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NEW --- second edition

PRINCIPLES OF TRANSISTOR CIRCUITS

by S. W. AMOS

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PRINCIPLES OF TRANSISTOR CIRCUITS

Introduction to the Design of Amplifiers, Receivers and other Circuits

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SECOND EDITION

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CONTENTS

PREFACE TO SECOND EDITION

SINCE the first edition of this book was written a number of significant developments have occurred in the science of semiconduction. The chief of these have been the disappearance of the point-contact transistor and the adoption of the drift transistor for r.f. applications. To keep the book up-to-date, appropriate changes have been made to the text of this second edition: all references to point-contact transistors have been omitted and the space saved has been devoted to drift transistors and their applications in pulse amplifiers and v.h.f. receivers. Other new topics included in the second edition are voltage-reference diodes and controlled rectifiers. The sections on transistor multivibrators and on miscellaneous applications of semiconductor devices have been considerably expanded and are now included in separate chapters.

The opportunity has been taken of rewriting parts of Chapters 1 and 2 which describe the physics of the junction diode and the transistor. The author would like to express his gratitude to Dr. J. R. Tillman for many valuable suggestions incorporated in the revised text of these two chapters.

Thanks are also due to Dr. T. B. Tomlinson for information which proved helpful in the preparation of the new chapter on transistor multivibrators.

Finally, mention must also be made of a number of useful comments (embodied in this new edition) which have been received from Dr. Charles M:son Gewertz, Mr. B. C. Jones, of Murphy Radio Limited, and the author's colleagues in the B.B.C. Engineering Training Department.

Stanmore
December, 1960

PREFACE TO FIRST EDITION

The introduction of junction transistors has made possible the construction of electronic equipment of small size and extreme economy in running costs. In particular, transistors are well suited for use in miniature radio receivers and amplifiers and are thus likely to stimulate interest among amateur constructors. To help designers of transistorised equipment there is a need for a book giving the basic principles of transistor circuits: this small volume was written in an effort to provide the elementary principles required.

The book begins with introductory chapters on the physical processes occurring in transistors but the emphasis is on applications rather than physics and the bulk of the book is devoted to determinations of such quantities as input resistance, stage gain, optimum load, power output, values of coupling capacitors and transformer winding inductances. The mathematics is confined to algebraic manipulation and is illustrated by a large number of numerical examples which show the order of practical magnitudes of these quantities.

Some details of transistor relaxation oscillators and photo-sensitive devices are given in the final chapter and the book ends with a short account of a number of new types of transistor which may become important in the future by extending the frequency range over which semiconducting devices can operate satisfactorily.

Stanmore
January, 1959

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SEMICONDUCTORS AND JUNCTION DIODES

Introduction

ONE of the most outstanding developments in electronics since the Second World War has been the introduction of the transistor, a minute semiconducting device which can perform most functions of a thermionic valve but with far greater efficiency. A transistor may be smaller than a $\frac{1}{4}$ in. cube, has no heater, may consume less than 1 mA from a 3-volt source, yet can be as effective as an amplifier or oscillator as a thermionic valve consuming, say, 50 mA of filament current at 1·4 volts and 2 mA of anode current at 60 volts. In addition to its greater efficiency, the transistor is non-microphonic and has a life many times that of a valve. Early junction transistors operated satisfactorily up to a few Mc/s only, but modern (drift) types can be used above 100 Mc/s. The output power of transistors has also steadily increased and a modern power transistor can deliver several watts in class-A operation.

One disadvantage of transistors compared with thermionic valves is that the current they draw from the h.t. supply increases with temperature. The dissipation produced by this current in the transistor itself generates heat and a regenerative action is possible, ending in a rapid rise of current which can damage or even destroy the transistor. This phenomenon is termed *thermal runaway* and, where danger of it exists, the circuits associated with the transistor must be designed to prevent runaway by making the current drawn from the supply less dependent of ambient and of transistor temperature. Stabilisation circuits of this type are discussed fully in Chapter 6. The danger of thermal runaway is a very real one in germanium transistors, and most of the transistors in use at present employ this element. For silicon transistors the usable range of temperature is much greater—and the danger of thermal runaway at normal operating temperatures is much less.

Although a transistor can perform nearly all the functions of a valve, such as detection, amplification, oscillation, pulse-generation and frequency changing, its properties are different from those of a valve and it therefore requires specialised circuitry. The principal difference is in the input and output resistances and in the internal feedback of the transistor. For a transistor the input is small compared with the output resistance, and to obtain efficient operation the associated circuit must be designed to allow for this. The internal feedback must in many circuits be neutralised or otherwise allowed for.

Possibly the most obvious application for transistors is in deaf-aid equipment where their small size makes possible the construction of unobtrusive amplifiers which can, in fact, be built into the framework of spectacles. Such amplifiers can operate for long periods on a single 1·5-volt cell and have very low running costs. Transistors have already supplanted valves in portable radio receivers and here too they have made possible a substantial reduction in running costs. A receiver may need five or six transistors to give a performance comparable with that of a four-valve battery-driven superhet but the average current consumption is not likely to exceed 25 mA at 6 volts, which represents less power than is consumed by the filaments alone of the valve receiver. Transistors are well suited for use in car radio receivers because they operate well on a 6-volt or 12-volt supply such as a car battery. Thus a completely-transistorised car radio requires only a 6-volt or 12-volt supply and there is no need for the bulky and noisy vibrator or rotary unit required in an all-valve car radio to supply the h.t. Thermionic valves can give a useful performance as voltage amplifiers and frequency changers even when operated from an h.t. supply as low as 12 volts but they cannot deliver the power output expected from a car radio from such a low-voltage supply. A transistor can, however, deliver the required a.f. output from such a supply and it is therefore possible to construct a 12-volt car radio with valves in all stages except the last which is transistorised. Such a radio is termed a *hybrid* type and it has the advantage of the fully-transistorised type that it eliminates the need for an h.t. unit. Some care is needed in the design of a car radio including transistors to protect them from thermal runaway because it is possible to obtain high temperatures inside a closed car on a hot summer day.

Transistors are well suited for use in portable public-address amplifiers, and it is possible to construct equipment to deliver, say, 20 watts of power, which operates from a 6-volt or 12-volt supply.

Transistors have been used in telephone equipment such as repeaters and also in walkie-talkie equipment. Transistor transmitters and receivers have even been successfully operated from solar cells (i.e. cells which derive their power from the sun): the running cost of such equipment is literally nothing.

DEFINITION OF A SEMICONDUCTOR

The heart of a transistor consists of semiconducting material, e.g., germanium or silicon and the behaviour of the transistor largely depends on the properties of this material. As the name suggests a semiconducting material is one with a conductivity lying between that of an insulator and that of a conductor: that is to say one for which the resistivity lies between, say, 10^{12} ohm-cm (a value typical of glass) and 10^{-6} ohm-cm (approximately the value for copper). Typical values for the resistivity of a semiconducting material lie between 1 and 100 ohm-cm.

Such a value of resistivity could, of course, be obtained by mixing a conductor and an insulator in suitable proportions but the resulting material would not be a semiconductor. Another essential feature of a semiconducting material is that its electrical

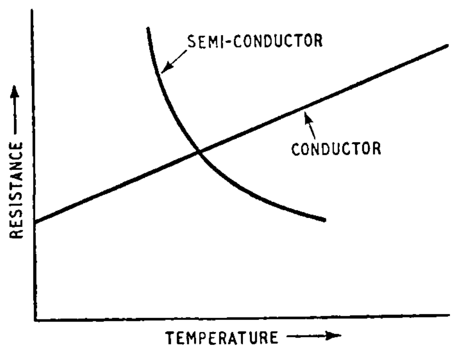


Fig. 1.1. Resistance-temperature relationship for a conductor and a semiconductor

resistance decreases with increase in temperature over a particular temperature range which is characteristic of the semiconductor. This behaviour contrasts with that of elemental metallic conductors for which the resistance increases with rise in temperature. This is illustrated in Fig. 1.1, which gives curves for a conductor and a semiconductor. The curve for the conductor shows the resistance increasing linearly with increase in temperature, whereas that for the semiconductor shows the resistance decreasing exponentially with increase in temperature. Over the significant temperature range the relationship between resistance and temperature for a semiconductor could be written

$$R_t = ae^{b/T}$$

where R_t is the resistance at an absolute temperature T, a and b being constants characteristic of the semiconducting material. The two curves in Fig. 1.1 are not to the same vertical scale of resistance.

All semiconducting materials exhibit the temperature dependence discussed in the paragraphs above in the pure state: the addition of impurities raises the temperature at which the material exhibits this behaviour, i.e. the region of negative temperature coefficient.

Germanium in its pure state is a poor conductor, the resistivity being 46 ohm-cm at 27° C, and is of little direct use in transistor manufacture. However, by the addition of a very small but definite amount of a particular type of impurity, the resistivity can be reduced and the material made suitable for transistors. Germanium and silicon so treated are extensively employed in the manufacture of transistors.

The behaviour of semiconductors can be explained in terms of atomic theory. The atom is assumed to have a central nucleus which carries most of the mass of the atom and has a positive charge. A number of electrons carrying a negative charge revolve around the nucleus. The total number of electrons revolving around a particular nucleus is sufficient to offset the positive nuclear charge, leaving the atom electrically neutral. The number of electrons associated with a given nucleus is equal to the atomic number of the element. The electrons revolve in a number of orbits and, for the purpose of this discussion, the orbits may be regarded as concentric, the nucleus being at the centre, as shown in Fig. 1.2. This diagram is greatly simplified; the orbits are in practice neither concentric nor co-planar.

The first orbit (sometimes called a ring or a shell) is complete when it contains 2 electrons, and an atom with a single complete shell is that of the inert gas, helium. The second ring is complete when it has 8 electrons, and the atom with the first 2 rings complete is that of the inert gas, neon. The third ring is stable when it has 8 or 18 electrons, and the atom having 2, 8 and 8 electrons in the 1st, 2nd and 3rd rings is that of the inert gas, argon. All the inert gases have their outermost shells stable. It is difficult to remove any electrons from a stable ring or to insert others into it. Atoms combine by virtue of the electrons in the outermost rings: for example, an atom with one electron in the outermost ring will willingly combine with another whose outermost ring requires one electron for completion.

The inert gases, having their outer shells stable, cannot combine with other atoms or with each other. The number of electrons in the outermost ring or the number of electrons required to make the outermost ring complete has a bearing on the chemical valency of the element and the outermost ring is often called the valence ring.

Now consider the copper atom: it has 4 rings of electrons, the first 3 being complete and the 4th containing 1 electron, compared with the 32 needed for completion. Similarly the silver atom has

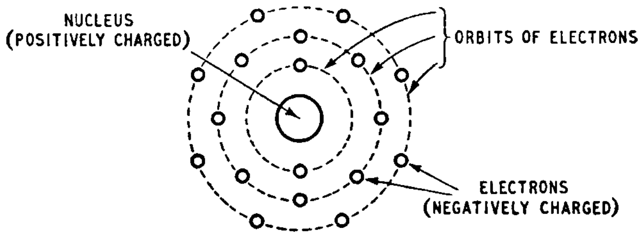


Fig. 1.2. Simplified diagram of structure of atom: for simplicity, electron orbits are shown as circular and co-planar

5 rings, 4 stable and the 5th also containing 1 out of 50 needed for completion. The atoms of both elements thus contain a single electron and this is loosely bound to the nucleus. It can be removed with little effort and is termed a free electron. A small e.m.f. applied to a collection of these atoms can set up a stream of free electrons, i.e., an electric current through the metal. Elements in which such free electrons are available are good electrical conductors.

It might be thought that an atom with 17 electrons in the outermost orbit would be an even better conductor, but this is not so. If one electron is added to such an orbit it becomes complete and a great effort is needed to remove it again.

The arrangement of orbital electrons in a germanium atom is pictured in Fig. 1.3. There are 4 rings, the first containing 2 electrons, the second 8, the third 18, and the 4th (final) 4. The total number of electrons is 32, the atomic number of germanium. The corresponding diagram for the silicon atom is given in Fig. 1.4, the three rings containing 2, 8 and 4 electrons respectively. The total number of electrons per atom is 14, the atomic number for silicon. A significant feature of these two atomic structures is that

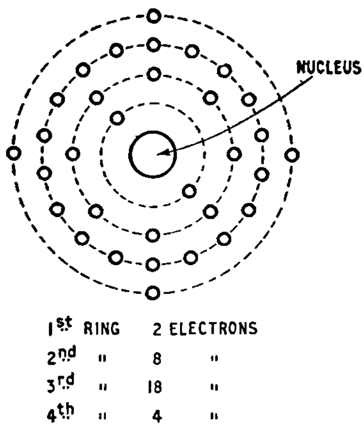


Fig. 1.3. Structure of germanium atom

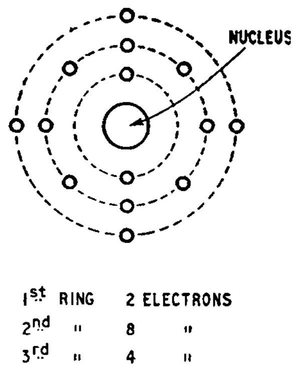


Fig. 1.4. Structure of silicon atom

the outermost ring contains 4 electrons: both elements belong to Group IV of the Periodic Table.

COVALENT BONDS

It might be thought that some of the 4 electrons in the valence ring of the germanium or silicon atom could easily be displaced and that these elements would therefore be good conductors. In fact, crystals of pure germanium and pure silicon are very poor conductors. To understand this we must consider the relationships between the valence electrons of neighbouring atoms when these are arranged in a regular geometric pattern as in a crystal. The valence electrons of each atom form bonds, termed *covalent bonds*, with those of neighbouring atoms as suggested in Fig. 1.5. It is difficult to portray a 3-dimensional phenomenon in a 2-dimensional diagram, but the diagram does show the valence electrons oscillating between two neighbouring atoms. The atoms behave in some respects as though each outer ring had 8 electrons and was stable. There are no free electrons and such a crystal is therefore an insulator: this is true of pure germanium and pure silicon at a very low temperature.

At room temperatures, however, germanium and silicon crystals do have a small conductivity even when they are as pure as modern chemical methods can make them. This is partly due to the presence of minute traces of impurities (the way in which these increase conductivity is explained below) and partly because thermal agitation enables some valence electrons to escape from their covalent bonds and thus becomes available as current carriers. They are able to do this by virtue of their kinetic energy which, at normal temperatures, is sufficient to allow a very small number to break these bonds. If their kinetic energy is increased by the addition of light or by increase in temperature, more valence electrons escape and the conductivity increases. When the temperature of germanium is raised to 100° C, the conductivity is so great that it swamps normal transistor action. Moreover, if a reasonable life is required, it is recommended that germanium transistors should not be operated above say 80° C. The life of a

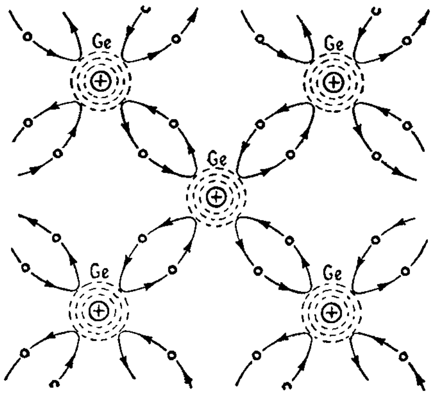


Fig. 1.5. Illustrating covalent bonds in a crystal of pure germanium: for simplicity, only electrons in the valence rings are shown

germanium transistor is shortened if it is operated above this temperature but a silicon transistor will give a satisfactory life even when operated at 150° C.

DONOR IMPURITIES

Suppose an atom of a Group V element such as arsenic is introduced into a crystal of pure germanium. The atom enters into the

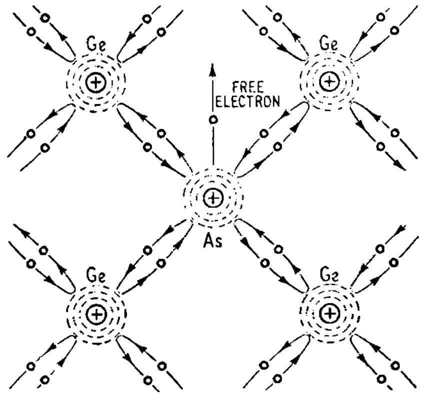


Fig. 1.6. Illustrating covalent bonds in the neighbourhood of an atom of a Group V element introduced into a crystal of pure germanium. For simplicity, only electrons in the valence rings are shown

lattice structure, taking the place of a germanium atom. Now the arsenic atom has 5 electrons in its outermost orbit and 4 of these form covalent bonds with the electrons of neighbouring atoms as shown in Fig. 1.6. The remaining (5th) electron is left unattached; it is a free electron which can be made to move through the crystal by an e.m.f., leaving a positively-charged ion. These added electrons give the crystal much better conductivity than pure germanium and the added element is termed a *donor* because it gives free electrons to the crystal. Germanium so treated with a Group V element is termed n-type because negatively-charged particles are available to carry current through the crystal. It is significant that the addition of the arsenic or some other Group V element was necessary to give this improvement in conductivity. The added element is often called an impurity and in the language of the chemist it undoubtedly is. However, the word is unfortunate in this context because it suggests that the pentavalent element is unwanted; in fact, it is essential.

When a battery is connected across a crystal of n-type germanium the free electrons are attracted towards the battery positive terminal and repelled from the negative terminal. These forces cause a drift of electrons through the crystal from the negative to the positive terminal: for every electron leaving the crystal to enter the positive terminal another must be liberated from the negative terminal to enter the crystal. The stream of electrons through the crystal constitutes an electric current. If the battery connections are reversed the direction of the current through the crystal also reverses but it does not change in amplitude; that is to say the crystal is a *linear* conductor. Some part of a transistor is always of n-type material.

ACCEPTOR IMPURITIES

Now suppose an atom of a Group III element such as indium is introduced into a crystal of pure germanium. It enters the lattice structure, taking the place of a germanium atom, and the 3 electrons in the valence ring of the indium atom form covalent bonds with the valence electrons of the neighbouring germanium atoms. To make up the number of covalent bonds to four, each indium atom competes with a neighbouring germanium atom and may leave this deficient of one electron as shown in Fig. 1.7. A group of covalent bonds, which is deficient of one electron, behaves in much the same way as a positively-charged particle with a charge equal in magnitude to that of an electron. Such a particle is called a *hole* in semiconductor theory, and we may say that the introduction

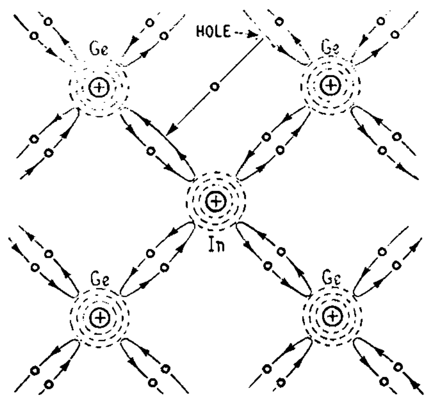


Fig. 1.7. Illustrating covalent bonds in the neighbourhood of an atom of a Group III element, introduced into a crystal of pure germanium. For simplicity, only electrons in the valence rings are shown

of the Group III impurity gives rise to holes in a crystal of pure germanium. These can carry current through the crystal and, because these current-carriers have a positive charge, germanium treated with a Group III impurity is termed p-type. Such an impurity is termed an *acceptor* impurity because it takes electrons from the germanium atoms. Thus the introduction of the Group III element into a crystal lattice of pure germanium also increases the conductivity considerably and, when a battery is connected across a crystal of p-type germanium, a current can flow through it in the following manner.

The holes have an effective positive charge, and are therefore attracted towards the negative terminal of the battery and repulsed by the positive terminal. They therefore drift through the crystal from the positive to the negative terminal. Each time a hole reaches the negative terminal, an electron is emitted from this terminal into the hole in the crystal to neutralise it. At the same time an electron from a covalent bond enters the positive terminal to leave another hole in the crystal. This immediately moves towards the negative terminal, and thus a stream of holes flows through the crystal from the positive to the negative terminal. The battery thus loses a steady stream of electrons from the negative terminal and receives a similar stream at its positive terminal. It may be said that a stream of electrons has passed through the crystal from the negative to the positive terminal. A flow of holes is thus equivalent to a flow of electrons in the opposite direction.

If the battery connections are reversed, the current in the crystal also reverses in direction but has the same amplitude: thus p-type germanium is also a linear conductor.

It is astonishing how small the impurity concentration must be to make germanium suitable for use in transistors. A concentration of 1 part in 10^6 may be too large, and concentrations commonly used are of a few parts in 10^8. A concentration of 1 part in 10^8 increases the conductivity by 16 times. Before such a concentration can be introduced, the germanium must first be purified to such an extent that any impurities still remaining represent concentrations very much less than this. Purification was one of the most difficult processes in the manufacture of transistors.

INTRINSIC AND EXTRINSIC GERMANIUM

If a germanium crystal contains no impurities, the only current carriers present are those produced by thermal breakdown of the covalent bonds. The conducting properties are thus characteristic of pure germanium. Such a crystal is termed an *intrinsic* semiconductor.

In general, however, germanium crystals contain some trivalent and some pentavalent impurities, i.e. some donors and some acceptors are present. Some free electrons fit into some holes and neutralise them but there are some residual current carriers left. If these are mainly electrons they are termed *majority carriers* (the holes being *minority carriers*), and the material is n-type. If the residual current carriers are mainly holes, these are majority carriers (the electrons being minority carriers) and the germanium is termed p-type. In an n-type or p-type crystal the impurities are chiefly responsible for the conduction, and the material is termed an *extrinsic semiconductor*.

PN JUNCTIONS

As already mentioned, n-type or p-type germanium is a linear conductor, but if a crystal of germanium has n-type conductivity at one end and p-type at the other end, as indicated in Fig. 1.8,

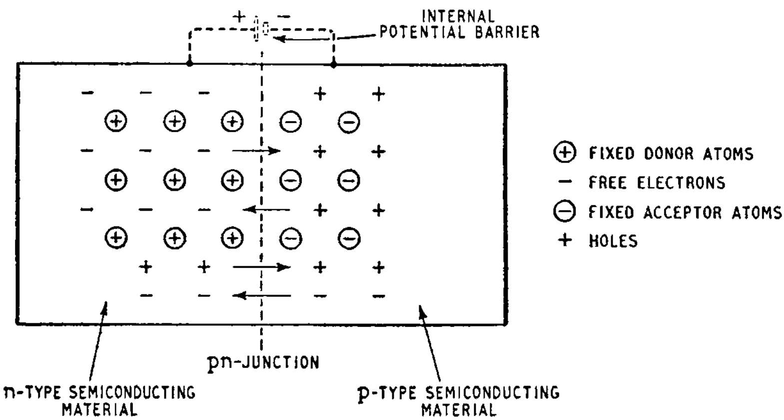


Fig. 1.8. Pattern of fixed and mobile charges in the region of a pn junction

the crystal so produced has asymmetrical conducting properties. That is to say, the current which flows in the crystal when an e.m.f. is applied between the ends depends on the polarity of the e.m.f., being small when the e.m.f. is in one direction and large when it is reversed. Crystals with such conductive properties have obvious applications as detectors or rectifiers.

It is not possible, however, to produce a structure of this type by placing a crystal of n-type germanium in contact with a crystal of p-type germanium. No matter how well the surfaces to be placed together are planed, or how perfect the contact between the two appears, the asymmetrical conductive properties are not properly obtained. The usual way of achieving a structure of this type is by treating one end of a single crystal of n-type germanium with a Group III impurity so as to offset the n-type conductivity at this end and to produce p-type conductivity instead at this point. Alternatively, of course, one end of a p-type crystal could be treated with a Group V impurity to give n-type conductivity at this end. The semiconducting device so obtained is termed a junction diode, and the non-linear conducting properties can be explained in the following way.

Behaviour of a pn or np Junction

Fig. 1.8 represents the pattern of charges in a crystal containing an np junction. The ringed signs represent charges due to the impurity atoms and are fixed in position in the crystal lattice: the unringed signs represent the charges of the free electrons and holes (majority carriers) which are liberated by the impurities. The n-region contains a few holes and the p-region a few free electrons: these are minority carriers liberated by thermal dissociation of the covalent bonds of the semiconducting element itself.

Even when no external connections are made to the crystal, there is a tendency, due to diffusion, for the free electrons of the n-region to cross the junction into the p-region: similarly the holes in the p-region tend to diffuse into the n-region. However the moment any of these majority carriers cross the junction, the electrical neutrality of the two regions is upset: the n-region loses electrons and gains holes, causing it to become positively charged with respect to the p-region. Thus a potential difference is established across the junction and this discourages further majority carriers from crossing the junction: indeed only the few majority carriers with sufficient energy succeed in crossing. The potential difference is, however, in the right direction to encourage minority carriers to cross the junction and these cross readily in just sufficient numbers to balance the subsequent small flow of majority carriers. Thus the balance of charge is preserved even though the crystal has a potential barrier across the junction. In Fig. 1.8 the internal potential barrier is represented as an external battery and is shown in dotted lines.

Reverse-bias Conditions

Suppose now an external battery is connected across the junction, the negative terminal being connected to the p-region and the positive terminal to the n-region as shown in Fig. 1.9. This connection gives a *reverse-biased* junction. The external battery is in parallel with and aiding the fictitious battery, increasing the potential barrier across the junction. Even the majority carriers with the greatest energy now find it almost impossible to cross the

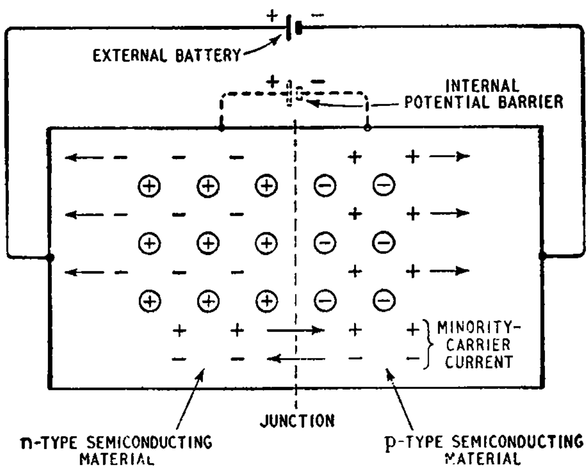
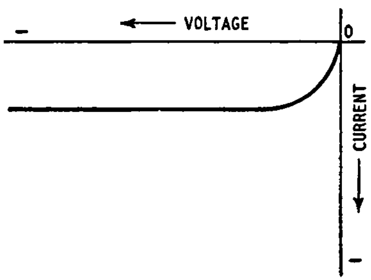


Fig. 1.9. Reverse-bias conditions in a pn junction



Left: Fig. 1.10. Current-voltage relationship for a reverse-biased pn junction

junction. On the other hand the minority carriers can cross the junction as easily as before and a steady stream of these flows across. When the minority carriers cross the junction they are attracted by the battery terminals and can then flow as a normal electric current in a conductor. Thus a current, carried by the minority carriers and known as the reverse current, flows across the junction. It is a small current because the number of minority carriers is small: it increases as the battery voltage is increased as shown in Fig. 1.10 but at a reverse voltage of less than 1 volt becomes constant: this is the voltage at which the rate of flow of minority carriers becomes equal to the rate of production of carriers by thermal breakdown of covalent bonds. Increase in the temperature of the crystal produces more minority carriers and an increase in reverse current.

Forward-bias conditions

If the external battery is connected as shown in Fig. 1.11, with the positive terminal connected to the p-region and the negative terminal to the n-region, the junction is said to be *forward-biased.* The external battery now opposes and reduces the potential barrier due to the fictitious battery and the majority carriers are now able to cross the junction more readily. A steady flow of majority

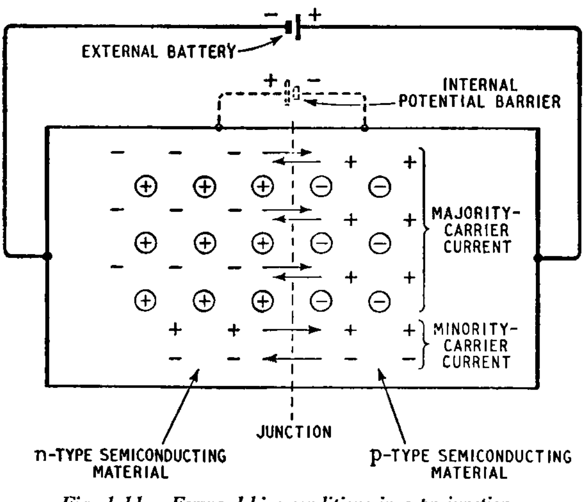


Fig. 1.11. Forward-bias conditions in a pn junction

electrons and majority holes can now flow across the junction and these together constitute a considerable current from the external battery. The flow of minority carriers across the junction also continues as in reverse-bias conditions but at a reduced scale and these give rise to a second current also taken from the battery but in the opposite direction to that carried by the majority carriers. Except for very small external battery voltages, however, the minority-carrier current is very small compared with the majority-carrier current and can normally be neglected in comparison with it.

