocking oscillator.

by the reverse bias between the base and emitter provided by battery voltage V_{BB}. The collector-base junction is reverse biased by battery V_{CC}. Transformer $T1$ provides regenerative feedback and couples the output signal to the load. Capacitor C_F acts as a dc blocking capacitor. Capacitor C_C couples the input signal to the base of the transistor.

(1) A negative trigger pulse is applied at the input and the transistor begins to conduct. Collector current flows through transformer winding 3–4 and produces a varying magnetic flux that induces a voltage of opposite polarity in transformer winding 1–2. This voltage is coupled through capacitor C_F to the base of the transistor and supplies regenerative feedback. This action continues until the transistor is driven into saturation and collector current ceases to increase.

(2) Since the collector current becomes constant, no feedback voltage is induced in transformer $T1$ and the reverse bias provided by battery V_{BB} cuts off the transistor. The collector current continues to flow for a short while because of minority carrier storage.

(3) When the collector current stops flowing, the collapsing field around transformer winding 1–2 induces a voltage in winding 3–4 which exceeds the value of battery voltage V_{CC} (B, fig. 196). In some instances, this backswing voltage may exceed the collector breakdown of the transistor. Driving the transistor into the saturation region introduces the undesired effect of minority carrier storage (par. 187*c*). A means of preventing transistor breakdown and operation in the saturation region is given in *b* below.

b. Diode Clamped Blocking Oscillator (Nonsaturating). A modified triggered blocking oscillator (fig. 197) prevents voltage breakdown and reduces the effects of minority carrier storage (*a*(3) above); this increases the maximum pulse repetition rate. Substitution of transistors is made less critical. The use of clamping diodes $CR1$ and $CR2$ prevents saturation and excessive induced voltage at the collector of the transistor. Circuit operation is similar to that described in *a* above except for the differences given in (1) through (5) below.

(1) The base-emitter junction of the transistor is reverse-biased by battery V_{BB}. The collector is reverse-biased by the combined voltages of V_{C1} and V_{C2}. Quiescently, diode $CR2$ has no bias applied and diode $CR1$ is reverse-biased (fig. 197).

(2) If a negative trigger pulse is applied, the transistor is driven from cutoff into conduction. Collector current increases rapidly as a result of the regeneration of the circuit. The collector voltage falls from the combined voltages of C_{C1} and V_{C2} to the value of V_{C1}.

(3) As the collector potential decreases, due to increase in collector current, the voltage across diode $CR2$ causes reverse bias. Therefore the effect of $CR2$ at this time may be neglected because it maintains reverse bias. The reverse-bias potential applied to diode $CR1$ decreases until the collector voltage reaches a value equal to V_{C1}. Further increase of collector current causes a further decrease in collector voltage, which results in diode $CR1$ becoming forward-biased. At this point diode $CR1$ conducts and carries away current from the collector to maintain collector voltage at the value of V_{C1}. Diode $CR1$ acts as a short circuit to transformer winding 3–4. Thus the transistor is prevented from going into saturation.

(4) When the magnetic field begins to collapse, the collector voltage becomes more negative (B, fig. 196), and the bias voltage across diode $CR1$ reverts to its nonconducting state when the value of V_{C1} is exceeded.

(5) The collector voltage increases negatively until it reaches the value of the combined voltages of V_{C1} and V_{C2}. Any attempt by the induced voltage ((4) above) to further increase the negative voltage at the collector is overcome by the conduction of diode $CR2$. The charge in transformer winding 3–4 is then discharged through this diode, and thus prevents the application of the excessive voltage between the collector and the emitter. The negative going voltage doesn't drop below voltage V_{CC} and the voltage swing below V_{CC} (fig. 196) has been eliminated by diode $CR2$.

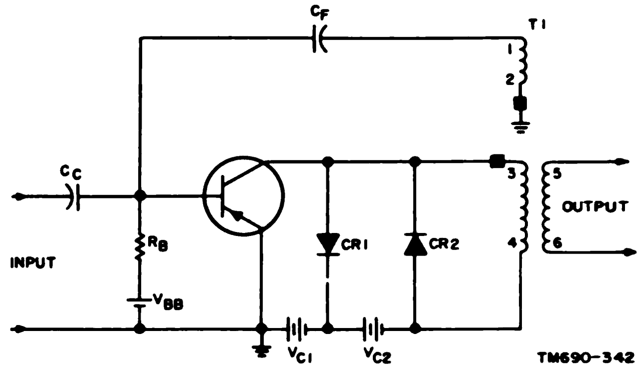


Figure 197. Triggered blocking oscillator, with clamping diodes.

193. Bistable Multivibrator Circuits

A bistable circuit is initially at rest in either one of the stable states. When triggered by an input pulse, the circuit switches to the second stable state where it remains until triggered by another pulse. This type of operation is particularly useful for providing a unit step voltage (par. 185*c*). The bistable multivibrator may be further classified as saturating (*a* below) or nonsaturating (par. 195). The conventional bistable multivibrator is described in *a* below, and a direct-coupled bistable multivibrator is described in *b* below.

a. Conventional Bistable Multivibrator. A basic Eccles-Jordan bistable multivibrator is shown in figure 198. This circuit differs from the astable multivibrator (par. 181) primarily in the base-emitter arrangement. In the stable state, either transistor is in the *on* state while the other is *off*. The states of the transistor are switched with the application of a properly applied trigger pulse ((2) below).

(1) *Cutoff and saturation conditions.* With the initial application of dc power, one transistor will be caused to turn on while the other will be cut off. Each transistor is held in its particular state by the condition of the other. A numerical example is illustrated in figure 199. Resistors R_L, R_F, and R_B correspond to the voltage divider networks in figure 198.

(*a*) When a transistor is cut off, its output resistance is high and its collector current is effectively zero. However, current flow through the voltage divider network (resistors R_L, R_F, and R_B) because of collector battery voltage V_{CC} (−28 volts) and base battery voltage V_{BB} (+2 volts), causes finite voltage drops across each resistor. The voltage drop across resistor R_L is 2 volts, leaving approximately the full value of collector battery voltage V_{CC} at point C. The voltage drops across resistors R_F and R_B are 21 volts and 7 volts, respectively. This provides a negative voltage (−5 volts) at point B and causes forward bias, and transistor $Q2$ conducts. (For NPN-type resistors, the polarities are reversed.) The resistive and bias values are chosen to drive the *on* transistor into saturation.

(*b*) The high collector current (saturation) of the *on* transistor flows through resistor R_L and causes a voltage drop equal to collector battery voltage V_{CC}, so that at point C, the voltage is effectively zero (B, fig. 199). Division of base battery voltage V_{BB} (+2 volts) by resistors R_F and R_B results in a positive voltage (+1.5 volts) at point B. This is a reverse bias and the transistor is held at cutoff.

(2) *Circuit operation* (fig. 198). The application of a negative trigger pulse to the base of the *off* transistor or a positive pulse to the base of the *on* transistor will switch the conducting state of the circuit. Collector triggering may be similarly accomplished. Two separate inputs are shown in figure 198. A trigger pulse at input A will change the state of the circuit. Once the state of the circuit is changed, an input of the same polarity at input B or an input of opposite polarity at input A will again trigger the circuit. Alternate methods for controlling the application of the trigger pulse are described in paragraph 194.

(*a*) Assume that transistor $Q1$ is cut off and transistor $Q2$ is conducting. A negative trigger pulse applied at input A causes transistor $Q1$ to conduct. The rise in collector current in transistor $Q1$ causes the collector voltage to fall. This change in voltage is coupled to the base of transistor $Q2$ and reduces its forward bias; conduction in transistor $Q2$ begins to decrease. The collector

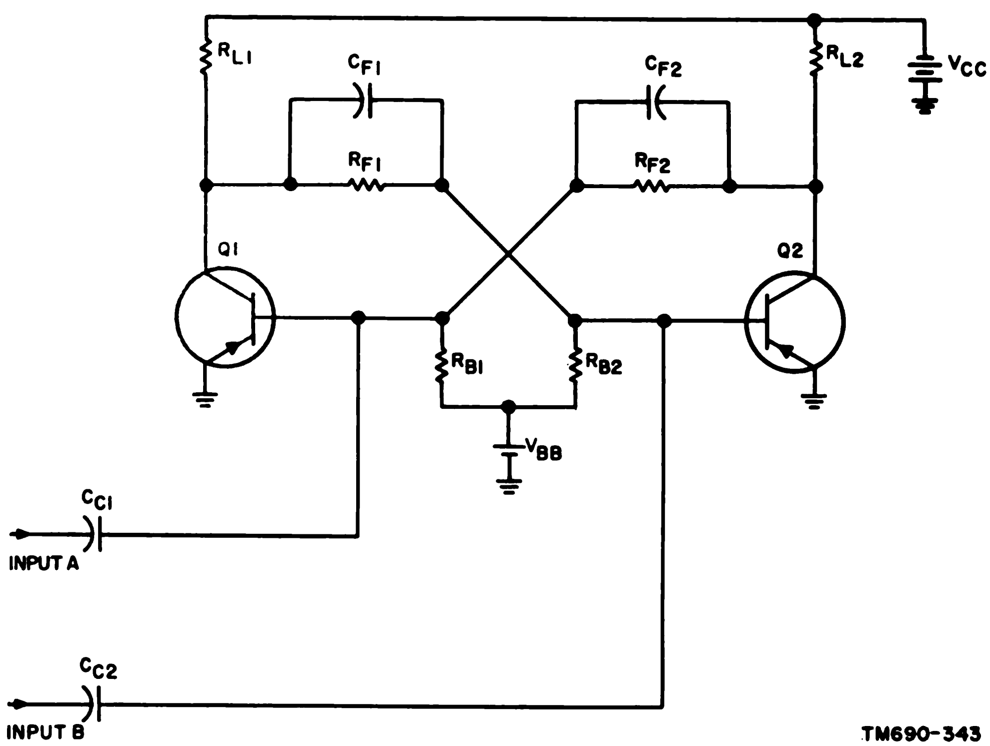


Figure 198. Conventional bistable multivibrator.

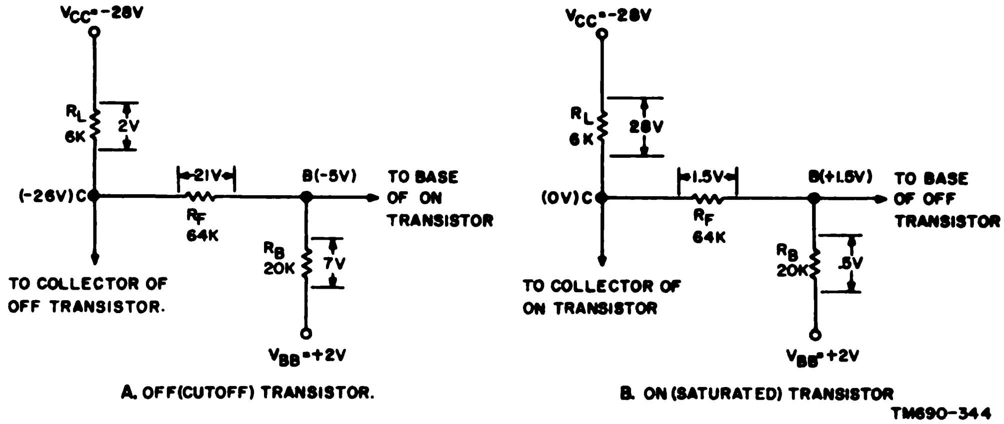


Figure 199. Voltage divider networks showing saturated and cutoff conditions of a bistable multivibrator.

current decreases, and the collector voltage changes from zero to a negative value (approaching the value of battery voltage V_{CC}). This change in voltage is coupled to the base of transistor $Q1$, making the base more negative and increasing the conduction of the transistor. The regenerative feedback continues until transistor $Q1$ is in saturation and transistor $Q2$ is cut off. The time constant of capacitor C_{F1} and resistor R_{F1} and that of capacitor C_{F2} and resistor R_{F2} essentially determines the fall time (from conduction to cutoff) of transistors $Q1$ and $Q2$, respectively. In addition, the capacitors rapidly couple the changing voltages to the bases to insure rapid switching action of the transistors.

Note. During the rapid transition period, both transistors conduct. In one transistor, the conduction is increasing and in the other, the conduction is decreasing.

(*b*) The output, taken between collector and ground, is a unit step voltage when one trigger is applied. A square-wave output could be obtained through continuous pulsing or triggering of the input. Frequency division is thus obtained, with a ratio of 1 to 2.

b. Direct-Coupled Bistable Multivibrator. The direct-coupled bistable multivibrator (A, fig. 200) is also referred to as a *binary* or count-by-two circuit when used in computing and counting applications. In the circuit illustrated, transistors Q1 and Q4 are not considered part of the bistable multivibrator. These are used to provide input control of the conducting state of the circuit. The load for each transistor of the direct-coupled binary consists of its load resistor (R_{L1} or R_{L2}) in parallel with the base input resistance of the other transistor. The callector to emitter voltage for the *off* condition of the transistor is the low base-emitter potential of the *on* transistor. Operation of the circuit is as follows:

(1) Assume that transistor $Q2$ is cut off and transistor $Q3$ is conducting prior to time $t1$ (B, fig. 200). With no input signal, collector voltage V_{C2} of *off* transistor $Q2$ is the same as base-emitter voltage V_{B3} of *on* transistor $Q3$. Circuit triggering is accomplished by effectively grounding the collector of the *off* transistor. At time $t1$, a negative pulse (set pulse) is applied at input 1 which causes transistor $Q1$ to conduct through common load resistor R_{L1}. Collector potential V_{C2} and base potential V_{B3} rise toward saturation voltage V_{SAT} of the transistor. Potential V_{B3} (although slightly negative with respect to ground) cuts off transistor $Q3$ and its collector voltage (V_{C3}) increases negatively, driving transistor $Q2$ further into conduction. The positive feedback between collectors and bases of the transistors results in the rapid change of state of the circuit. Circuit equilibrium is reached, transistor $Q2$ is in saturation, and transistor $Q3$ is at cutoff.

(2) At time $t2$, a negative pulse is applied at input 2, causing transistor $Q4$ to conduct through common load resistor R_{L2}. The

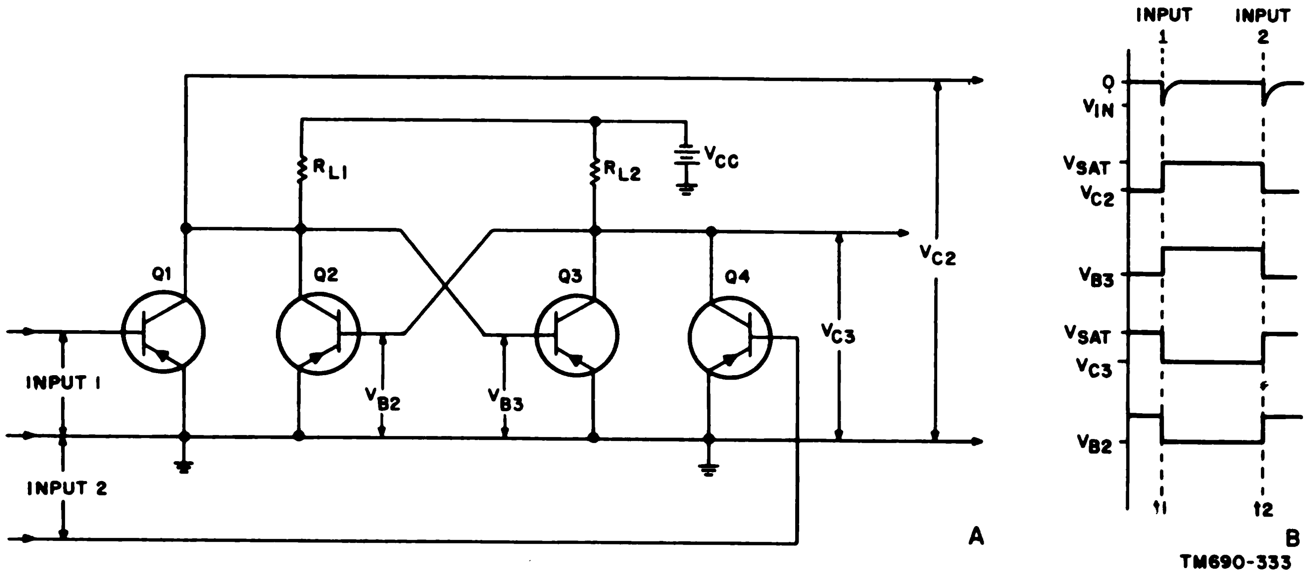


Figure 200. Direct-coupled bistable multivibrator.

process of regeneration duplicates that described in (*a*) above except that transistor Q_3 is switched *on* while transistor Q_2 is switched *off*. The negative pulse at input 2 may be considered a reset pulse that brings the circuit back to the original state.

194. Trigger Pulse Steering

Several methods are available for applying the triggering pulse or pulses to a bistable multivibrator circuit. Fundamentally, the method used is determined by the polarity and magnitude of the trigger pulse available and the desired repetition rate (*a* through *c* below). The trigger normally used is a pulse of extremely short duration or a unit step voltage or current. Each causes the same reaction of the circuit.

a. Triggering Methods. The trigger pulse may be applied or directed to the base or collector of either transistor in a bistable multivibrator circuit. Initial triggering of a bistable multivibrator could be accomplished in one of four ways ((1) through (4) below).

Note. PNP-type transistors are assumed in (1) through (4) below. All polarities must be reversed for NPN-type transistors.

(1) A negative trigger pulse may be applied to the base of the *off* transistor because the quiescent voltage at this point is positive (fig. 198). This pulse drives the transistor out of cutoff into conduction.

(2) A positive trigger pulse may be applied to the base of the *on* transistor because the quiescent voltage at this point is negative. This pulse drives the transistor out of saturation into the transition region.

(3) A negative trigger pulse may be applied to the collector of the *on* transistor because the quiescent voltage at this point is almost zero (figs. 198 and 200). This trigger pulse coupled to the base of the *off* transistor takes it out of cutoff.

(4) A positive trigger pulse may be applied to the collector of the *off* transistor because the quiescent voltage at this point is negative. This pulse coupled to the base of the *on* transistor takes it out of saturation.

Note. The trigger pulses must be of sufficient amplitude to reverse the stable state of the transistors.

b. Separate Input Triggering. The separate inputs at the bases of each transistor (fig. 198) provide more than one method of triggering the circuit. A negative trigger at input A or a positive trigger at input B will cause the initial transition. The second transition would be effected by a positive trigger at input A or a negative trigger at input B. An alternate method of triggering the circuit with separate inputs available would be the application of a negative pulse at input A and a second negative pulse at input B to restore the circuit to its original state.

c. Diode Steering, Single Input. In certain applications, it is desirable to trigger the bistable multivibrator on and off repeatedly with input pulses of the same polarity. If the pulse were applied simultaneously to both the *off* and *on* transistors, switching time would be delayed. With pulse steering diodes, the input pulse is directed to the proper transistor only, driving it *off* or *on* ((2) below). A bistable multivibrator using negative pulse steering is illustrated in figure 202. Trigger action of the multivibrator is identical with that given in paragraph 193*a* and *b*. A bistable multivibrator (par. 195*a*) illustrating positive pulse steering is shown in figure 203.

(1) Assume that the negative pulse is applied to the bases of both transistors from one input source (fig. 198). The tendency is to drive the *off* transistor into conduction and the *on* transistor (negative base) into saturation. Thus the feedthrough from the collector of the *off* transistor must overcome the initial bias condition of the *on* transistor and also the additional bias provided by the input pulse. Increased signal voltage is therefore required, and the process of regeneration and transition is slowed. The rise and fall times of the output signal are increased, thereby decreasing the pulse repetition rate of the circuit.

(2) Negative or positive pulse steering (fig. 201) may be provided in the bistable multivibrator circuit. In PNP-type transistors, a negative trigger pulse drives the *off* transistor *on;* a positive trigger pulse drives the *on* transistor *off*. In NPN-type transistors the polarities are reversed. In each method of steering illustrated, the base of the *on* transistor is negative and the base of the *off* transistor is positive with no signal applied.

(*a*) For negative pulse steering, diodes *CR*1 and *CR*2 are arranged as shown in A of figure 201. Resistors *R*1 and *R*2 form a voltage divider so that the cathodes of the diodes are at a positive potential. Diode *CR*1 is in its nonconducting condition because of the reverse bias, and diode *CR*2 is conducting (low-resistance short) because of forward bias. When a negative trigger pulse is applied through coupling capacitor C_C, the reverse bias diode *CR*1 prevents the pulse from being applied to the base of the *on* transistor. Diode *CR*2 is forward-biased, and the negative trigger pulse is applied to the base of the *off* transistor, thereby overcoming the positive bias on the base and driving the transistor *on*. Normal functioning of the bistable multivibrator takes place (par. 193) and the circuit switches its state. A second negative trigger pulse is directed through diode *CR*1 which would then be biased in the forward direction.

Note. **A positive trigger pulse applied to this steering circuit, regardless of the state of the bistable multivibrator, increases the reverse bias on the diodes and therefore has no effect on the circuit.**

(*b*) For positive pulse steering (B, fig. 201), the diodes are reversed and a negative potential is provided at the junction of the voltage divider which consists of resistors R1 and R2. Diode CR1 is forward-biased while diode CR2 is reverse-biased. A positive trigger pulse is therefore applied through diode CR1 to the base of the *on* transistor, driving it *off*. The re-

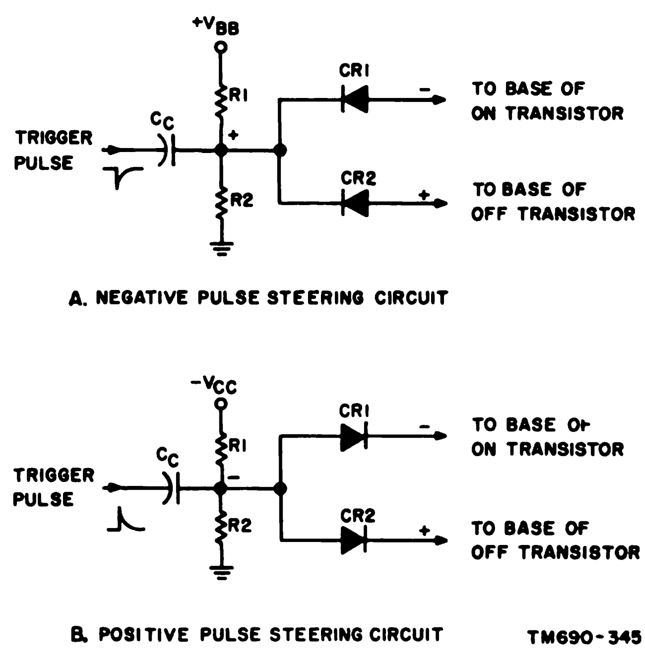


Figure 201. Pulse steering diodes.

verse bias on diode CR2 prevents the trigger pulse from being applied to the base of the *off* transistor. Once the circuit switches state, the bias conditions of the diodes are reversed, and a second positive trigger pulse is directed through forward-biased diode CR2 to the base of the *on* transistor.

Note. A negative trigger pulse applied to this arrangement increases the reverse bias on the diodes and therefore has no effect on the circuit.

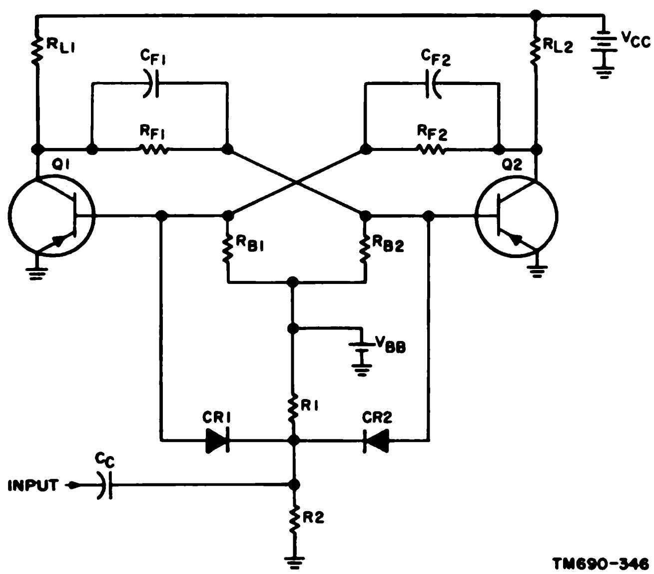


***Figure 202.** Bistable multivibrator with negative-pulse steering diodes.*

d. Transistor Pulse Steering. The transition time for the change of state of a bistate multivibrator may be decreased by using a symmetrical transistor for pulse steering (fig. 204). The symmetrical transistor (Q3) is one in which the emitter-base and collector-base junctions are identical. Reversal of bias conditions on each junction results in normal amplifier action except that the directions of current flow are reversed. That is, when the collector-base junction is forward-biased, and the emitter-base junction is reverse-biased, the collector functions as the emitter and the emitter functions as the collector. The reversal of current permits trigger pulse steering ((1), (2), and (3) below).

(1) If transistor $Q1$ is near saturation (collector voltage near zero) and transistor $Q2$ is at cutoff (collector voltage equal to battery voltage V_{CC}), transistor $Q3$ conducts normally. With no input signal, transistor $Q3$ is conducting with its operating point in the active region of its output characteristics. Transistor $Q3$ electron current (arrows) flows out of battery V_{CC} through resistor R_{L2}, to the collector from the emitter ($Q3$), into the collector of transistor $Q1$ and from the emitter and emitter resistor RE back to ground. This current causes an increase in the negative potential at the collector of transistor $Q1$. The latter voltage prevents deep voltage saturation of transistor $Q1$ and permits a higher pulse repetition rate.

(2) A negative trigger pulse applied at the input increases the conduction of transistor $Q3$. The increased collector current flow from battery V_{CC} through load resistor R_{L2} causes the collector potential of transistor $Q2$ to become more positive (less negative). Simultaneously, the increased emitter current (decreased current through load resistor R_{L1}) causes the collector potential of transistor $Q1$ to become more negative. The regenerative action of the multivibrator (par 195*b*) causes the state of transistor to move toward cutoff while transistor $Q2$ moves to saturation.

(3) When circuit equilibrium is reached, transistor $Q1$ is at cutoff while transistor $Q2$ is at saturation. Because of this, a reversal of transistor bias is established across the junctions of transistor $Q3$. Electron current through transistor $Q3$ flows out of the negative terminal of battery V_{CC}, through load resistor R_{L1}, through transistor $Q3$ collector-emitter (now reversed), through the collector of transistor $Q2$ to ground. The stable conditions that exist at this point are the same as explained in (1) above except for the reversal of states of transistors $Q1$ and $Q2$. A second negative pulse applied at this time results in similar action ((2) above) and the multivibrator and transistor $Q3$ revert to their initial conditions.

195. Nonsaturating Bistable Multivibrators

In the nonsaturating multivibrator, lower and upper limits are established for the output voltage in the stable or quiescent states. This is to increase the pulse repetition rate. Diode clamping (*a* below) or transistor clamping (*b* below) may be used to prevent saturation and cutoff.

a. Diode Clamping (fig. 203). Diodes *CR*1 and *CR*2, biased by batteries V_{CS} and V_{CO}, provide saturation and cutoff clamping for transistor *Q*1. Diodes *CR*3 and *CR*4, also biased by batteries V_{CS} and V_{CO}, provide saturation and cutoff clamping for transistor *Q*2. In each instance, the diodes function identically with those described in paragraph 189. Positive pulse steering diodes (par. 194*b*) are used at the input to the multivibrator. Multivibrator action is identical with that of the conventional bistable multivibrator (par. 193*a*) except for the prevention of saturation and cutoff conditions.

b. Transistor Clamping. The operation of transistor *Q*3 (fig. 204) as a pulse steering device is covered in paragraph 194*d*. It is shown that transistor *Q*3 shunts a portion of the collector current of the conducting transistor around its load resistor (R_{L1} or R_{L2}). This prevents zero voltage on the collector of the conducting transistor. Assume that transistor *Q*1 is conducting and the collector voltage is clamped at -3 volts. Assume further that the required base-emitted voltage is -0.5 volt. This voltage could be obtained by grounding the emitter and establishing -0.5 volt on the base with the voltage divider, consisting of resistors R_{L2}, R_{F2}, and R_{B1}. However, resistor R_{B1} would be very low in value and excessively shunt the regenerative pulses. To avoid excessive shunting, resistor R_{B1} can be increased so that the base voltage is -2 volts. To offset this high voltage, resistor R_E, bypassed by capacitor C_3, develops -1.5 volts at the emitter. The net forward bias between the base and emitter is -0.5 volt. Note that the collector-base junction remains reverse-biased.

196. Squaring Circuit (Schmitt Trigger)

The Schmitt circuit (A, fig. 205) differs from the conventional bistable multivibrator (par. 193*a*) in that one of the coupling networks is replaced by a common-emitter resistor. This arrangement provides additional regenerative feedback to obtain a faster switching time.

a. Assuming the quiescent condition (no input) with transistor *Q*1 at cutoff, the collector voltage is equal to battery voltage V_{CC}. This negative voltage is coupled to the base of transistor *Q*2 through resistor R_{F1} and the base voltage of transistor *Q*2 is equal to the voltage drop across resistor R_B minus battery voltage V_{BB}.

b. Current flow from the emitter of transistor *Q*2, through common-emitter resistor R_E, to battery V_{EE} maintains the emitter of transistor *Q*1 at a negative potential. The reverse bias now developed between the emitter and the base of transistor *Q*1 maintains the cutoff condition. The high negative voltage at the base of transistor *Q*2 produces forward bias for the base-emitter junction and causes it to operate in the saturation region.

c. A negative signal of sufficient amplitude (B, fig. 205) applied to the base of transistor *Q*1 will overcome the reverse bias and cause transistor *Q*1 to conduct. The potential at the collector decreases (becomes less negative); this change is coupled to the base of transistor *Q*2. The emitter current of transistor *Q*2 decreases, lowering the potential across resistor R_E. The emitter of transistor *Q*1 becomes less negative, reducing the reverse bias and increasing collector current. This regenerative action continues until transistor *Q*1 is operating in the saturation region and transistor *Q*2 is cut off. The output voltage is a maximum negative voltage (B, fig. 205).

d. The new stable condition continues until the input begins to rise (become more positive). This positive-going input decreases the base potential of transistor *Q*1 and increases the reverse bias. This causes the collector voltage to increase (become more negative), emitter current to decrease, and the potential across resistor R_E to decrease. Simultaneously, the increasing (negative) voltage at the collector of transistor *Q*1 is coupled to the base of transistor *Q*2, driving it negative; the decreasing voltage of resistor R_E causes the emitter of transistor *Q*2 to go more positive. Both actions reduce the reverse bias of the emitter-base junction and transistor *Q*2 again operates at saturation, cutting off transistor *Q*1, and returning the circuit to its original operating conditions. The output is a minimum negative voltage (B, fig. 205).

e. The rise and fall time of the output wave of this circuit is shorter than that of the conventional bistable multivibrator. The shape of the output wave does not depend on the shape of the input wave.

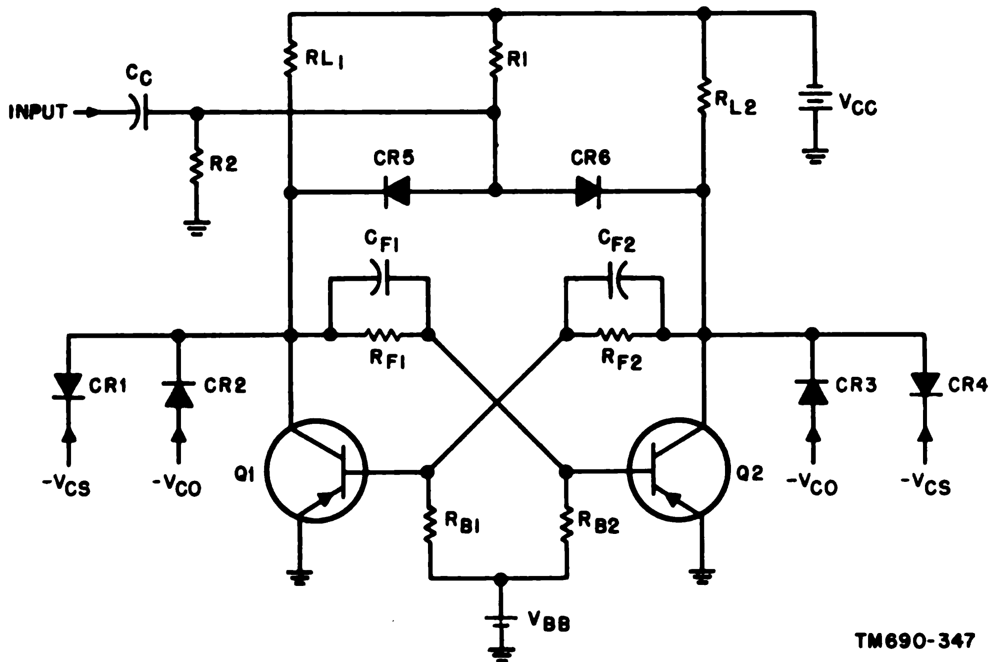


Figure 203. Nonsaturating bistable multivibrator, using diode clamping and positive pulse steering.

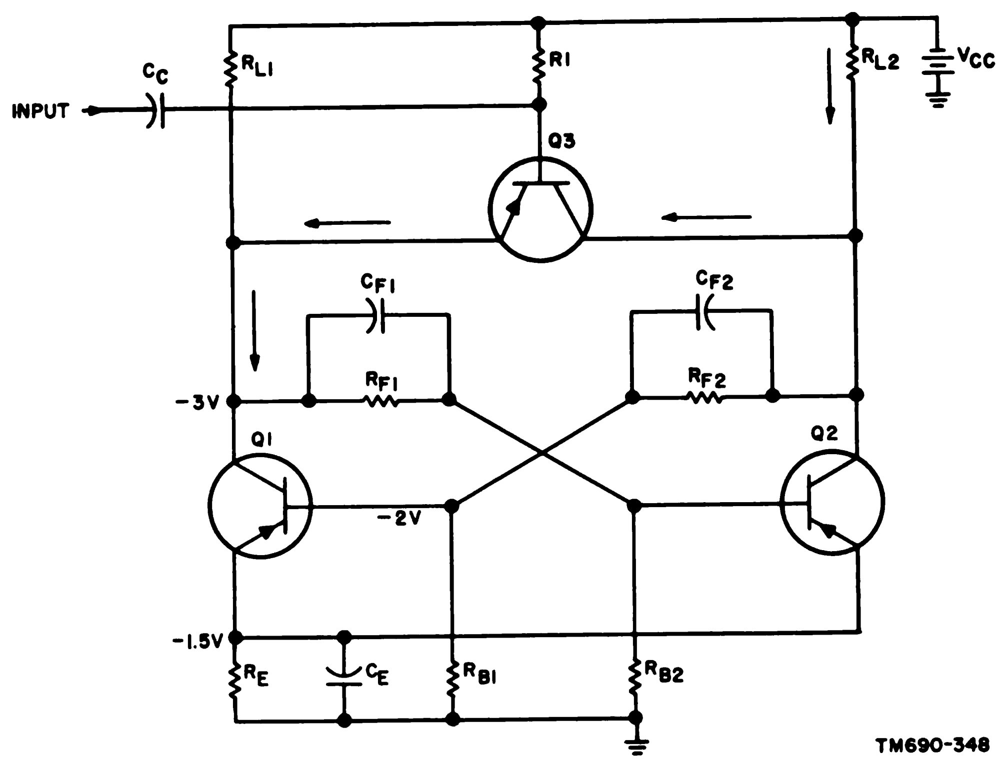


Figure 204. Bistable multivibrator, using transistor for trigger pulse steering and clamping.

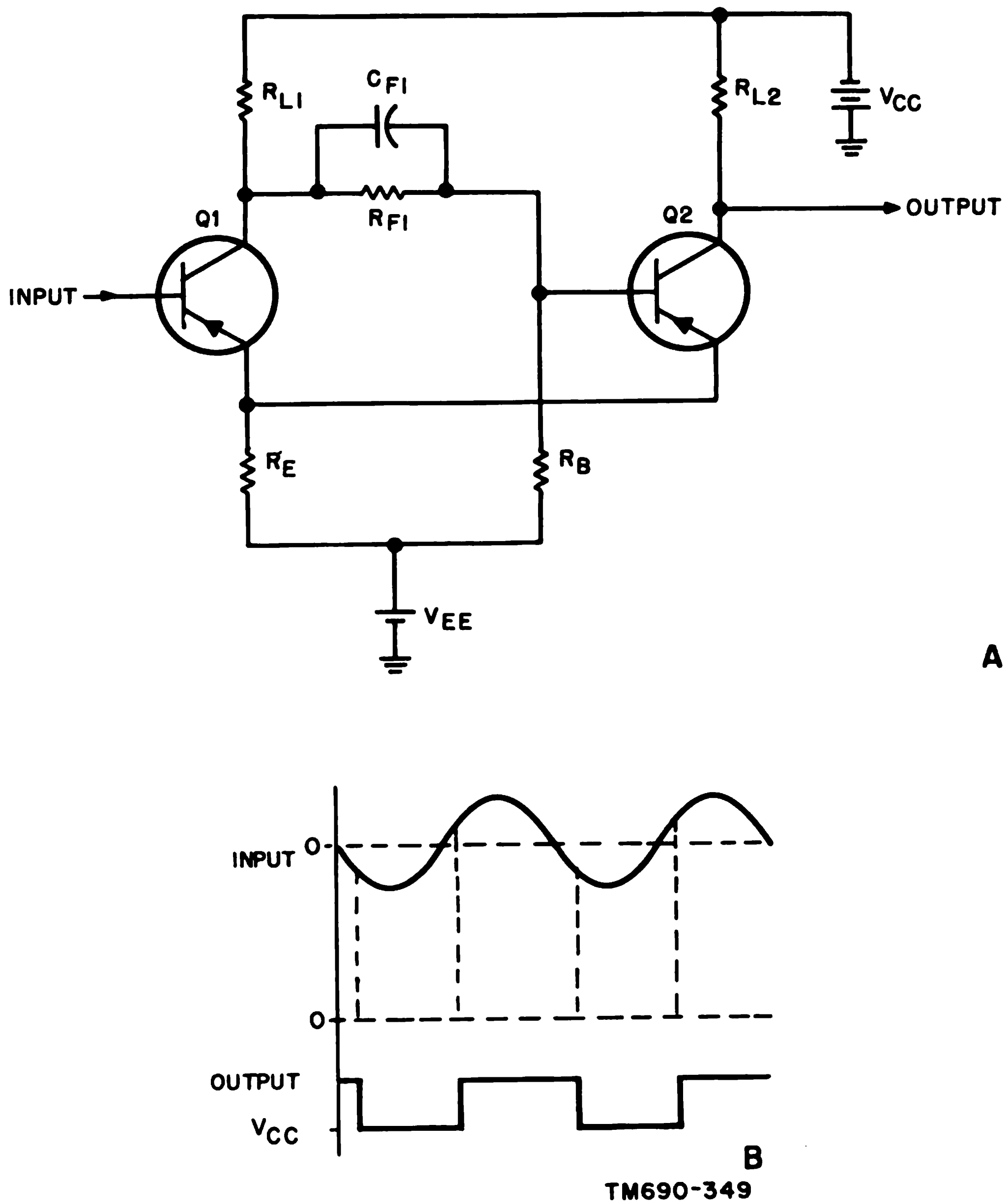


Figure 205. Squaring circuit (Schmitt trigger).

Section III. GATING CIRCUITS

197. General

a. Transistor gate circuits, used extensively (but not exclusively) in computer circuits, function as switches by making use of an effective open or short circuit between the emitter and collector (par. 186*b*). Transistors used in gate circuits may be connected in series, parallel, or series parallel to provide a variety of functions such as circuit triggering at prescribed intervals, and level and waveshape control.

b. Included in the overall category of gates are the *OR*, the *AND*, the series (or clamping), and the shunt or inhibitor circuits. The *NOT* or *negation* circuit refers to a circuit arrangement which provides pulse phase inversion which is an inherent characteristic of the common-emitter configuration. Combinations of these circuits result in a *NOT-OR* (NOR) gate or a *NOT-AND* gate. Because of the ability of the circuit to evaluate input conditions to provide a predetermined output, these circuits are also referred to as *logic* circuits. Typical gate circuits are described in the succeeding paragraphs. The gate circuits described make use of PNP-type transistors. NPN-type transistors function identically with all potentials of opposite polarity.

198. OR and NOR Gates

An *OR* gate has more than one input, but only one output. The *OR* gate provides a prescribed output condition when one or another prescribed input condition exists. When a negation feature (par. 197*b*) is directly incorporated in the circuit, through the use of the *CE* configuration, the *NOR* gate functions to provide a phrase inverted condition. Typical *OR* and *NOR* gates are described in *a* and *b* below.

a. Single-Transistor Circuits. Simple methods of obtaining *OR* and *NOR* gating functions are shown in figure 206. In both the *OR* gate ((1) below), using the *CB* configuration, and the *NOR* gate ((2) below), using the *CE* configuration (inverter), the high resistance of resistors *R*1 and *R*2 isolate one input source from the other. Resistors *R*1 and *R*2 are larger for the *CE* configuration because the input resistance is higher than that of the *CB* configuration. In each case resistor R_L develops the output wave form.

Note. Although only two input sources are illustrated for each configuration it is possible to have additional inputs.

(1) The *OR* gate (A, fig. 206) provides a positive output pulse (par. 185*c*) when a positive input pulse is applied at either input resistor. Initially, the transistor gate is in the cutoff condition with the emitter floating (par. 188*b*). The current flow through the collector load resistor consists only of the reverse-bias current (I_{CBO}). Since this small amount of current is negligible, the voltage level at the collector is effectively equal to battery voltage V_{CC}. When an input pulse of magnitude $+V_E$ is applied to either input, forward bias is established on the emitter-base junction and the transistor is driven into saturation. When the input pulse returns to the zero level, the transistor reverts from the saturated condition to the cutoff condition and the output voltage level swings back to $-V_{CC}$. Positive pulses applied to both inputs simultaneously results only in widening of the output pulse. This is due to the added emitter current which causes an increase in minority carrier storage in the base (par. 187*c*). An NPN-type transistor could be used as an *OR* gate with negative pulses and a negative output pulse.

(2) The *NOR* gate (B, fig. 206) functions similarly to the *OR* gate ((1) above). The *CE* configuration is used, providing the *NOT* or negation function. Base bias battery V_{BB} and resistor R_B provide the quiescent cutoff condition of the circuit. In the *CE* configuration, reverse current I_{CEO} is present (par. 188). The output voltage, in this condition, is nearly equal to the collector battery voltage ($-V_{CC}$). When a negative input pulse of magnitude $-V_B$ is applied to either input, the transistor is driven into saturation (output equals zero). When the input pulse is removed, the circuit reverts to its quiescent condition (output equals $-V_{CC}$). As in the *OR* gate ((1) above),

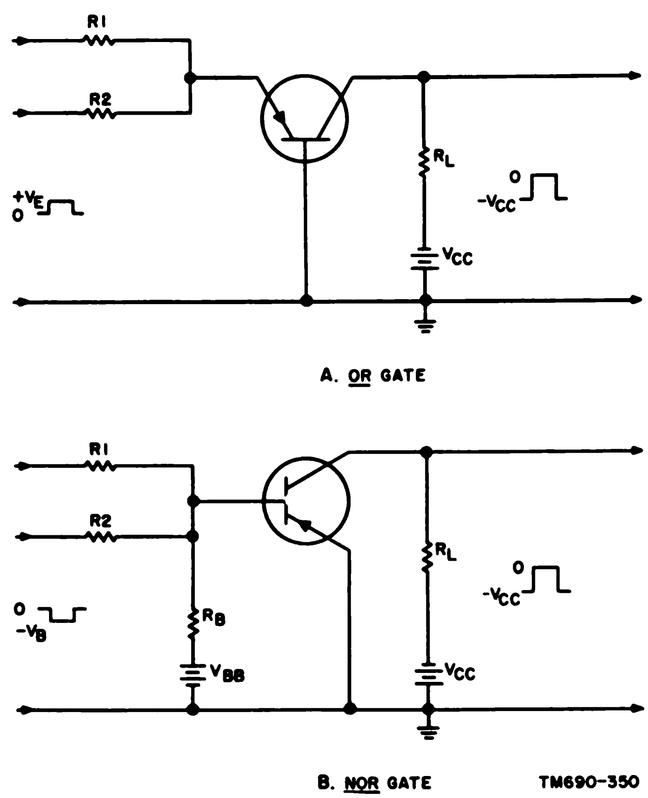
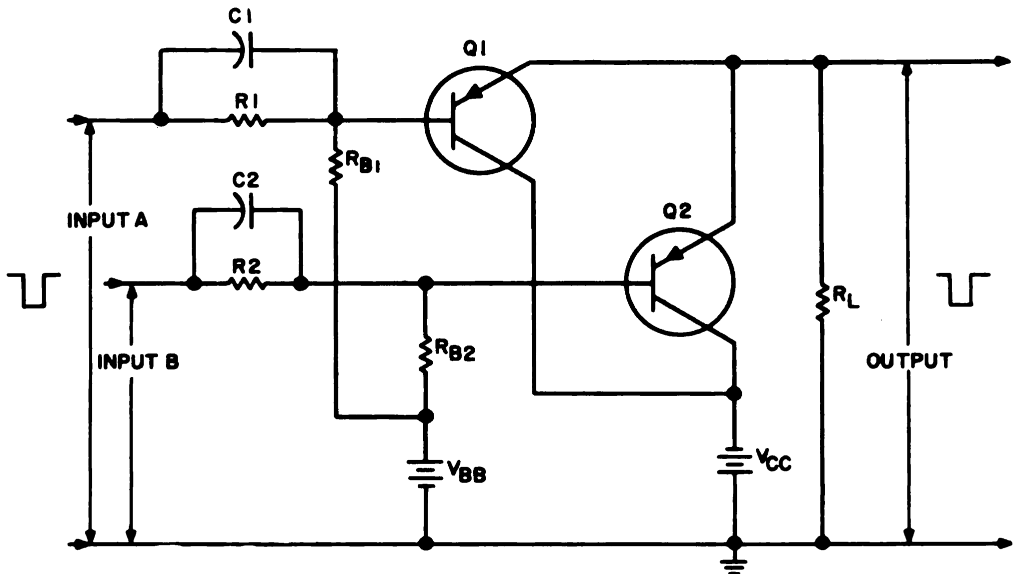


Figure 206. Single transistor OR gates with multiple inputs.

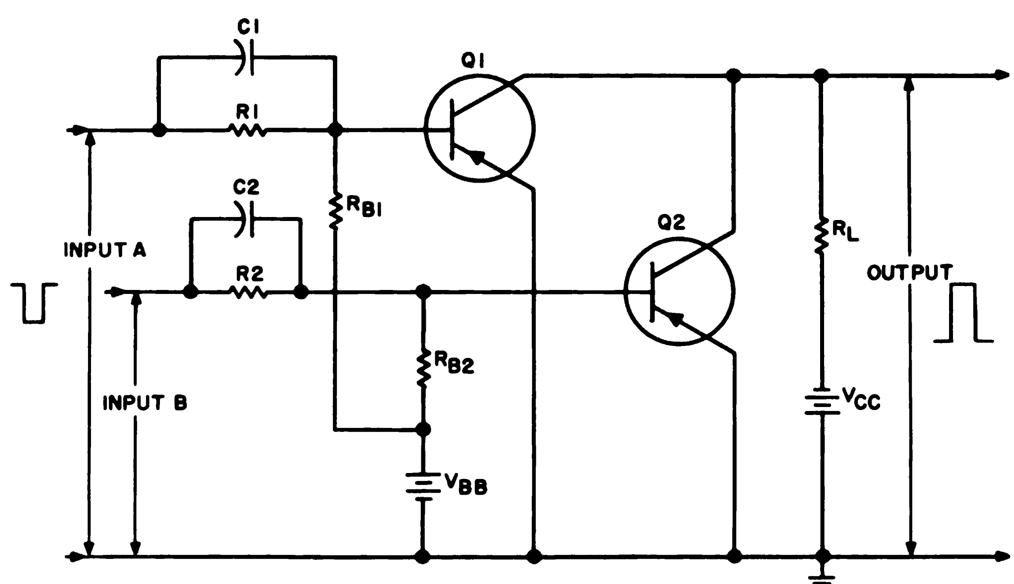
pulse widening is the result of more than one simultaneous input of proper polarity.

b. Two-Transistor Circuits. Additional gating circuits are shown in figure 207. In each case, the transistor outputs are connected in parallel and make use of a common load resistor R_L. At quiescence each transistor is held cutoff by battery V_{BB}. Resistors R_{B_1} and R_{B_2} aid resistors $R1$ and $R2$ in isolating the input sources from each other. Capacitors $C1$ and $C2$ provide a low impedance path for the transient currents (par. 193*a*) during turn-on and turn-off applications of the input pulses. This results in sufficient initial overdrive which reduces the rise and fall times of the output pulse.

(1) The *OR* gate (A, fig. 207) is a *CC* configuration (emitter follower) with high input and low output impedances. This circuit provides a negative output pulse when a negative input pulse is applied to *either* transistor $Q1$ or $Q2$. Battery V_{BB}



A. EMITTER FOLLOWER OR GATE



B. PARALLEL-CONNECTED NOR GATE TM690-351

Figure 207. Two transistor OR gates with multiple inputs.

provides reverse bias for both transistors and no current flows through load resistor R_L in the absence of an input signal. The output potential is at zero (ground). When a negative input pulse is applied at input A transistor $Q1$ is driven into conduction. The emitter voltage increases negatively due to the voltage drop across resistor R_L. The voltage drop across resistor R_L also increases the emitter-base reverse bias for transistor $Q2$, holding it at cutoff. If another negative pulse of equal amplitude is applied simultaneously at input B, it does not overcome the reverse bias on transistor $Q2$ and the transistor remains cutoff. If the input pulses are of unequal amplitude, the most negative pulse causes conduction of the transistor to which it is applied and the output voltage will follow the most negative input pulse. The emitter follower is rarely driven into saturation because of the high negative voltage feedback. Because saturation does not occur, the response of the emitter follower *OR* gate is more rapid than that of other configurations. This configuration is most desirable where speed of operation is of great importance.

(2) The *NOR* gate (B, fig. 207) provides a positive output pulse (par. 185*c*) when a negative input pulse is applied to *either* transistor $Q1$ or $Q2$. With no input applied, transistors $Q1$ and $Q2$ are cutoff by bias battery V_{BB}; the potential at the collectors equals collector battery voltage V_{CC}. When a negative pulse is applied at input A causing conduction of transistor $Q1$, the voltage drop across resistor R_L subtracts from battery voltage V_{CC}, and the output goes positive. When the signal is removed the output reverts to $-V_{CC}$. Pulse widening is determined by the amount of overdrive provided. Application of an additional pulse at input B does not cause large current flow in transistor $Q2$, because of the low collector voltage established by transistor $Q1$. Transistor $Q1$ collector voltage drops negligibly, therefore, as a result of the second pulse. Deep saturation of the collectors is thus avoided.