The relationship between current and forward bias voltage is illustrated in Fig. 1.12. The curve has a small slope for small

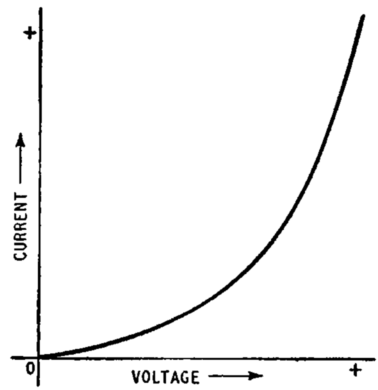


Fig. 1.12. Current-voltage relationship for a forward-biased pn junction

voltages because the internal potential barrier discourages movements of majority carriers across the junction. Increase in applied voltage tends to offset the internal barrier and current increases at a greater rate. Further increase in voltage almost completely offsets the barrier and gives a steeply-rising current. The curve is, in fact, closely exponential in form.

Junction Diodes

Junction diodes are very efficient and therefore little heat is generated in them in operation. Such diodes can therefore rectify surprisingly large currents: for example a silicon diode with a junction area less than ½ inch in diameter can supply 50 A at 100 volts. The capacitance between the terminals of a small-area junction is low enough for it to make an efficient detector.

Avalanche Effect

When a pn junction is reverse-biased the current is carried solely by the minority carriers, and at a given temperature the number of minority carriers is fixed. Ideally, therefore, we would expect the reverse current for a pn junction to rise to a saturation value as the voltage is increased from zero and then to remain constant and independent of voltage, as shown in Fig. 1.10. In practice, when the reverse voltage reaches a particular value which can be 100 volts or more the reverse current increases very sharply as shown in Fig. 1.13, an effect known as breakdown. The effect is reproducible,

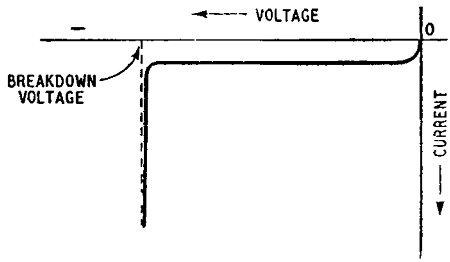


Fig. 1.13. Breakdown in a reverse-biased pn junction

breakdown in a particular junction always occurring at the same value of reverse voltage. This is known as the Avalanche effect and reversed-biased diodes known as Avalanche diodes (sometimes called – perhaps incorrectly – Zener diodes) can be used as the basis of a voltage stabiliser circuit. The junction diodes used for this purpose are usually silicon types and examples of voltage stabilising circuits employing such diodes are given in Chapter 13.

The explanation of the Avalanche effect is thought to be as follows. The reverse voltage applied to a junction diode establishes an electric field across the junction and minority electrons entering if from the p-region are accelerated to the n-region as illustrated in Fig. 1.9. When this field exceeds a certain value some of these electrons collide with valence electrons of the atoms fixed in the crystal lattice and liberate them, thus creating further hole-electron pairs. Some new carriers are themselves accelerated by the electric field due to the reverse bias and in turn collide with other atoms, liberating still further holes and electrons. In this way the number of current carriers increased very rapidly: the process is, in fact, regenerative. This multiplication in the number of current carriers produces the sharp increase in reverse current shown in Fig. 1.13. Once the breakdown voltage is exceeded, a very large reverse current can flow and unless precautions are taken to limit this current the junction can be damaged by the heat generated in it. Voltage stabilising circuits using Avalanche diodes must therefore include protective measures to avoid damage due to this cause.

Zener Effect

Some reverse-biased junction diodes exhibit breakdown at a very low voltage, say below 5 volts. In such examples breakdown is thought to be due, not to Avalanche effect, but to Zener effect which does not involve ionization by collision. Zener breakdown is attributed to spontaneous generation of hole-electron pairs within the junction region from the inner electron shells. Normally this region is carrier-free but the intense field established across the region by the reverse bias can produce carriers which are then accelerated away from the junction by the field, so producing a reverse current.

Capacitance of a Junction Diode

As pointed out above, the application of reverse bias to a pn junction discourages majority carriers from crossing the junction, and tends to produce a structure in which the carriers are separated by a relatively carrier-free region as suggested in Fig. 1.9. Such a structure is similar to that of a charged capacitor and, in fact, a reverse-biased junction diode has the nature of a capacitance shunted by a high resistance. The value of the capacitance is dependent on the reverse bias voltage and can be varied over wide

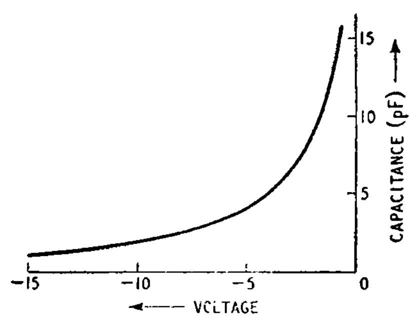


Fig. 1.14. Typical capacitance-voltage characteristic for a reverse-biased pn junction

limits by alteration in the bias voltage. This is illustrated in the curve of Fig. 1.14: the capacitance varies with the voltage according to a law of the type

$$C \simeq kV^{-1/n}$$

where k is a constant and n is between 2 and 3. When n is 2 the capacitance is inversely proportional to the square root of the voltage. A voltage-sensitive capacitance such as this has a number of useful applications: it can be used as a frequency modulator, as a means of remote tuning in receivers or for a.f.c. purposes in receivers. An example of one of these applications of the reverse-biased junction diode is given in Chapter 13.

Use of junction diode for voltage reference

The breakdown voltage of a reverse-biased junction diode can be placed within the range of a few volts to several hundred volts but for stabiliser and voltage reference applications it is unusual to employ a diode with a breakdown voltage exceeding a few tens of volts. Some of the reasons for this are given below.

Firstly the breakdown voltage varies with temperature, the coefficient of variation being negative for diodes with breakdown voltages less than approximately 5·3 volts and positive for diodes with breakdown voltages exceeding approximately 6·0 volts. Diodes with breakdown voltages between these two limits have very small coefficients of variation and are thus well suited for use in voltage stabilisers. However for voltage reference purposes the slope resistance of the breakdown characteristic must be very small (see p. 187) and the slope is less for diodes with breakdown voltages exceeding 6 volts than for those with lower breakdown voltages. Where variations in temperature are likely to occur it is probably best to use a diode with a breakdown voltage between 5·3 and 6·0 volts for voltage reference purposes but if means are available for stabilising the temperature it is probably better to use a diode with a higher breakdown voltage to obtain a lower slope resistance. Diodes with breakdown voltages around 6·8 volts have a temperature coefficient (2·5 mV/° C) which matches that of transistors; this can be useful in designing stabilised power supplies.

Voltage reference diodes are usually marketed with preferred values of breakdown voltage (4·7V, 5·6V, 6·8V etc.) with tolerances of 5 per cent or 10 per cent.

2

BASIC PRINCIPLES OF JUNCTION TRANSISTORS

Introduction

IN Chapter 1 we discussed semiconductors and showed that a junction between n-type and p-type materials has asymmetrical conducting properties enabling it to be used for rectification. A triode junction transistor includes two such junctions and they are arranged as shown in the theoretical diagram of Fig. 2.1. Fig. 2.1 (a) illustrates one basic type consisting of a layer of

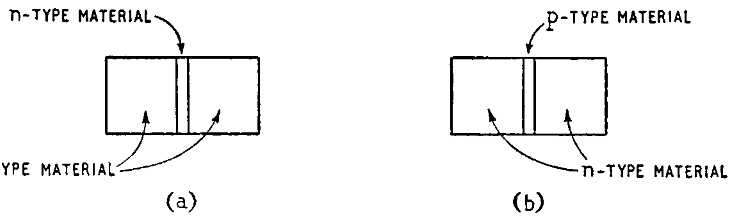


Fig. 2.1. Theoretical diagrams illustrating the structure of (a) a pnp and (b) an npn transistor

n-type material sandwiched between two layers of p-type material: such a transistor is referred to as a pnp type.

A second type, illustrated in Fig. 2.1 (b) has a layer of p-type material sandwiched between two layers of n-type semiconducting material: such a transistor is referred to as an npn type.

In both types, for successful operation, it is essential that the central layer should be thin. However, it is not possible to construct junction transistors by placing suitably-treated layers of semiconducting material in contact. One method which is employed is to start with a single crystal of, say, n-type germanium and to treat it so as to produce regions of p-type conductivity on either side of the remaining region of n-type conductivity.

Electrical connections are made to each of the three different regions of a triode transistor as suggested in Fig. 2.2. The thin central layer is known as the *base* of the transistor and corresponds with the control grid of a triode valve. One of the remaining two layers is known as the *emitter* and corresponds with the cathode of a triode valve. For efficient operation the emitter must have much greater conductivity than the base and must of course be of the opposite conductivity, i.e. must be p-type if the base is n-type. The

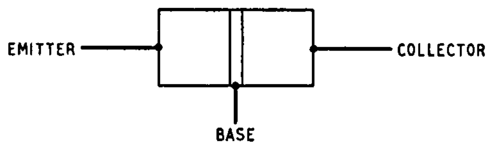


Fig. 2.2. Electrical connections to a junction transistor

remaining (third) layer is known as the *collector*: it corresponds with the anode of the triode. The transistor may be symmetrical and either of the outer layers may then be used as emitter: the operating conditions determine which of the outer layers behaves as emitter, because in normal operation the emitter-base junction is forward-biased whilst the base-collector junction is reverse-biased. In

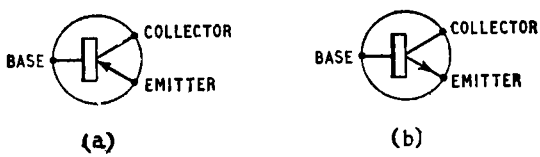


Fig. 2.3. Circuit diagram symbols for (a) a pnp and (b) an npn junction transistor

practice most junction transistors are unsymmetrical with the collector junction larger than the emitter junction and it is essential to adhere to the emitter and collector connections prescribed by the manufacturer.

The symbols used for junction transistors in circuit diagrams are given in Fig. 2.3. The symbol shown at (a), in which the emitter arrow is directed towards the base, is used for a pnp transistor and the symbol shown at (b), in which the emitter arrow is directed away from the base, is used for an npn transistor.

Construction of an alloy junction transistor

A practical form of construction for a pnp junction transistor is illustrated in Fig. 2.4. The basic element is a crystal or wafer of n-type germanium about 0·1 in. square and 0·005 in. thick. A small pellet of a trivalent element such as indium is placed near the centre of each of the two opposite faces of the wafer and the combination is heated in an oven to a temperature between the melting point of indium and that of germanium. This causes the indium pellets to alloy with the germanium to form regions of p-type conductivity separated from each other by a thin layer having the n-type conductivity of the original crystal. Connecting leads are soldered to each of the indium pellets and to the germanium crystal and the assembly is hermetically sealed in a light-proof container. For best performance from the transistor the indium pellet used as the collector electrode must usually be larger than the other. In a symmetrical transistor the indium pellets are of the same size;

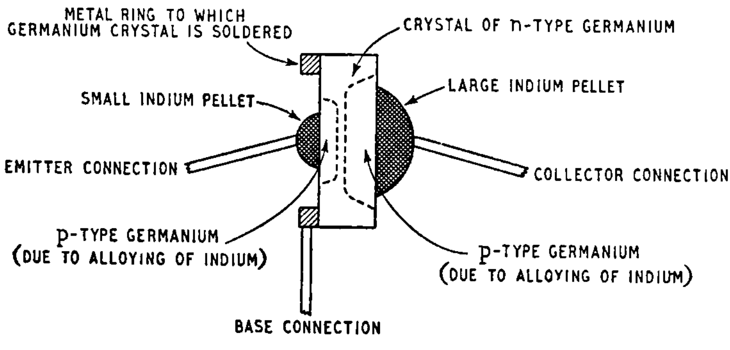


Fig. 2.4. Construction of alloy junction pnp germanium transistor

this produces a transistor more suitable for switching applications than for amplification.

A similar technique can be used to prepare npn transistors: a crystal of p-type germanium is used and pellets of a pentavalent element such as arsenic or antimony, carried perhaps in a neutral (Group IV) element such as lead, are alloyed to it. This alloying process is only one of a number of methods which can be used in the manufacture of transistors.

Operation of a pnp Transistor

Fig. 2.5 illustrates the polarity of the potentials which are neces sary in a pnp-transistor amplifying circuit. The emitter is biased slightly positively with respect to the base: this is an example of forward bias and the external battery opposes the internal potential barrier associated with the emitter-base junction. A considerable

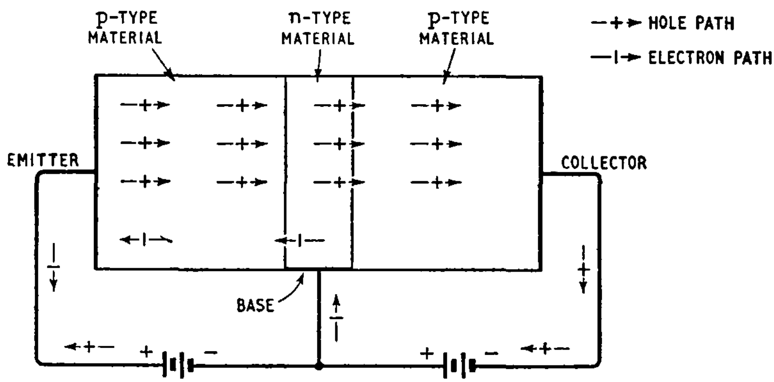


Fig. 2.5. Hole and electron paths in a pnp junction transistor connected for amplification

current therefore flows across this junction and this is carried by holes from the p-type emitter (which move to the right into the base) and by electrons from the n-type base (which move to the left into the emitter). However because the impurity concentration in the emitter is normally considerably greater than that of the base (this is adjusted during manufacture), the holes carrying the emitter-base current greatly outnumber the electrons and we can say with little error that the current flowing across the emitter-base junction is carried by holes moving from emitter to base.

The collector is biased negatively with respect to the base: this is an example of reverse bias and the external battery aids the internal potential barrier associated with the base-collector junction. If the emitter-base junction were also reverse-biased, no holes would be injected into the base region from the emitter and only a very small current would flow across the base-collector junction. This is the reverse current (described in Chapter 1): it is a saturation current independent of the collector-base voltage. However, when the emitter-base junction is forward-biased, the injected holes have a marked effect on the collector current: this is, in fact, the essence of transistor action. The holes are injected into the base, which is a particularly thin layer; most of them cross the base by diffusion and on reaching the collector-base junction are swept into the collector region. The reverse bias of the base-collector junction ensures the collection of all the holes crossing this junction, whether these are present in the base region as a result of breakdown of covalent bonds by virtue of thermal agitation or are injected into it by the action of the emitter. A few of the holes which leave the emitter combine with electrons in the base and so cease to exist but the majority of the holes (commonly more than 95 per cent) succeed in reaching the collector. Thus the increase in collector current due to hole-injection by the emitter is nearly equal to the current flowing across the emitter-base junction. The balance of the emitter carriers (equal to, say, 5 per cent) are neutralised by electrons in the base region and to maintain charge neutrality more electrons flow into the base, constituting a base current. The collector current, even though it may be considerably increased by forward bias of the emitter-base junction, is still independent of the collector voltage. This is another way of saying that the output resistance of the transistor is extremely high: it can in fact be several megohms. The input resistance is approximately that of a forward-biased junction diode and is commonly of the order of 25 ohms. A small change in the input (emitter) current of the transistor is faithfully reproduced in the output (collector) current even though the output resistance may be 40,000

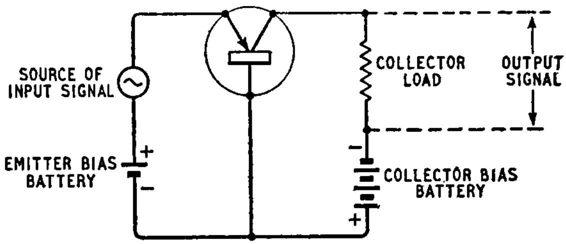


Fig. 2.6. Basic circuit for using a pnp junction transistor as an amplifier

times the input resistance: this marked disparity between input and output resistance enables the transistor to act as a very effective power amplifier. This point is illustrated by numerical examples later in this chapter.

Bias Supplies for a pnp Transistor

Fig. 2.6 shows a pnp transistor connected to supplies as required in one form of amplifying circuit. For forward bias of the emitter-base junction, the emitter is made positive with respect to the base; for reverse bias of the base-collector junction, the collector is made negative with respect to the base. Fig. 2.6 shows separate batteries used to provide these two bias supplies and it is significant that the

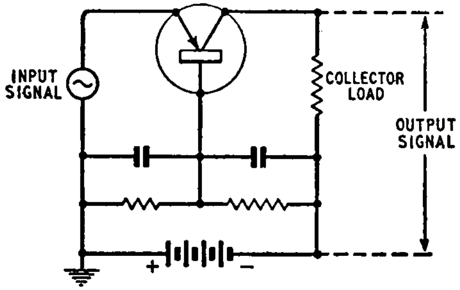


Fig. 2.7. The circuit of Fig. 2.6 using a single battery and a potential divider providing base bias

batteries are connected in series, the positive terminal of one being connected to the negative terminal of the other. Thus a single battery can be used to provide the two bias supplies by connecting it between emitter and collector, the base being returned to a tapping point on the battery or to a potential divider connected across the battery. The potential divider technique (illustrated in Fig. 2.7) is frequently used in transistor circuits and a pnp transistor operating with the emitter circuit earthed requires a negative collector voltage.

Operation of an npn Transistor

The action of an npn junction transistor is similar to that of a pnp type and is illustrated in Fig. 2.8 but the current carriers are largely electrons instead of holes. The emitter-base junction must be forward biased and in an npn transistor this requires the emitter to be biased negatively with respect to the base. This bias causes a considerable current to flow across the emitter-base junction, a current carried by the free electrons of the emitter and the holes of the base. The electrons are required to outnumber the holes and thus, as in a pnp transistor, the emitter region must have greater conductivity than the base region: this is ensured during manufacture. The base-collector junction is reverse-biased: in an npn transistor this requires the collector to be biased positively with respect to the base. Electrons entering the very thin base region from the emitter cross the base region by diffusion and on reaching the base-collector junction are swept into the collector region to give a collector current which is commonly about 95 per cent of the emitter current.

Thus the behaviour of an npn transistor is very similar to that of a pnp type with the bias polarities and directions of current flow reversed: in particular the collector supply for an npn transistor with the emitter circuit earthed must be positive.

High-frequency Performance of a Transistor

In the alloy transistor described above there is a uniform impurity concentration throughout the base region and the only changes in impurity which occur in the device do so very abruptly in the junctions separating the base from the emitter and from the collector. As a result of the uniform impurity concentration, the base region is field-free and the current carriers (holes in a pnp transistor and electrons in an npn type) cross this region to enter the collector region only by diffusion. Diffusion is a slow process and there is a

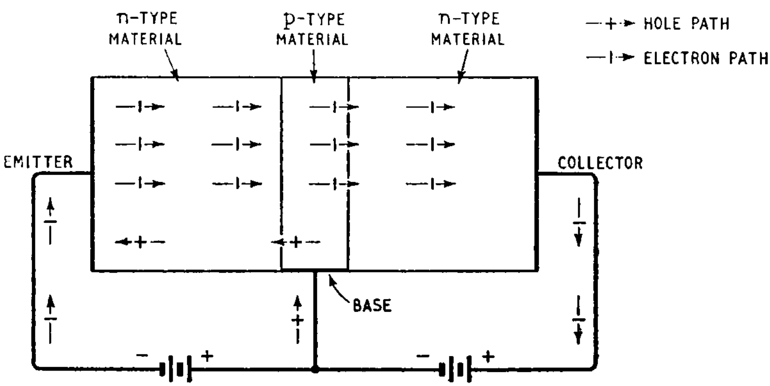


Fig. 2.8. Electron and hole paths in an npn junction transistor connected for amplification

significant time delay between the application of an input signal to the emitter and the receipt of the corresponding output signal from the collector. The effect of this on the transistor performance is discussed more fully later in this chapter but it is similar to the effect of electron transit time on the performance of a thermionic valve: that is to say it degrades the performance of the transistor as signal frequency is increased and in fact sets an upper limit to the frequency at which the transistor can be used satisfactorily. To obtain a good performance at high frequencies the time taken by the current carriers to cross the base must be reduced to a minimum. One method commonly employed to reduce transit time in a transistor is to reduce the thickness of the base region but this may limit the maximum permitted collector voltage to a few volts. However, it is scarcely practicable to raise the uppermost working frequency of a uniform-base transistor much above 20 Mc/s by this method alone.

Drift Transistor

Another method of reducing transit time in a transistor is to vary the impurity concentration through the base region so as to produce an electric field which aids the passage of current carriers across the base region. To give this effect, the impurity concentration must be a maximum near the emitter junction and a minimum near the collector junction and an exponential distribution is a suitable one to use.

Transistors with such a base region are known as drift transistors (because the motion of the current carriers from emitter to collector is mainly a drift imposed by the field in the base region) and have a much reduced transit time, giving a high-frequency performance much superior to that of a uniform-base type.

The technique of solid-state diffusion may be employed to produce the graded base region for drift transistors. One method is to expose the semiconductor material to a vapour of the desired impurity in a furnace. This causes impurity atoms to diffuse into the crystal structure to give an impurity concentration which falls off as the depth of penetration increases. By this means it is possible to produce a base region of accurately-controlled thickness and of graded impurity content. Drift transistors can be made using wafers of such material by alloying pellets to the two faces of the wafer as described at the beginning of this chapter. Theoretically the uppermost working frequency of a transistor can be improved by up to 8 times by grading the impurity concentration in the base region, and in practice, drift transistors of this type operate satisfactorily up to 50 Mc/s. With special care and attention in manufacture such transistors can be used at frequencies up to 150 Mc/s, making possible the construction of fully-transistorised v.h.f. equipment such as f.m. receivers.

Mesa Transistor

A form of construction often adopted for drift transistors is the so-called Mesa form illustrated in Fig. 2.9. This has the advantage of giving low collector-base capacitance and good collector dissipation. The collector region is a block of, say, p-type semiconductor material large enough to permit collector dissipation up to 0·5W,

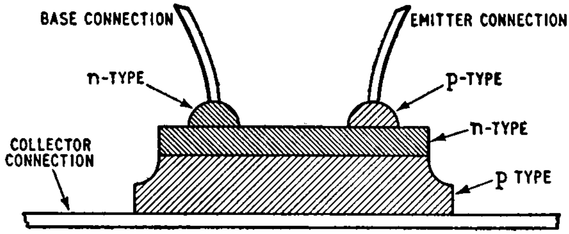


Fig. 2.9. Construction of a Mesa transistor

even at frequencies of hundreds of Mc/s. An n-type base region is formed on the upper surface of this block by diffusion and two pellets of semiconductor material are alloyed to the n-region; one pellet is of n-type material to form the base connection and the other is of p-type material to form the emitter connection. Mesa transistors of this type have operated satisfactorily at frequencies up to 1,000 Mc/s.

CURRENT AMPLIFICATION FACTOR

The ratio of a small change in collector current i_c to the small change in emitter current i_e which gives rise to it is known as the *current amplification factor* and is represented by α. It is measured with a short-circuited output. Thus we have

$$\alpha = \frac{i_c}{i_e}$$

for a junction transistor α cannot exceed unity and is usually between 0·95 and 0·99 corresponding to the 95% or greater capture of emitter current by the collector mentioned on p. 23.

The current amplification factor for a transistor can be compared with the voltage amplification factor μ for a thermionic valve, which can be defined as the ratio of the voltage acting in the anode circuit to the voltage applied to the grid which gives rise to it. It is characteristic of a valve that the voltage in the output circuit is directly related to the input voltage: for this reason valves are referred to as voltage-operated devices. Similarly it is characteristic of transistors that the current flowing in the output circuit is directly related to the input current: for this reason transistors are referred to as current-operated devices. This does not mean, of course, that transistors cannot be used as voltage amplifiers: they are, in fact, often so used and in the next section we shall calculate the voltage gain obtainable from transistor voltage amplifiers.

VOLTAGE GAIN OF A TRANSISTOR AMPLIFIER

The voltage gain obtainable from a number of commonly-employed transistor circuits is calculated in detail in Chapters 3 to 5. However, some simple introductory calculations are given at this point.

These apply to the simple circuit of Fig. 2.7: this is an example of the common-base type of amplifier described in Chapter 3. We know from the fundamental nature of transistors that the alternating component of the collector current is α times the alternating component of the emitter current. As a current amplifier, therefore, the gain has an upper limiting value of 1 for a junction transistor.

An approximate estimate of the voltage gain can be obtained by assuming the current gain to be unity, that is, by assuming that the alternating components of the emitter and collector currents are equal. In effect, therefore, the signal current passes straight through the transistor without significant change in value. Thus a signal voltage applied to the low-value input resistance gives an input current (readily calculated from Ohm's law) which appears at the output terminals as though it had originated from a high-resistance source. The high-resistance source is important for it means that high-value load resistors may be used without significantly affecting the output current: in this way high voltage gains may be achieved.

For example suppose the input resistance is 50 ohms and the load resistance 4·7 kΩ. An input voltage of 1 mV gives rise to an input current i_e where

$$i_e = \frac{1 \times 10^{-3}}{50}\text{ A}$$

$$= 20\ \mu\text{A}$$

If the current amplification factor is taken as unity this is also the collector current i_c. The output voltage V_{out} is given by the product of output current and load resistance.

$$\begin{aligned} V_{out} &= i_c R_l \\ &= 20 \times 10^{-6} \times 4{,}700 \text{ volts} \\ &= 94 \text{ mV} \end{aligned}$$

The voltage gain is thus 94, equal to the ratio of the load resistance to the input resistance.

A more accurate though still simple calculation of the voltage gain of a transistor amplifier can be performed as follows. The relationship between input voltage V_{in} and emitter current i_e is given by

$$V_{in} = i_e r_i$$

where r_i = input resistance of the transistor.

The relationship between output voltage V_{out} and collector current is a little more complicated because the effective resistance of this circuit is composed of the output resistance and the collector load resistance in parallel. Thus we have

$$V_{out} = \frac{R_l r_o}{R_l + r_o} \cdot i_c$$

The voltage gain of the amplifier is thus given by

$$\begin{aligned} \frac{V_{out}}{V_{in}} &= \frac{R_l r_o}{r_i(R_l + r_o)} \cdot \frac{i_c}{i_e} \\ &= \alpha \cdot \frac{R_l r_o}{r_i(R_l + r_o)} \end{aligned}$$

Usually transistor amplifiers are operated from low-voltage supplies such as 6 or 12 volts and the collector current (steady component) is often of the order of 1 mA. Thus the collector load resistance cannot exceed a few thousands of ohms and is commonly 4·7 kΩ. This is small compared with r_o for junction transistors and thus the above expression for the voltage gain can be simplified to

$$\frac{V_{out}}{V_{in}} = \alpha \cdot \frac{R_l}{r_i}$$

showing the gain to be directly proportional to the current amplification factor and the collector load resistance but inversely proportional to the input resistance. As a numerical example, suppose a junction transistor has $\alpha = 0 \cdot 98$, $r_i = 50$ ohms and $R_l = 4 \cdot 7$ kΩ. The voltage gain is given by

$$\frac{V_{out}}{V_{in}} = \alpha \cdot \frac{R_l}{r_i}$$

$$= 0 \cdot 98 \times \frac{4{,}700}{50}$$

$$= 92 \cdot 12$$

These calculations were made on the assumption that the effective collector load resistance is 4·7 kΩ. If a number of transistors are connected in cascade as in a multi-stage amplifier, even though the collector circuits may include a 4·7 kΩ resistor, the effective load resistance for any one transistor must be less than the input resistance of the following transistor. The stage gain for such a load value is limited to α, as shown by putting R_l equal to r_i in the above expression for the voltage gain.

It is, however, possible to obtain voltage gains much higher than this in cascaded amplifiers and some of the circuits which can be used for this purpose are described in Chapters 3 to 5.

Provided the load resistance for a transistor is small compared with the output resistance, the voltage gain is proportional to the load resistance. Thus very high voltage gains can be obtained by using high-value collector load resistances: such loads, however, necessitate high values of h.t. supply and moreover should not be shunted by any circuit which reduces the effective load resistance unduly.

Such conditions exist in a valve amplifier which uses a transistor as a pre-amplifying stage. If the amplifier has a 250 volt supply the transistor can have a collector load of 250 kΩ which allows a collector current of nearly 1 mA. If $\alpha = 0 \cdot 98$ and $r_i = 50$ ohms the voltage gain available is given by

$$\frac{V_{out}}{V_{in}} = \alpha \frac{R_l}{r_i}$$

$$= 0 \cdot 98 \times \frac{250{,}000}{50}$$

$$= 4{,}900$$

This is a surprisingly large gain, which prompts enquiry into the theoretical maximum. This can easily be calculated from the general expression

$$\frac{V_{out}}{V_{in}} = \alpha \cdot \frac{R_l r_o}{r_i(R_l + r_o)}$$

This expression approaches a maximum value as R_l approaches infinity and the maximum is given by

$$\frac{V_{out}}{V_{in}} = \alpha \cdot \frac{r_o}{r_i}$$

If $\alpha = 0{\cdot}98$, $r_o = 1$ megohm and $r_i = 50$ ohms, the maximum voltage gain is given by

$$\frac{V_{out}}{V_{in}} = 0{\cdot}98 \times \frac{1{,}000{,}000}{50}$$
$$= 19{,}600$$

Thus the theoretical maximum voltage gain from a junction transistor is very high indeed but it is largely of theoretical interest because it is possible to realise only a small fraction of it with low values of collector supply voltage.

POWER GAIN OF A TRANSISTOR

The signal input power applied to a transistor amplifier is given by

$$P_{in} = \frac{V_{in}^2}{r_i}$$

The power delivered by the transistor to the external collector load resistance is given by

$$P_{out} = \frac{V_{out}^2}{R_l}$$

The power gain of the amplifier is thus given by

$$\frac{P_{out}}{P_{in}} = \frac{V_{out}^2}{V_{in}^2} \cdot \frac{r_i}{R_l}$$

Substituting for V_{out}/V_{in} from the general expression for the voltage gain given above we have, for the common-base amplifier

$$\frac{P_{out}}{P_{in}} = \alpha^2 \cdot \frac{r_o^2 R_l}{r_i(r_o + R_l)^2}$$

If, as is usually true in practice, R_l is small compared with r_o, this simplifies to

$$\frac{P_{out}}{P_{in}} = \alpha^2 \cdot \frac{R_l}{r_i}$$

$$= \alpha \cdot \frac{V_{out}}{V_{in}}$$

a result which could have been deduced more simply from first principles. For if the voltage gain of an amplifier is given by V_{out}/V_{in} and if the current gain is given by α, the power gain must be the product of these two, namely $\alpha V_{out}/V_{in}$. This more rigorous deduction was, however, deliberately introduced because it reminds us of an essential qualification of this result, namely that it applies only when the collector load resistance is small compared with the amplifier output resistance. For such small loads, the power gain is α times the voltage gain. For a junction transistor the power gain is slightly less than the voltage gain and, to a first degree of approximation, may be taken as equal to it.

VARIATION OF CURRENT AMPLIFICATION FACTOR WITH FREQUENCY

It was pointed out in the earlier part of this chapter that electrons and holes subjected to electric fields move with a finite velocity in semiconductors. They therefore take an appreciable time to travel between the emitter and collector of a junction transistor and there is a significant time delay between the application of an input signal to the emitter and the receipt of the corresponding output signal from the collector. If this time were constant for all electrons or all holes travelling through the transistor, the output signal would be delayed but not distorted and the effect of the finite velocity of the current carriers would be unimportant in many applications. Unfortunately all the current carriers do not have the same velocity: there is an average velocity and most of the current carriers have a velocity which does not greatly differ from this average value. A small number of current carriers, however, have velocities markedly lower or higher than the average value. The distribution of velocities, in fact, follows the curve illustrated in Fig. 2.10 which is well known to statisticians as indicating so-called Gaussian distribution. The horizontal axis gives electron (or hole) velocities, and the ordinate at any particular value of

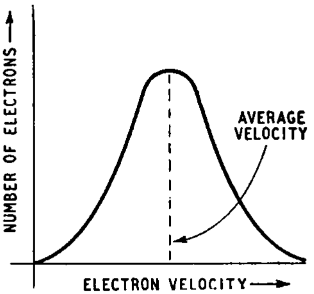
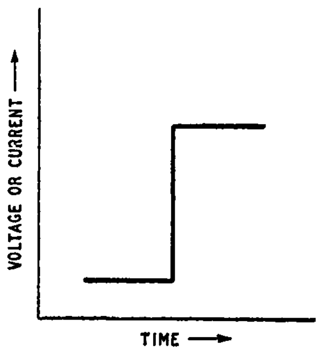


Fig. 2.10. Distribution of electron velocities about the average value

velocity indicates the likelihood of an electron (or hole) having that velocity. The curve naturally shows a maximum at the average value of velocity.

The variation in velocity about the mean value is analogous to dispersion and it causes distortion of the output signal. The precise form of the distortion is best illustrated by considering the amplification of a steep-sided signal such as the leading edge of a rectangular pulse illustrated in Fig. 2.11. Such a signal is usually

Fig. 2.11. A step signal



referred to as a *step signal* and its most important property is the very small time it takes to rise to its maximum value. This is measured by the *rise time* of the step signal, defined as the time taken for the voltage to rise from 10 per cent to 90 per cent of the final steady value. When a step signal is applied to the emitter circuit of a junction transistor it starts a flow of electrons (or holes) to the collector. The most rapid of the current carriers arrive at the collector very quickly, the majority arrive a little later, and the slowest arrive later still. Thus the output signal has the form shown in Fig. 2.12. The most significant feature of this waveform, apart from the fact that it is delayed with respect to the input

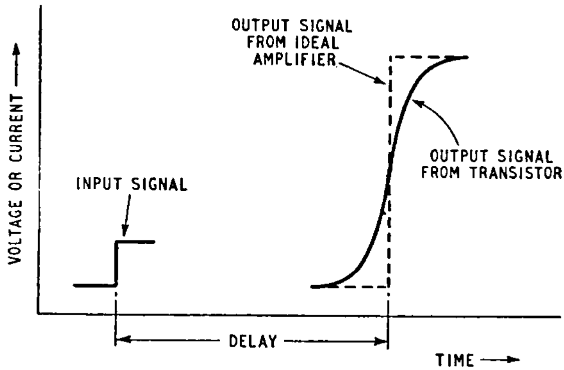


Fig. 2.12. Form of distortion due to variation in current-carrier velocity in a junction transistor

signal, is that it is not so steep as the input signal: in other words, the rise time has been increased by amplification in the transistor.

Fourier analysis of a single step signal shows that it has a continuous spectrum extending from zero frequency to an upper limit which depends on the steepness of the step. The steeper the step the more extended is the high-frequency response. In fact there is a simple but approximate relationship between the rise time of a step signal and the upper frequency limit of its spectrum. The relationship is

$$f_{max} = \frac{1}{2 \times \text{rise time}}$$

Thus an increase in rise time, such as occurs as a result of variation in current-carrier velocity, means that the output signal does not have such an extensive high-frequency response content as the input signal. A high-frequency loss has occurred in the transistor: the current amplification factor falls as frequency increases.

Although variation in current-carrier velocity is one of the chief causes of high-frequency loss in a junction transistor, it is certainly not the only one. Losses also occur as a result of the internal capacitance in the transistor which inevitably shunts the collector and emitter circuits. Further information on losses due to this source is given in Chapter 8.

ALPHA CUT-OFF FREQUENCY

For an alloy junction transistor with a uniform base region the rate of fall of current amplification factor (and the associated phase shift) are approximately expressed by the equation

$$\alpha = \frac{\alpha_0}{1 + jf/f_\alpha}$$

where α = current amplification factor at a frequency f
α_0 = current amplification factor at low frequencies
and f_α = a frequency characteristic of the transistor.

From this expression the numerical value of the current amplification factor is given by

$$\alpha = \frac{\alpha_0}{\sqrt{(1 + f^2/f_\alpha{}^2)}}$$

and at the frequency f_α the numerical value is given by

$$\alpha = \frac{\alpha_0}{\sqrt{2}}$$

i.e. f_α is the frequency at which the current amplification factor has fallen to $1/\sqrt{2}$ i.e. 0·707 of the low-frequency value: this corresponds to a loss of 3 dB. The alpha cut-off frequency is thus defined as the frequency at which the current amplification factor of the common-base amplifier (with short-circuited output) has fallen by 3 dB.

The value of the alpha cut-off frequency depends on the area of the emitter and collector junctions and on the thickness of the base region: in practice values for uniform-base alloy junction transistors lie between 10 kc/s for transistors intended for outputs of several watts to 20 Mc/s for transistors intended for r.f. applications.

FREQUENCY f_1

The expressions given in the above section represent the gain and phase shift in uniform-base alloy junction transistors with reasonable accuracy, but considerable errors can occur if the expressions are used for transistors containing a drift field in the base region. For such transistors it is better to express the high-frequency performance in terms of another frequency parameter f_1 which is defined as the frequency at which the real part of the current amplication factor has fallen to one half the low-frequency value.

For a non-drift transistor the current amplification factor has a value of $\alpha_0/\sqrt{2}$ at f_α and the phase angle is 45°: the real part of α is thus equal to $(\alpha_0 \cos 45)/\sqrt{2}$, i.e. $\alpha_0/2$. For such a transistor, therefore, f_α and f_1 are equal but for drift transistors, f_α can be as much as twice f_1 depending on the magnitude of the drift field.

The frequency f_1 gives a more useful indication of high-frequency performance than f_α. For example, f_1 is approximately equal to the frequency at which the current gain of a common-emitter amplifier (with short-circuited output) has fallen to unity. It thus measures the highest frequency at which the transistor can be used as an amplifier. It also measures the gain-bandwidth product of the transistor when used as a common-emitter amplifier.

3

COMMON-BASE AMPLIFIERS

Introduction

IN the two previous chapters we have discussed the mechanism of operation of junction transistors. We have shown that they can be used to amplify signals and have calculated the voltage gain and power gain of which they are capable. These quantities were calculated in terms of the emitter-base input resistance, the collector-base output resistance and the current amplification factor. However, in designing practical transistor amplifiers and in calculating the gain likely to be achieved, we must know the electrical characteristics of the transistor in greater detail.

In this chapter we shall discuss the parameters of a transistor and will derive an equivalent circuit which can be used to calculate the input and output resistances and the gain of the various types of transistor amplifier.

COMMON-BASE, COMMON-EMITTER AND COMMON-COLLECTOR AMPLIFIERS

A transistor has three terminals, one connected to the emitter, another to the base and the third to the collector. When the transistor is used as an amplifier, the input signal is applied between one pair of terminals and the output signal is taken from another pair.

It follows that one terminal must be common to the input and output circuits.

For example in the transistor amplifiers discussed in the previous chapter it was assumed that the input signal is applied between emitter and base and that the output signal is generated between collector and base. A simplified circuit illustrating a pnp junction transistor used in such an amplifier is given in Fig. 3.1 (a). The base terminal is common to input and output circuits and such amplifiers are referred to as *common-base* or less correctly *earthed-base* amplifiers.

The common-base amplifier is, however, not the only possible arrangement. Alternatively, with the input signal still applied between the emitter and the base, the output signal can be taken between collector and emitter as shown in Fig. 3.1 (b). This illustrates the *common-emitter* or *earthed-emitter* circuit, probably the most widely used of all transistor amplifier circuits.

A third possible arrangement is that illustrated in Fig. 3.1 (c) which shows the input signal applied between base and collector

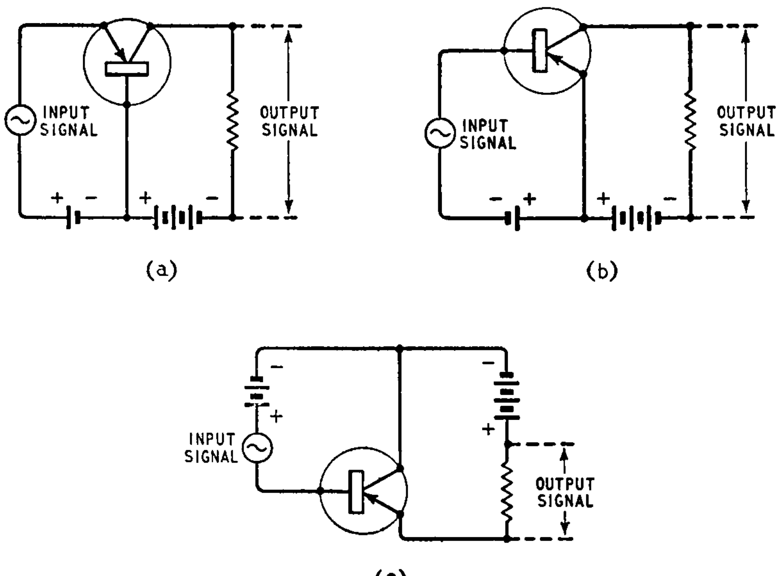


Fig. 3.1. The three basic forms of transistor amplifier: (a) common-base, (b) common-emitter, and (c) common-collector

and the output signal generated between collector and emitter. This is the *common-collector* or *earthed-collector* circuit.

EQUIVALENT CIRCUIT OF A TRANSISTOR

The calculations on the common-base amplifier in the previous chapter were carried out in terms of the emitter-base input resistance r_i, the collector-base output resistance r_o, and the current amplification factor α. The input and output resistance values and the current amplification factor assumed for these calculations do not apply to the common-emitter or the common-collector amplifiers and to perform calculations on these amplifier types we need to know their values of input resistance, output resistance and current amplification factor.

Alternatively, the performance of any kind of transistor circuit, not only the three basic amplifier circuits but other circuits such as those of oscillators and trigger circuits, can be carried out by first deriving the constants of a network which has a behaviour similar to that of the transistor. By applying Kirchhoff's laws or other network theorems to this equivalent circuit we can calculate the performance of the circuit.

A number of equivalent networks for a transistor have been suggested and several different parameters for expressing transistor properties have been developed including the so-called *h* or hybrid parameters and the *r* parameters (which include the input and output resistances used earlier) but in this book we shall represent the transistor as a T-network of resistances as shown in Fig. 3.2.

The transistor cannot, however, be perfectly represented by three resistances only because such a network cannot generate power (as a transistor can) but can only dissipate power. In other words the network illustrated in Fig. 3.2 is a *passive* network and to be accurate the equivalent network must include a source of power, i.e. must be *active*.

The source of power could be shown as a constant-current source connected in parallel with the collector resistance r_c as in Fig. 3.3 (a)

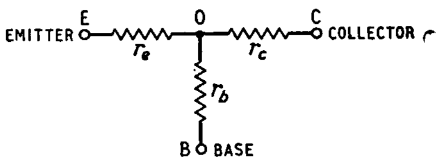


Fig. 3.2. A three-terminal passive network which can be used to build an equivalent circuit for a transistor

and the current so supplied is, as we have already seen, equal to αi_e, where α is the current amplification factor of the transistor and i_e is the current in the emitter circuit, i.e. in the emitter resistance r_e. Alternatively—and this is more convenient for calculation purposes—the source of power can be represented as a constant-voltage generator connected in series with r_c as shown in Fig. 3.3 (b).

Such a voltage has precisely the same effect as the constant-current generator, provided the voltage is given the correct value, and the value required is equal to $\alpha i_e r_c$ as can be shown by applying Thevenin's theorem to Fig. 3.3 (a).

In some books on transistors the voltage generator is given as $r_m i_e$, where r_m is known as the mutual or transfer resistance*

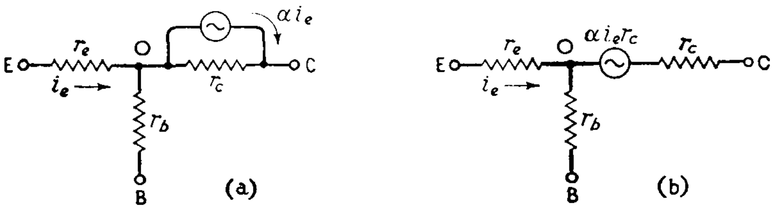


Fig. 3.3. A three-terminal active network which can be used as an equivalent circuit for a transistor (a) including a constant-current generator, and (b) a constant-voltage generator

and is equal to αr_c. The mutual resistance may be defined as the ratio of the e.m.f. in the collector circuit to the signal current in the emitter circuit which gives rise to it. This may be regarded as the dual of the mutual conductance g_m of a thermionic valve which is defined as the ratio of the current in the anode circuit to the signal voltage in the grid circuit which causes it.

This comparison is useful because it again reminds us that the transistor is a device operated by an input current rather than an input voltage.

This network is quite satisfactory for calculating the performance of transistor amplifiers at low frequencies such as audio amplifiers because the reactances of the internal capacitances have in general negligible effects on performance.

At higher frequencies, however, and in particular at radio frequencies, it is necessary to include such capacitances in the T-network to obtain accurate answers.

Values of r_e, r_b, and r_c for Junction Transistors

The values of r_e, r_b and r_c, the three resistances constituting the equivalent network of a transistor, cannot be measured directly because it is impossible to obtain a connection to the centre point O

* The word " transistor " is, in fact, derived from " transfer resistance ".

(Fig. 3.3), but their values can be deduced from measurements made at the terminals. Typical values for these resistances and also for the current amplification factor for junction transistors are as follows:

emitter resistance r_e	25 ohms
base resistance r_b	300 ohms
collector resistance r_c	1 megohm
current amplification factor α	0·98

These resistance values are those which apply when the transistor is handling a small alternating signal. They are differential or a.c. quantities and depend on the d.c. conditions in the transistor. If the steady emitter bias current or the collector bias current are changed, the values of r_e, r_b and r_c will, in general, also change.

COMMON-BASE AMPLIFIER

A simplified circuit of a transistor common-base amplifier is given in Fig. 3.4 (a) in which R_s represents the internal resistance of the signal source and R_l represents the collector load for the

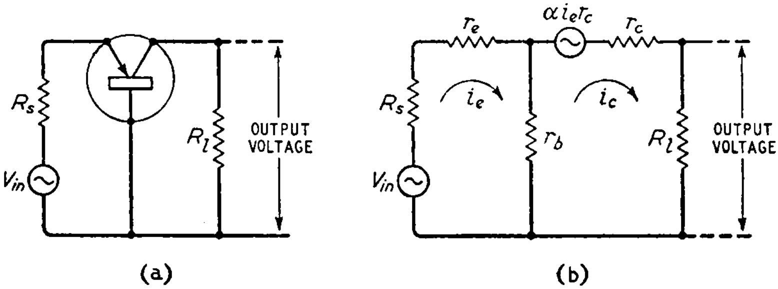


Fig. 3.4. Simplified circuit of a common-base amplifier (a), and its electrical equivalent (b)

amplifier. For the sake of simplicity, methods of applying bias are not indicated.

Fig. 3.4 (b) represents the same circuit in which the transistor is represented by its equivalent T-section network. The directions of the currents in the two meshes of this network can be determined by considering the physical processes occurring in a transistor.

From Chapter 2 we know that the emitter current i_e is equal to the sum of the collector current i_c and the base current i_b. If, therefore, i_e and i_c are given clockwise directions as shown in Fig. 3.4 (b), then the current in r_b is $(i_e - i_c)$ which is equal to i_b: all conventions are therefore satisfied in this diagram. By applying Kirchhoff's laws to this circuit we can calculate the performance of the amplifier. The features of the amplifier in which we are principally interested are the following:

(a) input resistance
(b) output resistance
(c) voltage gain.

Input Resistance

Applying Kirchhoff's laws to the circuit of Fig. 3.4 (b) we have

$$V_{in} = i_e(R_s + r_e + r_b) - i_c r_b \qquad \ldots.(1)$$

$$0 = i_c(r_b + r_c + R_l) - i_e r_b - \alpha i_e r_c \qquad \ldots.(2)$$

To obtain an expression for the input resistance we can eliminate i_c between these two equations to obtain a relationship between i_e and V_{in}.

From (2)

$$i_c = \frac{r_b + \alpha r_c}{r_b + r_c + R_l}\, i_e$$

Substituting for i_c in (1)

$$V_{in} = i_e(R_s + r_e + r_b) - \frac{r_b(r_b + \alpha r_c)}{r_b + r_c + R_l}\, i_e$$

$$\therefore\ i_e = \frac{V_{in}}{R_s + r_e + r_b - \dfrac{r_b(r_b + \alpha r_c)}{r_b + r_c + R_l}} \qquad \ldots.(3)$$

If we represent the input resistance of the transistor by r_i the circuit of Fig. 3.4 (b) takes the form shown in Fig. 3.5. For this circuit we have

$$i_1 = \frac{V_{in}}{R_s + r_i} \qquad \ldots.(4)$$

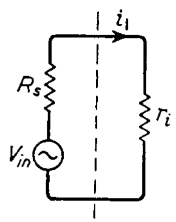
Comparing expressions (3) and (4)

$$r_i = r_e + r_b - \frac{r_b(r_b + \alpha r_c)}{r_b + r_c + R_l} \qquad \ldots.(5)$$

$$= r_e + r_b \cdot \frac{R_l + r_c(1 - \alpha)}{r_b + r_c + R_l} \qquad \ldots.(6)$$

Expressions (5) and (6) give the value of the input resistance in terms of the emitter resistance, base resistance, collector resistance, load resistance and current amplification factor. One respect in which the transistor differs from the valve is that the input resistance depends on the load resistance, this bearing out the internal

Fig. 3.5. Circuit illustrating the meaning of input resistance



feedback of the transistor mentioned earlier. For a given transistor operating under constant d.c. conditions, r_b, r_e, r_c and α are constant, and r_i therefore depends solely on R_l, increasing as R_l increases.

The range of input resistance can be calculated from expression (5). First consider expression (5) when R_l is made vanishingly small

$$r_i = r_e + r_b - \frac{r_b(r_b + \alpha r_c)}{r_b + r_c}$$

Now for junction transistors r_c and αr_c are both large compared with r_b and the input resistance is approximately given by

$$r_i = r_e + r_b - r_b \cdot \frac{\alpha r_c}{r_c}$$

$$= r_e + r_b(1 - \alpha) \qquad \ldots.(7)$$

and this gives the value of the input resistance of the common-base amplifier when the output terminals are short-circuited.

Now consider the input resistance when R_l is made very large. The final term in expression (5) vanishes leaving

$$r_i = r_e + r_b \qquad \ldots.(8)$$

which gives the value of the input resistance when the output terminals are open-circuited.

Input Resistance for a Junction Transistor as Common-base Amplifier

To obtain an estimate of practical values of input resistance for a junction transistor we can substitute $r_e = 25$ ohms, $r_b = 300$ ohms and $\alpha = 0{\cdot}98$ and from (7) we find that the input resistance for short-circuited output terminals is given by

$$\begin{aligned} r_i &= 25 + 300\,(1 - 0{\cdot}98) \\ &= 25 + 6 \\ &= 31 \text{ ohms} \end{aligned}$$

From (8) the input resistance for open-circuited output terminals is given by

$$\begin{aligned} r_i &= 25 + 300 \\ &= 325 \text{ ohms} \end{aligned}$$

Thus the input resistance of a junction transistor used as a common-base amplifier varies between a minimum value for

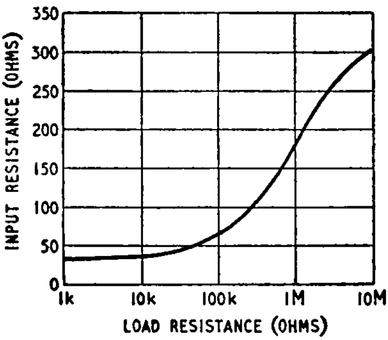


Fig. 3.6. Variation of input resistance with load resistance for a common-base amplifier

short-circuited output terminals to a maximum value for open-circuited terminals. The dependence of input resistance on output load is illustrated in Fig. 3.6.

Output Resistance

To obtain an expression for the output resistance of the common-base amplifier we can eliminate i_e between equations (1) and (2) to obtain a relationship between i_c and V_{in}.

From (2)

$$i_e = \frac{r_b + r_c + R_l}{r_b + \alpha r_c} \cdot i_c$$

Substituting for i_e in (1)

$$V_{in} = \frac{(r_b + r_c + R_l)\,(R_s + r_e + r_b)}{r_b + \alpha r_c}\, i_c - i_c r_b$$

$$\therefore\ i_c = \frac{V_{in}}{\dfrac{(r_b + r_c + R_l)\,(R_s + r_e + r_b)}{r_b + \alpha r_c} - r_b}$$

$$= \frac{(r_b + \alpha r_c)\,V_{in}}{(r_b + r_c + R_l)\,(R_s + r_e + r_b) - r_b(r_b + \alpha r_c)}$$

$$\therefore\ i_c = \frac{\dfrac{r_b + \alpha r_c}{R_s + r_e + r_b}\, V_{in}}{R_l + r_b + r_c - \dfrac{r_b\,(r_b + \alpha r_c)}{R_s + r_e + r_b}} \qquad \ldots.(9)$$

If we represent the output resistance of the transistor amplifier by r_o the circuit of Fig. 3.4 (b) has the form shown in Fig. 3.7. For this circuit we have

$$i_2 = \frac{V}{R_l + r_o} \qquad \ldots.(10)$$

Comparing expressions (9) and (10) we have that V, the e.m.f. effectively acting in the output circuit, is given by

$$V = \frac{r_b + \alpha r_c}{R_s + r_e + r_b}\, V_{in}$$

and the output resistance is given by

$$r_o = r_b + r_c - \frac{r_b\,(r_b + \alpha r_c)}{R_s + r_e + r_b} \qquad \ldots.(11)$$

$$= r_c + r_b \cdot \frac{R_s + r_e - r_c\alpha}{R_s + r_e + r_b} \qquad \ldots.(12)$$

This expression shows that the output resistance depends on the source resistance R_s as well as on the base resistance, emitter resistance, collector resistance and current amplification factor. For a given transistor operating with fixed d.c. conditions, r_b, r_e, r_c and α are fixed and r_o then depends on R_s, increasing as R_s is increased. The range over which r_o varies can be estimated as follows. First let $R_s = 0$. Expression (12) then becomes

$$r_o = r_c + r_b \cdot \frac{r_e - r_c\alpha}{r_e + r_b}$$

which, if the relatively small term $r_b r_e / r_c$ is neglected, gives

$$r_o = r_c \cdot \frac{r_e + r_b\,(1 - \alpha)}{r_e + r_b} \qquad \ldots .(13)$$

Now let R_s approach infinity. The final term in expression (11) then vanishes, leaving the output resistance as

$$r_o = r_c + r_b \qquad \ldots .(14)$$

Output Resistance of a Junction Transistor as Common-base Amplifier

To obtain a numerical estimate of the range of output resistance for a junction transistor let $r_e = 25$ ohms, $r_b = 300$ ohms, $r_c = 1$ megohm and $\alpha = 0{\cdot}98$. Substituting these values in expression (13) we

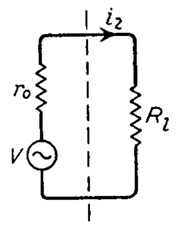


Fig. 3.7. Circuit illustrating the meaning of output resistance

find the output resistance for short-circuited input terminals is given by

$$r_o = r_c \cdot \frac{r_e + r_b\,(1 - \alpha)}{r_e + r_b}$$

$$= 1{,}000{,}000 \times \left[\frac{25 + 300\,(1 - 0{\cdot}98)}{25 + 300}\right] \text{ohms}$$

$$= 1{,}000{,}000 \times \left[\frac{25 + 6}{325}\right] \text{ohms}$$

$$= 100 \text{ k}\Omega \text{ approximately}$$

The output resistance for open-circuited input terminals is, from (14), given by

$$r_o = r_c + r_b$$

$$= 1{,}000{,}000 + 300 \text{ ohms}$$

$$= 1{,}000{,}000 \text{ ohms approximately}$$

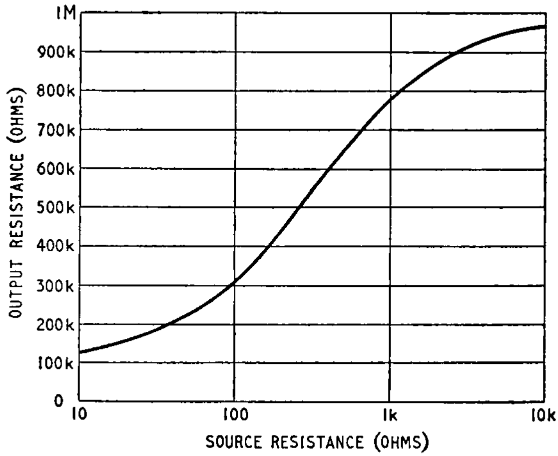


Fig. 3.8. Variation of output resistance with source resistance for a common-base amplifier

The variation of output resistance with source resistance for a common-base amplifier is illustrated in Fig. 3.8.

Voltage Gain

From Fig. 3.4 (b) we can see that the output voltage is given by $i_c R_l$. The voltage gain V_{out}/V_{in} is hence equal to $i_c R_l / V_{in}$. From (9) i_c is given by

$$i_c = \frac{(r_b + \alpha r_c)\, V_{in}}{(R_l + r_b + r_c)\,(R_s + r_e + r_b) - r_b\,(r_b + \alpha r_c)}$$

Hence

$$\frac{V_{out}}{V_{in}} = \frac{i_c R_l}{V_{in}} = \frac{(r_b + \alpha r_c) R_l}{(R_l + r_b + r_c)(R_s + r_e + r_b) - r_b (r_b + \alpha r_c)} \quad \ldots .(15)$$

which shows how the gain is related to R_l, R_s, r_e, r_b, r_c and α. The dependence of the voltage gain on the value of R_l is illustrated in Fig. 3.9, which shows gain increasing with R_l, linearly for small

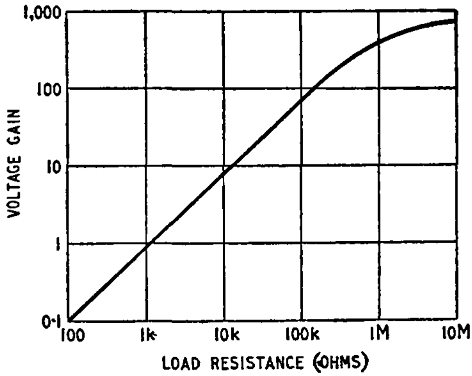


Fig. 3.9. Variation of voltage gain with load resistance for a common-base amplifier

values of R_l but becoming asymptotic to a limiting value of gain as R_l becomes large. The shape of the curve is, in fact, the same as that relating gain to load resistance for a thermionic valve.