199. AND and NOT AND Gates

The operation of the *AND* gate is similar to that of the *OR* gate (par. 198). However, the *AND* gate provides an output when all the inputs are applied simultaneously. The circuit is also referred to as a *coincidence* circuit or *all* circuit. Phase inversion, accomplished through the use of the *CE* configuration, provides the negation feature (*NOT*) in *AND* gates as it did in the *NOR* gate. Typical *AND* and *NOT AND* gates are described in *a* and *b* below.

a. Single-Transistor Circuits. Isolating resistors $R1$ and $R2$ (fig. 208) are used with a single transistor in the *CB* and *CE* configurations. For each gate forward (saturation) bias is provided by batteries V_{EE} (*AND* gate) and V_{BB} (*NOT AND*) gate. Values for these circuit components are chosen so that saturating current continues to flow when only one input pulse is applied. Two input pulses are required to change the conducting state of the transistor. Clamping the collector (par. 189*a*) at a small negative voltage prevents deep saturation.

(1) The *AND* gate (A, fig. 208) provides a negative output pulse when negative

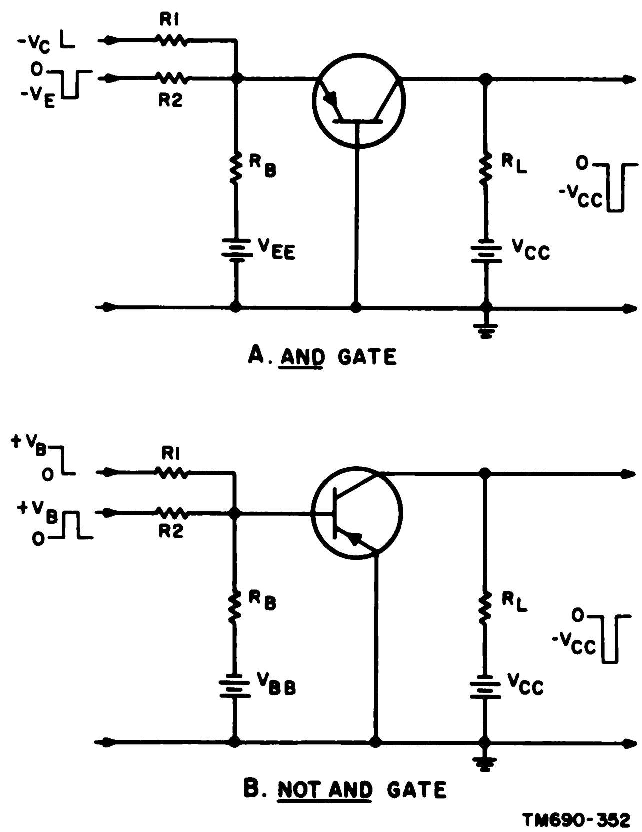


Figure 208. Single-transistor AND gates with multiple inputs.

pulses are applied at both inputs simultaneously. If a negative unit step voltage is applied to resistor $R1$ the decrease in base-emitter forward bias is not enough to cut off the transistor. With the application of the second negative voltage at the input of resistor $R2$, the base-emitter junction becomes reverse biased and the transistor is driven to cutoff. Note that the voltage drop across resistor R_B is caused by the *sum* of the two currents caused by the two negative input voltage drops. The output current falls to zero and the collector potential falls to its most negative value ($-V_{CC}$). When either pulse is removed the output potential rises to zero due to conduction of the transistor.

(2) The *NOT AND* gate (B, fig. 208) provides a negative output pulse when positive pulses are applied to both inputs simultaneously. The circuit operates in the same manner as the *AND* gate ((1) above) except that current amplification and phase inversion (*NOT*) are obtained.

b. Two-Transistor Circuits. Two gate circuits using two transistors with multiple inputs are shown in figure 209. Positive input pulses are required at inputs A and B to change the state of the output. The circuit configurations of the emitter follower *AND* gate and the parallel-connected *NOT AND* gates are similar to the *OR* and *NOR* gates respectively. The only difference is in the initial input bias condition provided by battery V_{BB}; forward bias is used in the *AND* ((1) below), and *NOT AND* ((2) below), gates. Positive input pulses are required to change the state from conduction to cutoff and back to conduction.

(1) When only one positive input pulse is applied to the *AND* gate (A, fig. 209) the transistor to which it is applied is cutoff. Since the other transistor remains in conduction, current continues to flow through the common load resistor R_L, and the output voltage level remains relatively constant. Only by applying a positive pulse to input A and input B simultaneously does the current flow through and the voltage drop across resistor R_L drop to near zero

(2) The *NOT AND* gate (B, fig. 209) operates in the same manner as the *AND* gate ((1) above) except that amplification and phase inversion (*NOT*) are obtained.

200. AND-OR Gate or Triggering

The *AND-OR* gate (fig. 210) illustrates the use of a direct-coupled transistor logic circuit (par. 193*b*) to trigger a bistable multivibrator. The CE configuration for each transistor causes pulse phase inversion (negation feature) and the gating function provided is *NOT AND-NOR*. For simplicity, however, the circuit is most often called the *AND-OR* gate. The gating functions and the triggering application are described in *a* and *b* below.

a. Gating Function. Transistors Q1, Q2, and Q3 perform the gating function; transistor Q4 is a part of a bistable multivibrator. The overall gating function consists of a *NOT AND* function and a *NOR* function described in (1) and (2) below.

(1) *NOT AND function.* Transistors Q1 and Q2 are series connected; transistors Q1 and Q3 are also series connected. Each combination forms a *NOT AND* gate (par. 199). Reverse collector bias is provided by battery V_{CC}.

(2) *NOR function.* Transistors $Q2$ and $Q3$ are parallel-connected and form a *NOR* gate (par. 198). The load for the *NOR* gate consists of transistor $Q1$ and load resistor R_L. Since transistor $Q1$ is a part of the *NOT AND* function, and represents an open switch when it is cutoff, the output is controlled by the *NOT AND* function (*b* below).

b. Triggering Application. Transistor $Q4$ is a switching element of a bistable multivibrator (par. 191). Collector bias for this transistor is provided by battery V_{CC}. Assuming all transistors are cutoff (quiescent condition), triggering of the bistable multivibrator is accomplished when the prescribed input conditions of the *AND-OR* gate are met ((1) and (2) below).

(1) *Off condition.* When all the transistors are cutoff, the current through load resistor R_L is zero and the output potential and the potential at the collector of transistor $Q4$ is negative and equal to battery

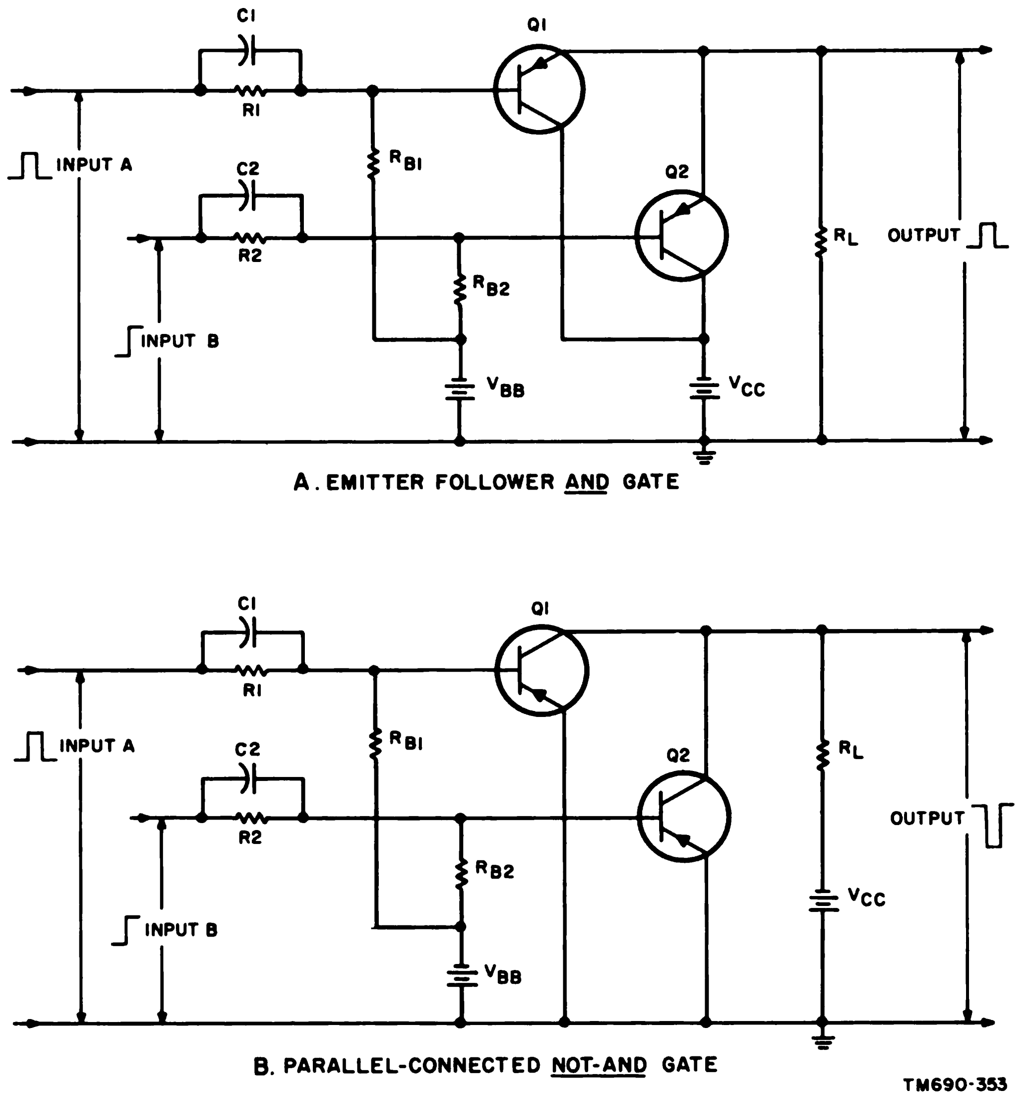


Figure 209. Two-transistor AND gates with multiple inputs.

voltage V_{CC}. The bases of transistors $Q1$, $Q2$ and $Q3$ are held at ground potential through connection to preceding saturated transistors. Transistor $Q4$ is held at cutoff by the second transistor (not shown) of the bistable multivibrator.

(2) *On condition.* The gate is opened when either transistors $Q1$ and $Q2$ or transistors $Q1$ and $Q3$ conduct. When the bases of transistors $Q1$ and $Q2$ are driven sufficiently negative, each transistor is driven to saturation. The series path through the conducting transistors is effectively short-circuited (low resistance) and current flows through load resistor R_L from battery V_{CC}. The output voltage rises from the high negative value ((1) above) to approximately zero. Since the potential at the collector of the *off* transistor $Q4$ provides the base drive for the *on* transistor (not shown) in the multivibrator, the positive going pulse at the output of the *AND-OR* gate results in driving transistor $Q4$ (the original *off* transistor in the multivibrator) into saturation.

201. Series Gating Circuits

Gating circuits may be used as amplitude discriminators (limiters), clippers, and clamping circuits. Some typical circuits, connected in the *CB* configuration, and acting as series switches, are described in *a* and *b* below.

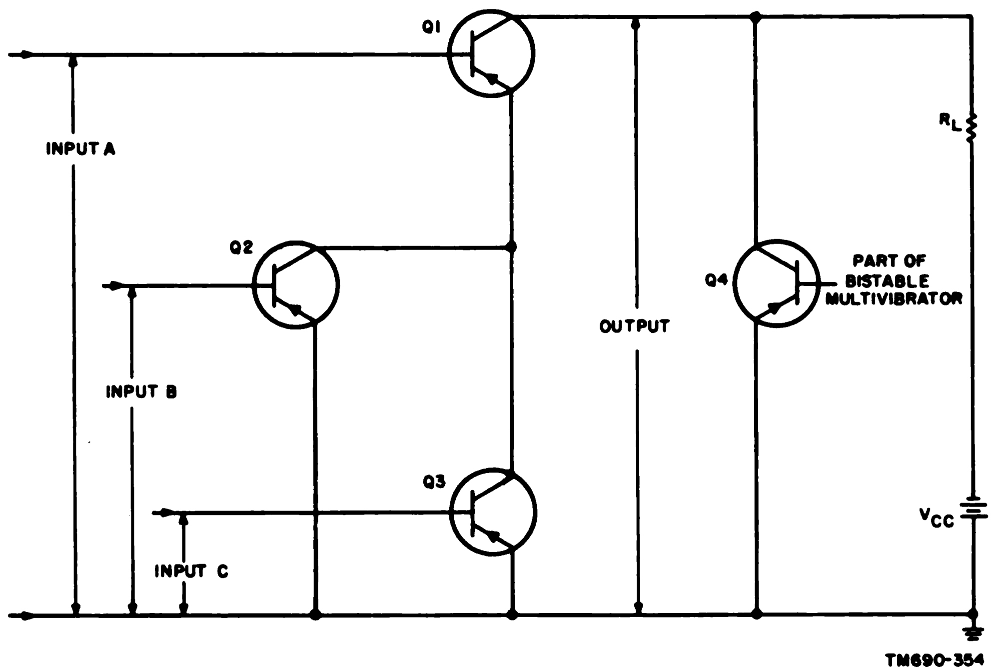


Figure 210. AND-OR combined gate.

a. Amplitude Discriminator. A, figure 211 shows a circuit that will produce an output pulse only when the input pulse is of sufficient magnitude (greater than V_{BB}) to drive the transistor into conduction.

(1) Battery V_{BB} reverse biases both the emitter and collector. In the quiescent condition no current flows through load resistor R_L and output voltage V_C is zero (prior to time $t1$ (B, fig. 211)).

(2) When a positive pulse, larger in magnitude than V_{BB} is applied, the base-emitter junction becomes forward biased and the transistor conducts. Current flows through resistor R_L and output voltage V_C rises at $t1$. When the input is removed, the output voltage tends to decrease to zero at $t2$; after a small storage delay the output voltage is zero.

(3) If the input pulse is too large, resistor R_B will limit the input current and resistor R_B will establish a small reverse bias for the base-emitter junction and deep saturation will be prevented. A shunting capacitor across resistor R_B and battery V_{BB} will improve the fall time of the pulse. Amplification may be obtained if load resistor R_L is made larger than input resistor R_B.

b. Clamping Circuits. In addition to providing amplitude discrimination (*a* above), the circuits shown in C and E, figure 211, provide constant high and low voltage levels for the output pulse. The collector may be negative (C, fig. 211) or positive (E, fig. 211). In either case battery V_{BB} is made large enough to reverse bias the collector and emitter. In the quiescent state, both circuits are at cutoff. A positive pulse, sufficient to overcome battery voltage V_{BB}, must be applied to produce an output.

Note. **High level clamping will occur only when the transistor is driven into saturation.**

(1) When the transistor (C, fig. 211) is cutoff (quiescent condition), no current flows through load resistor R_L and output voltage V_C is equal to battery voltage $-V_{CC}$. When a positive pulse is applied, sufficient to drive the transistor into saturation, the base-collector resistance is reduced to a very low value and output voltage V_C rises approximately to battery voltage V_{BB} ($t1$ D, fig. 211). When the input is removed ($t2$) the transistor returns to cutoff and output voltage V_C will again equal battery voltage $-V_{CC}$.

(2) The clamping circuit shown in E, figure 211, operates in the same manner as (1)

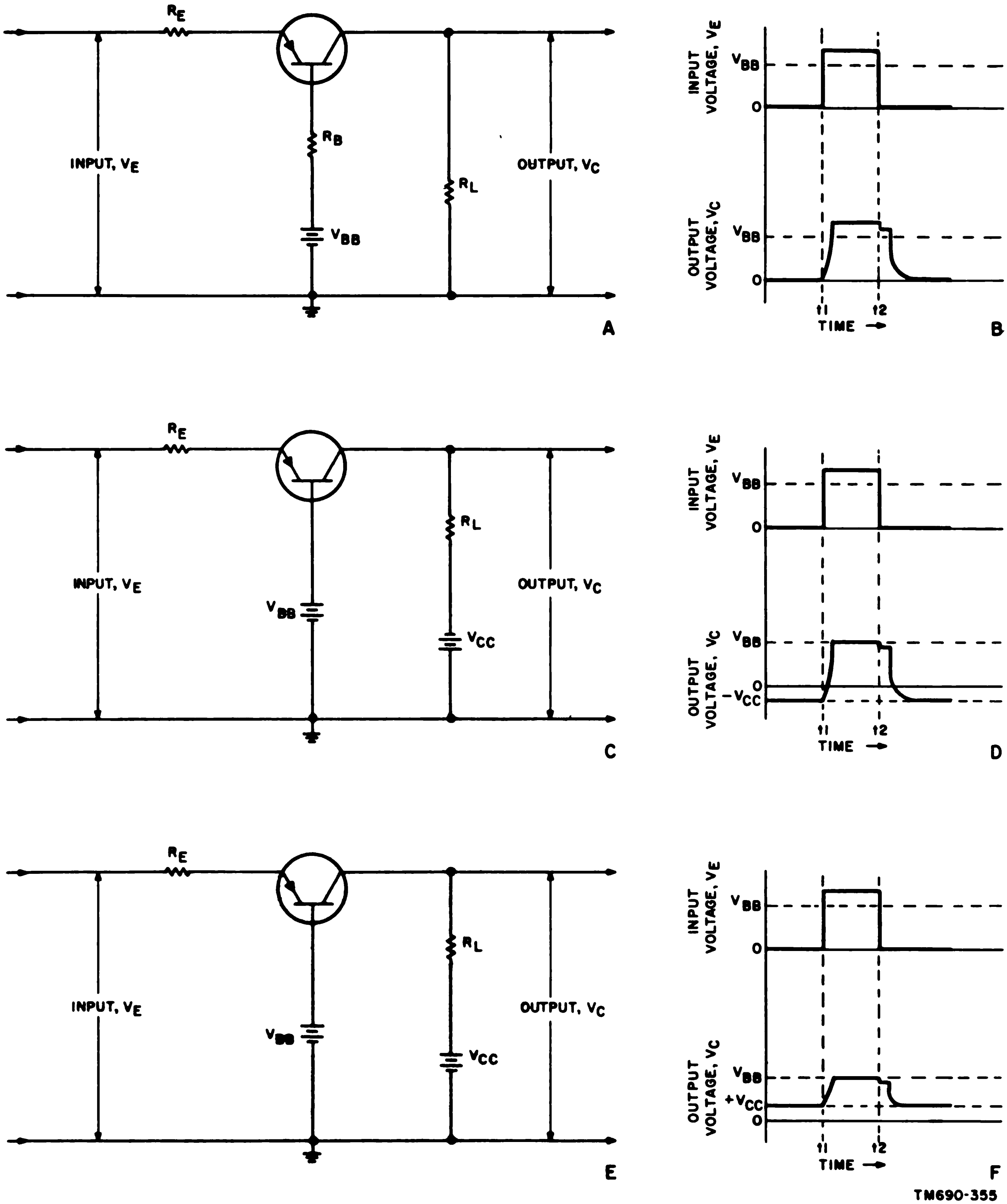


Figure 211. Series gating circuits.

above. The only difference is that when the transistor is at cutoff (quiescent condition) the output level is equal to $+V_{CC}$ and the amplitude of the output pulse is smaller than that of (1) above.

202. Shunt Gating Circuits (Inhibition Gates)

The *CE* configuration may be used as a signal shunting or transmission gate; this configuration introduces a minimum of transient distortion. A control voltage is used to actuate the transistor which acts as a shunt switch. The signal voltage is either bypassed around the switch to the load (switch open), or shunted around the load (switch closed). Signal amplification or phase inversion does not occur in the shunt gate (*a* and *b* below). Typical shunt gates (also called ***inhibition gates*** or ***inhibitors***) are shown in A and C, figure 212.

a. Reverse-Biased Shunt Gate. The emitter-base junction of the shunt gate (A, fig. 212) is

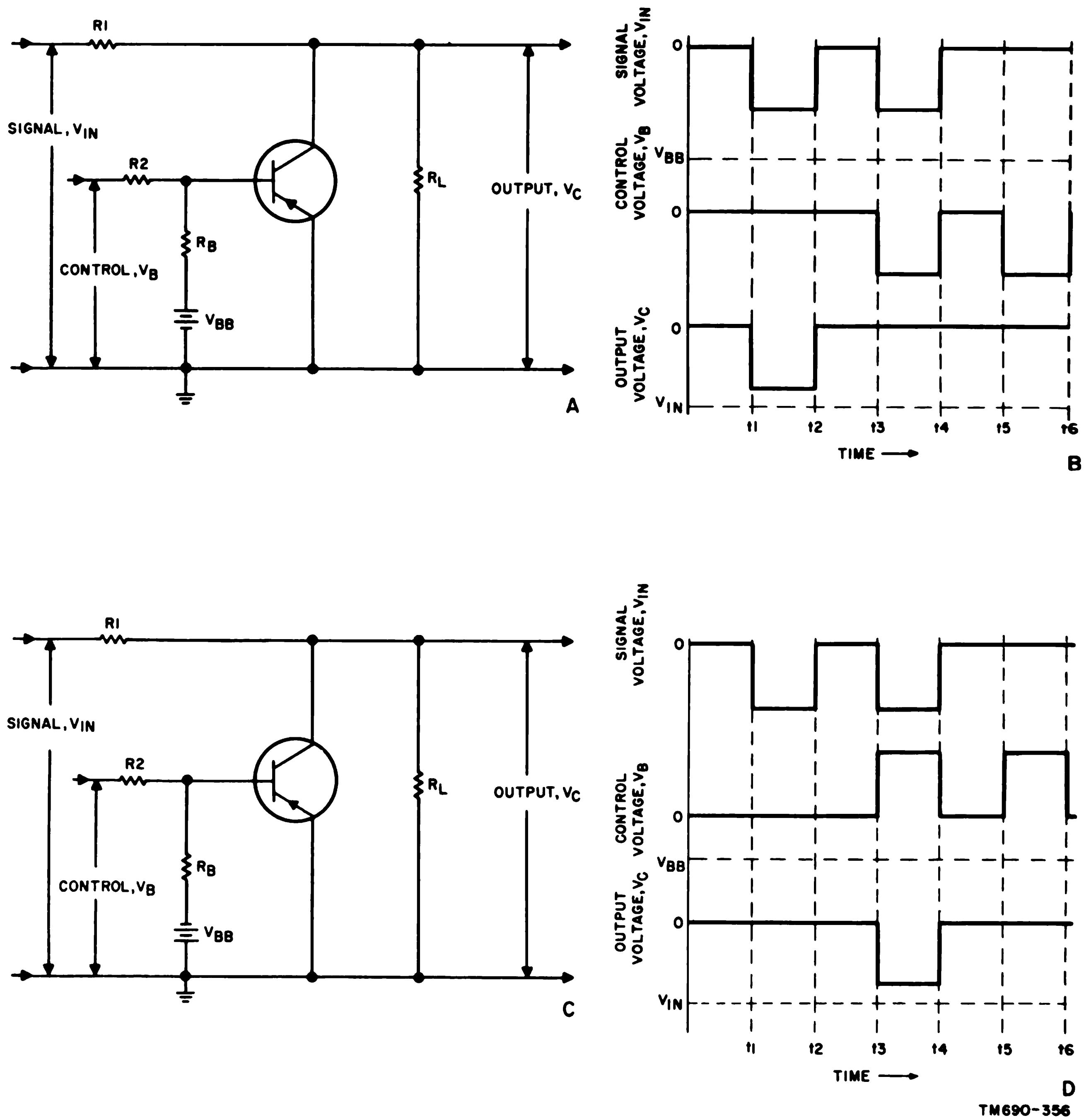


Figure 212. Reverse and forward-biased shunt (inhibition) gates.

reversed biased by battery V_{BB}; no collector bias voltage is applied.

(1) A signal occurs across resistor R_L (voltage V_C) only when a signal voltage V_{IN} is introduced and the transistor remains nonconducting (open switch); that is, no forward-biasing pulse (control voltage V_B) is introduced. This condition occurs between times $t1$ and $t2$ (B, fig. 212).

(2) No output signal (V_C) occurs when a forward-biasing pulse (V_B) is introduced even though a signal voltage V_{IN} is introduced. With the forward-biasing pulse assumed to be large enough to drive the transistor into saturation (closed switch), resistor R_L is effectively shorted to ground. This condition occurs between times $t3$ and $t4$.

(3) No output signal (V_C) occurs even if a forward-biasing pulse (V_B) is introduced and no signal pulse (V_{IN}) is introduced, because there is no bias voltage on the collector. This condition occurs between times $t5$ and $t6$.

b. Forward-Biased Shunt Gate. The emitter-base junction of the shunt gate (C, fig. 212) is forward biased by battery V_{BB}; no collector bias voltage is applied. With no input signal no collector current flows and the transistor acts as an open switch.

(1) When only signal voltage V_{IN} is introduced, this voltage acts as a collector voltage, the transistor conducts heavily (saturates) and acts as a closed switch across resistor R_L. No output (V_C) is produced. This condition occurs between times $t1$ and $t2$ (D, fig. 212).