For a junction transistor r_c is normally large compared with all other resistances in the expression, and it is possible to simplify expression (15) as follows:

$$\frac{V_{out}}{V_{in}} = \frac{\alpha r_c R_l}{r_c (R_s + r_e + r_b) - \alpha r_b r_c}$$

$$= \frac{\alpha R_l}{R_s + r_e + r_b (1 - \alpha)} \quad \ldots .(16)$$

As a numerical example suppose $\alpha = 0{\cdot}98$, $R_l = 10$ kΩ, $R_s = 1$ kΩ, $r_e = 25$ ohms and $r_b = 300$ ohms

$$\frac{V_{out}}{V_{in}} = \frac{0{\cdot}98 \times 10^4}{1{,}000 + 25 + 300 (1 - 0{\cdot}98)}$$

$$= \frac{9{,}800}{1{,}000 + 25 + 6}$$

$$= \frac{9{,}800}{1{,}031}$$

$$= 9{\cdot}77$$

Substituting r_i for $r_e + r_b(1 - \alpha)$ in expression (16) (see Equation 7) we have for small values of R_l

$$\frac{V_{out}}{V_{in}} = \frac{\alpha R_l}{R_s + r_i}$$

and if $R_s = 0$ we have

$$\frac{V_{out}}{V_{in}} = \alpha \cdot \frac{R_l}{r_i}$$

which confirms the result used in Chapter 2.

To obtain an impression of the limiting value of voltage gain available from a junction transistor, suppose R_l is made large compared with r_c. With a few other simplifications expression (15) now reduces to the form

$$\frac{V_{out}}{V_{in}} = \frac{\alpha\, r_c R_l}{R_l(R_s + r_e + r_b)}$$

$$= \frac{\alpha\, r_c}{R_s + r_e + r_b} \qquad \ldots.(17)$$

Substituting $\alpha = 0{\cdot}98$, $r_c = 1$ MΩ, $R_s = 1$ kΩ, $r_l = 25$ ohms and $r_b = 300$ ohms we have that the maximum stage gain is given by

$$\frac{V_{out}}{V_{in}} = \frac{0{\cdot}98 \times 10^6}{1{,}000 + 25 + 300}$$

$$= 740$$

Expression (17) shows the importance of the source resistance R_s in such determinations. If this could be reduced to negligible proportions relative to the values of r_e and r_b, the stage gain rises to

$$\frac{V_{out}}{V_{in}} = \frac{0{\cdot}98 \times 10^6}{325}$$

$$= 3{,}000$$

PRACTICAL CIRCUIT FOR A COMMON-BASE AMPLIFIER

Fig. 3.10 gives the circuit of a small-signal amplifier using a pnp junction transistor as a common-base amplifier. The base is earthed and two batteries are employed, one of 6 volts to bias the collector negatively and the other of 1·5 volts to bias the emitter positively.

Approximate values for the components used may be calculated in the following way. A suitable value for the collector voltage

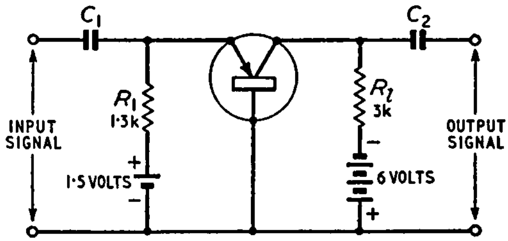


Fig. 3.10. Circuit for small-signal common-base transistor amplifier

in the absence of an input signal is −3 volts, for this permits excursions of collector voltage during amplification up to 3 volts in peak value. Under quiescent conditions there is a 3 volt drop in voltage across the collector load resistor. A suitable value of steady collector current is 1 mA and from Ohm's law the value of the load resistance is given by $3/(1 \times 10^{-3})$, i.e. 3 kΩ.

The steady emitter current in the absence of an input signal will be just over 1 mA but can be taken as 1 mA to give an approximate value for the emitter bias resistor. The emitter bias voltage is 1·5 and from Ohm's law the resistance of the emitter circuit to give a current of 1 mA is $1{\cdot}5/(1 \times 10^{-3})$, i.e., 1·5 kΩ. The d.c. emitter-base input resistance will contribute a few hundred ohms towards this and an emitter bias resistance of 1·3 kΩ will be suitable.

The capacitance C_1 depends on the lowest frequency it is required to amplify. For an audio-frequency amplifier this may be 50 c/s. The a.c. emitter-base input resistance shunts R_1 giving an effective input resistance of say 40 ohms (see Fig. 3.6). If the reactance of the capacitor is made equal to 40 ohms at 50 c/s, there is a loss of 3 dB at this frequency. The capacitance C_1 necessary is given by

$$C_1 = \frac{1}{2\pi f X}$$

where X is the reactance. Substituting $f = 50$ and $X = 40$ we have

$$C_1 = \frac{1}{2 \times 3{\cdot}142 \times 50 \times 40}\ \mathrm{F}$$

$$= 80\ \mu\mathrm{F}$$

In practice the source of input signal will have some resistance and this will permit the use of a lower capacitance for C_1.

The capacitance C_2 depends on the resistance which is to be connected across the output terminals.

COLLECTOR CURRENT – COLLECTOR VOLTAGE CHARACTERISTICS FOR COMMON-BASE CONNECTION

The operation of a circuit such as that of Fig. 3.10 can be illustrated by means of collector current – collector voltage characteristics. These are usually obtained by plotting the collector current against the collector voltage for a fixed value of emitter current and the resulting curves have the form shown in Fig. 3.11. The curves are plotted in terms of input current because the transistor is a

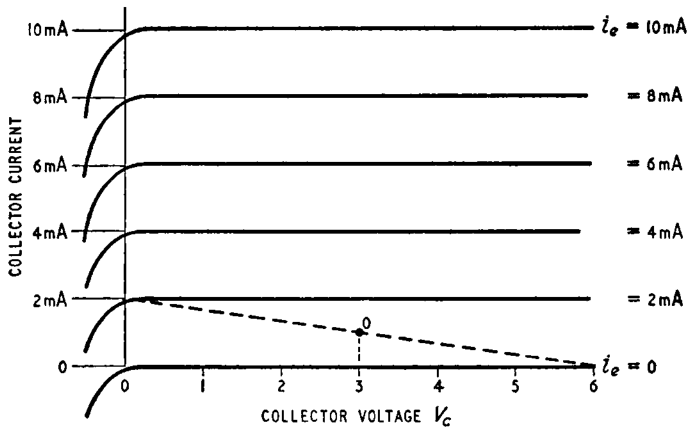


Fig. 3.11. Typical collector current–collector voltage characteristics for a junction transistor

current-operated device, and the curves show that the collector current is always slightly less than the emitter current.

The characteristics have a shape similar to that of the anode current – anode voltage curves for a pentode, the horizontal portions indicating the high collector a.c. resistance of the junction transistor. However, they represent a better approximation to the ideal characteristic than those of a pentode because the transistor characteristics have no knee to limit the swing of collector potential during amplification. The collector voltage can swing the whole extent of the collector supply voltage, giving the transistor an efficiency equal to the theoretical maximum. The load line shown superimposed on the curves illustrates operation of the amplifier of Fig. 3.10. The quiescent point O corresponds to half the collector supply voltage. The position of O relative to the characteristics gives the emitter current: the bias voltage together with the external emitter resistance (R_1 in Fig. 3.10) must be chosen to give this value of current.

Power Gain

The current gain α for a common-base amplifier is nearly equal to unity and the power gain is thus nearly equal to the voltage gain as shown on p. 31. A typical value for the voltage gain is 100 and this is also a typical value for the power gain (20 dB).

COMPARISON BETWEEN COMMON-BASE TRANSISTOR AMPLIFIER AND COMMON-GRID THERMIONIC-VALVE AMPLIFIER

An increase in the positive bias applied to the emitter (Fig. 3.10) gives a larger emitter current and hence a larger collector current. This increases the voltage drop across the collector load resistor. The collector bias is negative and this increased voltage drop makes the collector voltage less negative, i.e. more positive. Thus a positive voltage applied to the emitter results in a magnified positive pulse appearing at the collector: there is hence no phase inversion in a common-base transistor amplifier.

The behaviour of the transistor amplifier is in fact similar to that of a common-grid thermionic-valve amplifier. The input signal is applied to the cathode circuit, which presents a low input resistance, and the output signal is taken from the anode circuit which has a much higher output resistance. The relative resistances are hence of the same order as those of the common-base transistor amplifier. A positive pulse at the cathode of the valve reduces the anode current causing the anode voltage to rise: thus there is no phase inversion in a common-grid amplifier.

4

COMMON-EMITTER AMPLIFIERS

Introduction

IN the previous chapter we examined the common-base amplifier, a circuit arrangement characterised by very low input resistance, very high output resistance and less than unity current gain. Input resistances vary from 30 ohms to 300 ohms; output resistances range from 100 kΩ to 1 megohm. A noteworthy feature of the common-base amplifier is that the output signal is in phase with the input signal.

The majority of transistor amplifiers are, however, of the common-emitter type and we shall now analyse the performance of this

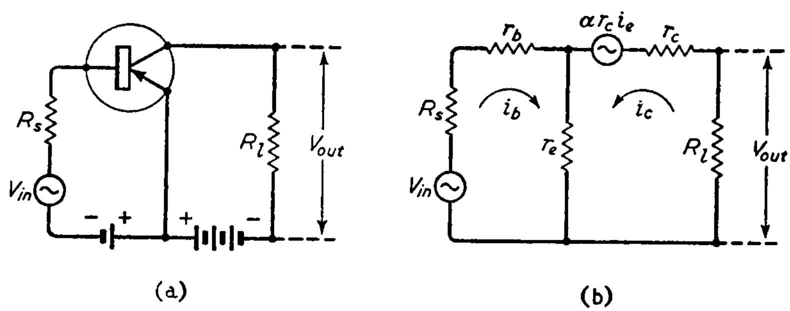


Fig. 4.1. Basic form of common-emitter amplifier (a), and its equivalent circuit (b)

circuit to account for its popularity. As before we shall do this by deriving expressions for the input resistance, output resistance and voltage gain.

The basic circuit for a common-emitter amplifier is given in Fig. 4.1 (a), in which R_s is the resistance of the signal source and R_l is the collector load resistance. Fig. 4.1 (b) gives the equivalent circuit in which the transistor is represented by the T-network of resistances r_e, r_b and r_c introduced in the previous chapter. If the base current i_b is shown acting in a clockwise direction and the collector current i_c in an anticlockwise direction as indicated in Fig. 4.1 (b), then the current in the common emitter resistance r_e is the sum $(i_b + i_c)$. From Chapter 2 we know that the sum of these two currents is the emitter current i_e: all conventions are therefore satisfied in this diagram.

Input Resistance

Applying Kirchhoff's laws to the circuit of Fig. 4.1 (b) we have

$$V_{in} = i_b\,(R_s + r_b + r_e) + i_c r_e \qquad \ldots.(1)$$

$$0 = i_c\,(R_l + r_c + r_e) + i_b r_e - \alpha r_c i_e \qquad \ldots.(2)$$

Now $i_e = i_b + i_c$ and equation (2) may therefore be written in the form

$$\begin{aligned} 0 &= i_c\,(R_l + r_c + r_e) + i_b r_e - \alpha r_c i_b - \alpha r_c i_c \\ &= i_c\,[R_l + r_e + r_c\,(1 - \alpha)] + i_b(r_e - \alpha r_c) \qquad \ldots.(3) \end{aligned}$$

As in the analysis of the common-base amplifier, we can obtain an expression for the input resistance of the amplifier by eliminating i_c between equations (1) and (3) to obtain a relationship between i_b and V_{in}. From (3)

$$i_c = -\frac{r_e - \alpha r_c}{R_l + r_e + (1 - \alpha)\,r_c}\,.\,i_b$$

Substituting for i_c in (1)

$$V_{in} = i_b\,(R_s + r_b + r_e) - \frac{r_e\,(r_e - \alpha r_c)}{R_l + r_e + (1 - \alpha)\,r_c}\,.\,i_b$$

$$\therefore\; i_b = \frac{V_{in}}{R_s + r_b + r_e - \dfrac{r_e\,(r_e - \alpha r_c)}{R_l + r_e + (1 - \alpha)\,r_c}}$$

Now αr_c is large compared with r_e and the fourth term in the denominator is best reversed in sign thus

$$i_b = \frac{V_{in}}{R_s + r_b + r_e + \dfrac{r_e\,(\alpha r_c - r_e)}{R_l + r_e + (1 - \alpha)\,r_c}} \qquad \ldots.(4)$$

If the input resistance of the amplifier is represented by r_i the input current i_1 can be expressed

$$i_1 = \frac{V_{in}}{R_s + r_i} \qquad \ldots.(5)$$

Comparing this with expression (4) we have

$$r_i = r_b + r_e + \frac{r_e\,(\alpha r_c - r_e)}{R_l + r_e + (1 - \alpha)\, r_c} \qquad \ldots.(6)$$

a useful expression because it shows that r_i increases as R_l decreases. Expression (6) can be simplified thus

$$r_i = r_b + \frac{r_e\,[R_l + r_e + (1 - \alpha)\, r_c] + r_e\,(\alpha r_c - r_e)}{R_l + r_e + (1 - \alpha)\, r_c}$$

$$= r_b + r_e \cdot \frac{R_l + r_c}{R_l + r_e + (1 - \alpha)\, r_c}$$

The denominator can be simplified by neglecting r_e in comparison with the other terms which are much larger in practice. We then have the final result

$$r_i = r_b + r_e \cdot \frac{R_l + r_c}{R_l + (1 - \alpha)\, r_c} \qquad \ldots.(7)$$

When R_l is small compared with r_c and $(1 - \alpha)\, r_c$ the input resistance has a value given by

$$r_i = r_b + r_e \cdot \frac{r_c}{r_c\,(1 - \alpha)}$$

$$= r_b + \frac{r_e}{1 - \alpha} \qquad \ldots.(8)$$

When R_l is large compared with r_c and $(1 - \alpha)\, r_c$ the input resistance has a value given by

$$r_i = r_b + r_e \cdot \frac{R_l}{R_l}$$

$$= r_b + r_e \qquad \ldots.(9)$$

Input Resistance of a Junction Transistor as Common-emitter Amplifier

To illustrate the range of values of input resistance likely to be encountered in practice in a common-emitter junction transistor amplifier, let us assume that $r_e = 25$ ohms, $r_b = 300$ ohms and

$\alpha = 0{\cdot}98$. Substituting in expression (8) to find the input resistance for short-circuited output terminals we have

$$r_i = r_b + \frac{r_e}{1 - \alpha}$$

$$= 300 + \frac{25}{1 - 0{\cdot}98} \text{ ohms}$$

$$= 300 + 1{,}250 \text{ ohms}$$

$$= 1{,}550 \text{ ohms}$$

Substituting in expression (9) to obtain the input resistance for open-circuited output terminals we have

$$r_i = r_b + r_e$$

$$= 300 + 25 \text{ ohms}$$

$$= 325 \text{ ohms}$$

These numerical examples show that the common-emitter amplifier has a higher input resistance than the common-base amplifier and varies with change in load resistance in the opposite way.

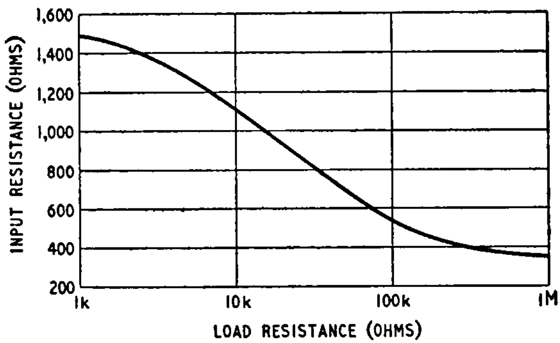


Fig. 4.2. Variation of input resistance with load resistance for a common-emitter amplifier

The variation of input resistance with output collector load for junction transistors is illustrated in Fig. 4.2.

Output Resistance

We can obtain an expression for the output resistance of the common-emitter amplifier by eliminating i_b between equations (1) and (3) to obtain a relationship between V_{in} and i_c. From (3) we have

$$i_b = -\frac{R_l + r_e + r_c(1-\alpha)}{r_e - \alpha r_c} i_c$$

Substituting in (1)

$$V_{in} = -\frac{R_l + r_e + r_c(1-\alpha)}{r_e - \alpha r_c}(R_s + r_b + r_e)\, i_c + i_c r_e$$

$$\therefore\ i_c = \frac{V_{in}}{-\dfrac{[R_l + r_e + r_c(1-\alpha)]\,(R_s + r_b + r_e)}{r_e - \alpha r_c} + r_e}$$

$$= \frac{V_{in}\,(r_e - \alpha r_c)}{-[R_l + r_e + r_c(1' - \alpha)]\,(R_s + r_b + r_e) + r_e\,(r_e - \alpha r_c)}$$

$$= \frac{\dfrac{r_e - \alpha r_c}{R_s + r_b + r_e} \cdot V_{in}}{-[R_l + r_e + r_c(1-\alpha)] + \dfrac{r_e\,(r_e - \alpha r_c)}{R_s + r_b + r_e}}$$

This can be more conveniently written in the form

$$i_c = \frac{\dfrac{\alpha r_c - r_e}{R_s + r_b + r_e} \cdot V_{in}}{R_l + r_e + r_c(1-\alpha) + \dfrac{r_e\,(\alpha r_c - r_e)}{R_s + r_b + r_e}} \qquad \ldots.(10)$$

In a simple circuit containing a signal source of voltage V and an internal resistance r_o feeding a load resistance R_l the current i_2 is given by

$$i_2 = \frac{V}{R_l + r_o} \qquad \ldots.(11)$$

If we compare expressions (10) and (11) we obtain the following expression for the output resistance of the common-emitter amplifier:

$$r_o = r_e + r_c (1 - \alpha) + \frac{r_e (\alpha r_c - r_e)}{R_s + r_b + r_e} \qquad \ldots .(12)$$

This expression shows that the output resistance depends on the source resistance (for a given transistor), decreasing as the source resistance is increased.

This expression can be simplified by combining the first term with the third; this gives the result

$$r_o = r_c (1 - \alpha) + r_e \,.\, \frac{R_s + r_b + \alpha r_c}{R_s + r_b + r_e} \qquad \ldots .(13)$$

To find an expression for the output resistance for short-circuited input terminals, let R_s approach zero in expression (13). We then have

$$r_o = r_c (1 - \alpha) + r_e \,.\, \frac{r_b + \alpha r_c}{r_b + r_e}$$

$$= r_c \,.\, \frac{r_e + r_b (1 - \alpha)}{r_e + r_b} \qquad \ldots .(14)$$

if we neglect the term $r_e r_b / r_c$ which is small compared with the others.

To find an expression for the output resistance for open-circuited input terminals let R_s approach infinity in expression (12). We then have

$$r_o = r_e + r_c (1 - \alpha)$$

and as r_e is small compared with $r_c (1 - \alpha)$ we can say

$$r_o = r_c (1 - \alpha) \qquad \ldots .(15)$$

Output Resistance of a Junction Transistor as Common-emitter Amplifier

We can determine the range of values of output resistance for a common-emitter junction transistor amplifier by substituting the typical practical values $r_e = 25$ ohms, $r_b = 300$ ohms, $r_c = 1$ megohm and $\alpha = 0{\cdot}98$ in expression (14) and (15).

From (14) the output resistance for short-circuited input terminals is given by

$$r_o = r_c \,.\, \frac{r_e + r_b (1 - \alpha)}{r_e + r_b}$$

$$= 1{,}000{,}000\ .\ \frac{25 + 300\ (1 - 0{\cdot}98)}{25 + 300}\ \text{ohms}$$

$$= 1{,}000{,}000\ .\ \frac{25 + 6}{325}\ \text{ohms}$$

$$= 1{,}000{,}000\ .\ \frac{31}{325}\ \text{ohms}$$

$$= 95\ \text{k}\Omega$$

From expression (15) the output resistance for open-circuited input terminals is given by

$$r_o = r_c\ (1 - \alpha)$$

$$= 1{,}000{,}000\ (1 - 0{\cdot}98)\ \text{ohms}$$

$$= 1{,}000{,}000 \times 0{\cdot}02\ \text{ohms}$$

$$= 20\ \text{k}\Omega$$

The variation in output resistance with source resistance for junction transistors is illustrated in Fig. 4.3.

Voltage Gain

From Fig. 4.1 (b) we can see that the output voltage is given by i_cR_l. The voltage gain V_{out}/V_{in} is thus given by i_cR_l/V_{in}. From (10) i_c is given by

$$i_c = \frac{\dfrac{\alpha r_c - r_e}{R_s + r_b + r_e}\ .\ V_{in}}{R_l + r_e + r_c\ (1 - \alpha) + \dfrac{r_e\ (\alpha r_c - r_e)}{R_s + r_b + r_e}}$$

Hence

$$\frac{V_{out}}{V_{in}} = \frac{\dfrac{\alpha r_c - r_e}{R_s + r_b + r_e}\ .\ R_l}{R_l + r_e + r_c\ (1 - \alpha) + \dfrac{r_e\ (\alpha r_c - r_e)}{R_s + r_b + r_e}}$$

$$= \frac{(\alpha r_c - r_e)\ R_l}{[R_l + r_e + r_c\ (1 - \alpha)]\ (R_s + r_b + r_e) + r_e\ (\alpha r_c - r_e)} \qquad \ldots.(16)$$

Normally αr_c greatly exceeds r_e and this expression therefore gives a positive value for the voltage gain. But, to agree with the physics of the transistor, the collector current in Fig. 4.1(b) was assumed to be flowing in an anticlockwise direction whereas the base current was shown flowing in a clockwise direction. A positive value for the voltage gain thus implies that the output voltage is inverted with respect to the input voltage. In this respect the common-emitter amplifier is similar to a common-cathode thermionic valve amplifier. Expression (16) can be simplified by neglecting r_e in comparison with αr_c in numerator and denominator.

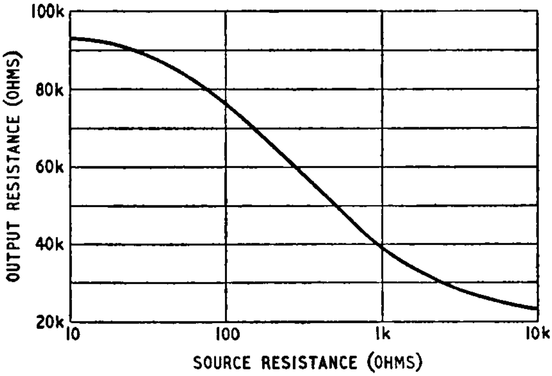


Fig. 4.3. Variation of output resistance with source resistance for a common-emitter amplifier

We then have

$$\frac{V_{out}}{V_{in}} = \frac{\alpha r_c R_l}{[R_l + r_c(1-\alpha)](R_s + r_b + r_e) + \alpha r_e r_c}$$

For values of R_l and r_e small compared with $r_c(1-\alpha)$ expression (16) may be written

$$\frac{V_{out}}{V_{in}} = \frac{\alpha r_c R_l}{r_c(1-\alpha)(R_s + r_b + r_e) + \alpha r_e r_c}$$

$$= \frac{\alpha R_l}{R_s(1-\alpha) + r_b(1-\alpha) + r_e} \qquad \ldots.(17)$$

As a numerical example suppose $\alpha = 0{\cdot}98$, $R_l = 10$ kΩ, $R_s = 1$ kΩ, $r_e = 25$ ohms and $r_b = 300$ ohms. We have

$$\frac{V_{out}}{V_{in}} = \frac{0{\cdot}98 \times 10^4}{1{,}000\,(1 - 0{\cdot}98) + 300\,(1 - 0{\cdot}98) + 25}$$

$$= \frac{0{\cdot}98 \times 10^4}{20 + 6 + 25}$$

$$= \frac{0{\cdot}98 \times 10^4}{51}$$

$$= 192$$

more than 19 times that available from the same transistor, the same source resistance and the same load resistance in a common-base amplifying circuit (see p. 49). However, expression (17) is similar to that for the common-base amplifier (expression 16 of Chapter 3) but contains $R_s\,(1 - \alpha)$ compared with R_s in the earlier formula. Moreover, if R_s is very small or—and this amounts to the same thing—if V_{in} is taken as the voltage between the input terminals of the amplifier, the two expressions for the voltage gain become equal.

Consider now the voltage gain of the common-emitter amplifier for load resistor values which are very large compared with $r_c\,(1 - \alpha)$. If r_e is neglected in comparison with r_c, expression (16) becomes

$$\frac{V_{out}}{V_{in}} = \frac{\alpha r_c R_l}{R_l\,(R_s + r_b + r_e) + \alpha r_e r_c}$$

The second term in the denominator can be neglected in comparison with the others, giving

$$\frac{V_{out}}{V_{in}} = \frac{\alpha r_c}{R_s + r_b + r_e} \qquad \ldots.(18)$$

which is identical with expression (17) for the common-base amplifier. Thus for small source resistances or large values of load resistance the voltage gain of a given transistor with a given value of collector load resistance is approximately the same, no matter whether the transistor is connected up as a common base or a common-emitter amplifier. The curve relating voltage gain with load resistance is similar to that for the common-base amplifier and is given in Fig. 4.4.

Expression (16) may also be written

$$\frac{V_{out}}{V_{in}} = \frac{\dfrac{\alpha r_c - r_e}{R_l + r_e + r_c(1-\alpha)} \cdot R_l}{R_s + r_b + r_e + \dfrac{r_e(\alpha r_c - r_e)}{R_l + r_e + r_c(1-\alpha)}}$$

Simplifying the numerator by omitting the negligible term r_e and substituting for r_i from expression (6) we have

$$\frac{V_{out}}{V_{in}} = \frac{\dfrac{\alpha r_c}{R_l + r_c(1-\alpha)} \cdot R_l}{R_s + r_i}$$

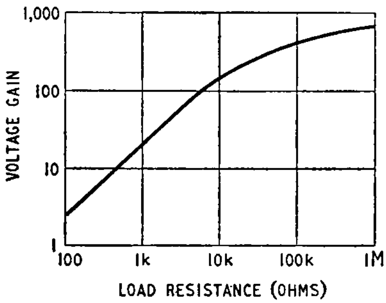


Fig. 4.4. Variation of voltage gain with load resistance for a common-emitter amplifier

If R_l is small compared with $r_c(1-\alpha)$ and R_s is small compared with r_i this becomes

$$\frac{V_{out}}{V_{in}} = \frac{\dfrac{\alpha}{1-\alpha} \cdot R_l}{r_i}$$

This is usually written

$$\frac{V_{out}}{V_{in}} = \frac{\alpha' R_l}{r_i} \qquad \ldots .(19)$$

where

$$\alpha' = \frac{\alpha}{1-\alpha} \qquad \ldots .(20)$$

Expression (19) is of the type used in Chapter 2 in the simplified calculation of stage gains of transistors and it suggests that the current gain of the common-emitter circuit is equal to $\alpha/(1 - \alpha)$. This can be confirmed from equation (3)

$$\dot{i}_c = -\frac{r_e - \alpha r_c}{R_l + r_e + (1 - \alpha)\, r_c} \cdot \dot{i}_b$$

which on simplification and for small collector load values gives the result

$$\text{current gain} = \frac{\dot{i}_c}{\dot{i}_b} = \frac{\alpha}{1 - \alpha}$$

In this book the current gain of the common-emitter amplifier is represented by α' but in some publications it is represented by β.

Current Gain of Common-emitter Circuit

Values of α are commonly near unity and the current gain of the common-emitter circuit can be considerable. For example, if $\alpha = 0{\cdot}98$ we have

$$\alpha' = \frac{0{\cdot}98}{1 - 0{\cdot}98}$$

$$= \frac{0{\cdot}98}{0{\cdot}02}$$

$$= 49$$

The common-emitter amplifier has an input resistance many times that of the common-base amplifier. For a given voltage across the input terminals, therefore, it takes a smaller input current.

The current gain of the common-emitter amplifier is, however, many times that of the common-base amplifier (which is always less than unity) and it is therefore not surprising that there is negligible difference between their voltage gains when the source resistance is low. If the source resistance is high, the input current has a tendency to be independent of the input resistance and is the same for both common-base and common-emitter amplifiers. The voltage gain of the common-emitter amplifier is now greater than that of the common-base amplifier because of its greater current gain.

Practical Circuit for a Common-emitter Amplifier

Fig. 4.5 gives the circuit diagram for a practical common-emitter amplifier. For simplicity this shows a single battery used for biasing the base and the collector relative to the emitter. There are alternative circuits employing a potential divider which also

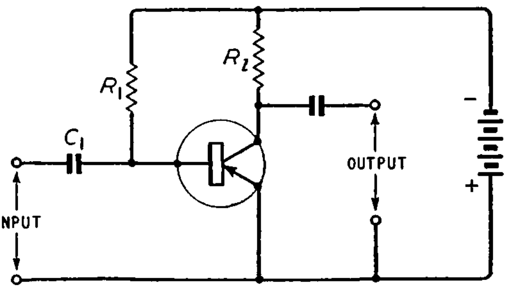


Fig. 4.5. Practical circuit for a common-emitter amplifier

give protection against thermal runaway: these are discussed more fully in Chapter 6. At the moment we are interested in calculating suitable values for C_1, R_1 and R_l. We will assume that the battery is 6 volts and that the mean collector current is to be 1 mA. If the transistor is to be a class-A amplifier in which the collector voltage can swing above and below its quiescent value, a suitable quiescent voltage is — 3 volts. The voltage drop across R_l is then 3 volts and, if the current is 1 mA, the value of R_l must be 3 kΩ.

We will assume α to be 0·98 giving α' as 49. When the quiescent collector current is 1 mA, the quiescent base current is 1/49 mA, approximately 20 μA. There is normally little difference between the potentials of emitter and base, and the voltage across the base bias resistor, R_1, is 6 volts. The value of R_1 is thus, from Ohm's law, $6/(20 \times 10^{-6})$, i.e. 300 kΩ. The value of C_1 depends on the input resistance of the transistor and the lowest frequency of operation. As shown earlier, the input resistance depends on the load resistance, but may be taken as approximately 1 kΩ. If we make the reactance of C_1 equal to 1 kΩ at the lowest frequency, there will be a 3 dB loss at this frequency. For an audio-frequency amplifier the low-frequency limit may be 50 c/s and we have

$$C_1 = \frac{1}{2\pi f r_i}$$

$$= \frac{1}{6{\cdot}284 \times 50 \times 1{,}000} \text{ F}$$

$$= \frac{20}{6{\cdot}284} \mu\text{F}$$

$$= 3 \ \mu\text{F approximately}$$

Collector Current – Collector Voltage Characteristics for Common emitter Operation

The behaviour of a common-emitter circuit such as that illustrated in Fig. 4.5 can be deduced from a set of collector current – collector

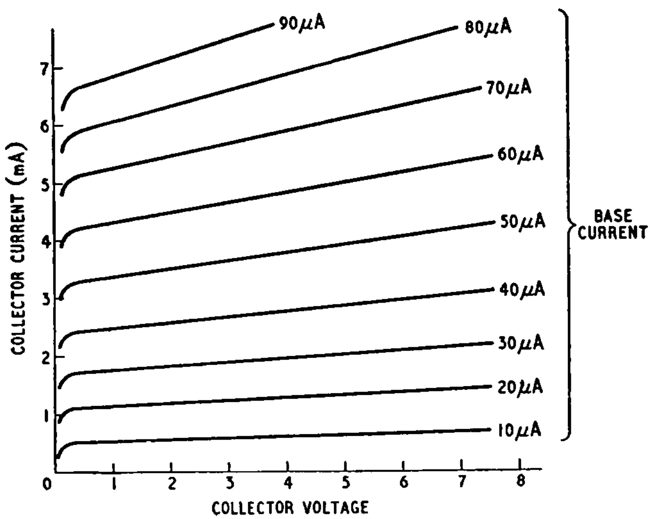


Fig. 4.6. Typical collector current–collector voltage characteristics for common-emitter connection

voltage characteristics plotted for various values of base current. Such a set of curves is given in Fig. 4.6. They are similar in shape to the anode current – anode voltage curves for a pentode valve but the knee of the characteristics occurs at a very low collector voltage, permitting large swings in collector voltage during amplification.

A number of parameters of the transistor can be obtained from these characteristics. For example the slope of the characteristics for collector voltages exceeding that of the knee is not as low as that for the common-base connection, showing that the effective collector a.c. resistance is smaller than for the common-base connection.

In fact, from Fig. 4.6 the effective collector a.c. resistance is approximately 30 kΩ.

Power Gain

We have seen that the voltage gain of a common-emitter amplifier is of the same order as that of the common-base amplifier and a typical value is 100. Unlike the common-base amplifier, however, the common-emitter type has considerable current gain, a typical value of α' being 50. Thus the power gain of a common-emitter amplifier can easily amount to 5,000 (37 dB), 17 dB more than is available from a common-base amplifier.

Comparison of Common-emitter and Common-base Amplifiers

We have now discussed the properties of common-base and common-emitter amplifiers and can assess their relative advantages and disadvantages. The common-base amplifier has a very low input resistance and a very high output resistance which make the design of inter-stage matching networks or transformers difficult. Nevertheless this circuit is used, particularly in head amplifiers where the input resistance provides a reasonable match for microphones or pick-ups of the moving-coil type. The input resistance of the common-emitter amplifier is higher and the output resistance is lower than that of the common-base amplifier, simplifying the design of inter-stage networks and transformers. On the other hand the thermal stability of the common-emitter circuit is inferior to that of the common-base amplifier and protective circuits are essential in common-emitter amplifiers (see Chapter 6). The greater power gain of the common-emitter circuit is usually regarded as more important than the disadvantages and this circuit is extensively employed in amplifiers and receivers.

5

COMMON-COLLECTOR AMPLIFIERS

Introduction

COMMON-BASE amplifiers were described in Chapter 3 and common-emitter amplifiers in Chapter 4. This chapter is devoted to the third basic form of transistor amplifier circuit, the common-collector circuit. As in the earlier chapters we shall analyse the performance of the amplifier by deriving expressions for the input resistance, output resistance and voltage gain.

The fundamental circuit of a common-collector amplifier is given in Fig. 5.1 (a) in which R_s represents the resistance of the signal source and R_l represents the load resistance. For simplicity,

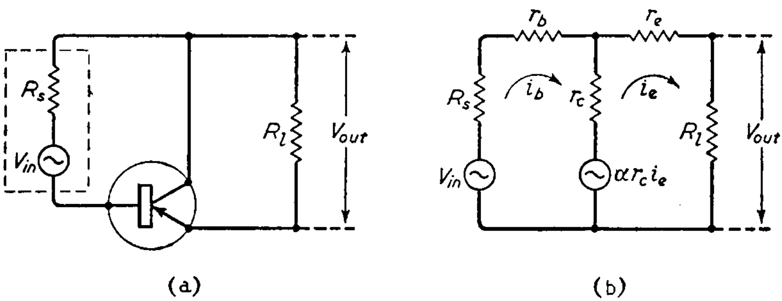


Fig. 5.1. The basic circuit for the common-collector transistor amplifier is given at (a), and the equivalent circuit at (b)

bias sources are omitted from this diagram. In Fig. 5.1 (b) the transistor is represented by an equivalent network of resistances r_e, r_b and r_c together with a voltage generator of e.m.f. $\alpha i_e r_c$ where i_e is the alternating current in the emitter circuit. If the base current i_b and emitter current i_e are both shown acting in clockwise directions as in Fig. 5·1 (b), then the current in the common collector resistance r_c is the difference $(i_e - i_b)$. From Chapter 2 we know that this difference is the collector current i_c: all conventions are therefore satisfied in this diagram.

Input Resistance

Applying Kirchhoff's laws to the circuit of Fig. 5.1 (b) we have

$$\begin{aligned} V_{in} &= i_b(r_b + r_c + R_s) + \alpha r_c i_e - i_e r_c \\ &= i_b(r_b + r_c + R_s) - i_e r_c(1 - \alpha) \qquad \ldots .(1) \\ 0 &= i_e(r_c + r_e + R_l) - \alpha r_c i_e - r_c i_b \\ &= i_e[r_c(1 - \alpha) + r_e + R_l] - r_c i_b \qquad \ldots .(2) \end{aligned}$$

To find an expression for the input resistance we must eliminate i_e between expressions (1) and (2) to obtain a relationship between V_{in} and i_b. From (2)

$$i_e = i_b \cdot \frac{r_c}{r_c(1 - \alpha) + r_e + R_l}$$

Substituting for i_e in (1)

$$V_{in} = i_b(r_b + r_c + R_s) - r_c(1 - \alpha) \cdot \frac{i_b r_c}{r_c(1 - \alpha) + r_e + R_l}$$

$$\therefore \quad i_b = \frac{V_{in}}{r_b + r_c + R_s - \dfrac{r_c^2(1 - \alpha)}{r_c(1 - \alpha) + r_e + R_l}} \qquad \ldots .(3)$$

In a simple circuit containing a generator of resistance R_s and a load of resistance r_i the current is given by

$$i_1 = \frac{V_{in}}{R_s + r_i} \qquad \ldots .(4)$$

Comparison of (3) and (4) gives the input resistance of the network as

$$r_i = r_b + r_c - \frac{r_c^2(1 - \alpha)}{r_c(1 - \alpha) + r_e + R_l} \qquad \ldots .(5)$$

which shows that the input resistance of the common-collector amplifier depends on the parameters of the transistor (r_b, r_c, r_e and α) and on the load resistance R_l.

This expression can be simplified slightly by combining the second and third terms. We then have

$$r_i = r_b + \frac{r_c^2(1-\alpha) + r_e r_c + r_c R_l - r_c^2(1-\alpha)}{r_c(1-\alpha) + r_e + R_l}$$

$$= r_b + r_c \frac{r_e + R_l}{r_c(1-\alpha) + r_e + R_l} \quad \ldots.(6)$$

If R_l and r_e are small compared with $r_c(1-\alpha)$ the input resistance is given by

$$r_i = r_b + \frac{r_c r_e}{r_c(1-\alpha)}$$

$$= r_b + \frac{r_e}{1-\alpha} \quad \ldots.(7)$$

If R_l is large compared with $r_c(1-\alpha)$ the input resistance is given by

$$r_i = r_b + r_c \frac{R_l}{R_l}$$

$$= r_b + r_c \quad \ldots.(8)$$

Input Resistance for Junction Transistor as Common-collector Amplifier

For a junction transistor having $r_b = 300$ ohms, $r_e = 25$ ohms, $r_c = 1$ megohm, and $\alpha = 0{\cdot}98$, the input resistance for short-circuited output terminals is given by expression (7), namely

$$r_i = r_b + \frac{r_e}{1-\alpha}$$

$$= 300 + \frac{25}{1 - 0{\cdot}98} \text{ ohms}$$

$$= 300 + \frac{25}{0{\cdot}02} \text{ ohms}$$

$$= 300 + 1{,}250 \text{ ohms}$$

$$= 1{\cdot}55 \text{ k}\Omega$$

The input resistance for open-circuited output terminals is given by expression (8)

$$r_i = r_b + r_c$$

$$= 300 + 1{,}000{,}000 \text{ ohms}$$

$$= 1 \text{ megohm approximately}$$

The input resistance therefore increases considerably as the load resistance is increased, the curve having the shape shown in Fig. 5.2. For large values of load resistance the input resistance approaches the value of r_c, which can be as high as 2 megohms for junction transistors. This is by far the highest value of input

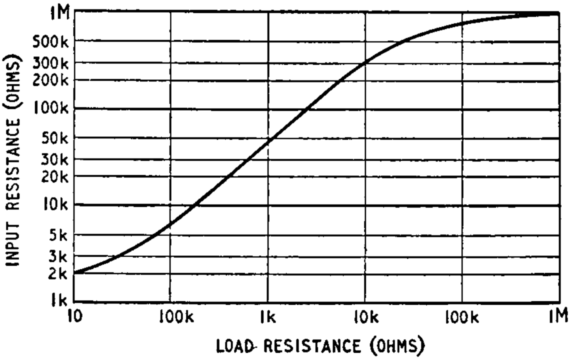


Fig. 5.2. Variation of input resistance with load resistance for a common-collector amplifier

resistance so far encountered for a circuit employing a junction transistor; it is, in fact, comparable with the input resistance of a thermionic valve amplifier which commonly has grid resistors of this value.

This high input resistance is the most useful property of the common-collector amplifier and the chief reason for its use.

Output Resistance

We can obtain an expression for the output resistance of a common-collector transistor amplifier by eliminating i_b between (1) and (2), so as to obtain a relationship between i_e and V_{in}. From (2) we have

$$i_b = \frac{r_c(1 - \alpha) + r_e + R_l}{r_c} \cdot i_e$$

Substituting for i_b in (1)

$$V_{in} = \frac{(r_b + r_c + R_s)\,[r_c(1 - \alpha) + r_e + R_l]}{r_c}\, i_e - i_e r_c(1 - \alpha)$$

$$\therefore i_e = \frac{V_{in}}{\dfrac{(r_b + r_c + R_s)\,[r_c(1-\alpha) + r_e + R_l]}{r_c} - r_c(1-\alpha)}$$

Multiplying numerator and denominator by $r_c/(r_b + r_c + R_s)$ we have

$$i_e = \frac{\dfrac{r_c}{r_b + r_c + R_s}\,V_{in}}{R_l + r_e + r_c(1-\alpha) - \dfrac{r_c^2(1-\alpha)}{r_b + r_c + R_s}} \qquad \ldots.(9)$$

In a simple circuit containing a signal source of voltage V and internal resistance r_o feeding a load resistance R_l, the current i_2 is given by

$$i_2 = \frac{V}{R_l + r_o} \qquad \ldots.(10)$$

Comparing (9) with (10) we obtain the following expression for the output resistance r_o of the common-emitter amplifier

$$r_o = r_e + r_c(1-\alpha) - \frac{r_c^2(1-\alpha)}{r_b + r_c + R_s} \qquad \ldots.(11)$$

Combining the second and third terms

$$r_o = r_e + r_c(1-\alpha) \cdot \frac{r_b + R_s}{r_b + r_c + R_s} \qquad \ldots.(12)$$

This expression shows that the output resistance depends on r_e, r_b, r_c, α and the source resistance R_s.

When the source resistance is very small we have

$$r_o = r_e + r_c(1-\alpha) \cdot \frac{r_b}{r_b + r_c}$$

For junction transistors r_b is small compared with r_c and this expression simplifies to

$$r_o = r_e + (1-\alpha)\,\frac{r_b r_c}{r_c}$$

$$= r_e + (1-\alpha) r_b \qquad \ldots.(13)$$

When the source resistance R_s is very large compared with r_b and r_c, expression (12) simplifies to

$$r_o = r_e + r_c(1 - \alpha) \qquad \ldots.(14)$$

Output Resistance for a Junction Transistor as Common-collector Amplifier

For a junction transistor having $r_b = 300$ ohms, $r_e = 25$ ohms, $r_c = 1$ megohm, and $\alpha = 0{\cdot}98$, the output resistance for short-

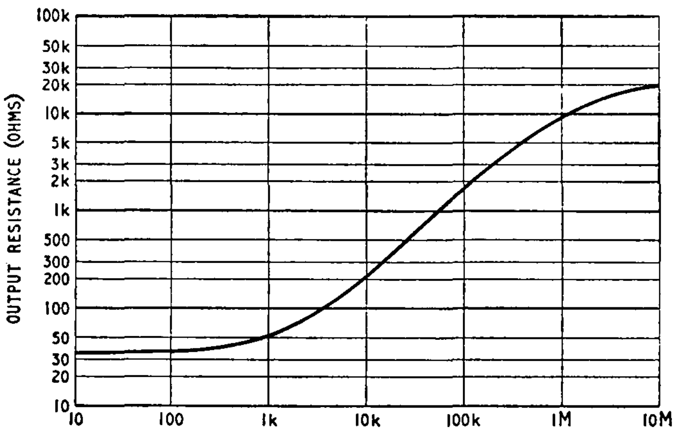


Fig. 5.3. Variation of output resistance with source resistance for a common-collector amplifier

circuited input terminals is, from expression (13), given by the following

$$\begin{aligned} r_o &= r_e + (1 - \alpha)r_b \\ &= 25 + (1 - 0{\cdot}98)\ 300 \text{ ohms} \\ &= 25 + 300 \times 0{\cdot}02 \text{ ohms} \\ &= 25 + 6 \text{ ohms} \\ &= 31 \text{ ohms} \end{aligned}$$

For open-circuited input terminals, the output resistance is given by expression (14)

$$
\begin{aligned}
r_o &= r_e + (1 - \alpha)r_c \\
&= 25 + (1 - 0{\cdot}98) \times 1{,}000{,}000 \text{ ohms} \\
&= 25 + 1{,}000{,}000 \times 0{\cdot}02 \text{ ohms} \\
&= 20 \text{ k}\Omega \text{ approximately}
\end{aligned}
$$

The variation of output resistance with generator resistance is illustrated in Fig. 5.3. For small values of generator resistance, the output resistance is very low, being only slightly greater than the emitter resistance r_e. It is easily possible, for example, to have an output resistance of less than 50 ohms: a value as low as this is impossible from common-base or common-emitter transistor amplifiers.

Thus a common-collector amplifier with a low value of generator resistance and a high value of collector load resistance can have a high value of input resistance and a low value of output resistance, conditions opposite to those normally encountered in transistor amplifiers and similar in fact to those which apply to a cathode follower using a thermionic valve.

Voltage Gain

From Fig. 5.1 (b) we can see that the output voltage is given by $i_e R_l$. The voltage gain V_{out}/V_{in} is thus given by $i_e R_l/V_{in}$. From (9) i_e is given by

$$
i_e = \frac{\dfrac{r_c}{r_b + r_c + R_s} V_{in}}{R_l + r_e + r_c(1 - \alpha) - \dfrac{r_c^2(1 - \alpha)}{r_b + r_c + R_s}}
$$

Hence

$$
\frac{V_{out}}{V_{in}} = \frac{\dfrac{r_c R_l}{r_b + r_c + R_s}}{R_l + r_e + r_c(1 - \alpha) - \dfrac{r_c^2(1 - \alpha)}{r_b + r_c + R_s}}
$$

$$
= \frac{r_c R_l}{[R_l + r_e + r_c(1 - \alpha)](r_b + r_c + R_s) - r_c^2(1 - \alpha)} \qquad \ldots .(15)
$$

This can be simplified by ignoring r_e in comparison with the other terms in the first bracket of the denominator and by ignoring r_b in the second bracket. After further evaluation this gives

$$\frac{V_{out}}{V_{in}} = \frac{r_c R_l}{R_l r_c + R_l R_s + r_c(1-\alpha) R_s}$$

$$= \frac{R_l}{R_l + (1-\alpha) R_s + R_l R_s / r_c} \qquad \ldots.(16)$$

For a junction transistor α is always less than unity, and hence the denominator always exceeds R_l, giving a voltage gain of less than unity. If R_s is small, the gain is very nearly equal to unity, another respect in which the common-collector amplifier is similar to a thermionic-valve cathode-follower circuit.

Consider a junction transistor having a source resistance (R_s) of 1 kΩ and a load resistance (R_l) of 100 kΩ. If $\alpha = 0{\cdot}96$ and $r_c = 1$ megohm we have, from expression (16)

voltage gain

$$= \frac{100{,}000}{100{,}000 + (1 - 0{\cdot}96)\ 1{,}000 + 100{,}000 \times 1{,}000/1{,}000{,}000}$$

$$= \frac{100{,}000}{100{,}000 + 40 + 100}$$

$$= 100{,}000/100{,}140$$

$$= 1 \text{ very nearly}$$

Current Gain of Common-collector Amplifier

Expression (15) may be written

$$\frac{V_{out}}{V_{in}} = \frac{\dfrac{r_c R_l}{R_l + r_e + r_c(1-\alpha)}}{R_s + r_b + r_c - \dfrac{r_c^2(1-\alpha)}{R_l + r_e + r_c(1-\alpha)}}$$

and this can be expressed in the form

$$\frac{V_{out}}{V_{in}} = \frac{\dfrac{r_c R_l}{R_l + r_e + r_c(1-\alpha)}}{R_s + r_i}$$

where r_i is the input resistance (see expression 5). This may be written

$$\frac{V_{out}}{V_{in}} = \frac{R_l}{R_s + r_i} \cdot \frac{r_c}{R_l + r_e + r_c(1 - \alpha)}$$

If R_l and r_e are small compared with $r_c(1 - \alpha)$ and R_s is small compared with r_i, the voltage gain becomes

$$\frac{V_{out}}{V_{in}} = \frac{R_l}{r_i} \cdot \frac{1}{1 - \alpha}$$

$$= \frac{\text{collector load resistance}}{\text{input resistance}} \times \text{current gain} \quad \ldots .(17)$$

This expression is of the type used in Chapter 2 in the simplified calculations of stage gains of transistors and it suggests that the current gain of the common-collector circuit is equal to $1/(1 - \alpha)$. This can be confirmed from equation (2).

$$\dot{i}_e = \frac{r_c}{R_l + r_e + r_c(1 - \alpha)} \cdot \dot{i}_b$$

which on simplification and for small collector load values gives the result

$$\text{current gain} = \frac{\dot{i}_e}{\dot{i}_b} = \frac{1}{1 - \alpha} \quad \ldots .(18)$$

When α is nearly unity, as in junction transistors, the value of current gain can be very high. For example if $\alpha = 0{\cdot}98$, we have

$$\text{current gain} = \frac{1}{1 - \alpha}$$

$$= \frac{1}{1 - 0{\cdot}98}$$

$$= \frac{1}{0{\cdot}02}$$

$$= 50$$

The value of current gain for a common-collector amplifier is not very different from that for a common-emitter amplifier: when α

is nearly unity, $\alpha/(1 - \alpha)$ is not very different from $1/(1 - \alpha)$. In spite of this very high value of current gain, however, the common-collector amplifier gives less than unity voltage gain because the input resistance is so high compared with the load resistance. In fact if we equate expression (18) to unity we have the following approximate relationship

$$\text{input resistance} = \frac{1}{1 - \alpha} \times \text{load resistance}$$

$$= \text{current gain} \times \text{load resistance} \quad \ldots . (19)$$

Practical Circuit for a Common-collector Amplifier

A circuit diagram for a common-collector amplifier is given in Fig. 5.4 and the component values required in it can be calculated by methods similar to those employed for other amplifiers. For example, if the supply is 6 volts and the mean current in the load resistance is to be 1 mA, a convenient value for the load resistance is 3 kΩ, this permitting a peak swing of emitter potential of 3 volts. The quiescent emitter potential is thus -3 volts and the quiescent

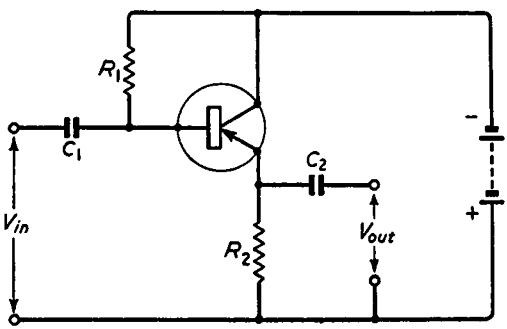


Fig. 5.4. Practical circuit of a common-collector amplifier

base potential will not differ very greatly from this value. The potential difference across the base bias resistor R_1 is thus approximately 3 volts and if the base current is 20 μA (corresponding to a value of α' of 50) the bias resistance is given by $3/(20 \times 10^{-6})$, i.e., 150 kΩ.

The capacitance C_2 depends on the resistance to which the output terminals of the common-collector amplifier is connected. The capacitance C_1 depends on R_1 and the input resistance of the amplifier which, as shown above, also depends on the load resistance. If the load resistance is high, the net input resistance can be as much as 100 kΩ and, for an a.f. amplifier, C_1 can be 0·1 μF which has a reactance of approximately 31 kΩ at 50 c/s and gives a loss of less than 1 dB at this frequency.

An interesting feature of the common-collector amplifier is that signals applied to the output terminals give a signal output at the input terminals. In this respect, such amplifiers, when used as buffer stages, are not as perfect as thermionic-valve cathode followers. This property of the common-collector amplifier can be anticipated from Fig. 5.1 (b): the equivalent generator $\alpha r_c i_e$ is in the common shunt arm of the network and the two resistances r_e and r_b are of approximately the same magnitude. Thus the network is very nearly symmetrical and a signal applied across R_l can produce an output across R_s.

6

BIAS STABILISATION

Introduction

THE collector current of a junction transistor consists of two components, one controlled by the base current and the other independent of it. The first component is useful because the amplifying properties of the transistor depend on it: the second is useless and is usually known as the leakage current. The leakage current is caused by thermal dissociation of covalent bonds as described in Chapter 1 and is strongly dependent on temperature; at high temperatures it may become comparable with the useful current, causing severe limitations in the performance of the transistor. For example, if the leakage current is an appreciable fraction of the total collector current, it is impossible for the collector voltage to swing up to the collector supply voltage under the action of an applied signal: thus the undistorted output-voltage swing is reduced by the presence of the leakage current.

Leakage current can, under certain conditions, cause damage to or even destruction of a transistor. For such a current heats up the collector junction, causing further increase in collector current which accelerates the heating process. Unless precautions are taken to prevent the collector current rising, a regenerative process can occur, resulting in an abnormally large collector current which, if it exceeds the safe collector dissipation, can cause damage to the transistor.

Transistor circuits must be designed to prevent this thermal runaway by limiting the collector current to a safe value. Such protective circuits also tend to make the performance of the equipment less dependent on the parameters of the transistors, permitting the exchange of one transistor for another.

LEAKAGE CURRENT IN A COMMON-BASE AMPLIFIER

Consider a junction transistor used as a common-base amplifier such as that shown in Fig. 6.1. If the emitter battery circuit is broken, the collector current falls to a low value which is due to the reverse current of the diode formed by the collector-base junction. This current is the leakage current referred to above and is represented by I_{c0}. When the emitter circuit is restored, the collector current rises due to the addition of the useful current αI_e from the emitter circuit. Thus the total collector current I_c is given by

$$I_c = \alpha I_e + I_{c0} \qquad \ldots .(1)$$

For a small junction transistor the leakage current is very small, a typical value being 5 μA at a temperature of 25° C. It increases substantially exponentially with temperature and may reach 50 μA at 55° C, but even so is still small compared with αI_e which may be 1 mA. Thus it may be said that leakage current has a negligible effect on the performance of a common-base amplifier, and provided

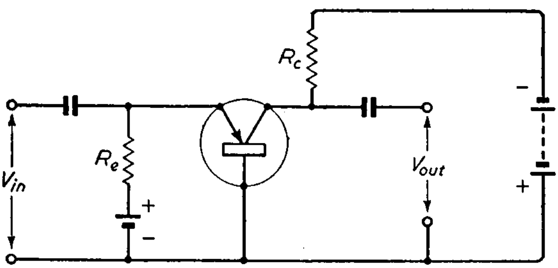


Fig. 6.1. Simple uncompensated common-base amplifier which has good thermal stability

R_e and R_c are chosen to keep the collector dissipation below the maximum safe value, the likelihood of thermal runaway does not exist.

Protective circuits are thus unnecessary in a common-base amplifier. Such an amplifier is said to have good d.c. stability and, in fact, it is the aim in the design of common-emitter transistor amplifiers to achieve d.c. stability comparable with that of a common-base circuit.

LEAKAGE CURRENT IN A COMMON-EMITTER AMPLIFIER

For a number of reasons which were listed earlier, the common-emitter amplifier is preferred in transistor circuitry to the common-base amplifier. The basic circuit for a common-emitter amplifier is given in Fig. 6.2.

If the base resistor circuit is broken in this amplifier, there is a residual current flowing between collector and emitter. However, this current arises from a process somewhat different from that producing the leakage current in a common-base amplifier. The collector-base junction is reverse-biased as in a common-base amplifier and there is hence a collector-base leakage current of

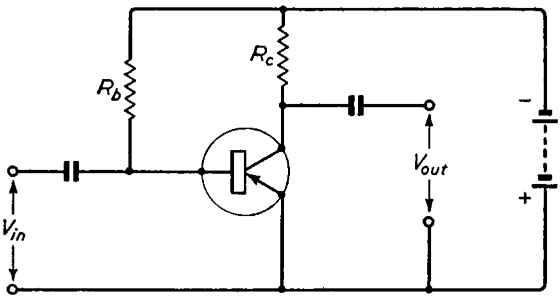


Fig. 6.2. Simple uncompensated common-emitter amplifier which has poor thermal stability

magnitude I_{c0} as before. There is, however, no net external base current (because the circuit is disconnected) and it follows that there must be an emitter-base current, also equal to I_{c0}, but flowing in the opposite direction to the collector-leakage current.

This current behaves as an input current and is magnified by transistor action, appearing as a current of $\alpha' I_{c0}$ in the collector circuit where $\alpha' = \alpha/(1 - \alpha)$. Thus the total leakage current in the collector circuit is

$$I_{c0} + \alpha' I_{c0} = (1 + \alpha')I_{c0}$$

This is usually represented by I_{c0}'

$$\therefore \; I_{c0}' = (1 + \alpha')I_{c0}$$

$$= \frac{1}{1 - \alpha} I_{c0}$$

α' is not so large as in signal calculations but I_{c0}' can be considerably greater than I_{c0}. In fact at 25° C for a small junction transistor I_{c0}' can be as much as 250 μA rising to 2·5 mA at 55° C;

this is more than twice normal values of signal current and can cause great deterioration in circuit performance. It is essential, therefore, in common-emitter transistor amplifiers to include protective circuits to avoid undue rise in collector current.

On the other hand the very large leakage current of the common-emitter amplifier is of great value in photo-transistors as explained in Chapter 13.

In general, for a common-emitter circuit

$$I_c = \alpha' I_b + I_{c0}' \qquad \ldots .(2)$$

where $\alpha' I_b$ is the useful component of the collector current and I_{c0}' is the leakage current.

STABILITY FACTOR

Suppose there is a change of leakage current $\Delta I_{c0}'$ due to a change of temperature in an unstabilised circuit. If the useful current $\alpha' I_b$ remains constant, this causes an equal change in the total collector current I_c.

Now suppose that a stabilising circuit is applied to the transistor amplifier. Over the same temperature range, this has the effect of reducing the change in collector current to a smaller value than that which occurred in the unstabilised circuit. The ratio of the two changes is known as the *stability factor K*.

Hence:

$$K = \frac{\text{change in total collector current in stabilised circuit}}{\text{change in total collector current in unstabilised circuit}}$$

The change in collector current in the unstabilised circuit is due entirely to a change in the leakage current, if, as assumed here, the useful component of collector current is constant. Thus K is given by

$$K = \frac{\Delta I_c}{\Delta I_{c0}'}$$

For an unstabilised circuit $\Delta I_c = \Delta I_{c0}'$ and $K = 1$. For a stabilised circuit ΔI_c is smaller than $\Delta I_{c0}'$ and K is less than unity, the smallness of K being a measure of the success of the stabilising circuit.