(2) An output voltage (V_C) occurs when a signal voltage (V_{IN}) and a control voltage (V_B) are introduced simultaneously. The control voltage (V_B) drives the transistor into cutoff (open switch) and a portion of signal voltage (V_{IN}) is developed across resistor R_L. This condition occurs between times $t3$ and $t4$.

(3) When only control voltage (V_B) is introduced, no output voltage (V_C) is developed. The control voltage reverse biases the base-emitter junction and the transistor continues to appear as an open switch (no collector current). This condition occurs between times $t5$ and $t6$.

203. Summary

a. Switching circuits operate as generators, pulse amplifiers, inverters, and wave shapers.

b. Switching circuits perform limiting, triggering, gating, and signal routing functions.

c. Pulse and switching circuits are characterized by large-signal operation.

d. A unit step voltage (or current) represents a voltage (or current) which undergoes an instantaneous change in amplitude from one established level to another.

e. Large-signal operation involves operation in the cutoff, active and saturation regions of the transistor characteristics.

f. The cutoff and saturation regions of a switching circuit are the quiescent and stable regions of operation.

g. The active region of a switching circuit is the unstable or transient region of operation.

h. When the emitter-base and collector-base junctions are reverse-biased, the transistor is in the nonconducting or *off* (cutoff) state. When both junctions are forward-biased, the transistor is in a conducting or *on* (saturated) state.

i. When a switching circuit is *on*, both input and output resistances are at a minimum. When a switching circuit is *off*, both input and output resistances are at a maximum.

j. The important characteristics of an output pulse are the rise time, pulse time, storage time, and fall time.

k. Clamping diodes are used to reduce the undesirable effects on the output wave form of a switching circuit. The undesirable effects are produced by driving a transistor to saturation and cutoff.

l. The monostable multivibrator has only one stable or quiescent state (cutoff *or* saturation). The bistable multivibrator has two stable or quiescent states (cutoff *and* saturation).

m. The Schmitt trigger provides a square wave output for a sine wave input.

n. Gating circuits, provide *OR* and *AND* functions (and variations of both). These circuits are commonly called logic circuits.

o. The CB configuration is used to provide series gating functions; the CE and CC configurations are used to provide shunt gating functions.

CHAPTER 12

MODULATION, MIXING, AND DEMODULATION

Section I. MODULATION

204. General

The three basic types of modulation are *amplitude modulation*, *frequency modulation*, and *phase modulation*. Since frequency and phase modulation are very similar, only amplitude modulation (am) and frequency modulation (fm) will be discussed in this chapter. Amplitude modulation can be established by applying a carrier signal to an amplifier and varying the amplifier bias at the modulating rate (par. 205*a*).

a. Amplitude modulation can also be established by varying the bias of an oscillator at the modulating rate (par. 205*b*). Various methods of amplitude modulating an amplifier (par. 206) or an oscillator (par. 207) may be employed.

b. Frequency modulation can be established by varying the frequency of an oscillator at the modulating rate (par. 209). Various methods of frequency modulating an oscillator (par. 210) may be employed.

205. Establishing Amplitude Modulation

a. Amplitude-Modulated Amplifier. The waveform analysis of an amplitude-modulated amplifier is shown in A, figure 213. The carrier signal is applied to the amplifier and the bias, which varies the gain of the amplifier, is varied by the modulating signal input. The output of the amplifier is an amplitude-modulated carrier. As the carrier signal passes through the amplifier, the gain of the amplifier is increased or decreased. When the amplifier gain is increased, the output is increased and when the amplifier gain is decreased, the output is decreased. Thus, the amplitude of the carrier signal is varied at the modulating rate.

b. Amplitude-Modulated Oscillator. The waveform analysis of an amplitude-modulated oscillator is shown in B, figure 213. The carrier signal is generated by the oscillator and the modulating signal varies the bias. Changing the bias of the oscillator changes the gain and the operating point of the transistor. Changing the bias provides amplitude changes and frequency changes in the output of the oscillator. An amplitude-modulated oscillator may be used in an am transmitter (par. 208*b*) or an fm transmitter (par. 210).

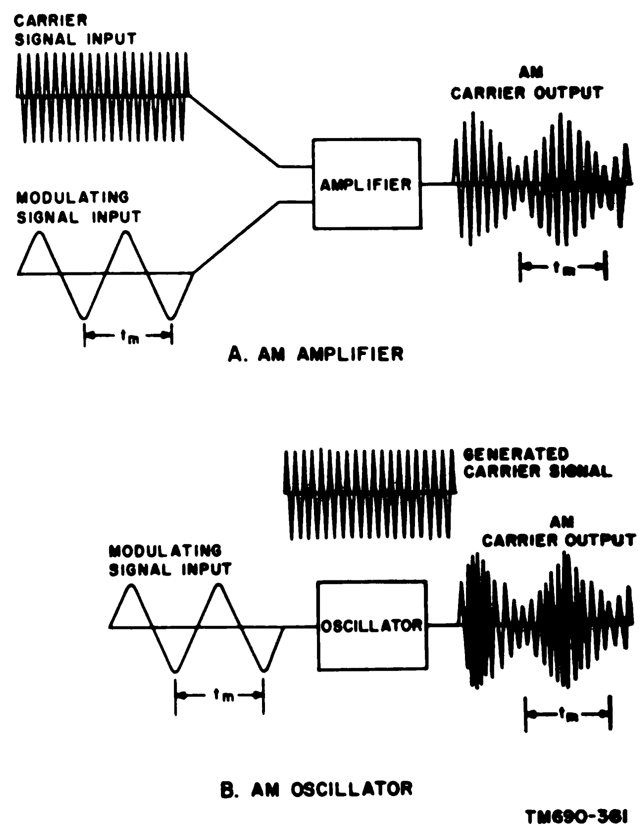


Figure 213. Amplitude modulation of an amplifier and an oscillator.

206. Methods of Amplitude Modulating an Amplifier

a. General. An amplifier may be am modulated by injecting a modulating signal into the base, the emitter, or the collector (fig. 214). Each method will vary the gain on the transistor. Transformer T1A, T1B, or T1C provides base, emitter, or collector injection, respectively. Any change in bias changes the gain of the transistor which results in amplitude variations in the output signal. The carrier signal, coupled through transformer T1, is applied to the base circuit of amplifier Q1. Capacitor C1 and the primary of transformer T2 form a parallel resonant circuit for the amplified carrier signal. Any of the modulating methods described above vary the gain of the transistor with respect to the modulating signal. The signal on the collector is increased in amplitude when the bias is increased and decreased in amplitude when the bias is decreased. The result is an amplitude-modulated carrier coupled through transformer T2 to the following stage. If the strength of the rf carrier signal were greater than the emitter-base bias, when the amplifier is modulated by base injection or emitter injection, the modulated output would be distorted. Under these conditions, the rf carrier signal would determine the emitter-base bias instead of the modulating signal. Thus, the strength of the rf carrier must always be smaller than the emitter-base bias when an amplifier is modulated with base injection or emitter injection. Satisfactory modulation with base injection (*b* below) or emitter injection (*c* below) is obtained only with low-level modulation. Satisfactory modulation with collector injection (*d* below) is obtained with low-level or high-level modulation.

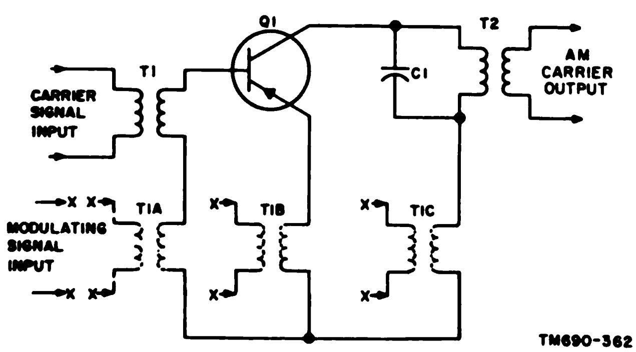


Figure 214. Methods of injecting amplitude modulating signal into amplifier circuit.

b. Amplitude-Modulated Amplifier (Base Injection)

(1) Figure 215 shows an amplifier circuit with the modulating signal injected into the base circuit. The carrier signal input is coupled through transformer T1 to the base circuit of amplifier Q1. The modulating signal, coupled through capacitor C1, is developed across resistor R1. The bias developed across resistor R1 is increased when the modulating signal increases and decreases when the modulating signal decreases. Thus, the base-emitter bias is increased and decreased as the modulating signal increases and decreases. Since the base-emitter bias is changing at the modulating rate, the collector current and the gain are changing at the modulating rate. As the gain increases, the amplitude of the carrier signal increases and as the gain decreases, the amplitude of the carrier signal decreases. The amplitude-changing carrier signal (modulated carrier signal) present in the parallel resonant circuit of capacitor C4 and the primary of transformer T2 is coupled through the transformer to the output.

(2) Base-emitter bias is developed across resistor R1. Resistor R2 is a voltage dropping resistor. Capacitor C2 is a bypass capacitor for the carrier signal. Resistor R3 is the emitter swamping resistor and capacitor C3 is a bypass capacitor. Capacitor C4 and the primary of transformer T2 form a parallel resonant circuit for the amplified carrier signal. Capacitor C1 is a coupling capacitor for the modulating signal.

c. Amplitude-Modulated Amplifier (Emitter Injection).

(1) Figure 216 shows an amplifier circuit with the modulating signal injected into the emitter circuit. Emitter injection is very similar to base injection, since both methods vary the emitter-base bias. The carrier signal input is coupled through transformer T1 to the base circuit of amplifier Q1. The modulating signal, coupled through capacitor C1, is developed across swamping resistor R3. When

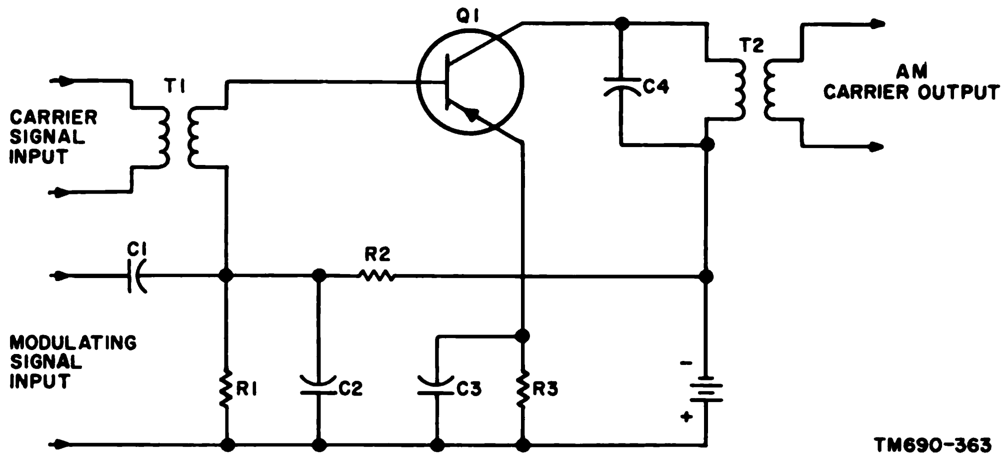


Figure 215. Amplitude-modulated amplifier with base injection.

the voltage drop (modulating signal voltage) across resistor R3 aids the voltage drop across series resistor R1, the emitter-base bias increases. When the voltage drop (modulating signal voltage) across resistor R3 opposes the voltage drop across series resistor R1, the emitter-base bias decreases. When the emitter-base bias increases, the gain of amplifier Q1 increases and when the bias decreases, the gain decreases. As the gain increases, the amplitude of the carrier signal on the collector increases and when the gain decreases, the amplitude of the carrier signal decreases. The modulated carrier signal present in the parallel resonant circuit of capacitor C4 and the primary of transformer T2 is coupled through the transformer to the output.

(2) The emitter-base bias is developed across resistor R1. Capacitor C2 is an audio and rf bypass capacitor. Capacitor C1 couples the modulating signal to the emitter circuit. Resistor R3 is the emitter swamping resistor and capacitor C3 is a bypass capacitor for the carrier signal. Capacitor C4 and the primary of transformer T2 form a parallel resonant circuit for the carrier signal.

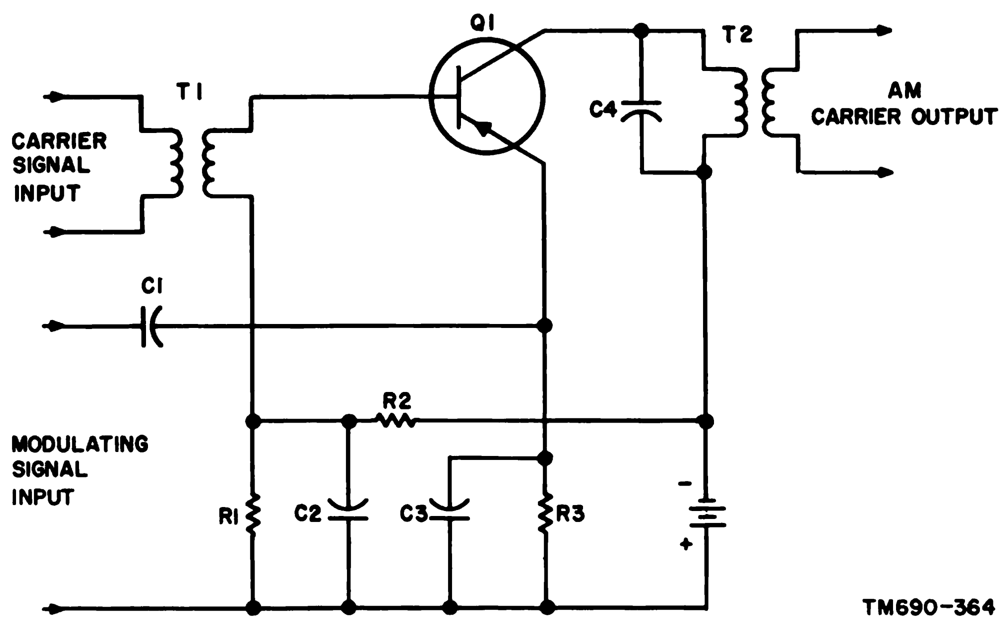


Figure 216. Amplitude-modulated amplifier with emitter injection.

d. Amplitude-Modulated Amplifier (Collector Injection).

(1) Figure 217 shows an amplifier circuit with the modulating signal injected into the collector circuit. In this circuit, the emitter-base bias is constant and the voltage between the emitter and the collector is varied at the modulating rate. This condition permits satisfactory low-level or high-level modulation. The carrier signal is applied through transformer $T1$ to the base circuit. Amplifier $Q1$ amplifies the carrier signal. The carrier signal is developed across the resonant circuit consisting of capacitor $C2$ and the primary of transformer $T2$. The modulating signal is applied through transformer $T3$ to the collector circuit. The modulating voltage across the secondary of transformer $T3$ is in series with collector battery voltage V_{CC}. The modulating voltage, series aiding battery voltage V_{CC}, increases the emitter-collector voltage. The modulating voltage, series opposing battery voltage V_{CC}, decreases the emitter-collector voltage. As the emitter-collector voltage increases, the gain of amplifier $Q1$ increases and as the emitter-collector voltage decreases, the gain decreases. Thus, when the gain of amplifier $Q1$ increases, the amplitude of the carrier signal increases, and when the gain decreases, the amplitude decreases. The amplitude-modulated carrier signal present in the parallel resonant circuit (capacitor $C2$ and the primary of transformer $T2$) is coupled through transformer $T2$ to the output.

(2) Transformer $T1$ is the input coupling transformer. Resistor $R1$ is the emitter swamping resistor and capacitor $C1$ is a bypass capacitor. Battery V_{EE} provides emitter-base bias. Capacitor $C2$ and the primary of transformer $T2$ form a parallel resonant circuit for the carrier signal. Transformer $T2$ provides output coupling. Transformer $T3$ provides input coupling for the modulating signal. Capacitor $C3$ is a bypass for the carrier signal and battery V_{CC} provides emitter-collector bias.

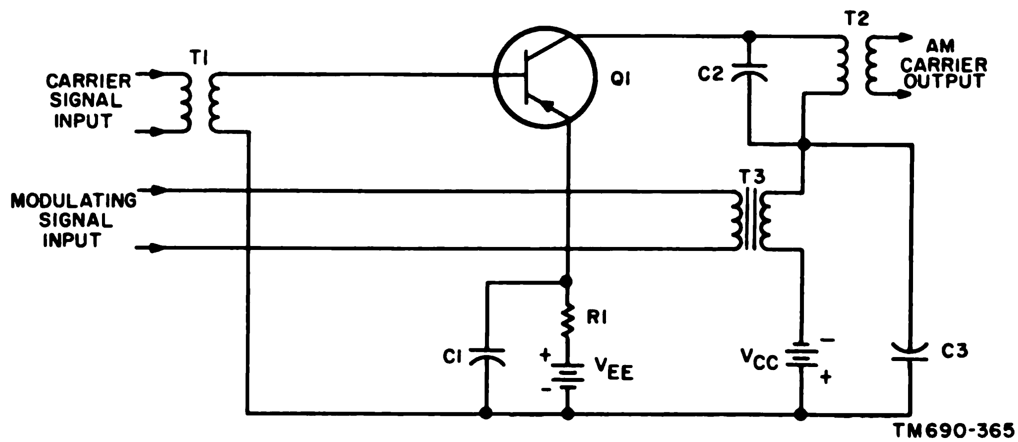


Figure 217. Amplitude-Modulated amplifier with collector injection.

207. Methods of Amplitude Modulating an Oscillator

a. General. An oscillator may be modulated by injecting a modulating signal into the base, the emitter, or the collector (fig. 218). Each method will change the gain and the operating point of the transistor. This results in amplitude changes and frequency changes in the output of the oscillator (par. 171). Transformer T1A, T1B, or T1C provides base, emitter, or collector injection, respectively. Amplifier Q1 provides the amplification required for oscillation. Winding 3–4 of transformer T2 provides the required feedback. Capacitor C1 and winding 1–2 of transformer T2 provide the required resonant circuit. Winding 5–6 of transformer T2 couples the output signal to the following stage. The dc circuits are not shown. Detailed circuit analysis

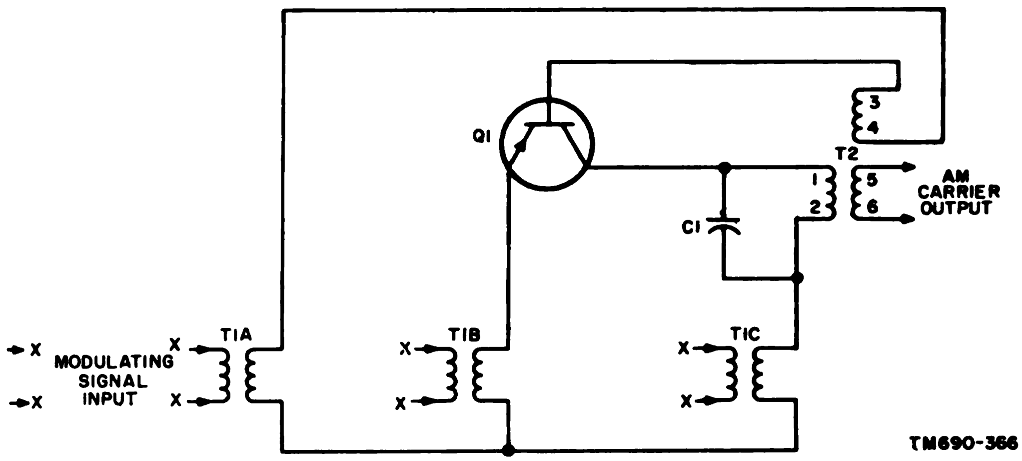


Figure 218. Methods of injecting amplitude-modulating signal into oscillator circuit.

of oscillator operation is covered in paragraph 171.

b. Amplitude-Modulated Oscillator.

(1) An amplitude-modulated oscillator with the modulating signal injected into the base circuit is shown in figure 219. The modulating signal voltage from microphone $M1$ is developed across variable resistor $R1$. The amount of modulating signal voltage, determined by the setting of variable resistor $R1$, is applied to the base circuit. Choke $L1$ and capacitor $C1$, series resonant at the carrier frequency, act as an open circuit for the modulating signal voltage and bypasses the carrier signal around resistor $R1$. The modulating signal voltage in series with the feedback voltage (winding 3–4 of transformer $T1$) is in parallel with emitter-base bias resistor $R2$. An aiding modulating signal voltage increases the emitter-base bias and an opposing modulating signal voltage decreases the emitter-base bias. An increase in the bias increases the gain of oscillator $Q1$ and a decrease in the bias decreases the gain. An increase in the gain of transistor $Q1$ increases the amplification of the generated signal and a decrease in the gain decreases the amplification. The generated signal is therefore amplitude modulated at the modulating rate. An increase in gain also decreases the collector voltage, which increases capacitance C_{CB} and a decrease in the gain increases the collector voltage which decreases capacitance C_{CB}. Capacitance C_{CB} is in parallel with the resonant circuit consisting of capacitor $C3$ and winding 1–2 of transformer $T1$. Increasing capacitance

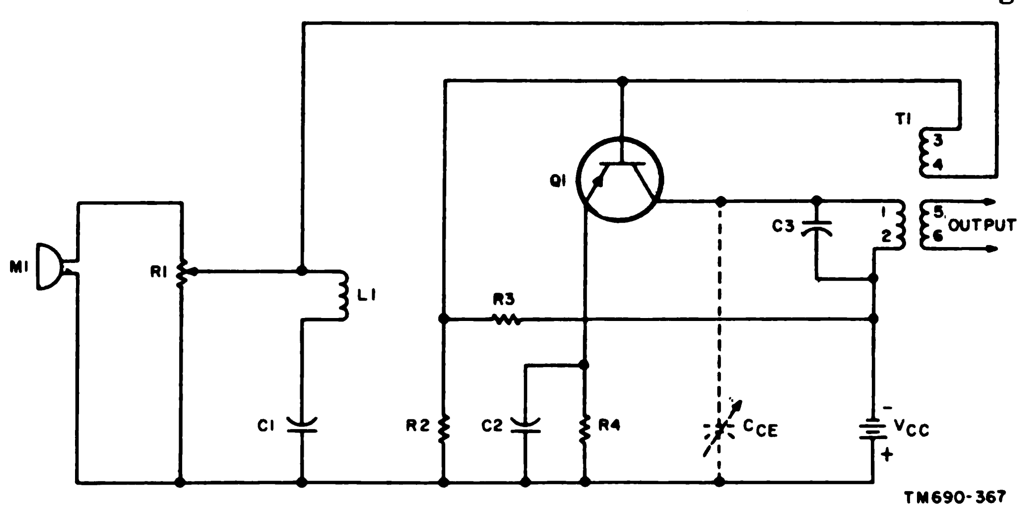


Figure 219. Amplitude-modulated oscillator with base injection.

C_{CE} decreases the resonant frequency of the resonant circuit and decreasing capacitance C_{CE} increases the resonant frequency. Since the frequency of oscillation is determined by this resonant circuit, the output frequency of the oscillator is changing at the modulating rate. Thus, the output signal of the amplitude-modulated oscillator is amplitude and frequency modulated at the modulating rate.

(2) Microphone $M1$ converts sound energy into an electrical signal. Variable resistor $R1$ determines the amount of audio signal (modulating signal voltage) applied to the base circuit. Choke $L1$ and capacitor $C1$ form a series resonant circuit for the oscillator frequency. Resistors $R2$ and $R3$ form a voltage divider network to establish emitter-base bias. Resistor $R4$ is the emitter swamping resistor and capacitor $C2$ is an audio bypass capacitor. Transistor $Q1$ provides the amplification necessary for oscillations. Winding 3–4 of transformer $T1$ provides the required feedback. Capacitor $C3$ and winding 1–2 of transformer $T1$ provides the resonant circuit. Capacitor C_{CE} represents the output capacitance between the emitter and collector (par. 103). Winding 5–6 of transformer $T1$ provides output coupling. Battery V_{CC} provides the operating voltage.

208. Use of Amplitude-Modulated Amplifier or Oscillator

a. Amplitude-Modulated Amplifier. The requirements for transmitting an amplitude modulated carrier using transistors are the same as those for electron tubes (TM 11–665). Figure 220 shows a block diagram of a typical transmitter using amplitude-modulated amplified waveforms of each stage. This type of modulation is normally employed in transmitters requiring a large power output. The oscillator generates the fundamental carrier frequency which is increased by the multiplier stage. The driver stage increases the amplitude of the carrier frequency to provide sufficient driving power for final amplification. The amplitude modulated amplifier stage is employed as the power amplifier. The microphone converts sound energy into an electrical signal. The audio amplifier amplifies the signal sufficiently to drive the modulator stage. The modulator stage amplifies the audio sufficiently to modulate the power amplifier stage.