Values of K of 0·1 or less can quite easily be achieved by conventional circuits, and this shows that the variations in collector current caused by, say, a given temperature change, are less than 1/10th the change which would occur in an unstabilised circuit.

PROTECTIVE CIRCUITS FOR COMMON-EMITTER AMPLIFIERS

Use of a Collector-base Resistor

A simple method which can be employed to increase the d.c. stability of a common-emitter amplifier is that illustrated in Fig. 6.3. This is similar to the circuit of Fig. 6.2 but the difference is that the resistor R_b is returned to the collector instead of to the negative terminal of the supply.

The improvement can be explained qualitatively in the following way. Suppose the leakage current increases as a result of an increase in temperature. This causes an increase in the voltage across the collector load R_c and the collector potential becomes more positive. The base input current is primarily determined

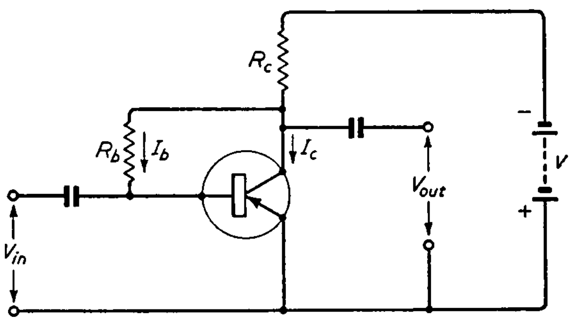


Fig. 6.3. A simple method of improving the thermal stability of a common-emitter amplifier

by the resistance R_b and the collector voltage, and this change in collector voltage causes a reduction in base current and hence in the useful component of collector current. This in turn causes a negative movement of collector potential and a process of re-adjustment occurs, ending with a total collector current greater than the original value but not so great as in an uncompensated circuit such as that shown in Fig. 6.2.

The precise improvement in d.c. stability afforded by this simple circuit can be calculated in the following way. From equation (2) we have

$$I_c = \alpha' I_b + I_{c0}' \qquad \ldots.(3)$$

The base-emitter voltage is usually small compared with the collector supply voltage V and thus we may say

$$I_bR_b + (I_b + I_c)R_c = V \qquad \ldots .(4)$$

From (3)

$$I_b = \frac{I_c - I_{c0}'}{\alpha'}$$

Substituting for I_b in (4)

$$\frac{(I_c - I_{c0}')R_b}{\alpha'} + \left(\frac{I_c - I_{c0}'}{\alpha'} + I_c\right)R_c = V$$

Rearranging this we have

$$I_c[R_b + (\alpha' + 1)R_c] = I_{c0}'(R_b + R_c) + \alpha'V$$

Differentiating with respect to I_{c0}'

$$K = \frac{\mathrm{d}I_c}{\mathrm{d}I_{c0}'} = \frac{R_b + R_c}{R_b + R_c + \alpha'R_c}$$

$$= \frac{1}{1 + \alpha'R_c/(R_b + R_c)}$$

R_c, R_b and α' are all positive quantities and this expression shows that the stability factor K is less than unity. The aim, of course, is to get the best stability, i.e. the lowest possible value of K. This implies a high value of α', a large value of R_c or a small value of R_b. There is not usually much choice of value of α' and this is not likely to exceed 50. R_c and R_b cannot be varied at will to secure good stability because their values have to satisfy other and more important considerations.

For example, R_c is the collector load resistance and its value is determined by the collector supply voltage and the mean collector current of the transistor. R_b supplies base bias and its value, together with that of the collector supply voltage, determines the bias current. Thus the d.c. stability obtainable from this circuit is to a large extent determined automatically by other design requirements.

To indicate the value of stability likely to be obtained, we will consider a typical common-emitter amplifier and will calculate likely values for R_c and R_b. If the collector supply is -6 volts and the mean collector current is 1 mA, a suitable value for R_c is 3 kΩ, for this gives a mean collector voltage of $-$ 3 volts which permits upward and downward swings of collector voltage of approximately 3 volts peak value.

If the mean collector current is 1 mA and α' is 50, the mean value of the base current is approximately 1/50 mA, i.e. 20 μA. The base-collector voltage is 3 approximately and to give the required value of base current, R_b must be $3/(20 \times 10^{-6})$ ohms, i.e. 150 kΩ.

The stability factor is thus given by

$$K = \frac{1}{1 + \alpha' R_c/(R_b + R_c)}$$

$$= \frac{1}{1 + 50 \times 3/(150 + 3)}$$

$$= \frac{1}{1 + 1}$$

$$= 0{\cdot}5 \text{ approximately}$$

This is not a great improvement over the unstabilised circuit. A better performance is possible if the amplifier is not required to give a large signal output. For example, if the output voltage swings never exceed 1 volt it would be possible to operate with a mean collector voltage of -1 volt. For 1 mA mean collector current the collector load-resistance must now be 5 kΩ, and for 20 μA mean base current R_b is given by $1/(20 \times 10^{-6}) = 50$ kΩ.

The d.c. stability is now given by

$$K = \frac{1}{1 + 50 \times 5/(50 + 5)}$$

$$= \frac{1}{1 + 4{\cdot}5} \text{ approximately}$$

$$= \frac{1}{5{\cdot}5} \text{ approximately}$$

The resistor R_b forms with the resistance of the signal source (not indicated in Figs. 6.2 and 6.3) a potential divider which returns a fraction of the output voltage of the amplifier to its input. This causes a reduction in gain which, if not desirable, can be eliminated by constructing R_b of two resistors R_1 and R_2 in series, the junction being connected to earth by a low-reactance capacitor as shown in Fig. 6.4.

If R_1 is small, it shunts the collector load resistance of the amplifier, and if R_2 is small, it shunts the input terminals and lowers the input resistance. Usually, therefore, R_1 and R_2 are made approximately equal and the capacitor is chosen to have a

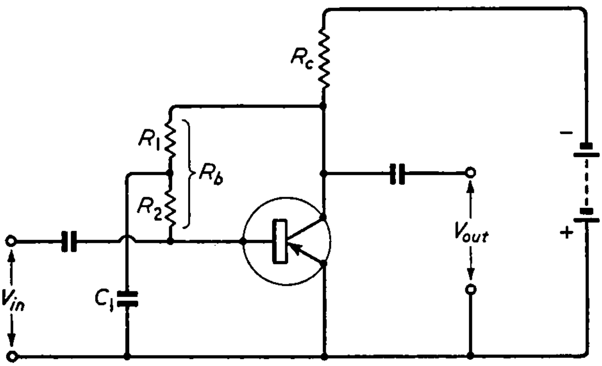


Fig. 6.4. To avoid negative feedback due to R_b, decoupling may be introduced as shown here

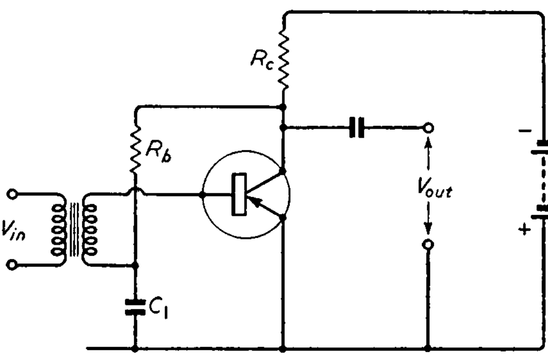


Fig. 6.5. Method of decoupling which can be used with transformer coupling

reactance which is small compared with the resistance value at the lowest frequency it is required to amplify. For example, if R_b is 100 kΩ, R_1 and R_2 can be 50 kΩ and, for an a.f. amplifier,

C_1 can be 2 μF which has a reactance of approximately 1·6 kΩ at 50 c/s.

If this simple method of stabilisation is employed in a common-emitter circuit which is coupled to the previous stage by a transformer, then the base end of the stabilising resistor R_b can be decoupled to the emitter directly as shown in Fig. 6.5.

Use of a Potential Divider and Emitter Resistor

A better method of ensuring good d.c. stability in a common-emitter transistor amplifier is illustrated in Fig. 6.6. It employs three resistors, two forming a fixed potential divider across the collector supply battery and a third included in the emitter circuit of the transistor. The emitter resistor gives negative feedback

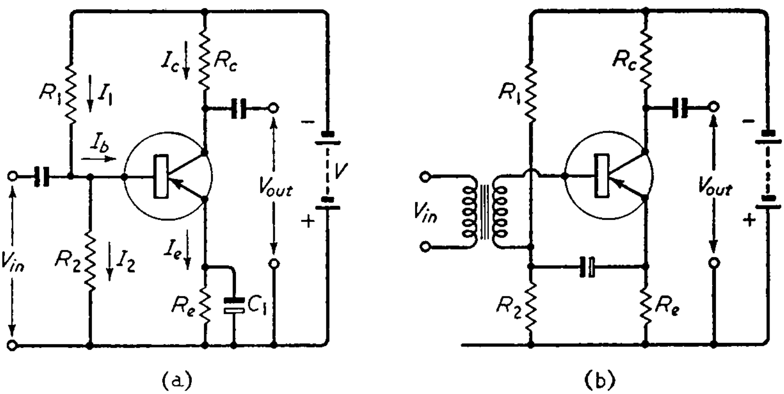


Fig. 6.6. Potential divider method of stabilising the d.c. conditions in a common-emitter amplifier (a) in an RC-coupled amplifier and (b) in a transformer coupled amplifier. Arrows indicate electron flow

and a consequent reduction of gain: if this is undesirable, the emitter resistor can be decoupled by a low-reactance capacitor, this constituting the fourth additional component required in this method of stabilisation.

Two basic versions of this circuit exist. In Fig. 6.6 (a) the circuit is arranged for RC coupling from the previous stage: a feature of this arrangement is that the resistance of R_1 and R_2 in parallel is effectively shunted across the input circuit of the transistor. This parallel resistance should not, therefore, be too small. In Fig. 6.6 (b) the circuit is arranged for transformer coupling from the previous stage: the parallel resistance of R_1 and R_2 does not now enter into input-resistance considerations.

The improvement in stability due to this circuit can be explained in the following way. The base is connected to the junction point of the potential divider and, if we neglect any variations in base current, has a constant voltage. The transistor takes an emitter current such that the emitter voltage is practically equal to the base voltage. If now the leakage current increases, due to a rise in temperature, the emitter voltage tends to approach that of the base and the base-emitter voltage therefore tends towards zero; tends, that is, to reduce collector and emitter currents. This offsets the increase in leakage current, and in the resulting state of equilibrium the collector current is greater than it was initially but is less than in an unstabilised circuit. In practice, variations in base current cause the base voltage to vary slightly and complete stabilisation of the collector current cannot be achieved. Nevertheless, the circuit is capable of reducing variations in collector current to say 1/15th of their value in an unstabilised circuit. This is illustrated in the following analysis of the circuit.

From expression (2) we have

$$I_c = \alpha' I_b + I_{c0}' \qquad \ldots.(5)$$

But from Fig. 6.6 (a)

$$I_b = I_1 - I_2.$$

Substituting for I_b in (5)

$$I_c = \alpha' I_1 - \alpha' I_2 + I_{c0}' \qquad \ldots.(6)$$

Moreover

$$V = I_1 R_1 + I_2 R_2 \qquad \ldots.(7)$$

Eliminating I_1 between (6) and (7) by multiplying (6) by R_1, (7) by α' and subtracting, we have

$$\alpha' V - I_c R_1 = \alpha' I_2 (R_1 + R_2) - I_{c0}' R_1$$

$$\therefore \quad \alpha' V - R_1(I_c - I_{c0}') = \alpha' I_2 (R_1 + R_2) \qquad \ldots.(8)$$

The voltage drop between base and emitter is very small. Thus

$$I_e R_e = I_2 R_2$$

giving

$$I_2 = \frac{I_e R_e}{R_2}$$

But $I_e = I_b + I_c$

$$\therefore\ I_2 = \frac{(I_b + I_c)R_e}{R_2} \qquad \ldots.(9)$$

Substituting for I_2 in (8)

$$\alpha' V - R_1(I_c - I_{c0}') = \frac{\alpha' R_e(R_1 + R_2)}{R_2} \cdot (I_b + I_c)$$

$$= \frac{R_e(R_1 + R_2)}{R_2} \cdot \alpha' I_b + \frac{\alpha' R_e(R_1 + R_2)}{R_2} \cdot I_c$$

From (5) $\alpha' I_b = I_c - I_{c0}'$

$$\therefore\ \alpha' V - R_1(I_c - I_{c0}') = \frac{R_e(R_1 + R_2)}{R_2} \cdot (I_c - I_{c0}') + \frac{\alpha' R_e(R_1 + R_2)}{R_2} \cdot I_c$$

Dividing by R_1 and rearranging

$$I_c\left[1 + \frac{R_e(R_1 + R_2)}{R_1 R_2} + \frac{\alpha' R_e(R_1 + R_2)}{R_1 R_2}\right] = I_{c0}'\left[1 + \frac{R_e(R_1 + R_2)}{R_1 R_2}\right] + \frac{\alpha' V}{R_1}$$

Let $R_b = R_1 R_2/(R_1 + R_2)$. R_b is thus the parallel resistance of R_1 and R_2.

$$I_c\left[1 + (\alpha' + 1) \cdot \frac{R_e}{R_b}\right] = I_{c0}'\left(1 + \frac{R_e}{R_b}\right) + \frac{\alpha' V}{R_1}$$

The stability factor K is obtained by differentiating this expression with respect to I_{c0}'.

$$K = \frac{dI_c}{dI_{c0}'} = \frac{1 + \dfrac{R_e}{R_b}}{1 + (1 + \alpha') \cdot \dfrac{R_e}{R_b}}$$

$$= \frac{R_e + R_b}{R_e + R_b + \alpha' R_e}$$

$$= \frac{1}{1 + \alpha' R_e/(R_b + R_e)} \qquad \ldots.(10)$$

an expression similar to that for the previous circuit, with R_e in place of R_c. In both circuits R_b is the resistance in the external base circuit.

To obtain good stability K must be small and this requires a high value of α', a high value of R_e and a small value of R_b. In the circuit previously described the stability similarly depended on R_b and R_c but the values which could be used were to a large extent dependent on other design considerations. In this circuit R_b and R_e can be given any values necessary to achieve the desired stability within wide limits without significant effect on the performance of the circuit as an amplifier. Naturally there are some limitations on permissible values of R_b and R_e in the potential-divider circuit: for example, if R_b is made very small it makes the input resistance very small in an RC-coupled amplifier such as that illustrated in Fig. 6.6 (a). Moreover when R_b is small it is possible that the potential divider may take a very large current from the collector-supply voltage. This applies to the circuits of Figs. 6.6 (a) and 6.6 (b). As R_b is decreased towards zero, the circuit degenerates into a common-base type and, as shown in expression (10), K tends to $1/(1 + \alpha')$, which is its value for a common-base amplifier.

The stability factor also improves as R_e is increased, but if this is made too large it will limit the collector current which can be achieved, unless a large value of collector supply voltage is used. At the other extreme, when R_e approaches zero and if R_b is high, the circuit becomes a simple common-emitter type with its particularly poor d.c. stability.

Design of a Potential Divider Circuit

Let us assume R_e is to be 1 kΩ: this is a convenient value because, for an emitter current of 1 mA, it loses only 1 volt of the collector supply voltage, leaving in a typical circuit with 6 volts supply, 5 volts for the transistor and its load resistor. If $\alpha' = 50$, the base current is 1/50 mA, i.e. 20 μA. This flows through R_1 in addition to the bleed current which flows through R_1 and R_2 from the collector supply. For good d.c. stability the potential at the junction of R_1 and R_2 must be steady in spite of variations in base current and this is achieved by making the parallel resistance of R_1 and R_2 small: this implies that the bleed current must be large compared with the base current. The bleed current can therefore be 200 μA which is ten times the base current but is still only one-fifth of the collector current. Thus the total current in R_1 is 220 μA. The voltage across R_1 is 5 volts because the potential at the junction of R_1 and R_2 does not differ appreciably from the emitter potential. Hence R_1 is given by

$$R_1 = \frac{5}{220 \times 10^{-6}} \text{ ohms}$$

$$= 23 \text{ k}\Omega \text{ approximately}$$

The voltage drop across R_2 is 1 volt and the current in it is 200 μA, giving the value of R_2 as 5 kΩ. C_1 should have a reactance small enough to avoid negative feedback and consequent fall in gain even at the lowest operating frequency. To achieve this the reactance must be small compared with the internal emitter resistance of the transistor: a typical resistance value is 25 ohms. In an a.f. amplifier C_1 may be 500 μF which has a reactance of 6·5 ohms at 50 c/s.

For this circuit R_b is given by

$$R_b = \frac{R_1 R_2}{R_1 + R_2}$$

$$= \frac{5{,}000 \times 23{,}000}{28{,}000} \text{ ohms}$$

$$= 4{\cdot}1 \text{ k}\Omega$$

The stability factor is thus given by

$$K = \frac{1}{1 + \dfrac{\alpha' R_e}{R_b + R_e}}$$

$$= \frac{1}{1 + \dfrac{50 \times 1{,}000}{4{,}100 + 1{,}000}}$$

$$= \frac{1}{1 + \dfrac{50 \times 1{,}000}{5{,}100}}$$

$$= \frac{1}{1 + 10} \text{ approximately}$$

$$= \frac{1}{11} \text{ approximately}$$

a considerable improvement over the stability factor obtainable from the circuit described before. This value of stability factor is, in fact, only 4·5 times worse than that for the common-base amplifier (for which K is $1/(1 + \alpha')$ i.e. $1/51$).

In a common-emitter amplifier so stabilised when temperature increases the collector current increase is approximately 1/10th that which would occur if the amplifier were unstabilised. To improve stability still further a means is required of making the potential at the junction of R_1 and R_2 slightly more positive as temperature rises. One method of achieving this is to connect in parallel with R_2 a resistor with a value which decreases as temperature rises. There are a number of different types of temperature-sensitive resistors having such a negative coefficient: one example is the thermistor. The coefficient required for good stabilisation can be obtained by choosing the correct ratio of R_2 to thermistor resistance. Alternatively the stabilisation can be improved by using for R_e a resistor with a positive temperature coefficient: an example of such a circuit is given on p. 92.

Use of a Two-battery Supply

The potential-divider circuit reduces variations in collector current by stabilising the base potential. However the base current flows through R_1 and variations in base current must cause some variations in base potential and hence in collector current, although these can be made slight. Thus the circuit is not completely successful although, as we have shown, it can give a considerable improvement in d.c. stability.

Instead of using a potential divider to stabilise the base potential of a common-emitter amplifier, a battery may be used instead. In pnp transistors the base must be biased negatively with respect to the emitter and it is thus unnecessary to use a separate battery for this purpose; the base can be returned to a tapping point on the collector supply battery as indicated in Fig. 6.7. This method of stabilisation is shown applied to an RC-coupled common-emitter amplifier at (a) and to a transformer-coupled amplifier at (b).

In an RC-coupled circuit it is not possible to connect the base directly to the tapping point without short-circuiting the amplifier input. A base resistor R_b must be included, and this necessarily degrades the d.c. stability by permitting variations in base potential. R_b should therefore be given the lowest value consistent with reasonable gain from the previous stage. Better stability (equal in fact to that of a common-base amplifier) is obtainable from the transformer-coupled circuit because no base resistor is necessary. The only external resistance in the base circuit is that of the transformer secondary winding, and this can be very small, usually less than 100 ohms.

The two-battery circuit is very simple to design. If it is required to stabilise the collector current at, say, 0·5 mA, and if the tapping

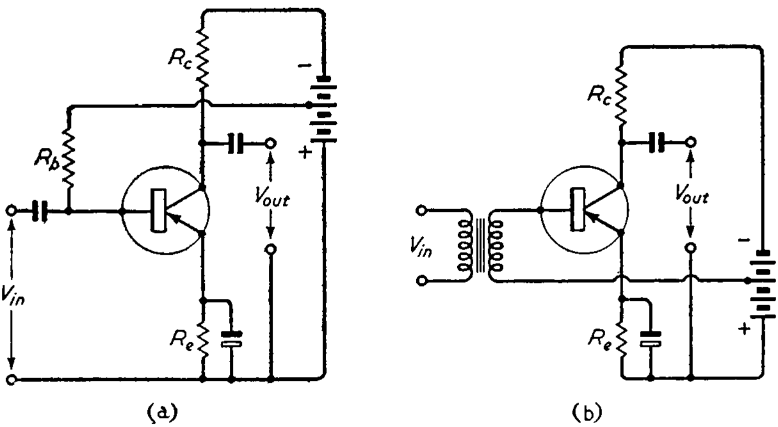


Fig. 6.7. Two-battery method of stabilising the d.c. conditions in a common-emitter amplifier (a) in an RC-coupled amplifier and (b) in a transformer-coupled amplifier

point is chosen to give a base-emitter voltage of $-1{\cdot}5$, the emitter resistance R_e is given by $1{\cdot}5/(0{\cdot}5 \times 10^{-3})$, i.e. 3 kΩ. This circuit does, however, require a slightly larger battery than other circuits. The stabilising voltage is, in effect, subtracted from the collector supply voltage, and a larger battery is needed to maintain adequate voltage across the transistor and its load resistance.

Use of a Temperature-dependent Emitter Resistor

The variations in collector current so far discussed are due to changes in leakage current: as explained these variations set an upper limit to the temperature at which a transistor can give a satisfactory performance. These variations occur even in the absence of the useful component αI_e, i.e. even if the transistor is cut off.

The useful component αI_e is also strongly dependent on temperature and, if temperature falls low enough, can become so small that the required collector current swing is impossible to achieve.

This variation thus sets a lower limit to the temperature at which satisfactory operation can be obtained. The variations in leakage current and useful component together define the temperature range within which satisfactory operation of the transistor is possible.

Both components of the collector current of a transistor thus vary with temperature but the total change in collector current can be reduced by using the methods of d.c. stabilisation discussed above.

In a power transistor designed for collector dissipations of several watts, the collector current may be of the order of 0·5 ampere. This is large compared with the leakage current and variations in collector current are almost entirely due to changes in the useful component. The mean value of the useful component can be kept constant by adjustment of the base-emitter voltage and a change of approximately 2·5 mV per °C is required by silicon and germanium transistors. Thus the stabilising circuits for power transistors should be designed to apply a correction of this value to the base-emitter voltage.

A simple method of effecting this compensation is to use an external emitter resistance of pure metal. Such resistors have a positive temperature coefficient and, if the base potential is constant, a rise in temperature causes the external emitter resistance to increase, thus increasing the voltage across this resistance. This in turn reduces the emitter-base voltage and thus tends to maintain the collector current constant.

The temperature coefficient of electrical resistance of copper is approximately 0·004 per °C: if the emitter current is assumed constant the voltage across a copper emitter resistance therefore increases by 0·004 of its initial value per °C. If the initial voltage is unity, the increase in emitter potential is 4 mV per °C. To offset a 2·5 mV change in emitter-base voltage, an initial emitter voltage of 2·5/4, i.e., approximately 0·6 volts, is needed. If the emitter current is 0·5 A (as is likely in a transistor of 5 W dissipation) the emitter resistance should be 0·6/0·5, i.e. 1·2 ohms, a convenient value to construct of copper wire. The fixed base potential is usually achieved by use of a resistive potential divider as discussed above.

7

SMALL-SIGNAL A.F. AMPLIFIERS

Introduction

WE have so far discussed the properties of simple amplifiers containing only a single transistor. Sometimes the power output and the gain of a single transistor are adequate but where greater gain is required, it is necessary to use a number of transistors connected in cascade to form a multi-stage amplifier.

Some multi-stage amplifiers are required to deliver an output voltage or output current which is small compared with the maximum that the final stage could deliver: usually this output is used as the input voltage or current for a following amplifier. In the design of amplifiers of this type the aim is to minimise distortion in the output voltage or current and the power output of the final stage is of little consequence. Such amplifiers are termed small-signal amplifiers, voltage amplifiers or current amplifiers: typical examples are microphone head amplifiers and the r.f. and early i.f. amplifiers in a receiver.

The final stage of a multi-stage amplifier may, however, be required to drive a loudspeaker or a recording head or some other load requiring appreciable power for its operation. Such stages must deliver undistorted power and their design principles differ from those of small-signal amplifiers. Amplifiers with a final stage of this type are termed large-signal or power amplifiers. The early stages of a large-signal amplifier can, of course, be regarded as constituting a small-signal amplifier.

In this chapter we shall consider the basic principles of small-signal a.f. amplifiers using junction transistors. We shall assume common-emitter operation throughout and all the circuits require protective measures to stabilise d.c. operating conditions.

Current and Voltage Amplifiers

It is usual in the design of valve amplifiers to regard each class-A stage as a voltage amplifier, i.e., as a stage whose input signal is the output voltage of the previous stage and whose output signal applied to the following stage is a voltage which is a substantially faithful copy of the input voltage. If a stage is to be regarded as a voltage amplifier, it is essential that the voltage of the signal source should not be affected by the connection of the amplifier across it: this requires that the input resistance of the amplifier should be very high compared with the resistance of the signal source. The output voltage of the amplifier should be substantially unaffected by variations in the value of the load resistance: this requires that the output resistance should be low compared with the load resistance. Thus a voltage amplifier must have a high input resistance and a low output resistance as shown in Fig. 7.1 (a). A class-A amplifying valve satisfies this requirement because the input resistance is nearly

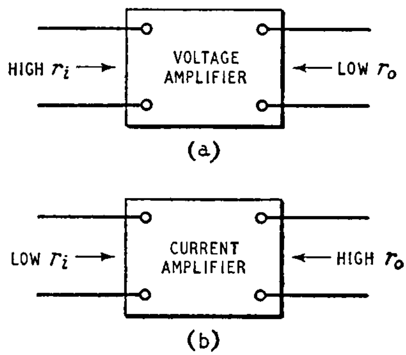


Fig. 7.1. Fundamental properties of (a) a voltage amplifier and (b) a current amplifier

infinite, whereas the output resistance can be less than 1 kΩ. Provided these resistance requirements are satisfied the gains of the individual stages of a voltage amplifier can be multiplied together to give the overall gain of the amplifier or (and this is another way of expressing the same fact) the gains of individual stages, when expressed in decibels, can simply be added to give the overall gain of the amplifier.

Similarly it is possible to have a current amplifier. When such an amplifier is connected to a signal source it is essential that the current flowing in the signal source should not be affected by the connection of the amplifier: this requires that the input resistance of the amplifier should be small compared with the resistance of the signal source. The output current from a current amplifier should ideally be independent of the load resistance: this requires that the output resistance should be high compared with the load resistance. Thus a current amplifier must have a low input resistance and a high output resistance as shown in Fig. 7.1 (b). Provided these resistance requirements are satisfied, the gains of the individual stages of a current amplifier can be multiplied together to give the overall gain of the amplifier or (and this is another way of expressing the same fact) the gains of the individual stages, when expressed in decibels, can simply be added to give the overall current gain of the amplifier.

As we have seen in previous chapters, a common-base or a common-emitter transistor amplifier stage has a small input resistance and a high output resistance. Both types of amplifier are therefore best regarded as current amplifiers. The common-collector amplifier, on the other hand, has a high input resistance and a low output resistance: this is best regarded, therefore, as a voltage amplifier.

The majority of transistor amplifiers are of the common-emitter type and the current gain of a particular stage is usually computed from the input (base-emitter) circuit to the input (base-emitter) circuit of the following stage. If, as commonly occurs, both stages include similar types of transistor operating under similar d.c. conditions, these input circuits are likely to have equal input resistances. If this is so then the current gain from one input circuit to the next is numerically equal to the voltage gain from one input circuit to the next. Provided, therefore, the input circuits are of equal resistance, it is possible to calculate the gain in terms of voltage or current as desired. This point is illustrated in some of the calculations given later in this chapter.

RC-coupled a.f. Amplifiers

Two successive stages of a small-signal a.f. amplifier are illustrated in Fig. 7.2. RC-coupling is used and both stages are stabilised by the potential-divider method. Earlier chapters give the method of calculating the values of the emitter resistor, collector resistor and the potential-divider resistors.

The input resistance of a common-emitter transistor stage is commonly approximately 1 kΩ, but this is reduced by the arms of the potential divider which are effectively in parallel with it and we can take the net input resistance as 800 ohms. This is in parallel with the collector load resistance of the previous stage and if this is taken as 3·3 kΩ the effective load resistance is thus

$$\frac{3{,}300 \times 800}{3{,}300 + 800} = 640 \text{ ohms approximately}$$

The gain of an amplifier with such a value of collector load resistance can be calculated in the following manner. Assume a collector signal current of i_c in the transistor: the signal voltage across the load resistance is thus i_cR_c. If the current gain is α' the base current is i_c/α'. The input resistance of the transistor

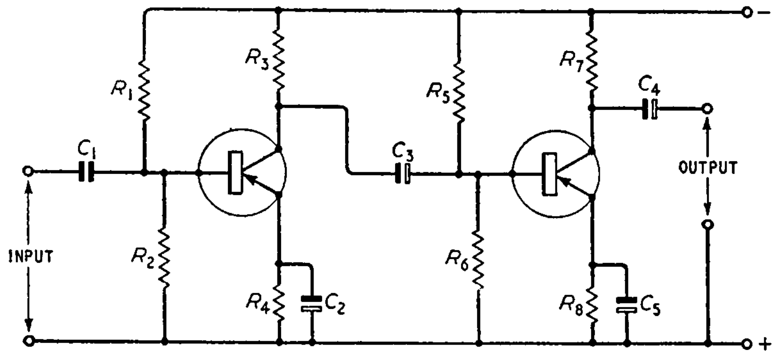


Fig. 7.2. A small-signal a.f. amplifier consisting of two RC-coupled common-emitter stages

is r_i and the input voltage which must be applied across r_i to produce a base current of i_c/α' is i_cr_i/α'. The voltage gain is thus given by

$$\frac{V_{out}}{V_{in}} = \frac{\alpha' i_c R_c}{i_c r_i}$$

$$= \frac{\alpha' R_c}{r_i}$$

Substituting $\alpha' = 50$, $R_c = 640$ ohms and $r_i = 1$ kΩ

$$\frac{V_{out}}{V_{in}} = \frac{50 \times 640}{1{,}000}$$

$$= 32$$

This calculation shows that the input resistance of the second stage is low compared with the load resistor of the first stage and it virtually determines the voltage gain of the first stage. The value of the collector load resistance for the first stage does not greatly affect its voltage gain.

If both stages of the amplifier of Fig. 7.2 have a voltage gain of 32 the overall voltage gain is 32^2, i.e., approximately 1,000.

The gain could alternatively be calculated from the mutual conductance of the transistor, i.e. the ratio of the signal-current output to the input voltage applied between base and emitter.

The mutual conductance of a transistor is directly proportional to the emitter current and inversely proportional to the absolute temperature. At 25°C it is given approximately by the expression

$$g_m = \frac{I_e}{V_e} \text{ where } V_e = 25 \text{ mV}$$

Thus for an emitter current of 1 mA the mutual conductance is 40 mA/V and for a load resistance of 640 ohms the voltage gain is 25·6, less than the value deduced on the previous page.

The gain of the amplifier can be written

$$A = g_m R_c$$
$$= \frac{I_e R_c}{V_e}$$

But, as shown on page 97, the gain is also equal to $\alpha' R_c/r_{in}$. Equating these two expressions for the gain we have

$$\frac{\alpha'}{r_{in}} = \frac{I_e}{V_e}$$

giving

$$r_{in} = \frac{\alpha' V_e}{I_e}$$

This is a useful expression because it shows that if $\alpha' = 50$ the emitter current must be 1·25 mA to give an input resistance of 1 kΩ. For 1·25 mA emitter current g_m is 50 mA/V and the gain for a load resistance of 640 ohms is 32, which agrees with the result deduced on the previous page.

Value of Coupling Capacitor

The capacitance C_3 is determined by the low-frequency limit of the pass-band, the response being 3 dB down at the frequency for which the capacitive reactance equals the resistance effectively in parallel with C_3. In this amplifier the resistance is, say, 800 ohms from the input of the following transistor and 3·3 kΩ from the output circuit of the preceding transistor. If the low-frequency limit is 50 c/s we have

$$\frac{1}{2\pi f C_3} = r$$
$$\therefore\ C_3 = \frac{1}{2\pi f r}$$
$$= \frac{1}{6{\cdot}284 \times 50 \times 4{,}100} \text{ F}$$
$$= 0{\cdot}8\ \mu\text{F approximately}$$

Small-signal Transformer-coupled a.f. Amplifier

A circuit giving greater gain is that shown in Fig. 7.3 in which a transformer is used to couple the collector circuit of the first transistor to the base circuit of the second. For maximum gain the turns ratio of the transformer should be chosen to match the output resistance of the first transistor (commonly approximately

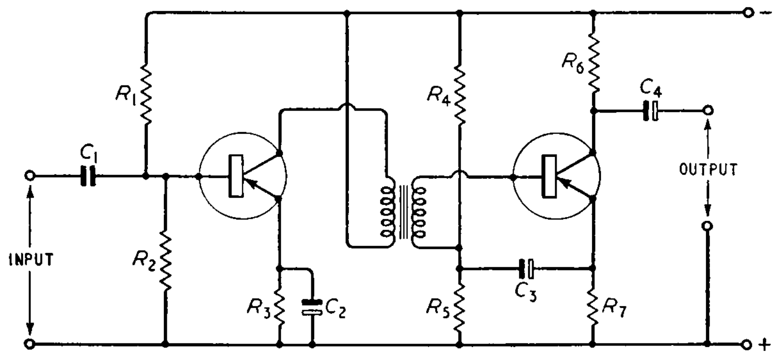


Fig. 7.3. A small-signal a.f. amplifier consisting of two transformer-coupled common-emitter stages

30 kΩ) to the input resistance of the second (1 kΩ). The turns ratio required is thus $\sqrt{}(30{,}000/1{,}000$: 1, i.e. 5·5 : 1.

As explained on page 145 perfect matching gives distortion because variations in input resistance affect the input current. Lower distortion (and lower gain) can be obtained by using a smaller turns ratio, thus tending to drive constant current into the following transistor.

The primary is, of course, the larger of the two windings and this is connected to the collector circuit so as to step down the voltage applied to the following base. There is, of course, an equal step up in current and as transistors are current-operated devices the transformer gives gain greater than is obtainable from the RC-coupled amplifier.

We can confirm this by calculation in the following manner. When the turns ratio is correct, the collector load presented to the first transistor by the transformer primary winding is 30 kΩ. This is equal to the output resistance of the first transistor, and the gain is thus half that which would be obtained if the output resistance were very large*. Thus the voltage gain from base to collector of the first transistor is equal to that of a high-impedance source

* A similar calculation for an r.f. amplifier is given on page 110.

feeding a load resistance of 30 kΩ/2, i.e. 15 kΩ, and is given by

$$g_m R_l = 50 \times 10^{-3} \times 15 \times 10^3$$
$$= 750$$

in which the mutual conductance of the transistor is taken as 50 mA/V. This gain is reduced in the ratio 5·5 : 1 by the transformer and hence the voltage gain from the base of the first transistor to the base of the second is given by

$$\frac{750}{5{\cdot}5} = 135 \text{ approximately}$$

This is more than 4 times the gain from the RC-coupled amplifier.

Calculation of Transformer Primary Inductance

The primary inductance of the transformer is determined by the lowest frequency it is desired to amplify. There is a loss of 3 dB at the frequency for which the inductive reactance is equal to the effective resistance of the collector circuit: this resistance is composed of the output resistance of the transistor and the load resistance at the transformer primary. Normally these two resistances are each equal to 30 kΩ and the effective resistance is 15 kΩ. If a loss of 3 dB at 50 c/s is acceptable, the primary inductance required can be calculated thus

$$2\pi f L = 15{,}000$$
$$\therefore\ L = \frac{15{,}000}{2\pi f}$$
$$= \frac{15{,}000}{6{\cdot}284 \times 50} \text{ H}$$
$$= 50 \text{ H approximately}$$

NEGATIVE FEEDBACK IN SMALL-SIGNAL AMPLIFIERS

Negative feedback is employed in transistor amplifiers for a number of reasons: it reduces distortion and makes the performance of the amplifier less dependent on transistor parameters and hence less dependent on temperature. In addition feedback can be used to extend the frequency response and to give the amplifier desired values of input or output resistance.

There are two basic feedback circuits and they are illustrated in Fig. 7.4. In circuit (a) feedback is applied by connecting a resistor R_b between the collector and the base. This circuit is also used to improve the d.c. stability of the common-emitter amplifier.

In the second basic circuit illustrated at (b) feedback is applied by inclusion of an undecoupled emitter resistance R_e.

Circuit (a) has the effect of reducing the input resistance of the amplifier and may be used as the input stage of a current amplifier where the input resistance must be small compared with the resistance of the signal source. This form of feedback decreases

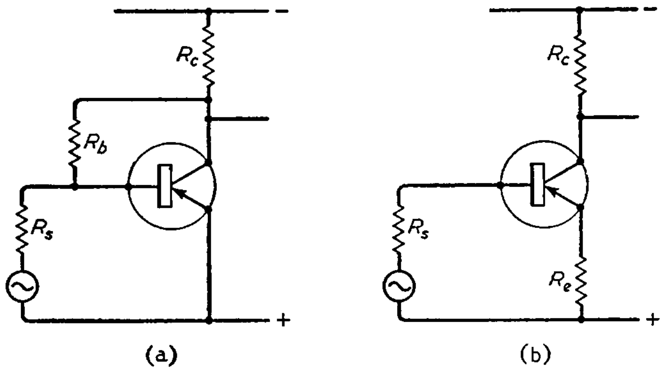


Fig. 7.4. Essential connections for two methods of applying negative feedback to common-emitter transistor amplifiers

the output resistance of the amplifier and is useful where the output is required in the form of a voltage and the output resistance must be small compared with that of the following load. This circuit is therefore likely to be used with a current input and a voltage output, and the gain measured in these terms can be evaluated in the following way.

If the current flowing into the base is i_b, the current in the collector circuit is $\alpha' i_b$ and this, in flowing through the collector load resistance R_c, gives rise to a collector voltage of $\alpha' i_b R_c$. This, in turn, gives rise to a current of $\alpha' i_b R_c/R_b$ in the resistor R_b. The total input current i_{in} required is thus given by

$$i_{in} = i_b + \frac{\alpha' i_b R_c}{R_b}$$

$$= i_b(1 + \alpha' R_c/R_b)$$

If R_b is removed, the input current is, of course, simply i_b and the effect of feedback is thus to reduce the input resistance to

$1/(1 + \alpha' R_c/R_b)$ of its former value. If R_b is small compared with $\alpha' R_c$ the input resistance is $R_b / \alpha' R_c$ of its former value. The gain* of the circuit is given by

$$\frac{\text{output voltage}}{\text{input current}} = \frac{\alpha' i_b R_c}{i_b(1 + \alpha' R_c/R_b)}$$

If R_b is small compared with $\alpha' R_c$ this reduces to

$$\frac{\text{output voltage}}{\text{input current}} = R_b$$

which is independent of the transistor parameters and of R_c.

Circuit (b) has the effect of increasing the input resistance of the amplifier and is used as the input stage of a voltage amplifier. This form of feedback increases the output resistance of the amplifier and is useful when the output is required in the form of a current and the output resistance must be large compared with the value of the following load resistance. Such a circuit is likely to be used with a voltage input and a current output, and its gain, measured in these terms, can be evaluated in the following manner.

If the signal current flowing into the base is i_b, the current in the collector circuit is $\alpha' i_b$. The current in the emitter circuit is $(\alpha' + 1)i_b$ and this, in flowing through the emitter resistor R_e, gives rise to an emitter voltage of $(\alpha' + 1)i_b R_e$. Provided R_e is large enough this voltage is large compared with the base-emitter voltage and the input signal voltage required is approximately $(\alpha' + 1)i_b R_e$, giving the input resistance as $(\alpha' + 1)R_e$. The gain† is given by

$$\frac{\text{output current}}{\text{input voltage}} = \frac{\alpha' i_b}{(\alpha' + 1)i_b R_e}$$

As α' is normally large compared with unity, $(\alpha' + 1)$ is approximately equal to α' and we have

$$\frac{\text{output current}}{\text{input voltage}} = \frac{1}{R_e}$$

which is independent of the transistor parameters and of the load resistance R_c.

Two-stage Current Amplifier

A transistor stage of the type illustrated in Fig. 7.4 (a) gives a voltage output for a current input, whereas a stage of the type shown

* Strictly mutual resistance. † Strictly mutual conductance.

in Fig. 7.4 (b) gives a current output for a voltage input. By combining a stage of each type it is possible to produce a two-stage current amplifier or a two-stage voltage amplifier. For example by combining a first stage of type (a) with a second stage of type (b) we obtain an amplifier with a low input resistance and a high output resistance, i.e. a current amplifier.

In such a combination of the circuits of Figs. 7.4 (a) and (b) the feedback resistor R_b bridges the base and collector of the first transistor. It is, however, more usual in two-stage current amplifiers for R_b to be connected between the base of the first transistor and the emitter of the second as shown in Fig. 7.5 which also includes

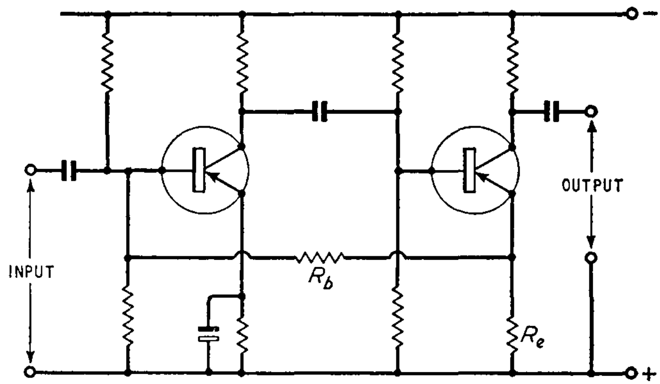


Fig. 7.5. A two-stage transistor current amplifier with low input resistance and high output resistance

components for d.c. stabilisation. The two circuits are strictly equivalent because the signal voltage at the collector of the first transistor is equal to that at the base of the second transistor which is, in turn, equal to that at the emitter of the second transistor. Thus the same feedback is obtained whether R_b is returned to the collector (as in Fig. 7.4 (a)) or to the emitter (as in Fig. 7.5). It is, in fact, preferable to use the circuit of Fig. 7.5 because the feedback then reduces any distortion arising in the inter-transistor coupling network.

A signal current i_{in} applied to the input of the amplifier of Fig. 7.5 gives a signal voltage of $i_{in}R_b$ at the collector of the first stage. The output resistance of this stage is low and the input resistance of the second stage is high. Hence this voltage is not reduced by the addition of the second stage and becomes the input for this stage. The (output current)/(input voltage) ratio for the second stage is equal to $1/R_e$ and the signal current output of the amplifier is thus $i_{in}R_b/R_e$. The overall current gain is thus given by R_b/R_e and is independent of the transistor parameters.

Two-stage Voltage Amplifier

By combining a first stage of the type shown in Fig. 7.4 (b) with a second of type Fig. 7.4 (a) we obtain an amplifier with a high input resistance and a low output resistance, i.e., a voltage amplifier. The circuit is given in Fig. 7.6 with components for d.c. stabilisation. As shown in Fig. 7.6 it is usual in two-stage amplifiers for

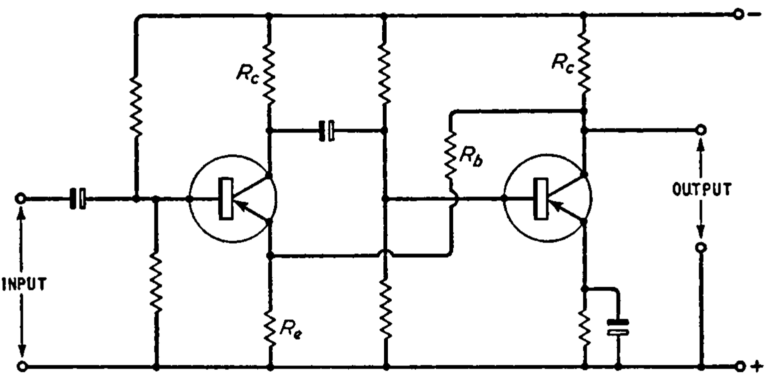


Fig. 7.6. A two-stage transistor voltage amplifier having high input resistance and low output resistance

the feedback resistor R_b to be returned to the emitter of the first stage rather than to the base of the second. This modification does not affect the action of the feedback because R_b provides a feedback *current* which must be connected in parallel with a signal current of the correct phase. Now the signal current in the base circuit of the second transistor is the collector current of the first transistor and also flows in R_e. Thus it does not matter whether R_b is returned to the base of the second stage or the emitter of the first. The second alternative is generally preferred because the feedback is then able to reduce any distortion occurring in the inter-transistor coupling network.

A signal voltage V_{in} applied to the input of the amplifier illustrated in Fig. 7.6 gives a current output of V_{in}/R_e from the first stage. The output resistance of this stage is high and the input resistance of the second stage is low: hence this current is not reduced by the connection of the second stage and is the input for the second stage. The (output voltage)/(input current) ratio for the second stage is equal to R_b and the output voltage is $V_{in}R_b/R_e$. The overall voltage gain is hence R_b/R_e. This is the same expression as for the current gain of the amplifier previously described and is independent of the transistor parameters and of the value of the collector load resistors.

As a numerical example consider a voltage amplifier of the type illustrated in Fig. 7.6. We have shown before that a suitable value for R_c is 3·3 kΩ and we will take α' as 50 for both transistors. $\alpha' R_c$ is thus 165 kΩ and R_b must be small compared with this for the second transistor to have adequate feedback. A suitable value for R_b is 20 kΩ. R_e should have a value such that the signal voltage across R_e is large compared with that across the base-emitter junction. If we take the resistance of this junction as r_i, a voltage V_{in} across it gives rise to an input current of V_{in}/r_i which, in turn, gives an emitter current of approximately $\alpha' V_{in}/r_i$ and a voltage across the emitter resistor of $\alpha' V_{in}R_e/r_i$. The ratio of the voltage across R_e to the base-emitter voltage is thus $\alpha' R_e/r_i$, i.e., $g_m R_e$, and this should be large compared with unity. If g_m is 50 mA/V, R_e can be 100 ohms, giving $g_m R_e$ as 5. The overall voltage gain of the amplifier is given by R_b/R_e, i.e. 20,000/100, which is equal to 200.

8

I.F. AMPLIFIERS

Introduction

WE have so far confined the discussion of transistor amplifiers to those used at low frequencies, particularly audio frequencies. In this frequency range transit-time effects and the effects of internal capacitances within the transistor are in general negligible. At higher frequencies, and in particular at radio frequencies, these effects are significant and cause the gain to fall below the values obtainable at lower frequencies.

The effects of transit time within the transistor were discussed in Chapter 2 where it was mentioned that this effect can be allowed for by assuming that the current amplification factor falls with frequency according to the expression

$$\alpha = \frac{\alpha_0}{\sqrt{(1 + f^2/f_\alpha{}^2)}}$$

in which f_α is the alpha cut-off frequency.

Of the various internal capacitances within a junction transistor, that between the collector and the base has the greatest effect on the high-frequency performance. In practice the effects of this capacitance are often more important than the fall in current amplification factor.

In a transistor r.f. amplifier the capacitance between collector and base provides feedback from the output to the input circuit. This is illustrated in Fig. 8.1 (a) in which the capacitance c_{bc} is shown connected directly between collector and base terminals. However, a better approximation to the performance of a transistor at high frequencies is obtained by assuming that the collector capacitance is returned to a tapping point b' on the base resistance as shown in Fig. 8.1 (b).

This modifies the feedback which now occurs via $c_{b'c}$ and $r_{bb'}$ in series and the high-frequency performance of the transistor is now dependent on the time constant $r_{bb'}\ c_{b'c}$ which is probably the most important characteristic of a transistor intended for high-frequency use.

The effect of this internal feedback connection is similar to that of the anode-grid capacitance of a triode valve; that is to say, if a resonant circuit is included in the base circuit (as is likely in an r.f. or i.f. amplifier) $r_{bb'}$ $c_{b'c}$ give positive feedback at frequencies

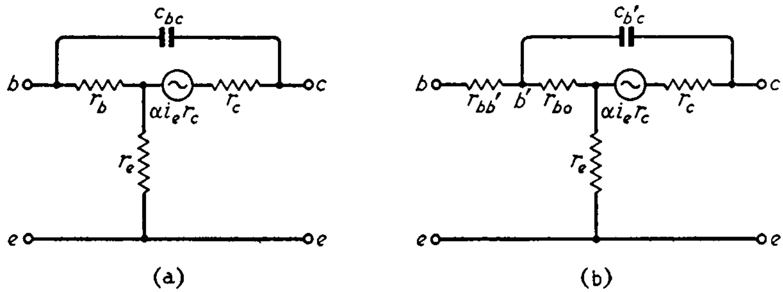


Fig. 8.1. T-section equivalent circuit of a transistor showing collector capacitance, (a) returned to base input terminal and (b) returned to a tapping point on the base resistance

to one side of the resonance value and negative feedback at frequencies to the other side. In a triode valve the positive feedback is usually sufficient to cause oscillation but in a transistor amplifier this does not always occur because of the low input resistance. However, the change in the nature of the feedback around the resonance frequency makes the response curve of the amplifier asymmetrical and it is therefore desirable to include neutralising components in order to minimise the effects of the internal feedback.

Transistor r.f. and i.f. amplifiers can be connected in common-base or in common-emitter modes. Chapters 3 and 4 show that the common-emitter amplifier has higher power gain than the common-base type. This advantage of the common-emitter amplifier holds for all frequencies below the v.h.f. range. Maximum gain can only be achieved, of course, by perfect matching between the output circuit of one transistor and the input circuit of the succeeding one.

Such matching is difficult, particularly at radio frequencies, when there is a marked difference in value between the input and output resistances as for a common-base amplifier. The problem is not so acute for a common-emitter amplifier, for which the difference between the two resistances is not so great, and in practice this type of amplifier usually gives 10 dB or more power gain than the common-base type.

Neutralising is not difficult at frequencies that are well below the alpha cut-off value but becomes progressively more difficult as the cut-off frequency is approached. It is usually advisable,

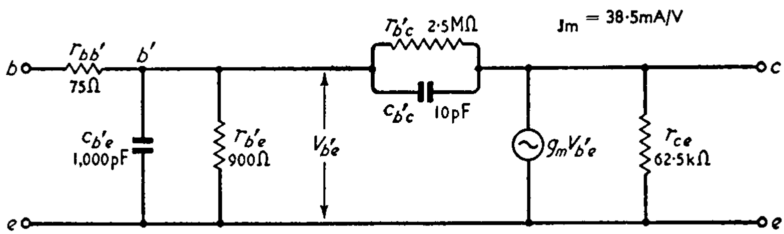


Fig. 8.2. A π-section equivalent circuit of a uniform-base junction transistor

therefore, that the operating frequency should not be above $f_\alpha/10$ or at worst $f_\alpha/5$: for example, the transistors used in 465 kc/s i.f. amplifiers generally have alpha cut-off frequencies of not less than 5 Mc/s.

The T-section equivalent circuits of Fig. 8.1 are not well suited for calculating the performance of an r.f. or i.f. amplifier which uses resonant circuits as coupling elements and the equivalent π-section is preferred. This is illustrated in Fig. 8.2 in which $g_m V_{b'e}$

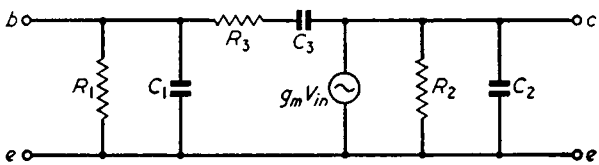


Fig. 8.3. At a single frequency the π-section equivalent circuit of Fig. 8.2 can be reduced to this form

represents a constant-current generator. Also indicated on Fig. 8.2 are typical capacitances and resistances for a uniform-base junction transistor. This diagram may be used at any radio frequency but it is possible to simplify it to the form shown in Fig. 8.3 when operation is confined to a very small frequency range as in a transistor i.f. amplifier. This is a particularly useful equivalent circuit because it shows that the signal source is connected to a parallel RC combination $R_1 C_1$ and the output load is connected to the parallel RC combination $R_2 C_2$. As the source and load impedances are both likely to contain resonant circuits the values of R_1 and R_2 are important because they damp these circuits, reducing the working Q value and degrading selectivity. If C_1 and C_2 were constant they would be comparatively unimportant because any effect they had on the tuned circuit could be offset by an adjustment of the inductance or the capacitance of the tuned circuits.

Unfortunately C_1 and C_2 both depend on the d.c. operating conditions of the transistor and vary with changes in these conditions such as those produced by a.g.c. action. Their values and variations are thus important.

Unilateralisation

R_3 and C_3 are the components which provide the internal feedback path between collector and base. As already pointed out, this feedback is undesirable because it causes asymmetry in the amplifier

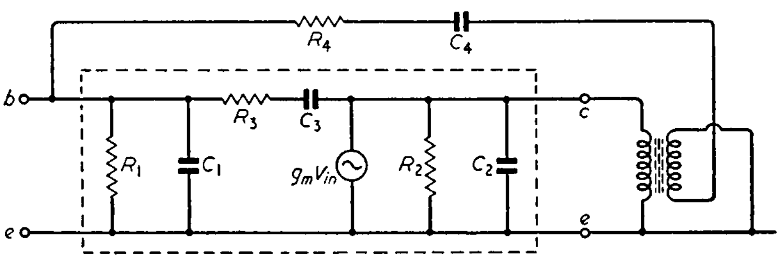


Fig. 8.4. Method of unilateralising a junction transistor by means of a phase-inverting transformer and components R_4 *and* C_4

response curve or possibly instability. It also causes interaction between the input and output tuned circuits and can make alignment of the amplifier difficult.

To avoid these effects, feedback due to R_3 and C_3 must be eliminated and the usual method is to apply an equal degree of feedback of opposite phase to the amplifier by means of an external RC network. From a point where the signal voltage is in antiphase to that at the collector a series RC network is returned to the base as shown in Fig. 8.4. If this signal voltage is equal to the collector voltage, the values of R_4 and C_4 required in the external feedback loop are equal to R_3 and C_3 (1,800 ohms and 10 pF are typical values for a uniform-base transistor). In a practical circuit, as shown later, the antiphase voltage is usually smaller, say $1/n$th of the collector voltage, and the external feedback components should then be R_3/n and nC_3 for equality of external and internal feedback.

When the effects of R_3 and C_3 are thus eliminated, the input and output circuits of the transistor are completely divorced and the equivalent circuit reduces to the simple form shown in Fig. 8.5.

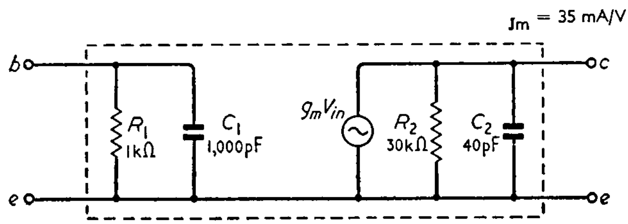


Fig. 8.5. For perfect unilateralisation the equivalent circuit for a uniform-base transistor reduces to this form

The transistor now is truly a " one-way " device in which the output circuit has no effect on the input circuit: for this reason the transistor is described as *unilateralised.* Unilateralisation is essential when it is required to obtain without instability the maximum gain of which the transistor is capable.

It is possible to reduce (though not eliminate) the effects of R_3 and C_3 by omitting R_4 and using a capacitor only in the external feedback circuit: this technique is known as *neutralisation.*

Maximum Gain of r.f. Transistor

Typical values of R_1, R_2, C_1, C_2 and the constant-current generator for a uniform-base transistor operating at the intermediate frequency of 465 kc/s are indicated on Fig. 8.5. Not surprisingly, the input resistance is approximately 1 kΩ and the output resistance 30 kΩ: these are the values which apply to an a.f. amplifier. From these component values we can calculate the maximum gain in the following way. C_1 and C_2 are considered absorbed into the input and output tuned circuits and the gain is a maximum when the inter-stage coupling is designed to match the output resistance (R_2) of one stage to the input resistance (R_1) of the following stage. If a transformer is used its turns ratio must be $\sqrt{(30{,}000)/(1{,}000)} : 1 = \sqrt{30} : 1 = 5{\cdot}5 : 1$ as before. The constant-current generator then operates into a load of 15 kΩ and the voltage gain is given by

$$g_mR_l = \frac{35}{1{,}000} \times 15{,}000$$
$$= 525$$

The gain from the base of one transistor to the base of the following one is hence $525/5\cdot5 = 95$. This is the maximum gain which can be achieved and it is not reached in practice because, as explained later, the tuned circuits give an insertion loss.

If we repeat this calculation in general terms, taking the input resistance as r_i and the output resistance as r_o, we have for perfect matching that the constant-current generator operates into a resistance or $r_o/2$ and the gain, from base to collector, is $g_m r_o/2$. The matching transformer has, however, a step-down turns ratio of $\sqrt{(r_o/r_i)}$: 1 and the maximum gain (losses in the transformer being ignored) is given by

$$\text{maximum gain} = g_m \cdot \frac{r_o}{2}\sqrt{\left(\frac{r_i}{r_o}\right)}$$
$$= \frac{g_m\sqrt{(r_i r_o)}}{2}$$

The gain hence depends on the mutual conductance, the input resistance and the output resistance and can be improved by increase in any of these quantities. This expression gives the maximum gain from the base of the transistor to the base of the succeeding one. If the input resistances of both transistors are assumed equal, the maximum gain can be expressed in decibels thus

$$\text{maximum gain} = 20 \log_{10} \frac{g_m\sqrt{(r_i r_o)}}{2} \text{ dB}$$
$$= 10 \log_{10} \frac{g_m{}^2 r_i r_o}{4} \text{ dB}$$

For a uniform-base junction transistor having $g_m = 35$ mA/V, $r_i = 1$ kΩ and $r_o = 30$ kΩ the maximum gain is given by

$$\text{maximum gain} = 10 \log_{10} \frac{35^2 \times 10^{-6} \times 10^3 \times 30 \times 10^3}{4} \text{ dB}$$
$$= 40 \text{ dB nearly}$$

To realise this gain it is essential that the transistor should be unilateralised. For without unilateralisation there is interaction between the input and output circuits which has the effect of lowering both input and output resistances, thus lowering the gain as indicated in the above expression.