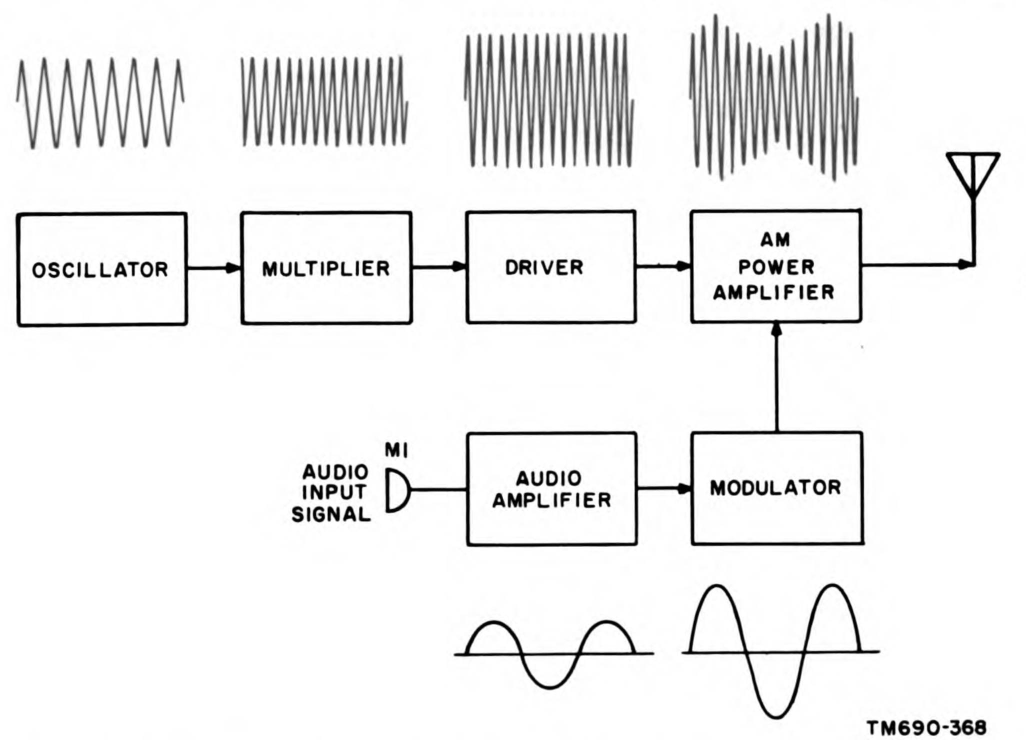


Figure 220. Block diagram of transmitter using an amplitude-modulated power amplifier.

495909 O—59——15

The modulated carrier signal from the amplitude-modulated power amplifier stage is applied to the antenna. The modulating signal would normally be injected into the collector circuit of the amplitude-modulated power amplifier for high-level modulation.

b. Amplitude-Modulated Oscillator. An amplitude-modulated oscillator (par. 207) modulates the rf carrier while it is being generated. Figure 221 shows a block diagram of a transmitter using an amplitude-modulated oscillator. Since an amplitude-modulated oscillator produces both amplitude and frequency modulation of the rf carrier, frequency multiplication is normally not used. Frequency multiplication would increase the frequency modulation as well as the carrier frequency. An amplitude-modulated oscillator normally has high amplification of the rf carrier and does not require a great deal of amplification for transmission. In this application, the audio signal is amplified through two stages of audio amplification before being applied to the oscillator. The output of the modulated oscillator is amplitude and frequency modulated. The percentage of frequency modulation is low and therefore tolerable. The power amplifier stage amplifies the amplitude modulated signal sufficiently to drive the antenna.

209. Establishing Frequency Modulation

a. General. The input and output waveforms of a frequency-modulated oscillator are shown in figure 222. In an fm transmitter, the modulation is accomplished at the oscillator stage. A transistor oscillator can be frequency modulated in the same manner as an electron tube oscillator (TM 11–668) or by varying the oscillator gain at the modulating rate (par. 207*b*). The same amplitude-modulated oscillator used in an am transmitter can be used in an fm transmitter. The unwanted amplitude changes can be removed by a limiter stage before the carrier signal is increased in frequency and magnitude.

b. Frequency-Modulated Oscillator.

(1) A typical frequency-modulated oscillator stage is shown in figure 223. In this application, the frequency modulation is established by reactance modulation. The modulating signal, coupled through transformer $T2$, varies the emitter-base bias of reactance modulator $Q2$. Since the bias is increasing and decreasing at the modulating rate, the collector voltage also increases and decreases at the modulating rate. As the collector voltage increases, output capacitance C_{CE} decreases (par. 103) and as the collector voltage decreases, output capacitance C_{CE} increases. When output capacitance C_{CE} decreases, the resonant frequency of the oscillator $Q1$ tank circuit (capacitor $C1$ and winding 1–3 of transformer $T1$) increases. When output capacitance C_{CE} increases, the resonant frequency of the oscillator tank circuit decreases. The resonant frequency of the oscillator tank circuit is therefore increasing and decreasing at the modulating rate. Thus, the frequency of the signal generated by the oscillator is increasing and decreasing at the modulating rate. The output of the oscillator is therefore a frequency-modulated carrier signal.

(2) Transistor $Q1$ provides the oscillator signal. Capacitor $C1$ and winding 1–3 of transformer $T1$ form a parallel resonant circuit for the oscillator frequency. Winding 4–5 of transformer $T1$ provides

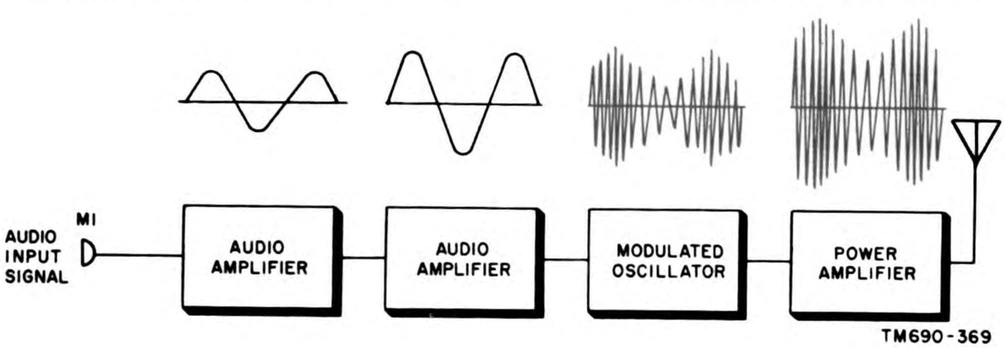


Figure 221. Block diagram of transmitter employing amplitude-modulated oscillator.

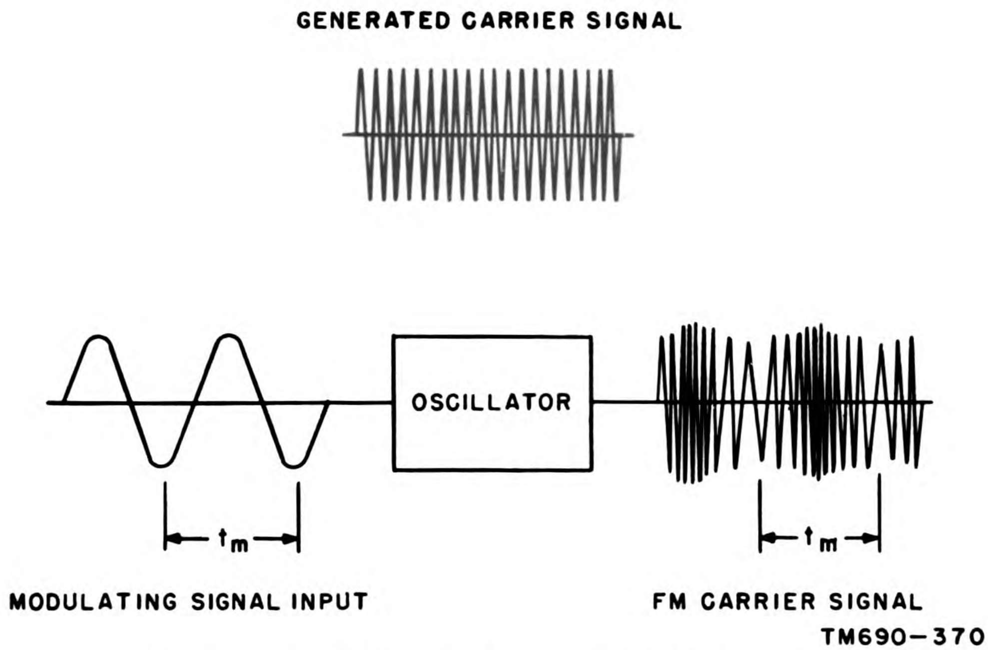


Figure 222. Block diagram of frequency-modulated oscillator.

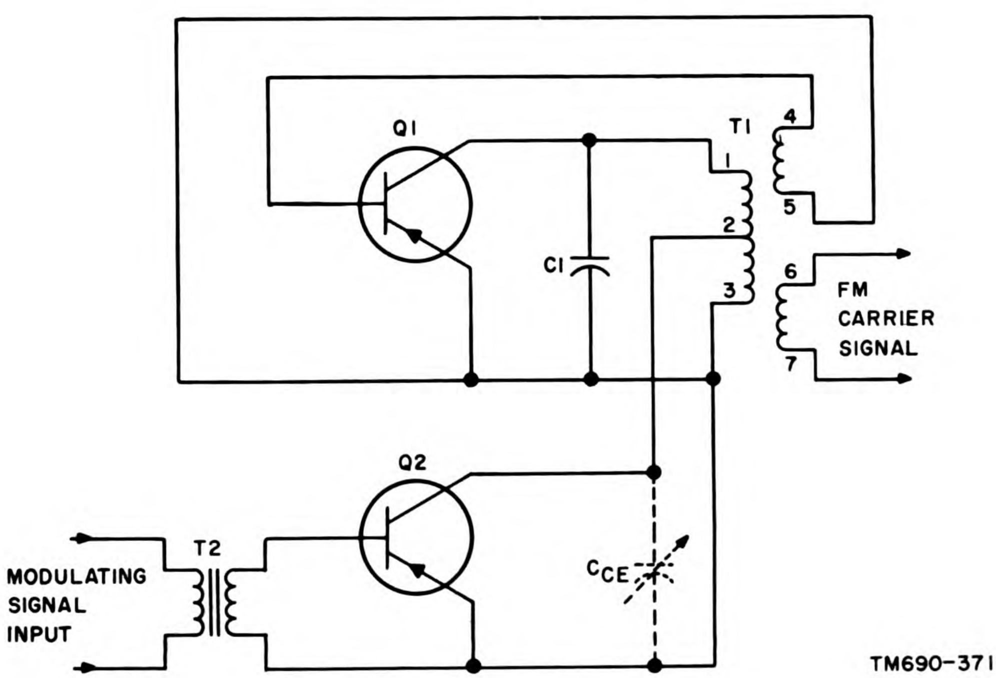


Figure 223. Oscillator circuit, frequency modulated by a reactance modulator.

the required feedback and winding 6–7 couples the oscillator signal to the following stage. Transformer $T2$ couples the modulating signal to reactance modulator $Q2$. The reactance of output capacitance C_{CE} across winding 2–3 of transformer $T1$ varies the resonant frequency of the oscillator tank circuit (capacitor $C1$ and winding 1–3 of transformer $T1$).

210. Use of Frequency-Modulated Oscillator

A frequency-modulated oscillator establishes the fundamental frequency-modulated carrier signal necessary for fm transmission. The requirements for transmitting a frequency-modulated carrier using transistors are the same as those for electron tubes (TM 11–668). A block diagram of a typical fm transmitter with waveforms is shown in figure 224. The audio signal (modulating signal) is applied to the oscillator stage. The output of the oscillator stage is an amplitude- and frequency-modulated carrier signal. The limiter stage removes the amplitude modulation and its output is a frequency-modulated carrier signal with constant amplitude. The multiplier stage increases the frequency to the desired transmitting frequency. The power amplifier stage increases the magnitude of the frequency-modulated carrier signal sufficiently to drive the antenna.

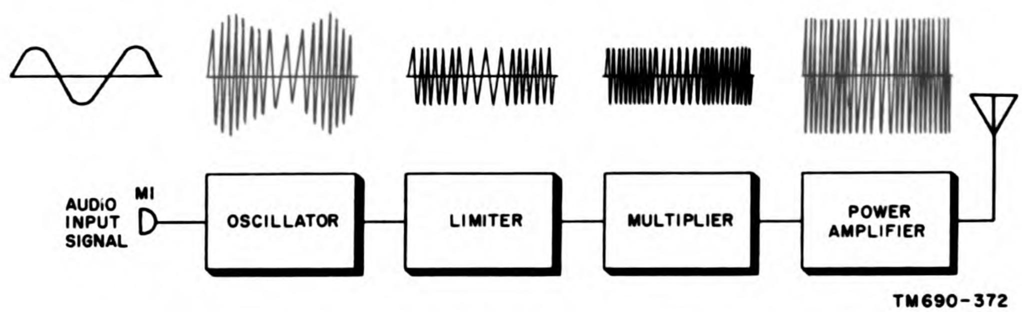


Figure 224. Block diagram of fm transmitter showing waveforms.

Section II. MIXING

211. General

a. The process of combining a radio frequency with an oscillator frequency to produce an intermediate frequency is called *mixing* or *frequency conversion* (TM 11–665). The two basic methods of frequency conversion employed with electron tubes are also employed with transistors. The first method, a transistor that combines an oscillator frequency and a radio frequency, is called a mixer (*b* below). In the second method, only one transistor (functioning as oscillator and mixer), known as a converter (*c* below), is used. When a transistor is used as a mixer or a converter, it is operated on the curved portion of the dynamic transfer characteristic curve (par. 101). Under these conditions, when two frequencies are applied to the transistor input, four major frequencies are produced in the output. Two of the output frequencies are the original frequencies that were present in the input. Another one of the output frequencies is a frequency that is equal to the sum of the two original frequencies. The remaining frequency that is present in the output is a frequency that is equal to the difference of the two original frequencies. In most superheterodyne receivers, only the difference frequency is of interest, and all of the other frequencies must be filtered out. The difference frequency used in a receiver is referred to as the *intermediate frequency* (if).

b. The input and output waveforms of a mixer stage are shown in A, figure 225. The input radio frequency is represented by $F1$ and the input oscillator frequency is represented by $F2$. The two input frequencies applied to the mixer stage produce four prominent output frequencies (*a* above). The original radio frequency present in the output is represented by $F1_1$. The original oscillator frequency present in the output is represented by $F2_1$. The sum of the two original frequencies is represented by $F1$ plus $F2$. The difference of the two original frequencies is represented by $F1$ minus $F2$.

c. The input and output waveforms of a converter stage are shown in B, figure 225. One frequency is applied to the converter stage and the other frequency is generated in the converter stage. The radio frequency applied to the converter stage is represented by $F2$. The oscillator frequency generated by the oscillator is represented by $F1$. The two frequencies present in the converter stage produce four prominent frequencies in the output (*a* above). The original radio frequency present in the output is represented by $F2_1$. The original oscillator frequency present in the output is represented by $F1_1$. The sum of the two frequencies is represented by $F1$ plus $F2$. The difference of the two frequencies is represented by $F1$ minus $F2$.

212. Injection of Oscillator Frequency

The oscillator frequency may be fed to the base, the emitter, or the collector circuit of a mixer stage (fig. 226). The oscillator frequency is coupled through transformer $T3A$, $T3B$, *or* $T3C$ to the base, the emitter, or the collector, respectively. Capacitor $C1$ and the primary of transformer $T1$ form a parallel resonant circuit for the radio frequency which is coupled to the base circuit of mixer $Q1$. Capacitor $C3$ and the primary of transformer

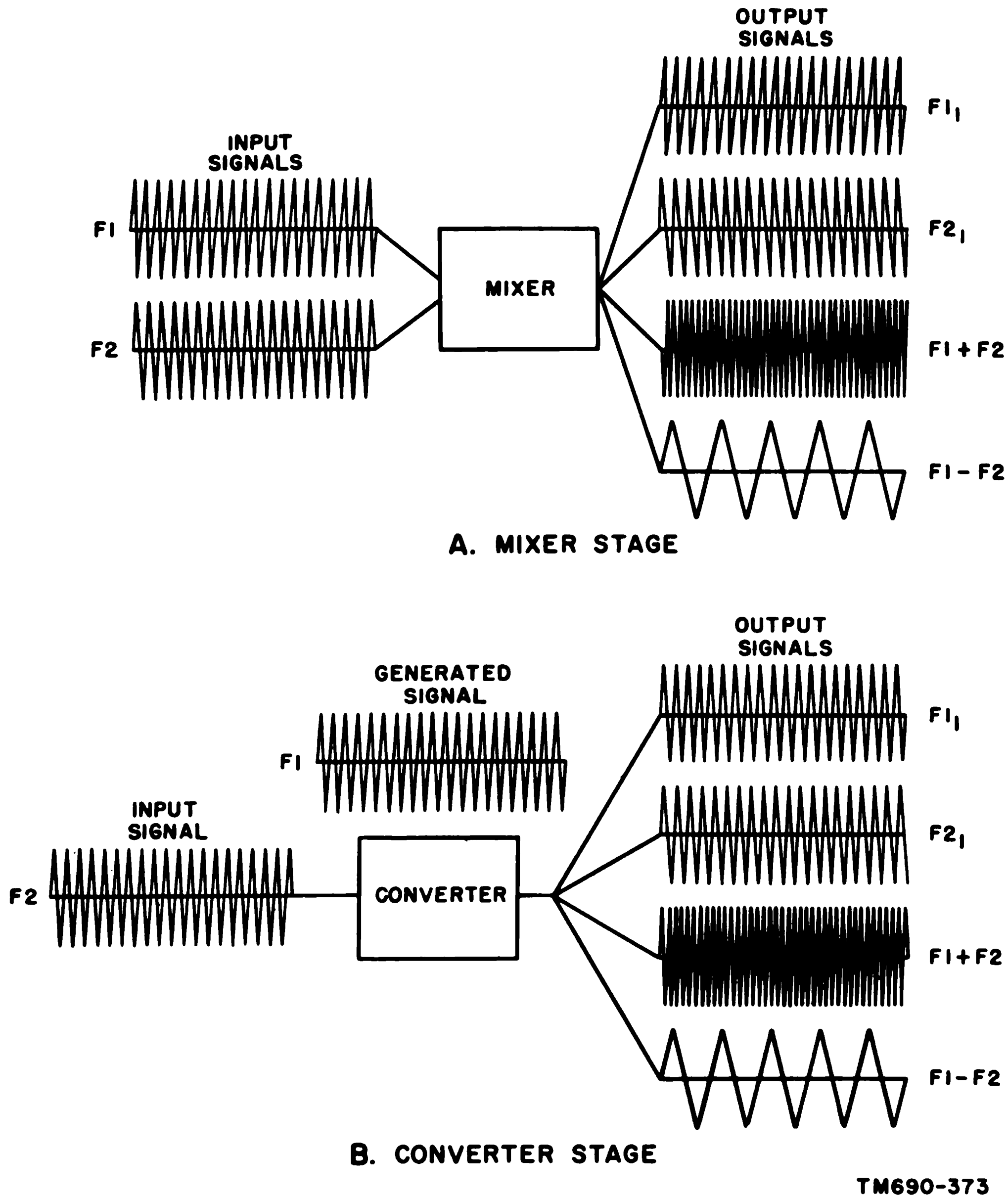


Figure 225. ***Input and output waveforms of a mixer and a converter.***

T3A, *T3B*, or *T3C* form a parallel resonant circuit for the oscillator frequency. Capacitor *C2* and the primary of transformer *T2* form a parallel resonant circuit for the intermediate frequency which is coupled through the transformer to the following stage.

213. Mixer

a. A schematic diagram of a mixer stage with typical values of components is shown in figure 227. The rf signal is 1 megacycle; the oscillator signal, 1.5 megacycles; the if signal, 500 kc. Emitter injection is employed to couple the oscillator signal into the mixer circuit. The radio frequency injected into the base circuit and the oscillator frequency injected into the emitter circuit are heterodyned in mixer Q1. The intermediate frequency is selected by the collector tank circuit. The intermediate frequency is then coupled through transformer T3 to the following stage.

b. Capacitor C1 and the primary of transformer T1 form a parallel resonant circuit for the rf signal which is coupled through the transformer to the base circuit of mixer Q1. Capacitor C2 and the primary of transformer T2 form a parallel resonant circuit for the oscillator frequency which is coupled through the transformer to the emitter circuit of mixer Q1. Resistor R1 provides emitter-base bias and resistor R3 is a voltage dropping resistor. Capacitor C4 is a bypass capacitor. Resistor R2 is the emitter swamping resistor and capacitor C3 is a bypass capacitor. Capacitor C5 and the primary of transformer T3 form a parallel resonant circuit for the intermediate frequency which is coupled through the transformer to the following stage.

214. Converter

a. A schematic diagram of a converter stage is shown in figure 228. The radio frequency injected into the base circuit and the oscillator frequency generated by converter Q1 are heterodyned in the converter. The parallel resonant circuit, consisting of capacitor C3 and the primary of transformer T3, selects the desired intermediate frequency. The intermediate frequency is then coupled through transformer T3 to the following stage.

b. Capacitor C1 and the primary of transformer T1 form a parallel resonant circuit for the radio frequency which is coupled through the transformer to the base circuit of converter Q1. Resistor R1 develops the emitter-base bias and resistor R3 is a voltage dropping resistor. Resistor R2 is the emitter swamping resistor and capacitor C4 is a bypass for the radio frequencies. Capacitor C2 and the primary of transformer T2 form a parallel resonant circuit for the oscillator frequency. The secondary of transformer T2 provides the required feedback for the oscillator portion of converter Q1. Capacitor C3 and the primary of transformer T3 form a parallel resonant circuit for the intermediate frequency which is coupled through the transformer to the following stage. The primaries of transformers T2 and T3 are tapped to obtain the desired selectivity (par. 146).

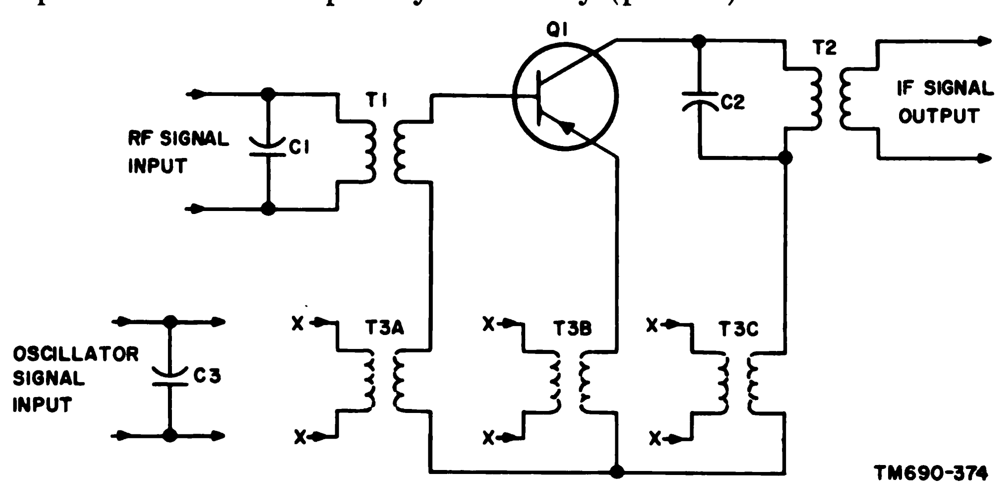


Figure 226. Methods of injecting an oscillator frequency into a mixer stage.

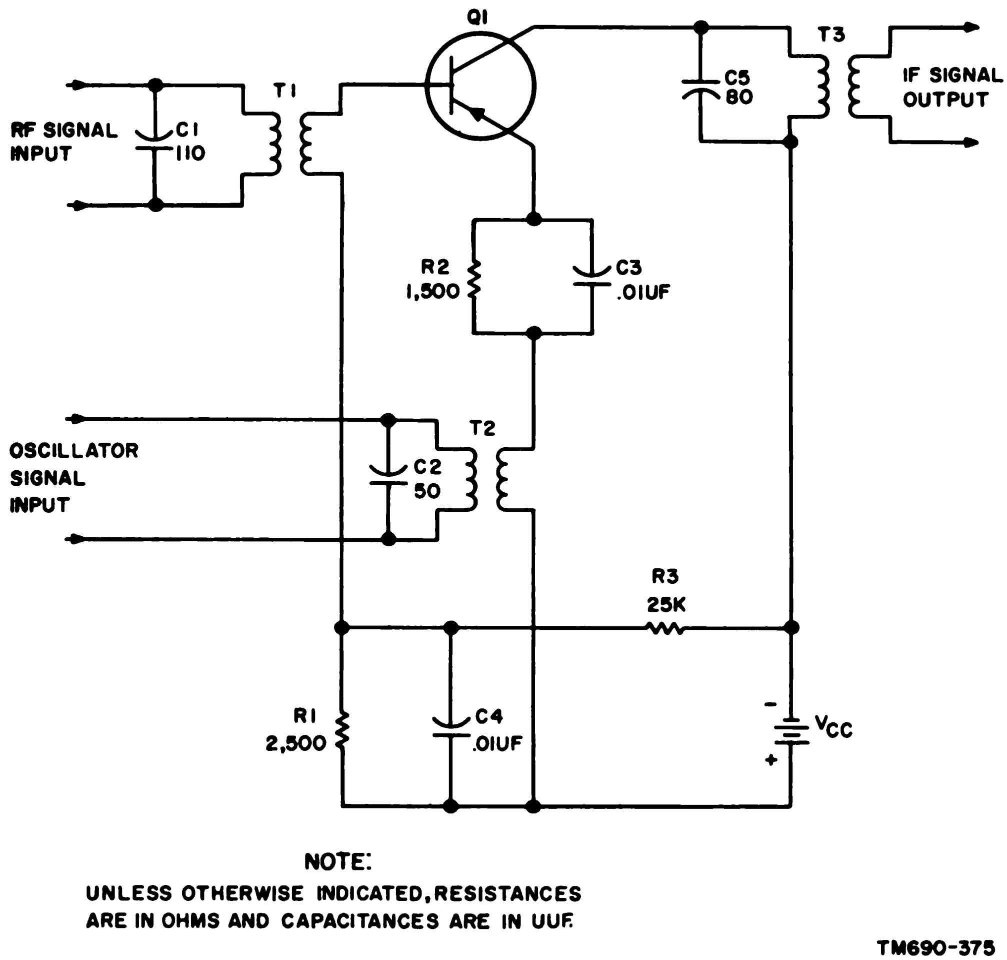


Figure 227. Schematic diagram of mixer showing typical values of components.

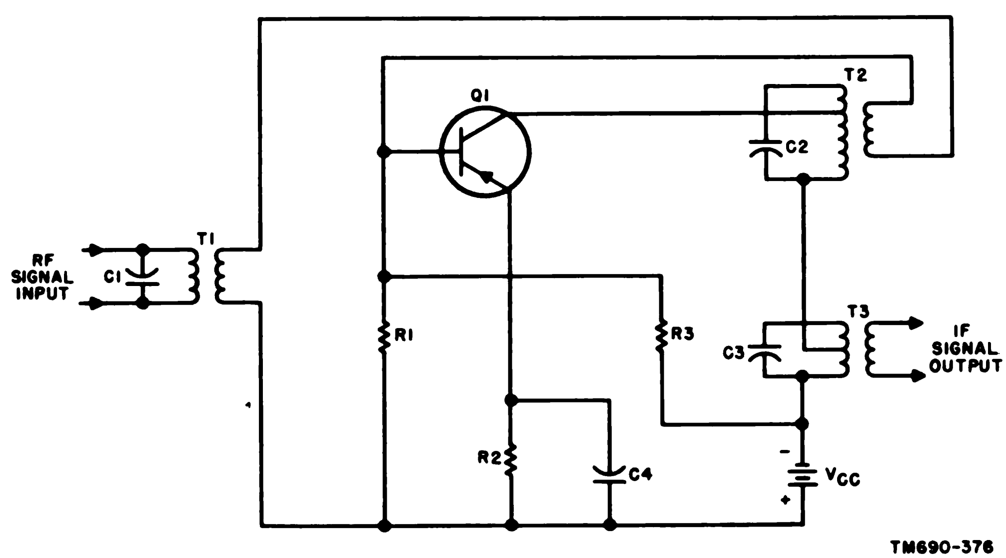


Figure 228. Schematic diagram of converter.

215. General

The three basic types of demodulation (detection) are *cw detection*, *am detection*, and *fm detection*.

a. Cw Detection. Cw detection is accomplished by heterodyning the carrier signal with the output of an oscillator to produce an audio signal. The oscillator stage in a superheterodyne receiver used for this purpose is called a *beat frequency oscillator* (bfo). The bfo frequency is adjusted to differ from the intermediate frequency of the receiver by a low audio frequency. For a complete theoretical analysis of cw detection, refer to TM 11–665. The principles of cw detection for electron tubes apply equally to transistors.

b. Am Detection. Am detection may be accomplished by employing a diode (par. 216) or a transistor (par. 217). The waveforms and the am detector are shown in A, figure 229. The amplitude-modulated carrier signal is applied to the am detector. The am detector rectifies and filters the amplitude variations of the carrier signal. The output of the am detector is an audio signal.

c. Fm Detection. Fm detection may be accomplished by employing a discriminator (par. 218) or a slope detector as an am detector (par. 219). The waveforms for fm detection are shown in B, figure 229. The frequency-modulated carrier signal is applied to the fm detector. The fm detector rectifies and filters the frequency variations of the carrier signal. The output of the fm detector is an audio signal.

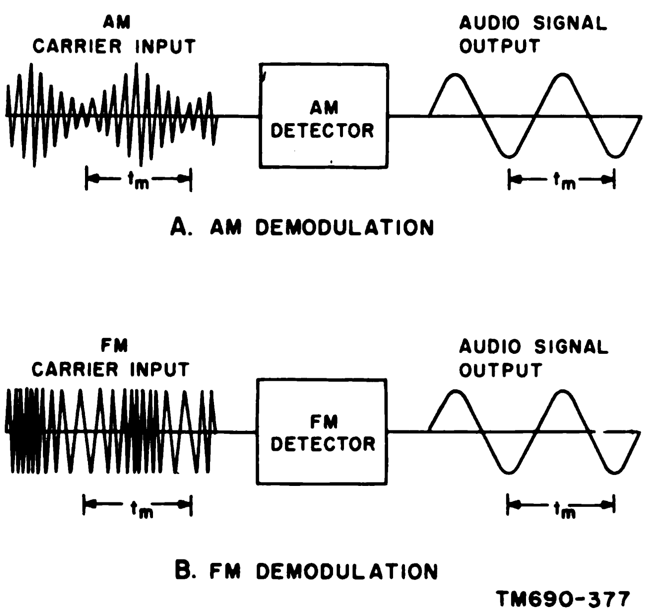


Figure 229. Input and output waveforms for am and fm detection.

216. Am Demodulation, Diode Detector

Diode detectors provide a voltage output (*a* below) or a current output (*b* below).

a. Voltage Output. A voltage output-type diode detector is shown in A, figure 230. This type of detector is normally used with electron tubes since electron tubes are voltage amplifying devices. Although this type of detector is called a voltage output type, there must also be a changing current to produce the changing voltage. Thus, with certain applications, this type of detector is also used in transistor circuits. In the circuit shown, diode CR1 is the rectifying device, resistor R1 is the load, and capacitor C1 is the filter. Diode CR1 rectifies the carrier signal since it only conducts on one-half of a cycle. Current flows through resistor R1 when diode CR1 conducts. Capacitor C1 charges to the voltage drop across resistor R1 and discharges through resistor R1 when diode CR1 is not conducting. The time constant of resistor R1 and capacitor C1 prevents capacitor C1 from discharging completely before the next alternation of the carrier signal. When diode CR1 again conducts, capacitor C1 charges to the voltage drop across resistor R1. The charging and discharging of capacitor C1 filters the output voltage which is essentially increasing and decreasing with the amplitude changes of the carrier signal. Thus, the amplitude changes of the carrier signal produce voltage changes in the output of the voltage detector. The output voltage is essentially the envelope of the carrier signal.

b. Current Output. A current output-type diode detector is shown in B, figure 230. This type of detector may be used with transistors since transistors are current amplifying devices. Since a changing current also produces a changing voltage, this type of detector is also used in electron tube circuits. In the circuit shown, diode CR1 is the rectifying device, resistor R1 is the load, and choke L1 is the filter. Diode CR1 only conducts on one-half of a cycle of the carrier signal. When diode CR1 conducts, resistor R1 is shorted by the diode and no current

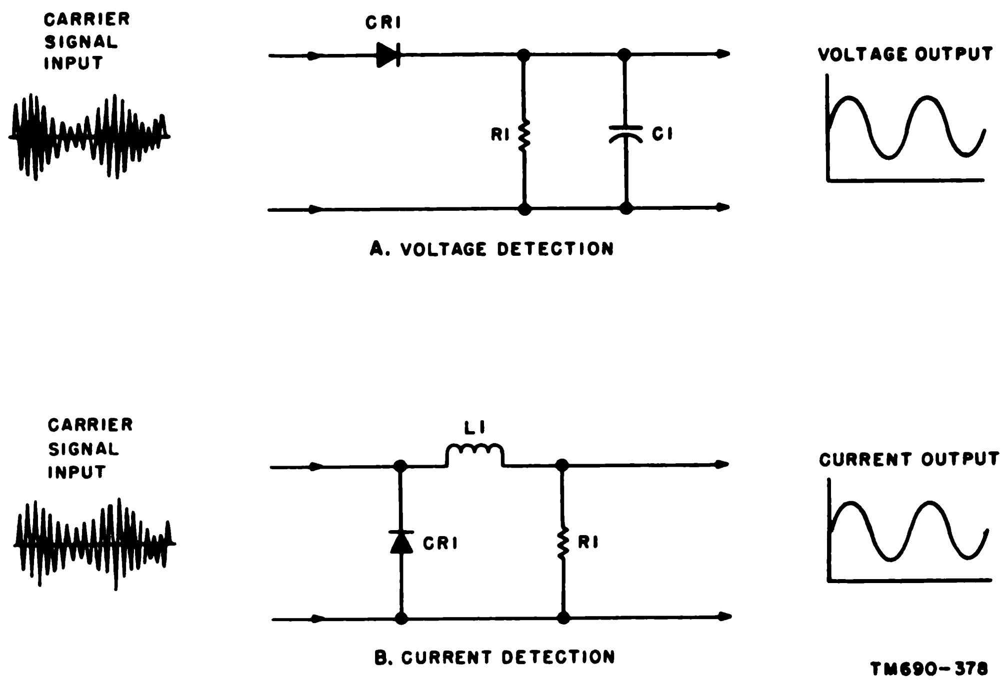


Figure 230. Input and output waveforms of voltage output and current output-type diode detectors.

flows through resistor R1. When diode CR1 does not conduct, all of the current flows through choke L1 and resistor R1. On the alternation that diode CR1 does not conduct, the collapsing field of choke L1 tends to keep current flowing through resistor R1. This action filters the carrier signal current flow through resistor R1. The result is that the output current is filtered and increasing and decreasing at a rate equivalent to the amplitude changes in the carrier signal. Thus, the amplitude changes in the carrier signal produce current changes in the output of the current detector.

217. Am Demodulation, Transistor Detector

A transistor detector is similar in operation to an electron tube grid leak detector (TM 11–665). The rectification for the electron tube grid leak detector takes place in the control grid to cathode portion of the electron tube. The amplification takes place in the control grid to the plate portion of the electron tube. Rectification for the transistor detector takes place in the emitter-base portion of the transistor. Amplification takes place in the emitter-collector portion (*a* below) or the base-collector portion (*b* below) of the transistor. Resistance-capacitance coupling (*a* below) or transformer coupling (*b* below) may be used to couple the output of the transistor detector to the following stage.

a. Common-Emitter.

(1) A common-emitter transistor detector is shown in figure 231. The if signal applied to the base-emitter circuit (biased for nonlinear operation) is rectified by the diode portion (base-emitter) of detector Q1. Resistor R1 acts as the diode load resistor and capacitor C2 filters the if voltage changes. The input circuit acts as a voltage output diode detector (par. 216*a*). The base-emitter bias (audio voltage) developed across resistor R1 causes the collector current to vary at the audio rate. The amplified audio signal developed across resistor R4 is coupled through capacitor C5 to the following stage.

(2) Capacitor C1 and the primary of transformer T1 form a parallel resonant circuit for the if signal which is coupled through the transformer to the base-emitter circuit of detector Q1. Resistor R1 is the emitter-base bias resistor and resistor R2 is a voltage dropping resistor. Capacitors C2 and C4 are bypass capacitors for the intermediate frequency.

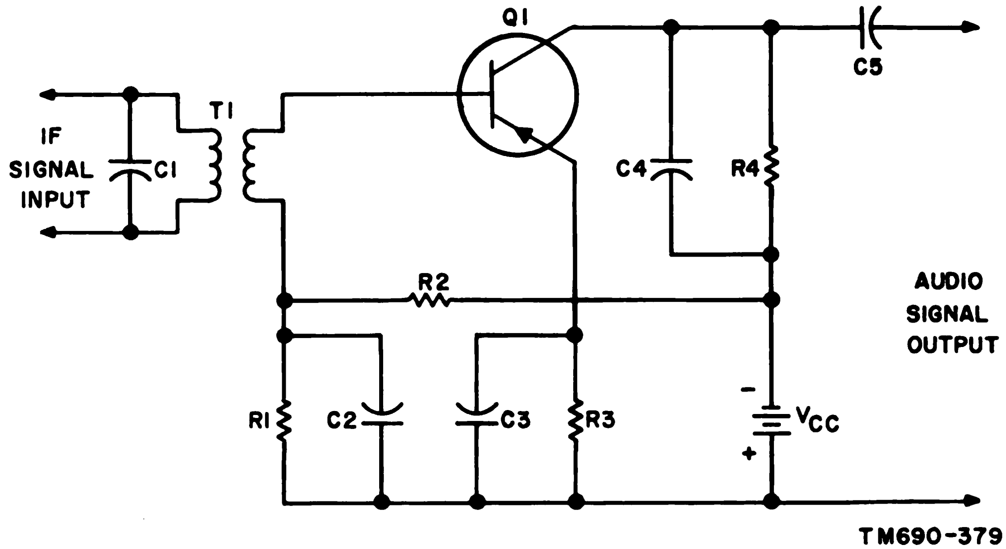


Figure 231. Common-emitter transistor detector.

Capacitor C3 is an audio bypass capacitor. Resistor R3 is the emitter swamping resistor. Resistor R4 is the collector load resistor. Capacitor C5 couples the audio output signal to and blocks the dc voltage from the following stage.

b. Common-Base.

(1) A common-base transistor detector is shown in figure 232. Detection takes place in the emitter-base circuit of transistor detector Q1 and amplification takes place in the collector-base circuit. When the polarity of the if signal is positive on the emitter, current flows through the emitter-base circuit. Capacitor C2 charges on the positive alternation. On the negative alternation, capacitor C2 discharges through resistor R1. The long time constant of capacitor C2 and resistor R1 does not allow capacitor C2 to discharge much during the negative alternation. The next positive alternation again charges capacitor C2. The result is that the peaks of the positive alternation are filtered by capacitor C2. The bias between the emitter and the base is therefore a dc voltage with an audio voltage component. Thus, the amplitude changes of the if signal are detected in the emitter-base circuit of transistor detector Q1. Since the bias of the emitter-base circuit is changing at the audio rate, the collector current also varies at the audio rate. Capacitor C3 filters out any variations of the if signal present in the collector circuit. The audio signal is coupled to the output through transformer T2.

(2) Capacitor C1 and the primary of transformer T1 form a parallel resonant circuit for the if signal which is coupled through the transformer to the emitter circuit of transistor detector Q1. Resistor R1 is the emitter-base bias resistor and capacitor C2 filters the amplitude changes. Capacitor C3 is an if bypass

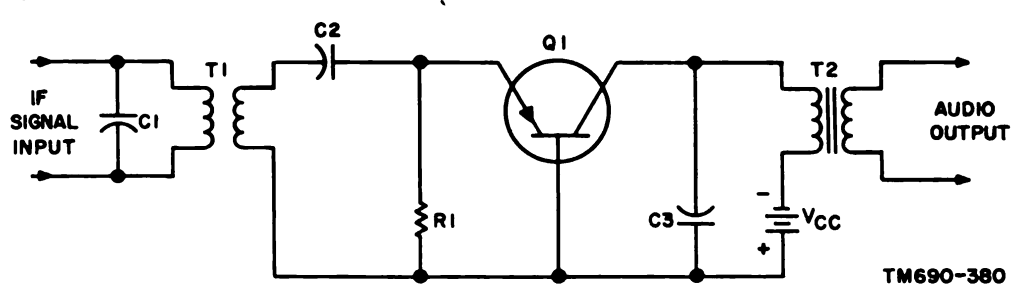


Figure 232. Common-base transistor detector.

and transformer T2 is the output coupling transformer.

218. Fm Demodulation, Discriminator

A discriminator in an fm receiver performs the same function as a detector in an am receiver. Figure 233 shows a transistorized version of an if stage and a discriminator. Detailed vector analysis of the discriminator operation is covered in TM 11-668. Amplifier Q1 amplifies the if signal applied to the discriminator. Resistor R1 is the emitter swamping resistor and capacitor C1 is an if bypass. Capacitor C2 and the primary of transformer T1 form a parallel resonant circuit for the if signal which is coupled through the transformer to the discriminator. Capacitor C3 couples the if signal to the secondary of transformer T1 for phase shift comparison. The if signal, coupled across capacitor C3, is developed across coil L1. Capacitor C4 and the secondary of transformer T1 form a resonant circuit for the if signal coupled through the transformer. The top half of transformer T1 secondary, diode CR1, coil L1, load resistor R2, and filter capacitor C5 form one half of the comparison network. The bottom half of transformer T1 secondary, diode CR2, coil L1, load resistor R3, and filter capacitor C6 form the second half of the comparison network. The audio output of the discriminator circuit is taken from the top of capacitor C5 and the bottom of capacitor C6. The audio output is coupled through capacitor C7 to the primary of transformer T2. The audio signal, coupled through transformer T2, is applied to the following stage.

219. Fm Demodulation, Slope Detector

a. A slope detector converts the frequency changes of a carrier signal into amplitude changes. The amplitude changes can then be detected by an am diode detector or an am transistor detector. The input and output waveforms of a slope detector and an am diode detector are shown in figure 234. The if signal with frequency deviations is applied to slope detector Q1. The output of slope detector Q1, the if signal with amplitude and frequency deviations, is applied to diode detector CR1. The resultant output is an audio signal which is equivalent to the frequency deviations of the if input signal.

b. The if signal coupled through transformer T1 is applied to the base circuit. The resonant circuit consisting of coil L1 and capacitor C2 (tuned slightly off the carrier frequency) develops a large amount of if signal when the frequency deviation is near the resonant frequency. As the frequency deviation of the if signal becomes lower than the resonant frequency of the resonant circuit, a smaller amount of if signal is developed. A large amount of if signal added to the bias voltage developed across resistor R1 increases the emitter-base bias and a small amount of if signal decreases the emitter-base bias. The emitter-base bias is therefore increasing and decreasing as the frequency of the if signal increases and decreases, respectively. Since the bias of slope detector Q1 changes at the frequency deviation rate, the gain also changes at the frequency deviation rate. Thus, the output of the slope detector is an if signal that is changing in amplitude and frequency. The if signal applied to diode detector CR1 is rectified, filtered by coil L2, and developed across resistor R4. The output of the current output-type diode detector (par. 216*b*) is an audio signal.

c. Capacitor C1 and the primary of transformer T1 form a parallel resonant circuit for the if signal which is coupled through the trans-

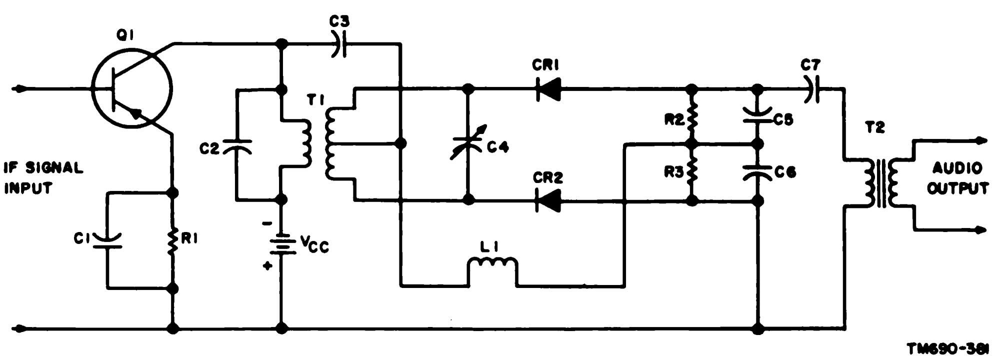


Figure 233. If amplifier and discriminator stage.

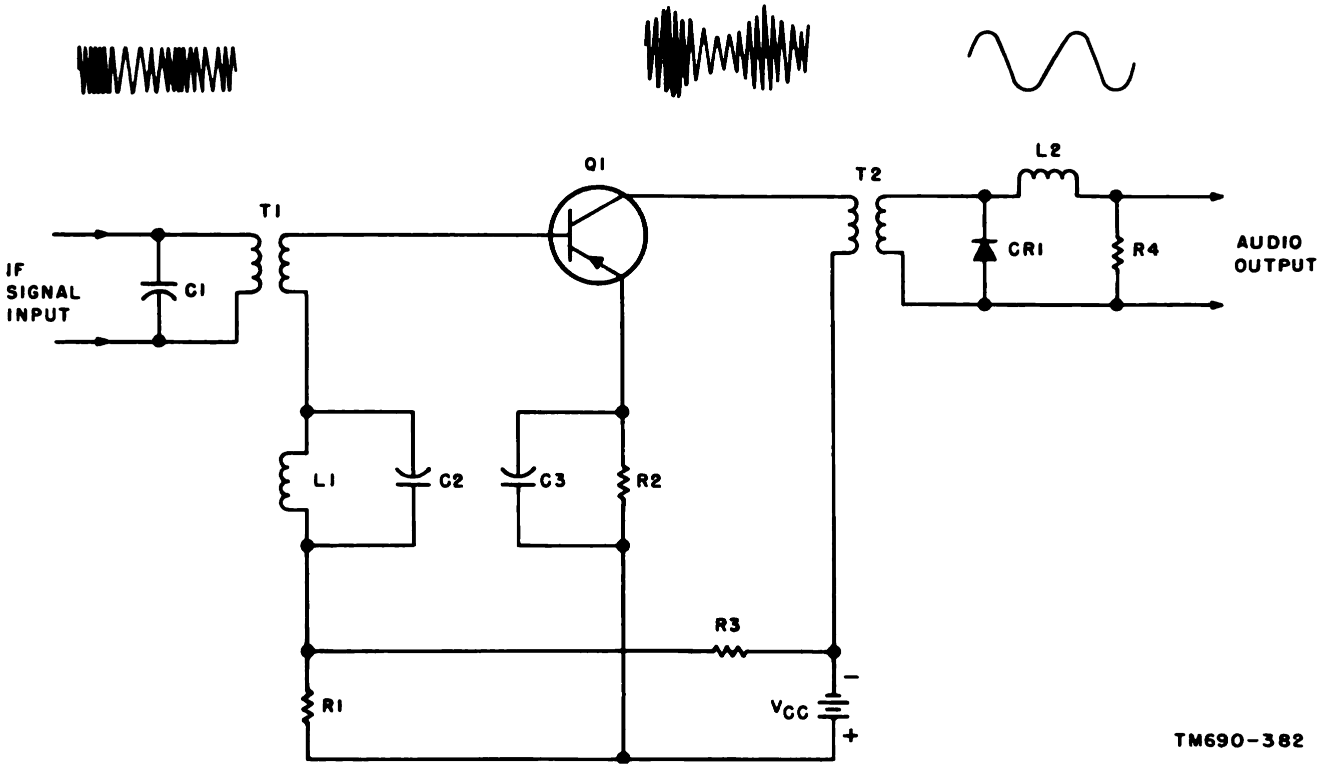


Figure 234. Slope detector and diode detector.

former to the base circuit of slope detector Q1. Capacitor C2 and coil L1 form a parallel resonant circuit for a frequency slightly higher than the maximum frequency deviation of the if signal. Resistor R1 is the base-emitter bias resistor and resistor R3 is a voltage dropping resistor. Resistor R2 is the emitter swamping resistor and capacitor C3 is a bypass capacitor for the if signal. Transformer T2 is an output coupling transformer for slope detector Q1. Diode CR1 is the am detector, resistor R4 is a load resistor, and coil L2 is a filter.

220. Summary

a. Amplitude modulation may be established by modulating an amplifier or an oscillator using base, emitter, or collector injection.

b. Satisfactory high-level modulation is established with collector injection.

c. When an oscillator is amplitude modulated, it also is frequency modulated.

d. The output frequency of an am oscillator does not usually have frequency multiplication before it is transmitted.

e. An amplitude-modulated oscillator may be used in an am transmitter or an fm transmitter.

f. The amplitude changes in the carrier frequency of an fm transmitter are removed by a limiter stage.

g. The distinguishing feature of a mixer stage is that it uses a separate transistor for generating the oscillator signal.

h. In a converter stage, heterodyning takes place in the same transistor that generates the oscillator signal.

i. The four frequencies present in the output of a mixer or a converter are the two original frequencies, the sum of the two original frequencies, and the difference of the two original frequencies.

j. A current output-type diode detector may be used with transistors.

k. A transistor detector detects and amplifies the audio component of a carrier signal.

l. A discriminator circuit in an fm receiver performs the same function as a detector in an am receiver.

m. The audio component of a frequency-modulated carrier may be detected by using a discriminator or a slope detector.

CHAPTER 13

ADDITIONAL SEMICONDUCTOR DEVICES

221. General

Since the development of the point-contact and the junction transistor, continuous research in the field of semiconductors has provided many additional semiconductor devices. Some of these semiconductor devices are now used in military and commercial equipments. Many of them may be used in future equipments. The previous chapters of this manual cover the theory and application of the junction transistor. The point-contact transistor, the first transistor developed, is covered briefly in paragraph 222. This chapter discusses briefly the theory and applications of some of the additional semiconductor devices. The additional semiconductor devices are as follows:

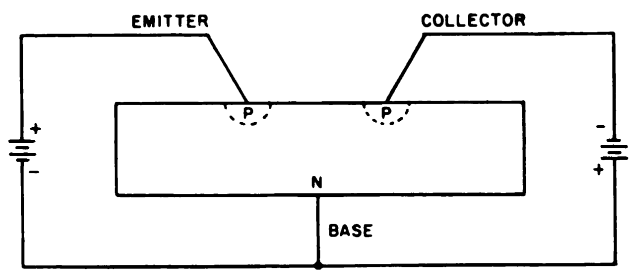
a. The photosensitive semiconductors (par. 223).
b. The tetrode transistor (par. 224).
c. The spacistor (par. 225).
d. The unijunction transistor (par. 226).
e. The four-layer diode (par. 227).

222. Point-Contact Transistor

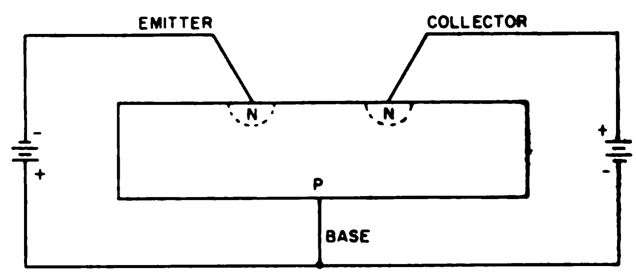
a. The point-contact transistor (fig. 4) consists of a single piece of P-type or N-type material. The base contact forms a large area (low resistance) connection, but the emitter and the collector contacts form small area (high resistance) connections.

b. Although the point-contact transistor employs an N-type or a P-type material, P-type or N-type layers respectively are present under the contact points. The P-type layers present under the contact points in the N-type material or the N-type layers present under the contact points in the P-type material are formed during the manufacturing process. The N-type point-contact transistor is shown in A, figure 235. For biasing polarities, the N-type point-contact transistor may be considered similar to a PNP junction transistor (par. 32); the P-type point-contact transistor (B, fig. 235) may be considered similar to an NPN junction transistor (par. 33). The emitter is biased in the forward direction with respect to the base and the collector is biased in the reverse direction with respect to the base.

c. The small area connections of the collector and the emitter to the N-type or P-type material provide a high resistance. Because of this high resistance, the point-contact transistor has a higher input and output impedance than the junction transistor. The base current in a junction transistor is very small and the emitter current is larger than the collector current (par. 32). In a point-contact transistor, the base current is fairly large and the collector current, except for



A. N-TYPE POINT-CONTACT TRANSISTOR



B. P-TYPE POINT-CONTACT TRANSISTOR

TM690-392

Figure 235. Bias arrangement for N-type and P-type point-contact transistors.

greater nonlinearity, is greater than the emitter current. The characteristic curves of the point-contact transistor are similar to those of the junction transistor (ch. 4).

223. Photosensitive Semiconductors

a. General. In a semiconductor material, the movement of the carriers may be caused by electrical, heat, or light energy. In a photosensitive semiconductor, light energy is used to control the movement of the carriers. To control the movement of carriers in a photosensitive semiconductor, the light energy must be concentrated on a sensitive area of the photosensitive semiconductor. When the concentrated beam of light strikes the sensitive area of the photosensitive semiconductor, the current flow through the semiconductor increases. When the light energy is decreased, the current flow through the photosensitive semiconductor decreases. The increase or decrease in light energy causes a proportional increase or decrease in the current flow through the semiconductor. The point-contact photosensitive semiconductor (*b* below) and the junction photosensitive semiconductor (*c* below) are used in various circuits for detection, translation, switching, etc.

b. Point-Contact. A point-contact photosensitive semiconductor is shown in A, of figure 236. The energy from the light source is concentrated into a beam through the lens. The concentrated beam of light energy strikes the sensitive area of the semiconductor. The light energy is increased or decreased by varying the intensity of the light source. An increase in light energy causes an increase in current flow and a decrease in light energy causes a decrease in current flow. The current flow through the load resistor varies proportionally to the intensity of the light source. Thus, the light energy is converted into electrical energy. The output signal is developed across the load resistor. The point-contact photosensitive semiconductor may be N-type or P-type material. The material is hollow ground on one side so that the light energy may be concentrated on the sensitive area.

c. Junction.

(1) *Single.* A single junction photosensitive semiconductor circuit is shown in B, figure 236. The operation of this circuit is identical with that described in *b* above except for the photosensitive semiconductor. The sensitive area in the junction photosensitive semiconductor is located in the PN junction. The sensitive area of a single junction photosensitive semiconductor may be located in either of the semiconductor materials.

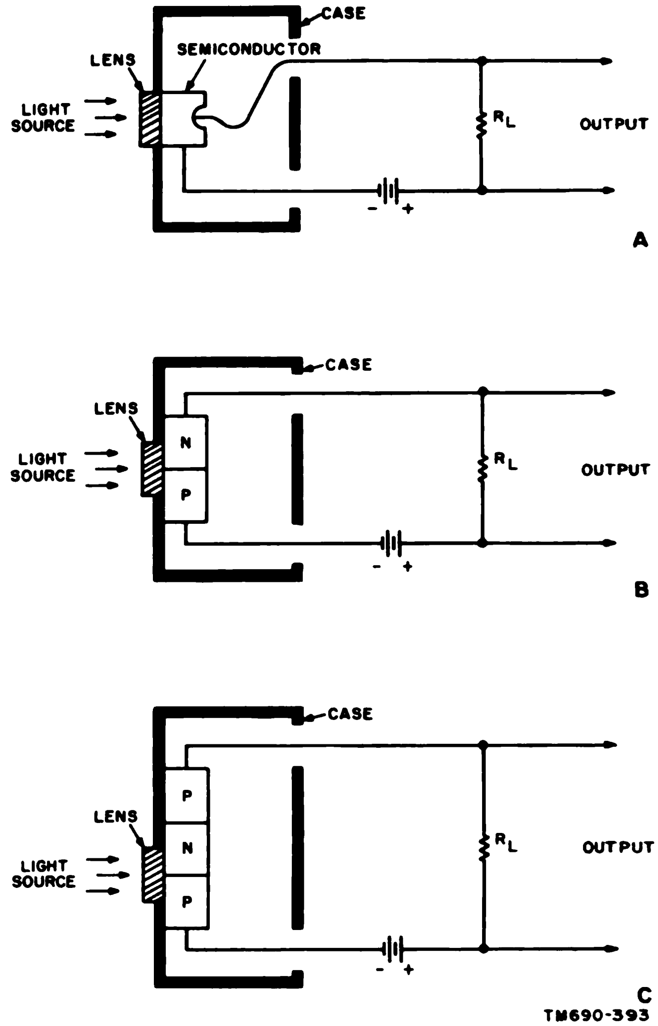


Figure 236. Photosensitive semiconductors.

(2) *Double.* A double junction photosensitive semiconductor circuit is shown in C, figure 236. The operation of this circuit is identical with that described in *b* above except for the photosensitive semiconductor. The operation of the double junction photosensitive semiconductor may be compared to the operation of the junction transistor amplifier (par. 37). The light source, which strikes the first PN junction, may be considered equivalent to a signal applied to the emitter-base junction of a transistor amplifier. Current amplification takes place in the

second PN junction, which is equivalent to the collector-base circuit of a transistor amplifier. In the double junction photosensitive semiconductor, the variation of collector current is greater than that of the point-contact and the single junction photosensitive semiconductors (*b* and (1) above).

224. Tetrode Transistor

a. Operation.

(1) The highest frequency that can be amplified by a three-terminal junction transistor is limited by the input (emitter-base junction) capacitance and particularly the output (collector-base junction) capacitance of the transistor. One method of reducing these capacitance values is to reduce the size of the transistor semiconductor material. However, this method is physically limited because the semiconductor material must be large enough so that leads may be attached to the three regions. A second method is to restrict transistor action to a small portion of the semiconductor material so that the effect of a small size transistor with low input and output capacitance is obtained. To employ the latter method, a tetrode (four-terminal) transistor must be used.

(2) A tetrode transistor is shown in A, figure 237. The tetrode transistor is constructed in the same manner as the three-terminal (PNP or NPN) junction transistor, except for the addition of a second terminal to the base region. Terminals 1, 2, and 3 are the conventional emitter, base, and collector terminals, respectively, and are biased in the same manner as the three-terminal transistor. Terminal 4 is the second connection to the base region. By placing a negative bias on terminal 4, using battery V_{BB}, the indicated voltage gradient occurs within the base region. Note that forward bias (emitter more negative than base) occurs only over a small portion of the emitter-base junction. Current flow between base and collector also occurs over a small portion of the collector-base junction. Since the effective input and output capacitances involve only the active portions of the emitter-base junction and the collector-base junction, respectively, these capacitance values are substantially reduced. The high-frequency response of the tetrode transistor then, is substantially greater than that of the comparable three-terminal transistor.

b. Application. A wide-band amplifier circuit employing a tetrode transistor is shown in B, figure 237. The input signal is coupled through capacitor $C1$ to the emitter-base circuit. The input signal aids or opposes the bias developed across resistor $R1$ and varies the emitter-base bias of wide-band amplifier $Q1$. Varying the bias at the rate of the input signal varies the collector current at the same rate. The variations of collector current are developed in the collector circuit and coupled through capacitor $C3$ to the following stage. Capacitor $C1$ is the input coupling capacitor. Resistor $R1$ develops the emitter-base bias and resistor $R2$ is a voltage dropping resistor. The bias across the base region is obtained from battery V_{EE}. Resistor $R3$ is a voltage dropping resistor. Resistor $R4$ and capacitor $C2$ form a low-frequency compensating network. Chokes $L1$ and $L2$ form a high-frequency peaking circuit (ch. 9). Capacitor $C3$ is the output coupling capacitor.

225. Spacistor

a. General. In addition to the input and output capacitances of the three-terminal junction transistor limiting the high-frequency amplification range (par. 224*a*(1)), another factor is the transit time of current carriers from emitter to collector. As the time of 1 cycle of the frequency to be amplified approaches the transit time, the gain falls off rapidly. In an electron tube, the transit time required for electrons to travel from cathode to plate also limits its high-frequency range. To reduce the transit time in the electron tube, the plate voltage is increased and the spacing between cathode and plate is decreased. In the conventional transistor, the transit time is long because very little voltage gradient exists in the base region so that current carriers in the base region diffuse (or migrate) from the emitter junction to the collector junction. Although physically limited, a partial solution is the reduction of the size of the base region. This method has increased the response of transistors to approximately 75 megacycles. Other methods involving

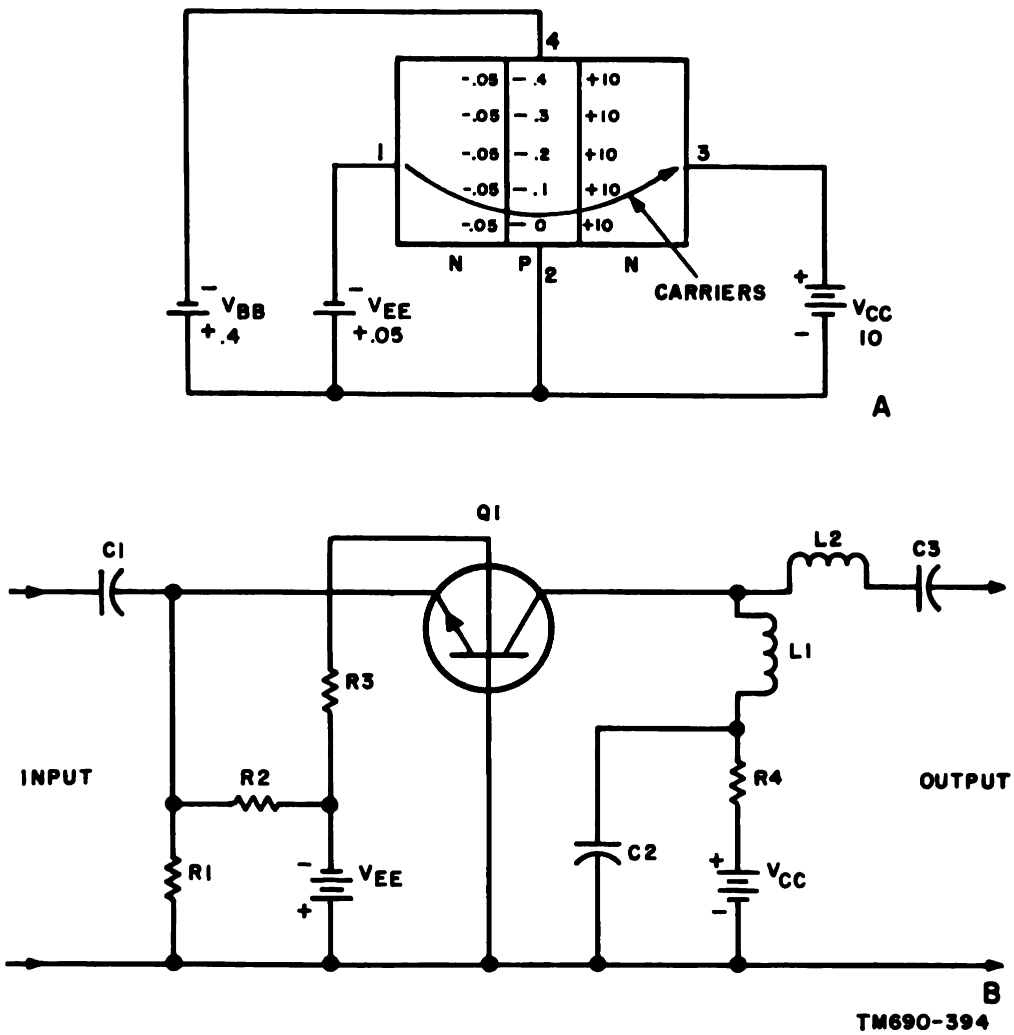


Figure 237. Tetrode transistor and application in wide-band amplifier.

nonuniform distribution of impurities in the base region have increased the response of transistors to 350 megacycles. The spacistor (fig. 238) eliminates the diffusion of carriers through a base region; this device may be operated up to several kilomegacycles because of the resultant low transit time.

b. Construction and Operation.

(1) A, figure 238, shows a portion of the spacistor; this is a reverse-biased PN junction. Note that the depletion region in the N-type material is larger than that in the P-type material. This is accomplished by using a lower percentage of donor impurities in the N-type material than the percentage of acceptor impurities used in the P-type material (par. 27*b*). The larger depletion region in one of the materials is required so that two connections ((1) and (2) below), can be made at the larger depletion region. With the bias indicated, very little current (only reverse-biased current) flows through the junction. However, a strong electric field exists on either side of the junction. Because of the very short distance of the depletion region (several mils), a reverse bias of 100 volts will produce a substantial electric field of 60,000 to 120,000 volts per centimeter. This intense field is used to reduce the transit time.

(2) A connection made to the upper depletion region (B, fig. 238) is called the injector. This terminal is biased by battery V_{II}. The voltage of battery V_{II} is less than that of battery V_{CC}, so that an electron current flows from the injector to the collector.

The magnitude of this current is proportional to the difference in voltage between battery V_{II} and battery V_{CC}. The transit time of electrons from the injector to the collector is very low because of the intense field through which the electrons flow.

(3) A second connection made to the depletion region on the same side of the junction and close to the injector (C, fig. 238) is called a modulator. The modulator is actually a small section of P-type material forming a PN junction with the collector region. Battery V_{MM} biases the modulator; the voltage of battery V_{MM} is less than that of battery V_{CC} so that the PN junction formed by modulator and collector regions is reverse biased. This reverse-biased PN junction produces a high input impedance at the modulator terminal. If battery voltage V_{MM} is varied, the current flow between the injector and the collector varies accordingly. This effect indicates that the field between the modulator and the injector affects the injector current. A varying voltage (such as a sine wave) introduced between the modulator and injector then will result in a similarly varying current to the collector.

(4) The spacistor, then, can be used as a very-high-frequency (or low-frequency) amplifier (*c* below). Since the modulator (acting as a control grid) draws no current, the input impedance between the modulator and injector (acting as a cathode) is extremely high (several megohms). The input capacitance (between modulator and injector) is very low (1 or 2 $\mu\mu$f) because of the small area each covers. The output capacitance (collector to base) is very low because of the large depletion region. This acts as a capacitor, the plates of which are far apart. Under these conditions, current, voltage, and power gain can be achieved.

c. Application. A practical amplifier using the spacistor is shown in D, figure 238. A signal is coupled to the modulator terminal through trans-

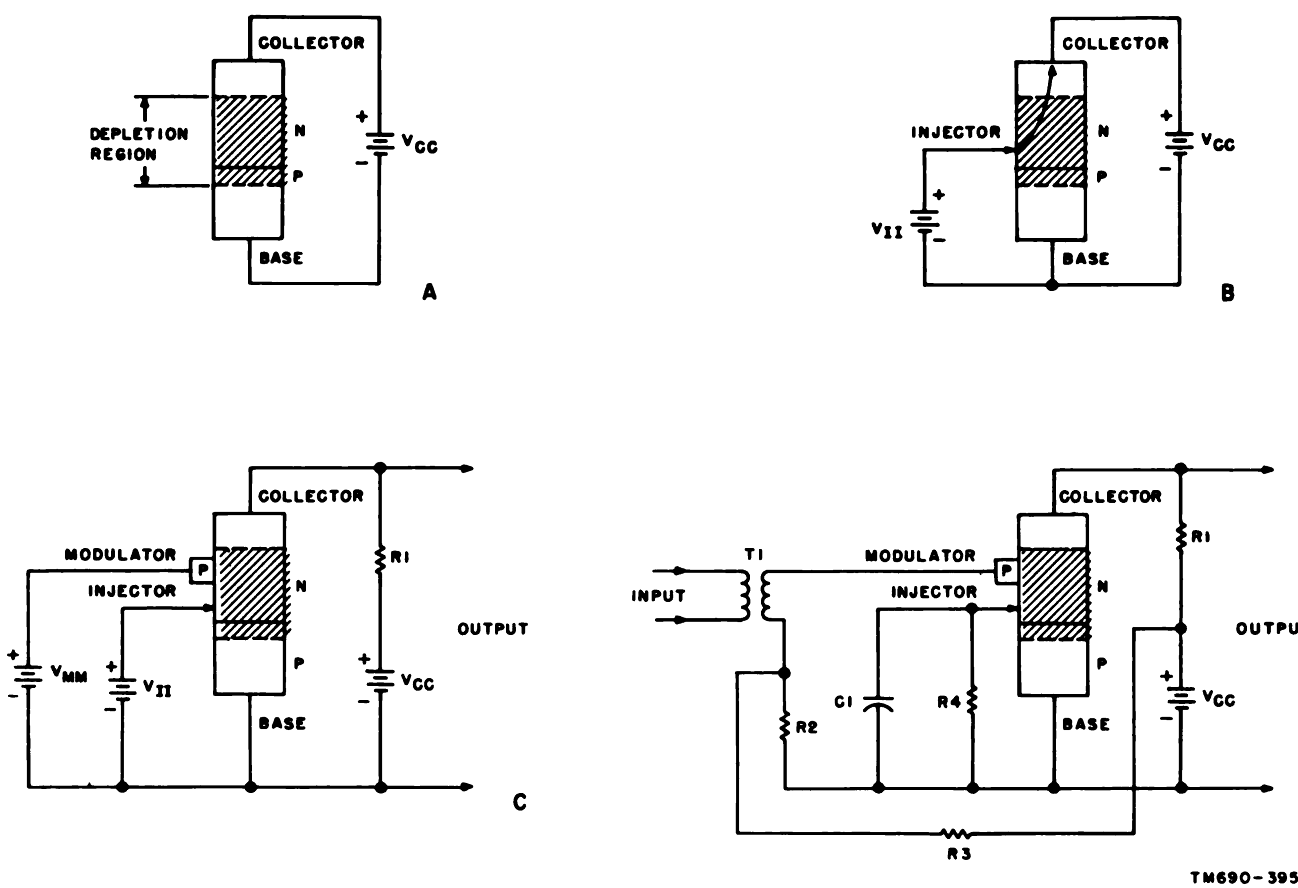


Figure 238. Spacistor and spacistor amplifier.

former T1. The voltage divider, consisting of resistors R2 and R3, establishes the modulator bias. Resistor R4, bypassed by capacitor C1, provides the injector bias. Collector load resistor R1 develops the output signal.

226. Unijunction Transistor

a. Operation. The unijunction (single junction) transistor (A, fig. 239) is actually a diode with two connections made to one portion of the semiconductor. The unijunction transistor is also referred to as a *double-base* diode. If terminals 2 and 3 are connected together, the resultant device would have the characteristics of a conventional junction diode.

(1) With battery V_{BB} connected as indicated, electron current flows from terminal 3 to terminal 2 as indicated. With terminal 1 connected to terminal 3, the PN junction is reverse biased. The only current flowing through terminal 1 is a reverse-bias current; this current consists of electron flow (solid-line arrow) from terminal 1 to the PN junction and holes (dashed-line arrow) from terminal 2 to the PN junction (ch. 3).

(2) Battery V_{BB} establishes within the N-type semiconductor the voltage gradient indicated from terminal 2 to terminal 3 (B, fig. 239). Battery V_{EE} is inserted in the circuit between terminals 1 and 3 with the polarity indicated. If the voltage of battery V_{EE} is less than the voltage gradient opposite the P-type material, the condition shown in A, figure 239 prevails. If the voltage of battery V_{EE} is greater than the voltage gradient opposite the P-type material, the PN junction becomes forward biased. Heavy electron current flow (solid-line arrows) occurs in the N-type material, and a heavy hole current flow (dashed-line arrows) occurs in the

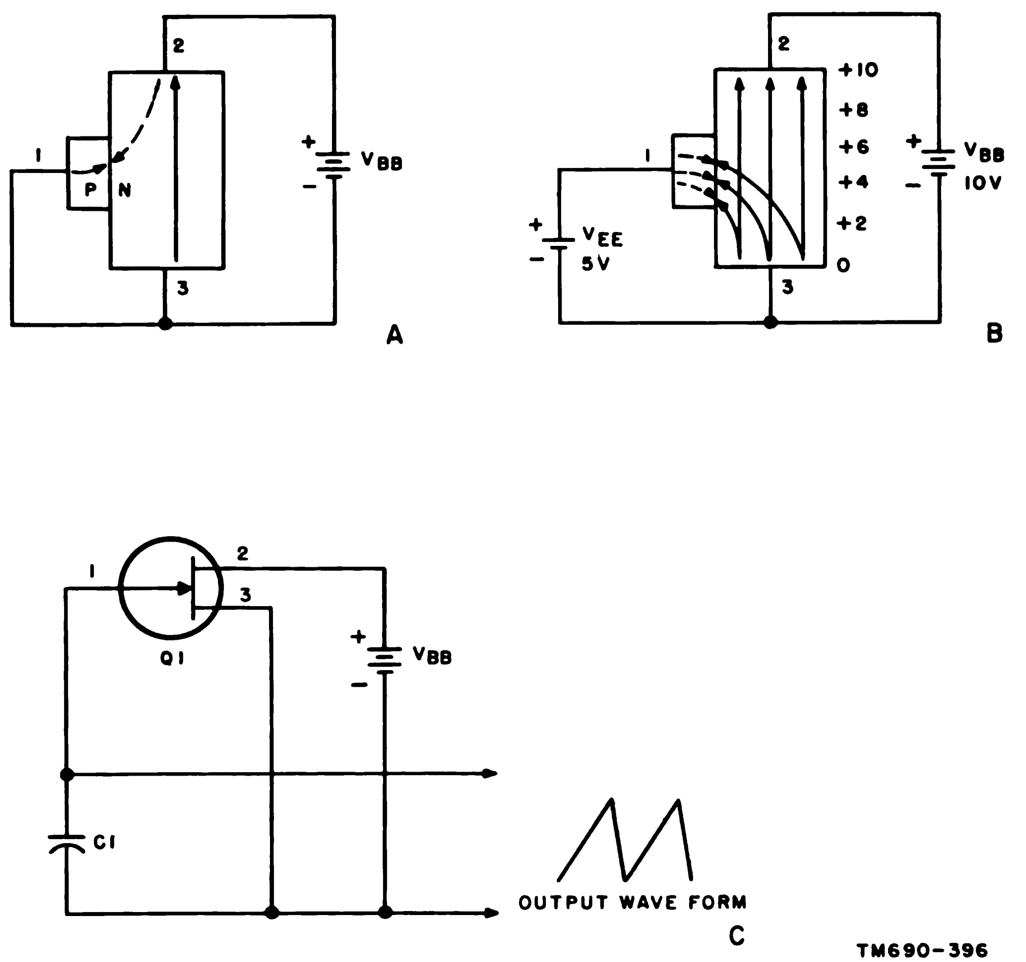


Figure 239. Unijunction transistor and sawtooth generator.

P-type material; electrons flow out of terminal 1.

(3) These characteristics ((1) and (2) above) of the unijunction transistor make it especially suitable for use in multivibrators and sawtooth generators (*b* below).

b. Application. A sawtooth generator using a unijunction transistor is shown in C, figure 239. When power is applied to the circuit, the conditions are the same as those described in *a*(1) above. Capacitor C1 charges very slowly through the small amount of current flowing through the reverse-biased PN junction. The charging time equals the product of the capacitance of capacitor C1 and the resistance of the reverse-biased PN junction. As capacitor C1 charges, the positive dc voltage across it rises. When this voltage is greater than the voltage gradient opposite the P-type material, the PN junction is forward biased (*a*(2) above), and the capacitor discharges very rapidly. The discharge time equals the product of the capacitance of capacitor C1 and the resistance of the forward-biased PN junction. After capacitor C1 is discharged, the conditions are again the same as those described in *a*(1) above. Capacitor C1 again charges and establishes a forward bias for the PN junction. This sequence continues and the slow charging and rapid discharging of capacitor C1 produces a sawtooth waveform in the output.

227. Four-Layer Diode

a. Operation. The four-layer diode (A, fig. 240) consists of four layers of semiconductor material. The four layers of N-type and P-type material form three PN junctions. When properly biased, the center PN junction is reverse biased and the outer PN junctions are forward biased. The emitter-base junction of a three-terminal transistor is always forward biased, and the collector-base junction is always reverse biased

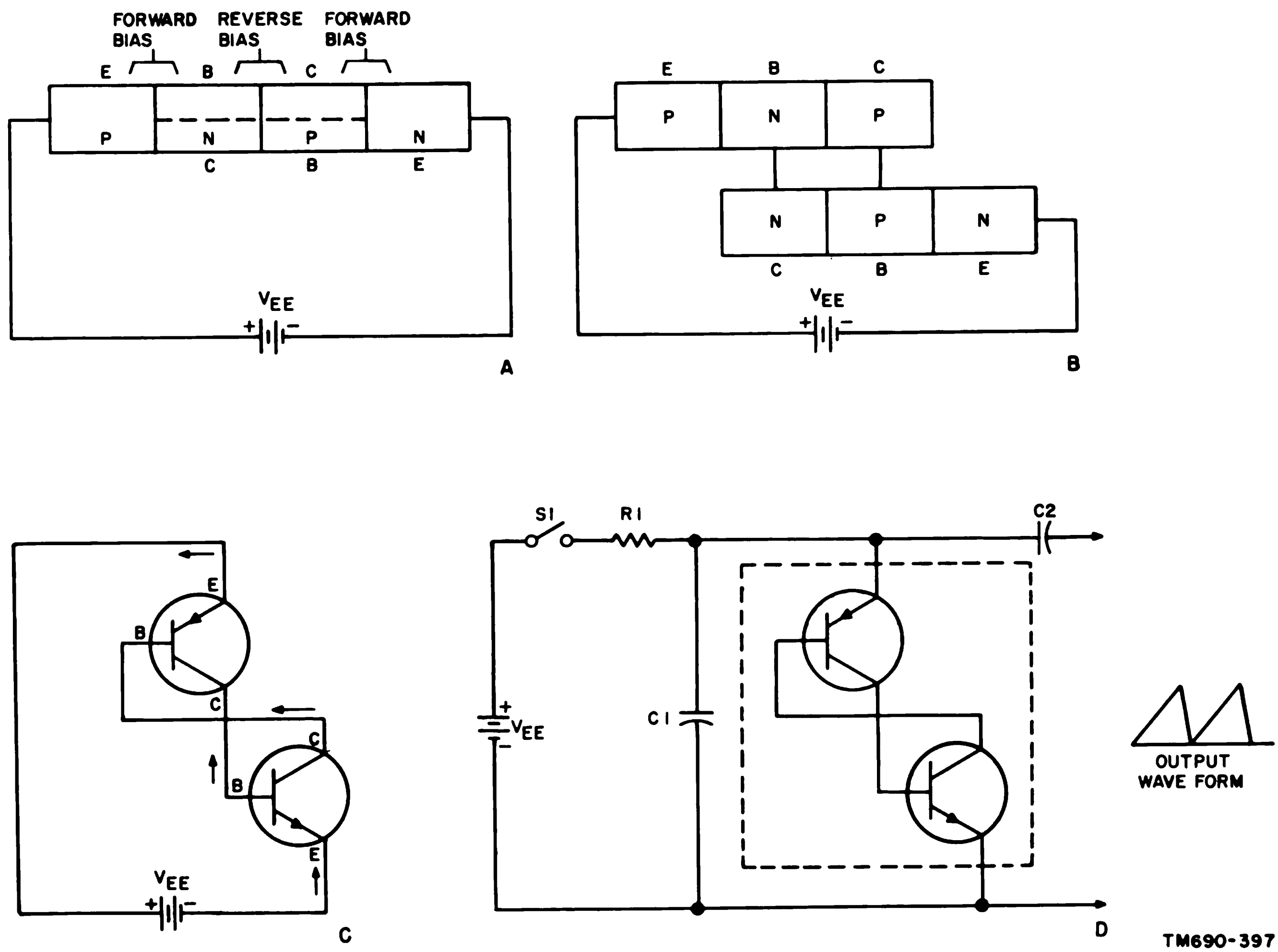


Figure 240. Four-layer diode, and sawtooth generator.

(pars. 32 and 33). The four-layer diode therefore can be analyzed as two separate junction transistors. The dotted line through the PN material in the center divides the four-layer diode into a PNP transistor and an NPN transistor. The PN junction of the PNP transistor (B. fig. 240) connected to the positive side of battery V_{BE} is forward biased and is therefore the emitter-base junction. The remaining PN junction of the PNP transistor is reverse biased and is therefore the collector-base junction. The PN junction of the NPN transistor connected to the negative side of battery V_{BE}, is forward biased and is therefore the emitter-base junction. The remaining PN junction of the NPN transistor is reverse biased and is therefore the collector-base junction. The base of the PNP transistor is connected directly to the collector of the NPN transistor. The collector of the PNP transistor is connected directly to the base of the NPN transistor. The schematic representation of the four-layer diode is shown in C, figure 240. Proper biasing is obtained from battery V_{BE}. The arrows represent electron current flow.

b. Application. The schematic diagram of a sawtooth oscillator employing the four-layer diode is shown in D, figure 240. Battery V_{BE} furnishes power for the oscillator circuit. Switch $S1$ completes the circuit and applies power to the oscillator circuit. The time constant of resistor $R1$ and capacitor $C1$ determines the frequency of oscillation. The four-layer diode may be considered the equivalent of a thyratron electron tube which provides a discharge path for capacitor $C1$. When power is applied to the circuit through switch $S1$, capacitor $C1$ appears as a short circuit. Current flows through resistor $R1$ and capacitor $C1$ charges. When the voltage drop across capacitor $C1$ provides sufficient bias, current flows through the four-layer diode. Capacitor $C1$ then discharges rapidly through the four-layer diode. As capacitor $C1$ discharges, the bias applied to the four-layer diode decreases. When the bias reaches a minimum value for conduction, the four-layer diode stops conducting. Capacitor $C1$ again charges to a point where sufficient bias causes the four-layer diode to conduct. This sequence keeps repeating, and the output coupled through capacitor $C2$ is a sawtooth waveform.

228. Summary

a. A P-type point-contact transistor operates similarly to an NPN junction transistor and an N-type point-contact transistor operates similarly to a PNP junction transistor.

b. In a photosensitive semiconductor, the light energy must be focused into a beam to fall on a sensitive area of the semiconductor.

c. The photosensitive semiconductor converts the light energy into electrical energy.

d. The tetrode transistor has low input and output capacitances and can be used in high-frequency applications.

e. The spacistor may be used to amplify frequencies in the kilomegacycle range, because of the low transit time required for current carriers to flow from injector to collector.

f. Input and output impedances of a spacistor are very high in comparison to the junction transistor.

g. The unijunction transistor is a single junction semiconductor with three electrodes. It may be used in multivibrators and sawtooth generators.

h. Operation of a four-layer diode can be considered the equivalent of two junction transistors.

APPENDIX I

LETTER SYMBOLS

1. Roman Letters

Symbols	Meaning or description	Appears first in paragraph
A_f	Band pass of a resonant circuit	146
agc	Automatic gain control	157
A_i	Current gain	54
am	Amplitude modulation	3
A_v	Voltage gain	55
bfo	Beat frequency oscillator	215
C	Constant (when following an equal (=) sign)	51
CB	Common-base configuration	38
C_{cb}	Interelement capacitance between base and collector	103
C_{eb}	Interelement capacitance between base and emitter	103
C_{ce}	Interelement capacitance between collector and emitter	103
CC	Common-collector configuration	38
CE	Common-emitter configuration	38
cm	Centimeter	15
C_N	Neutralizing capacitor	154
CR	Junction diode	86
C_{ep}	Reflected capacitance	146
cw	Continuous wave transmission	215
E_{IN}	Input voltage	94
E_{OUT}	Output voltage	94
fm	Frequency modulation	3
f_a	Cutoff frequency	186
f_r	Frequency of resonance	146
G	Power gain	56
g_{bc}	Base-collector conductance	70
g_{be}	Base-emitter conductance	70
g_{ce}	Collector-emittance conductance	70
g_{fe}	Forward transfer conductance	70
g_{ie}	Input conductance	70
G_m	Maximum power	60
g_{oe}	Output conductance	70
g_{re}	Reverse transfer conductance	70
h	Hybrid	51
h_{ib}	Input resistance of common-base configuration with output short-circuited	62
h_{ic}	Input resistance of common-collector configuration with output short-circuited.	63
h_{ie}	Input resistance of common-emitter configuration with output short-circuited.	51
h_{ob}	Output conductance of common-base configuration with input open.	62
h_{oc}	Output conductance of common-collector configuration with input open.	63
h_{oe}	Output conductance of common-emitter configuration with input open.	52
I_e	Total emitter current	43
I_b	Changing base current (ac)	47
I_B	Dc base current	51
i_c	Changing collector current (ac)	47
I_C	Dc collector current	52
if	Intermediate frequency	211
m	Turns ratio of a transformer	146
ma	Milliampere	30
mv	Millivolt	58
mw	Milliwatt	102
N	Number of turns in an indicator	146
N-type	Semiconductor with donor impurity	2
NPN	Transistor with one P-type and two N-type semiconductors.	2
P-type	Semiconductor with acceptor impurity.	2
P	Power	31
PN	Combination of N-type and P-type semiconductors.	2
PNP	Transistor with one N-type and two P-type semiconductors.	2
Q	Selectivity of a resonant circuit	146
RT	Thermistor	83
r_b	Ac base resistance	68
RC	A coupling circuit employing a resistor and a capacitor.	88
r_d	Ac collector resistance	68
r_e	Ac emitter resistance	68
r_{fe}	Forward transfer resistance with input open.	68
r_{ie}	Input resistance with output open	68
R_L	Load resistor	54
r_m	Mutual resistance	68
R_N	Neutralizing resistor	154
r_{oe}	Output resistance with input open	68

Symbols	Meaning or description	Appears first in paragraph
r_{re}	Reverse transfer resistance with input open	68
S_I	Current-stability factor	75
S_V	Voltage-stability factor	75
T	Transformer	83
TC	Time constant	180
t_f	Fall time	187
t_p	Pulse time	187
t_r	Rise time	187
t_s	Storage time	187
V_{BE}	Fixed base-emitter voltage	51
V_{CE}	Fixed collector-emitter voltage	51
V_s	Source voltage	54
v_{be}	Changing voltage (ac) between base and emitter	47
v_{ce}	Changing voltage (ac) between collector and emitter	47
V_{SAT}	Saturation voltage	186
Y	Admittance	70
Z	Impedance	70

2. Greek Letters

Symbols	Meaning or description	Appears first in paragraph
α	Alpha; current amplification factor	52
α_{fb}	Forward short circuit current amplification factor for the *CB* configuration	62
α_{fc}	Forward short circuit current amplification factor for the *CC* configuration	63
α_{fe}	Forward short circuit current amplification factor for the *CE* configuration	52
Δ	Delta; incremental change	51
∞	Infinity	77
μa	Microampere	30
μ_{rb}	Reverse open-circuited voltage amplification factor for *CB* configuration	62
μ_{rc}	Reverse open-circuited voltage amplification factor for *CC* configuration	63
μ_{re}	Reverse open-curcuited voltage amplification factor for *CE* configuration	51

APPENDIX II

REFERENCES

TM 11–661 Electrical Fundamentals (Direct Current)

TM 11–662 Basic Theory and Application of Electron Tubes

TM 11–664 Basic Theory and Use of Electronic Test Equipment

TM 11–665 C-w and A-m Radio Transmitters and Receivers

TM 11–668 F-m Transmitters and Receivers

TM 11–669 Transients and Waveforms

TM 11–670 Special-Purpose Oscillators and Amplifiers

TM 11–672 Pulse Techniques

TM 11–673 Generation and Transmission of Microwave Energy

TM 11–681 Electrical Fundamentals (Alternating Current)

GLOSSARY

DEFINITIONS OF UNUSUAL TERMS

Acceptor Impurity—A substance with three (3) electrons in the outer orbit of its atom which, when added to a semiconductor crystal, provides one hole in the lattice structure of the crystal.

Amplifier, Class A—An amplifier in which the swing of the input signal is always on the linear portion of the characteristic curves of the amplifying device.

Amplifier, Class AB—An amplifier which has the collector current or voltage at zero for less than half of a cycle of input signal.

Amplifier, Class B—An amplifier which operates at collector current cutoff or at zero collector voltage and remains in this condition for ½ cycle of the input signal.

Amplifier, Class C—An amplifier in which the collector voltage or current is zero for more than ½ cycle of the input signal.

AND Circuit (AND Gate)—A coincidence circuit that functions as a gate so that when all the inputs are applied simultaneously, a prescribed output condition exists.

AND-OR Circuit (AND-OR Gate)—A gating circuit that produces a prescribed output condition when several possible combined input signals are applied; exhibits the characteristics of the *AND* gate and the *OR* gate.

Astable Multivibrator—A multivibrator that can function in either of two semistable states switching rapidly from one to the other; referred to as free running.

Barrier—In a semiconductor, the electric field between the acceptor ions and the donor ions at a junction. (See *Depletion Layer*.)

Barrier Height—In a semiconductor, the difference in potential from one side of a barrier to the other.

Base (junction transistor)—The center semiconductor region of a double junction (NPN or PNP) transistor. The base is comparable to the grid of an electron tube.

Base Spreading Resistance—In a transistor, the resistance of the base region caused by the resistance of the bulk material of the base region.

Beat Frequency Oscillator—An oscillator that produces a signal which mixes with another signal to provide frequencies equal to the sum and difference of the combined frequencies.

Bistable Multivibrator—A circuit with two stable states requiring two input pulses to complete a cycle.

Blocking Oscillator—A relaxation type oscillator that conducts for a short period of time and is cut off for a relatively long period of time.

Clamping Circuit—A circuit that maintains either or both amplitude extremities of a wave form at a certain level or potential.

Collector—The end semiconductor material of a double junction (NPN or PNP) transistor that is normally reverse-biased with respect to the base. The collector is comparable to the plate of an electron tube.

Common-Base (CB) Amplifier—A transistor amplifier in which the base element is common to the input and the output circuit. This configuration is comparable to the grounded-grid triode electron tube.

Common-Collector (CC) Amplifier—A transistor amplifier in which the collector element is common to the input and the output circuit. This configuration is comparable to the electron tube cathode follower.

Common-Emitter (CE) Amplifier—A transistor amplifier in which the emitter element is common to the input and the output circuit. This configuration is comparable to the conventional electron tube amplifier.

Complementary Symmetry Circuit—An arrangement of PNP-type and NPN-type transistors that provides push-pull operation from one input signal.

Compound-Connected Transistor—A combination of two transistors to increase the current amplification factor at high emitter currents. This combination is generally employed in power amplifier circuits.

Configuration—The relative arrangement of parts (or components) in a circuit.

Constant Power Dissipation Line—A line (superimposed on the output static characteristic curves) representing the points of collector voltage and current, the products of which represent the maximum collector power rating of a particular transistor.

Cross-Over Distortion—Distortion that occurs at the points of operation in a push-pull amplifier where the input signals cross (go through) the zero reference points.

Current Stability Factor—In a transistor, the ratio of a change in emitter current to a change in reverse-bias current flow between the collector and the base.

Cutoff Frequency—The frequency at which the gain of an amplifier falls below .707 times the maximum gain.

Dependent Variable—In a transistor, one of four variable currents and voltages that is arbitrarily chosen and considered to vary in accordance with other currents and voltages (independent variable).

Depletion Region (or Layer)—The region in a semiconductor containing the uncompensated acceptor and donor ions; also referred to as the space-charge region or barrier region.

Differentiating Circuit—A circuit that produces an output voltage proportional to the rate of change of the input voltage.

Donor Impurity—A substance with electrons in the outer orbit of its atom; added to a semiconductor crystal, it provides one free electron.

Double-Junction Photosensitive Semiconductor—Three layers of semiconductor material with an electrode connection to each end layer. Light energy is used to control current flow.

Dynamic Transfer Characteristic Curve—In transistors, a curve that shows the variation of output current (dependent variable) with variation of input current under load conditions.

Electron-Pair Bond—A valence bond formed by two electrons one from each of two adjacent atoms.

Elemental Charge—The electrical charge on a single electron (megatron or positron).

Emitter-Follower Amplifier—See *Common-Collector Amplifier*.

Emitter (junction transistor)—The end semiconductor material of a double junction (PNP or NPN) transistor that is forward-biased with respect to the base. The emitter is comparable to the cathode of an electron tube.

Equivalent Circuit—A diagrammatic circuit representation of any device exhibiting two or more electrical parameters.

Fall Time—The length of time during which the amplitude of a pulse is decreasing from 90 percent to 10 percent of its maximum value.

Forward Bias—In a transistor, an external potential applied to a PN junction so that the depletion region is narrowed and relatively high current flows through the junction.

Forward Short Circuit Current Amplification Factor—In a transistor, the ratio of incremental values of output to input current when the output circuit is ac short-circuited.

Four-Layer Diode—A diode constructed of semiconductor materials resulting in three PN junctions. Electrode connections are made to each end layer.

Gating Circuit—A circuit operating as a switch, making use of a short or open circuit to apply or eliminate a signal.

Grounded Base Amplifier—See *Common-Base Amplifier*.

Hole—A mobile vacancy in the electronic valence structure of a semiconductor. The hole acts similarly to a positive electronic charge having a positive mass.

Hybrid Parameter—The parameters of an equivalent circuit of a transistor which are the result of selecting the input current and the output voltage as independent variables.

Increment—A small change in value.

Independent Variable—In a transistor, one of several voltages and currents chosen arbitrarily and considered to vary independently.

Inhibition Gate—A gate circuit used as a switch and placed in parallel with the circuit it is controlling.

Interelement Capacitance—The capacitance caused by the PN junctions between the regions of a transistor; measured between the external leads of the transistor.

Junction Transistor—A device having three alternate sections of P-type or N-type semiconductor material. See *PNP Transistor* and *NPN Transistor*.

Lattice Structure—In a crystal a stable arrangement of atoms and their electron-pair bonds.

Majority Carriers—The holes or free electrons in P-type or N-type semiconductors respectively.

Minority Carriers—The holes or excess electrons found in the N-type or P-type semiconductors respectively.

Monostable Multivibrator—A multivibrator having one stable and one semistable condition. A trigger is used to drive the unit into the semistable state where it remains for a predetermined time before returning to the stable condition.

Multivibrator—A type of relaxation oscillator for the generation of nonsinusoidal waves in which the output of each of two stages is coupled to the input of the other to sustain oscillations. See *Astable Multivibrator, Bistable Multivibrator, and Monostable Multivibrator.*

Neutralization—The prevention of oscillation of an amplifier by canceling possible changes in the reactive component of the input circuit caused by positive feedback.

NOR Circuit—An *OR* gating circuit that provides pulse phase inversion.

NOT AND Circuit—An *AND* gating circuit that provides pulse phase inversion.

NPN Transistor—A device consisting of a P-type section and two N-type sections of semiconductor material with the P-type in the center.

N-Type Semiconductor—A semiconductor into which a donor impurity has been introduced. It contains free electrons.

Open Circuit Parameters—The parameters of an equivalent circuit of a transistor which are the result of selecting the input current and output current as independent variables.

OR Circuit (OR Gate)—A gate circuit that produces the desired output with only one of several possible input signals applied.

Parameter—A derived or measured value which conveniently expresses performance; for use in calculations.

Photosensitive Semiconductor—A semiconductor material in which light energy controls current carrier movement.

PN Junction—The area of contact between N-type and P-type semiconductor materials.

PNP Transistor—A device consisting of an N-type section and two P-type sections of semiconductor material with the N-type in the center.

Point-Contact—In transistors, a physical connection made by a metallic wire on the surface of a semiconductor.

Polycrystalline Structure—The granular structure of crystals which are nonuniform in shape and irregularly arranged.

Preamplifier—A low level stage of amplification usually following a transducer.

P-Type Semiconductor—A semiconductor crystal into which an acceptor impurity has been introduced. It provides holes in the crystal lattice structure.

Pulse Amplifier—A wide-band amplifier used to amplify square wave forms.

Pulse Repetition Frequency—The number of nonsinusoidal cycles (square waves) that occur in 1 second.

Pulse Time—The length of time a pulse remains at its maximum value.

Quiescence—The operating condition that exists in a circuit when no input signal is applied to the circuit.

Reverse Bias—An external potential applied to a PN junction such as to widen the depletion region and prevent the movement of majority current carriers.

Reverse Open Circuit Voltage Amplification Factor—In a transistor, the ratio of incremental values of input voltage to output voltage measured with the input ac open-circuited.

Rise Time—The length of time during which the leading edge of a pulse increases from 10 percent to 90 percent of its maximum value.

Saturation (Leakage) Current—The current flow between the base and collector or between the emitter and collector measured with the emitter lead or the base lead, respectively, open.

Semiconductor—A conductor whose resistivity is between that of metals and insulators in which electrical charge carrier concentration increases with increasing temperature over a specific temperature range.

Short Circuit Parameters—The parameters of an equivalent circuit of a transistor which are the result of selecting the input and output voltages as independent variables.

Single-Junction Photosensitive Semiconductor—Two layers of semiconductor materials with an electrode connection to each material. Light energy controls current flow.

Spacistor—A semiconductor device consisting of one PN junction and four electrode connections characterized by a low transient time for carriers to flow from the input element to the output element.

Stabilization—The reduction of variations in voltage or current not due to prescribed conditions.

Storage Time—The time during which the output current or voltage of a pulse is falling from maximum to zero after the input current or voltage is removed.

Stray Capacitance—The capacitance introduced into a circuit by the leads and wires used to connect circuit components.

Surge Voltage (or Current)—A large sudden change of voltage (or current) usually caused by the collapsing of a magnetic field or the shorting or opening of circuit elements.

Swamping Resistor—In transistor circuits, a resistor placed in the emitter lead to mask (or minimize the effects of) variations in emitter-base junction resistance caused by variations in temperature.

Tetrode Transistor—A junction transistor with two electrode connections to the base (one to the emitter and one to the collector) to reduce the interelement capacitance.

Thermal Agitation—In a semiconductor, the random movement of holes and electrons within a crystal due to the thermal (heat) energy.

Thyratron—A gas-filled triode electron tube that is used as an electronic switch.

Transducer—A device that converts one type of power to another, such as acoustical power to electrical power.

Transistor—A semiconductor device capable of transferring a signal from one circuit to another and producing amplification. See *Junction Transistor*.

Triggered Circuit—A circuit that requires an input signal (trigger) to produce a desired output determined by the characteristics of the circuit.

Trigger Pulse Steering—In transistors, the routing or directing of trigger signals (usually pulses) through diodes or transistors (called steering diodes or steering transistors) so that the trigger signals affect only one circuit of several associated circuits.

Tuned-Base Oscillator—A transistor oscillator with the frequency-determining device (resonant circuit) located in the base circuit. It is comparable to the tuned grid electron tube oscillator.

Tuned-Collector Oscillator—A transistor oscillator with the frequency-determining device located in the collector circuit. It is comparable to the tuned plate electron tube oscillator.

Turnoff Time—The time that it takes a switching circuit (gate) to completely stop the flow of current in the circuit it is controlling.

Unijunction Transistor—A PN junction transistor with one electrode connection to one of the semiconductor materials and two connections to the other semiconductor material.

Unilateralization—The process by which an amplifier is prevented from going into oscillation by canceling the resistive and reactive component changes in the input circuit of an amplifier caused by positive feedback.

Unit Step Current (or Voltage)—A current (or voltage) which undergoes an instantaneous change in magnitude from one constant level to another.

Voltage Gain—The ratio of incremental values of output voltage to input voltage of an amplifier under load conditions.

Wide-Band Amplifier—An amplifier capable of passing a wide range of frequencies with equal gain.

Zener Diode—A PN junction diode reverse-biased into the breakdown region; used for voltage stabilization.

INDEX

495909 O—59——17

[AG 413.44 (30 Dec 58)]

By Order of *Wilber M. Brucker*, Secretary of the Army:

MAXWELL D. TAYLOR,
General, United States Army,
Chief of Staff.

Official:

R. V. LEE,
Major General, United States Army,
The Adjutant General.

Distribution:

Active Army:

CNGB (1)
Tech Stf, DA (1) except CSigO (30)
Tech Stf Bd (1)
USCONARC (5)
USA Arty Bd (1)
USA Armor Bd (1)
USA Inf Bd (1)
USA AD Bd (1)
USA Abn & Elct Bd (1)
USA Avn Bd (1)
USA Armor Bd Test Sec (1)
USA AD Bd Test Sec (1)
USA Arctic Test Bd (1)
US ARADCOM (2)
US ARADCOM Rgn (2)
OS Maj Comd (5)
OS Base Comd (5)
MDW (1)
Armies (5)
Corps (2)
Div (2)
USATC (2)
Svc Colleges (5)
Br Svc Sch (5) except USASCS (75)
Sig Sec, Gen Dep (10)
Sig Dep (17)
Army Pictorial Cen (2)
USA Sig Pub Agcy (500)
USA Sig Engr Agcy (1)
USA Comm Agcy (2)
USA Sig Eqp Spt Agcy (2)
USA Sig Msl Spt Agcy (2)
WRAMC (1)
AFIP (1)
Yuma Test Sta (2)
USA Elct PG (1)
Sig Lab (5)
Sig Fld Maint Shops (3)
Fld Comd, AFSWP (5)
Mil Dist (1)
USA Corps (Res) (1)
Sector Comd, USA Corps (Res) (1)
JBUSMC (2)

Units organized under following TOE's:

11–7 (2)	11–97 (2)
11–15 (2)	11–98 (2)
11–16 (2)	11–117 (2)
11–22 (2)	11–500 (AA–AE) (2)
11–25 (2)	11–537 (2)
11–32 (2)	11–557 (2)
11–39 (2)	11–587 (2)
11–57 (2)	11–592 (2)
11–95 (2)	11–597 (2)
11–96 (2)	

NG: State AG (3); units—same as Active Army except allowance is one copy to each unit.
USAR: None.
For explanation of abbreviations used, see AR 320–50.

For sale by the Superintendent of Documents, U. S. Government Printing Office
Washington 25, D. C. - Price $1.25

☆ U.S. GOVERNMENT PRINTING OFFICE: 1959 O—495909