Design of Small-signal r.f. and i.f. Amplifiers

Because of the low input and output resistances of a common-emitter amplifier the design of r.f. and i.f. transformers for transistors is a different problem from that of designing transformers for valve amplifiers.

In valve amplifiers the tuned circuit can be designed without regard to the input and output resistances because these are so high that they impose negligible damping on the circuits. The performance of the transformers is thus independent of the properties

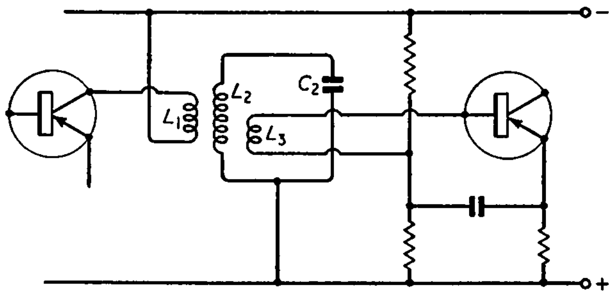


Fig. 8.6. Essential features of a transistor r.f. or i.f. amplifier: for simplicity neutralising components are omitted

of the valves preceding and following them. In a transistor amplifier, however, the performance of the transformer is almost completely determined by the properties of the transistor.

The design of a transistor r.f. or i.f. amplifier is, in fact, determined by two considerations:

(a) to secure maximum gain the inter-stage tuned circuit must match the feeding to the terminating resistance;

(b) to obtain the desired selectivity, the coupling between the two transistors and the tuned circuit must be such that the circuit has the required value of working Q.

Fig. 8.6 illustrates one circuit which can be used. The tuned circuit L_2C_2 is coupled to the collector circuit of the first transistor by the primary winding L_1 and to the base circuit of the second transistor by the tertiary winding L_3. We shall assume that all the windings are unity-coupled. The turns ratio between L_1 and

L_3 must match a generator of 30 kΩ to a load of 1 kΩ and must hence have a turns ratio of 5·5 : 1 as in transistor a.f. amplifiers. This satisfies requirement (a) above.

In mass production of receivers it may be desirable to depart from perfect matching to obtain satisfactory results in spite of spreads in transistor parameters and unilateralising components.

Insertion Loss

The damping due to the two transistors can be calculated separately but it is easier to combine their effects by omitting, say, the tertiary winding and to assume that the primary is connected to a 15-kΩ resistance, i.e. two 30-kΩ in parallel (one due to transistor 1 and the other due to transistor 2). We now have to calculate the turns ratio of L_1 to L_2 for this determines to what extent the tuned circuit is damped. If a working Q value of 100 is required, this could be obtained by using a circuit with an undamped Q of, say, 120 which is reduced to the required value by the damping due to the transistors: in this example the damping is very light. Alternatively, the tuned circuit could be designed to have a much higher undamped Q value—say 300—which is again damped by the transistors to the required value of 100: the damping in this example is very heavy. Both these alternatives yield a tuned circuit with a working Q value of 100 but the gain of the amplifier is not the same for both examples. We can show this in the following way.

Consider first the example of light damping. To reduce an undamped Q from 120 to 100, the resistance effectively transferred by the primary circuit to the tuned circuit must be five times the dynamic resistance. This also means, of course, that at the resonance frequency the tuned circuit effectively connects across the primary winding a resistance equal to one-fifth that due to transistor damping. This reduces the effective load resistance to one-sixth of its value before the tuned circuit was connected, thus reducing the voltage gain to one-sixth. This represents an insertion loss of 15·6 dB and if, as assumed earlier, the maximum gain is 40dB the gain realised in practice is only 24·4 dB.

Now consider the example of heavy damping. To reduce the Q to one-third, the resistance effectively transferred by the primary circuit to the tuned circuit must be one-half the dynamic resistance. This also means that at the resonance frequency the tuned circuit effectively connects across the primary winding a resistance equal to twice that due to transistor damping. This reduces the effective

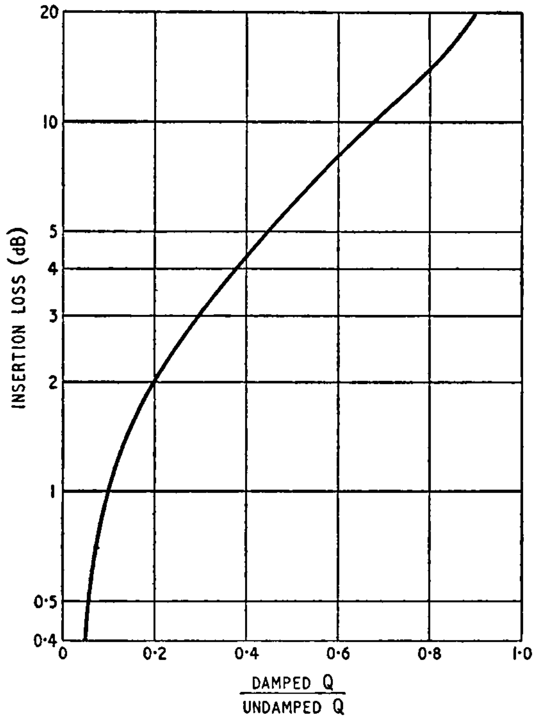


Fig. 8.7. Dependence of insertion loss on the ratio of damped to undamped Q values

load resistance to two-thirds the value before the tuned circuit was connected, thus reducing the voltage gain in the same ratio. The insertion loss due to the tuned circuit is now only 3·5 dB and a practical gain of 36·5 is possible, if the theoretical maximum gain is 40 dB. To avoid a great insertion loss, therefore, the damping of the tuned circuit must be heavy. On the other hand, with a great reduction in Q there is a danger of large and unwanted variations in Q and hence in selectivity if for any reason (such as a change in a.g.c. voltage) the damping due to either transistor alters. In the compromise solution sometimes adopted, the reduction in Q due to transistor damping is 2 : 1, giving an insertion loss of 6 dB. For transistors with the parameters assumed above the voltage gain is then approximately 45, i.e., 34 dB.

If the above calculation is repeated in general terms a curve can be prepared which expresses the insertion loss in terms of the ratio of working Q value to undamped Q value. This can be shown in the following way.

Basically an inter-transistor coupling which includes an LC circuit can be regarded at resonance as two resistances connected in parallel and fed from a constant-current source. One resistance, which we will call R_1, is that due to the collector circuit of the transistor feeding the tuned circuit and the base circuit of the transistor following the tuned circuit. The second resistance, which we will call R_2, is the resistance effectively connected in parallel with R_1 by the tuned circuit. R_2 depends on the dynamic resistance of the tuned circuit and, if the tuned circuit is connected directly in the collector circuit, is equal to the dynamic resistance.

When the tuned circuit is connected, the load for the transistor becomes $R_1R_2/(R_1 + R_2)$. It was formerly R_1 and thus the insertion loss is given by

$$\text{insertion loss} = 20 \log_{10} \frac{R_1R_2}{R_1 + R_2} \cdot \frac{1}{R_1} \text{ dB}$$

$$= 20 \log_{10} \frac{R_2}{R_1 + R_2} \text{ dB}$$

The resistance R_2 of the tuned circuit is effectively reduced to $R_1R_2/(R_1 + R_2)$ when it is connected in circuit and thus the ratio of damped Q to undamped Q is given by

$$\frac{\text{damped } Q}{\text{undamped } Q} = \frac{R_1R_2}{R_1 + R_2} \cdot \frac{1}{R_2}$$

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

We thus have

$$\text{insertion loss} = 20 \log_{10} \frac{R_2}{R_1 + R_2}$$

$$= 20 \log_{10} \left(1 - \frac{R_1}{R_1 + R_2}\right)$$

$$= 20 \log_{10} \left(1 - \frac{\text{damped } Q}{\text{undamped } Q}\right)$$

and this is the expression from which Fig. 8.7 was plotted.

Primary Capacitance

If the transistor damping is equivalent to a single resistance R_t across the primary winding, the turns ratio n : 1 required for a 2 : 1 reduction in Q is given by

$$n = \sqrt{\left(\frac{R_d}{R_t}\right)}$$

being a step-up ratio if the dynamic resistance R_d exceeds R_t as is usual in practice. The dynamic resistance is given by $QL\omega$ or $Q/\omega C$. Substituting $Q/\omega C$ for R_d

$$\therefore\ n = \frac{\sqrt{Q}}{\sqrt{(\omega CR_t)}}$$

from which

$$n^2 = \frac{Q}{\omega CR_t}$$

$$\therefore\ n^2C = \frac{Q}{\omega R_t}$$

Now n^2C is the effective capacitance across the primary winding and it is an important quantity because it must be large compared with the variations in transistor output capacitance which occur

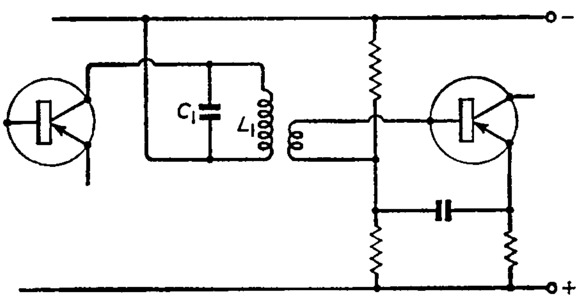


Fig. 8.8. Simple form of transistor r.f. or i.f. amplifier: neutralising components are again omitted

when the a.g.c. voltage changes. As shown in the equivalent circuit (Fig. 8.5), the output capacitance of a uniform-base r.f. transistor is of the order of 40 pF but it can change by as much as 20 pF due to changes in bias. To minimise mistuning errors due to this variation, the primary capacitance must be large. For example, if the primary capacitance is 1,000 pF, a change of 20 pF in output capacitance represents a 2 per cent variation in net capacitance and a 1 per cent variation in resonance frequency. At 465 kc/s this is equivalent to nearly 5 kc/s.

The above expression shows that the primary capacitance depends on Q, ω and R_t. Now, for a particular amplifier and transistor all these quantities are fixed and hence the primary capacitance is also fixed. If $R_t = 15$ kΩ, $Q = 200$ and $\omega = 2\pi \times 465$ kc/s we have

$$n^2C = \frac{200}{6{\cdot}284 \times 465 \times 10^3 \times 15 \times 10^3}\ \text{F}$$

$$= 4{,}560\ \text{pF}$$

For such a large value of effective primary capacitance the mistuning due to a 20 pF variation in transistor output capacitance will be of the order of only 1 kc/s, which is acceptable.

The primary capacitance is independent of the value of n and we can therefore make n any value we please. If we make n unity, the primary winding has the same number of turns as the secondary winding and there is thus no need for both windings. The tuning capacitance of 4,560 pF can be placed across the primary winding and the secondary winding can be omitted. This now reduces the inter-transistor transformer to the simple form shown in Fig. 8.8 which still, however, gives the required selectivity characteristic and transistor matching.

A value of C_1 (Fig. 8.8) as large as 4,560 cannot conveniently be made variable and in a circuit of this type, tuning is best achieved by adjusting the inductance L_1 by movement of a dust-iron or ferrite core. If it is considered essential to tune by means of a variable capacitor the circuit of Fig. 8.6 can be used, the value of n being so chosen that C_2 has a small capacitance such as 200 pF: a variable capacitor can then be used for C_2.

Unilateralisation

So far we have omitted components for unilateralisation in order to keep the circuits simple. To add these components we require a point on the transformer where the signal voltage is in antiphase to that at the collector. There is no need to add a winding specially for these components: the tertiary winding which feeds the base of the succeeding stage can be used, provided that the ends are chosen to give the required phase relationship. The circuit is shown in Fig. 8.9. The values of R_n and C_n are given by R_3/n and nC_3 (see Fig. 8.4 for R_3 and C_3) and the turns ratio between L_1 and L_3 is 5·5 : 1 for matching purposes. Thus for a transistor having internal feedback equivalent to 1·8 kΩ and 10 pF we have

$$\begin{aligned} R_n &= \frac{R_3}{n} \\ &= \frac{1{,}800}{5 \cdot 5} \text{ ohms} \\ &= 330 \text{ ohms} \\ C_n &= nC_3 \\ &= 5 \cdot 5 \times 10 \text{ pF} \\ &= 55 \text{ pF} \end{aligned}$$

Use of Drift Transistors as r.f. or i.f. Amplifiers

It is mentioned in Chapter 2 that drift transistors have a better performance at high frequencies than uniform-base transistors and can, in fact, be satisfactorily used at v.h.f. The adoption of the

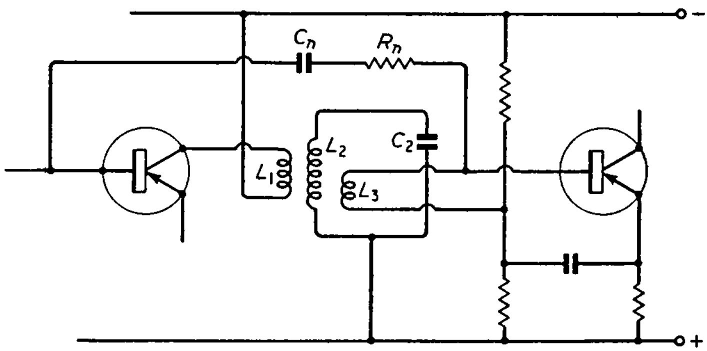


Fig. 8.9. Complete circuit of common-emitter transistor r.f. or i.f. amplifier including unilateralising components R_n and C_n

graded-base impurity content makes the output resistance of the drift transistor considerably higher than for the uniform-base type. For example at 465 kc/s a typical value for the output resistance of a drift transistor is 0·5 MΩ compared with 30 kΩ for the uniform-base transistor. The input resistance and the mutual conductance of the drift transistor do not differ greatly from the values for a uniform-base transistor. The increased output resistance results in a significant improvement in gain as illustrated by the expression for the maximum gain:

$$\text{maximum gain} = 10 \log_{10} \frac{g_m{}^2 r_i r_o}{4}$$

Substituting the numerical values for a drift transistor

$$\text{maximum gain} = 10 \log_{10} \frac{(35 \times 10^{-3})^2 \times 10^3 \times 0{\cdot}5 \times 10^6}{4}$$

$$= 52 \text{ dB nearly}$$

This represents an improvement of nearly 12 dB over the maximum gain available from a uniform-base transistor although, of course, the maximum gain cannot be realised in practice because of the insertion loss of the tuned circuits which must be used.

The capacitances of a drift transistor are less than those of the uniform-base type. Typical values are:

input capacitance = 100 pF

output capacitance = 5 pF

internal feedback capacitance = 1·5 pF

Use of Drift Transistors at 465 kc/s

The small value of the internal feedback capacitance in the drift transistor means that there is less danger of instability or of asymmetry of the frequency response curve than for a uniform-base transistor and, in fact, provided gain well below the theoretical maximum is accepted it is possible to dispense with unilateralisation or neutralising altogether both at 465 kc/s and at 10·7 Mc/s. However, if very high gain is essential unilateralisation is advisable. The circuit illustrated in Fig. 8.9 can then be used, the values of R_n and C_n being calculated as previously explained. For a drift transistor typical values for the internal feedback resistance and capacitance are 10 kΩ and 1·5 pF. It is not usual to attempt to match the very high output resistance to the input resistance of the following stage and transformer ratios of 7 : 1 are common. For this ratio the values of the unilateralising components are 1·4 kΩ and 10·5 pF. Usually, however, a greater value of capacitance is used to allow for the effects of stray capacitance in the wiring to the transistor which inevitably increases the effective value of the feed-back capacitance. The gain obtainable from such a stage of amplification at 465 kc/s is commonly 10 dB greater than is possible from a uniform-base transistor.

Use of Drift Transistors at 10·7 Mc/s

The introduction of drift transistors has made possible the construction of transistorised v.h.f. receivers, e.g., f.m. receivers for use in Band II (87·5 to 100 Mc/s in this country and 87·5 to 108 Mc/s in U.S.A.). Such receivers are usually superheterodyne types and require an i.f. amplifier with a bandwidth of approximately 200 kc/s (for transmissions with a deviation of 75 kc/s) centred on the standard intermediate frequency of 10·7 Mc/s.

Drift transistors can be used in such an i.f. amplifier but their characteristics at this frequency differ markedly from those which apply at the lower intermediate frequency of 465 kc/s. The following are typical values of the significant parameters at 10·7 Mc/s:

input resistance = 330 ohms

input capacitance = 65 pF

output resistance = 17 kΩ

output capacitance = 5 pF

internal feedback capacitance = 1·5 pF

mutual conductance = 30 mA/V

The input resistance, output resistance and mutual conductance are less than at 465 kc/s and limit the maximum gain at 10·7 Mc/s as indicated in the following calculation

$$\text{maximum gain} = 10 \log_{10} \frac{g_m{}^2 r_i r_o}{4}$$

$$= 10 \log_{10} \frac{(30 \times 10^{-3})^2 \times 330 \times 17 \times 10^3}{4}$$

$$= 31 \text{ dB}$$

The value of input resistance quoted above (330 ohms) applies when the output circuit of the transistor is short-circuited. If the transistor were unilateralised this would also be the input resistance of a practical amplifier. Commonly, however, drift transistors are not unilateralised or are deliberately mismatched and conditions in the output circuit affect the input circuit. For normal values of collector load the input resistance is less than for a short-circuited output circuit and a value of 150 ohms may be taken as typical at 10·7 Mc/s.

Similarly the value of the output resistance quoted above (17 kΩ) applies when the input circuit of the transistor is short-circuited. If unilateralisation is not used the output resistance for normal values of source resistance may be taken as 6kΩ at 10·7 Mc/s.

Substitution of these amended values of input and output resistance gives the maximum gain for non-unilateralised drift transistors at 10·7 Mc/s as

$$\begin{aligned}\text{maximum gain} &= 10 \log_{10} \frac{g_m{}^2 r_i r_o}{4} \\ &= 10 \log_{10} \frac{(30 \times 10^{-3})^2 \times 150 \times 6 \times 10^3}{4} \\ &= 23 \text{ dB}\end{aligned}$$

which is 8 dB less than the value for the unilateralised transistor. This is not a very high gain and practical gains must inevitably be lower because of the insertion losses of the tuned circuits employed. The insertion loss is therefore kept to a minimum by so designing the i.f. amplifier that the tuned circuits are very heavily damped by the input and output resistances of the transistor. It is possible in this way to reduce the insertion losses to 3 dB per stage thus giving a stage gain of 20 dB.

Very heavy damping of the primary and secondary tuned circuits is advocated in a 10·7 Mc/s amplifier to achieve good stability. It was not advised in a 465 kc/s i.f. amplifier because the variation in damping caused by a.g.c. action can upset the frequency characteristic. It is not usual to employ a.g.c. in an f.m. receiver and variations in damping are therefore unlikely to occur.

It is possible to construct a 10·7 Mc/s i.f. amplifier using a succession of single tuned circuits as described for the 465 kc/s i.f. amplifier earlier but, because of the wide bandwidth required in an f.m. receiver, it is preferable to use double-tuned transformers with the mutual inductance between primary and secondary windings slightly greater than optimum. In this way a good approximation to the ideal square-topped frequency response curve can be obtained.

The heavy damping of the primary winding can be achieved by connecting it directly in the collector circuit as shown in Fig. 8.10

and by arranging that the dynamic resistance of the undamped winding be very large compared with the 6-kΩ output resistance to which it is connected. The effective dynamic resistance of the damped winding is, of course, slightly less than 6 kΩ but is taken as 6 kΩ in the following approximate calculation of the inductance

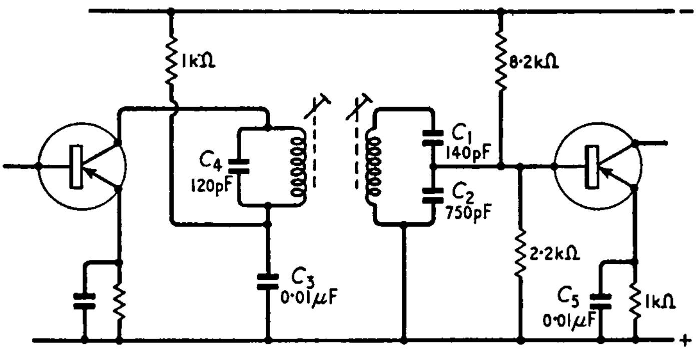


Fig. 8.10. A stage of a 10·7 Mc/s i.f. amplifier using drift transistors and double tuned transformers

required. A suitable value for the working Q is 50. Thus, for the primary winding:

$$QL\omega = R_d$$

$$L = \frac{R_d}{Q\omega}$$

$$= \frac{6 \times 10^3}{50 \times 6{\cdot}284 \times 10{\cdot}7 \times 10^6}\ \mathrm{H}$$

$$= 1{\cdot}8\ \mu\mathrm{H}$$

and for resonance at 10·7 Mc/s the tuning capacitance must be 120 pF.

To obtain a satisfactory i.f. transformer it is advisable to have an identical secondary winding also with an inductance of 1·8 μH and a tuning capacitance of 120 pF. This must be damped by the input resistance of the following transistor to give the desired working Q of 50 and also to match the output resistance of the previous transistor to this input. The input resistance of the transistor is 150 ohms and to obtain the equivalent of 6 kΩ connected across the whole of the secondary winding the transistor can be connected to a tapping point on the inductor, the turns ratio of the whole secondary inductor to the tapped part being given by

$$\sqrt{(6 \times 10^3/150)} : 1 = \sqrt{40} : 1$$
$$= 6{\cdot}3 : 1$$

Alternatively, as suggested in the circuit diagram of Fig. 8.10 the transistor can be tapped down the capacitive branch of the secondary circuit. The ratio of the two capacitances to give the required damping should be (6·3 — 1) : 1 i.e. 5·3 : 1. We thus require two capacitors for the secondary circuit whose ratio is 5·3 : 1 and which give a net capacitance of 120 pF. Their values can be calculated in the following way:

$$C_2 = 5{\cdot}3C_1$$
$$\frac{C_1C_2}{C_1 + C_2} = 120 \text{ pF}$$

Substituting for C_2 in the second equation

$$\frac{5{\cdot}3C_1^2}{5{\cdot}3C_1 + C_1} = 120 \text{ pF}$$
$$\frac{5{\cdot}3C_1}{6{\cdot}3} = 120 \text{ pF}$$
$$\therefore C_1 = 143 \text{ pF}$$
$$\therefore C_2 = 5{\cdot}3 \times 143 \text{ pF}$$
$$= 756 \text{ pF}$$

These values and also other practical values of components required in a 10·7 Mc/s i.f. amplifier stage are indicated in Fig. 8.10. A number of such stages are necessary to give the gain and selectivity required in an f.m. receiver and a description of a complete receiver using stages of this type is given in Chapter 11.

It was shown earlier that the gain of a 10·7 Mc/s i.f. stage which feeds into the base circuit of a similar stage is approximately 20 dB. This is the gain obtainable from each of the i.f. stages in a receiver with the exception of the last which operates into the detector. The input resistance of a ratio detector can easily be 10 kΩ and such a value permits gains approaching 100 from the previous amplifier. To show this we can use the general expression

$$\text{maximum gain} = 10 \log_{10} \frac{g_m{}^2 r_i r_o}{4}$$

Substituting $g_m = 30$ mA/V, $r_i = 10$ kΩ, $r_o = 6$ kΩ which apply when the previous stage is a drift transistor we have

$$\text{maximum gain} = 10 \log_{10} \frac{(30 \times 10^{-3})^2 \times 10^4 \times 6 \times 10^3}{4} \text{ dB}$$

$$= 41{\cdot}3 \text{ dB}$$

The insertion loss of the tuned circuits will, of course, reduce this figure and a more practical value is probably 36 dB corresponding to a voltage gain of 65.

Decoupling

In Fig. 8.10, as in most of the circuit diagrams in this book, the source of base input signal (i.e. the secondary winding of the i.f. transformer) and the emitter decoupling capacitor are both returned to the positive terminal of the battery supply. This minimises the impedance of the external base-emitter circuit which

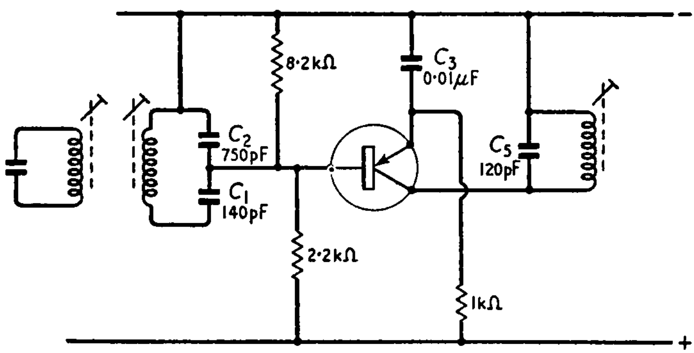


Fig. 8.11. A modified version of the wiring of Fig. 8.10 which eliminates the need for collector decoupling components

is essential for maximum performance. The primary winding of the i.f. transformer must be returned to the negative terminal of the supply to provide the necessary collector bias but should be returned to the positive terminal of the supply in order to minimise the impedance of the external collector-emitter circuit which is also essential for best results. This difficulty is usually overcome by connecting a low-reactance capacitor C_3 (the collector decoupling capacitor) between the primary winding and the emitter (or the positive terminal of the supply which is in turn connected to the emitter via the low-reactance emitter decoupling capacitor). If the collector decoupling capacitor is omitted, the signal output current of the transistor must flow through the impedance of the battery in addition to the primary winding of the transformer. In this way signal voltages are developed across the battery and these, if impressed upon other stages of the amplifier or receiver of which this stage is part, can distort the shape of the frequency response curve or even cause instability. To avoid this the collector decoupling capacitor C_3 is included to short-circuit the battery at signal frequencies.

However, by a simple alteration to the wiring such a capacitor becomes unnecessary. If the source of base input signal (i.e., the secondary winding) and the emitter decoupling capacitor are both returned to the negative terminal of the battery supply, as shown in Fig. 8.11, then the impedance of the external base-emitter circuit and of the external collector-emitter circuit are both minimised simultaneously and no additional decoupling capacitor is necessary.

9

PULSE AMPLIFIERS

Introduction

AMPLIFIERS such as oscilloscope Y-amplifiers and video amplifiers are required to handle signals which may have steep, almost vertical edges and may also have long, almost horizontal sections. A signal which has both features is a rectangular pulse and is commonly used in tests of Y-amplifiers and video amplifiers: in fact the amplifiers are usually known as pulse amplifiers.

The ability of an amplifier to reproduce rapid changes in a signal waveform such as a steep edge is determined by the high-frequency response of the amplifier: in fact such amplifiers must have a response good up to the frequency given by

$$f = \frac{1}{2t}$$

where t is the rise time of the steepest edge. If the rise time is $0{\cdot}1$ μsec the upper frequency limit is given by

$$\begin{aligned} f &= \frac{1}{2 \times 0{\cdot}1 \times 10^{-6}} \text{ c/s} \\ &= 5 \text{ Mc/s} \end{aligned}$$

The ability of an amplifier to reproduce very slow changes in a signal waveform such as an almost horizontal section is determined by the low-frequency response of the amplifier: the longer the section of the waveform, the better must be the low-frequency response of the amplifier to reproduce it without distortion. As a numerical example, an amplifier required to reproduce a 50 c/s square wave with less than 2 per cent sag in the horizontal sections must have a low-frequency response which is good down to at least 1 c/s. Sometimes in fact, pulse amplifiers are d.c. coupled to extend the low-frequency response down to zero frequency.

To summarise the above, we may say that pulse amplifiers are characterised by an extremely wide frequency response: for a video amplifier suitable for the British television standards the useful frequency response extends from very low frequencies up to 3 Mc/s. A statement, such as this, of the steady-state amplitude response of the amplifier does not, however, give complete information about its performance as a pulse amplifier. Pulse signals in general contain a large number of components and it is essential for distortionless reproduction of such signals that the components after amplification should reach the output with the same phase relationship that they had at the input to the amplifier. The shunt capacitance which is inevitable in any amplifier causes the phase of high-frequency components of a complex signal such as a pulse to lag behind that of low-frequency components. Such a lag would be comparatively unimportant in an a.f. amplifier but can seriously degrade the performance of a pulse amplifier by increasing the rise time. Phase response is thus important in pulse amplifiers and for a good performance both the amplitude and the phase response must satisfy certain stringent requirements over the spectrum occupied by the signals to be amplified.

For amplifiers using simple inter-transistor coupling networks such as are likely to be used for pulse amplification there is a mathematical relationship between the amplitude and the phase response and, given one, it is possible to deduce the other. In general the better the amplitude response the better is the phase response also and it is thus possible to ensure an adequate phase response by making the amplitude response of the amplifier sufficiently good. When a pulse amplifier is designed in this way the amplitude response must satisfy standards far more exacting than if phase response were also under consideration. For example to design a video amplifier with a passband extending to 3 Mc/s it may be necessary to make the amplitude response of the amplifier flat within 0·1 dB up to, say, 20 Mc/s.

α' Cut-off frequency

Most transistor pulse amplifiers can be regarded as simple common-emitter RC-coupled amplifiers which have been designed with special attention paid to the low-frequency and the high-frequency response so that the necessary wide bandwidth and phase response are achieved. Let us consider the factors which limit the performance of simple RC-coupled amplifiers at the high-frequency end of the spectrum. The two most important factors are shunt capacitances (chiefly the input and output capacitances of the transistors) and transit-time effects in the transistors (measured by the alpha cut-off frequency). Usually the second factor predominates and its effect on the frequency response of a common-emitter amplifier can be assessed in the following way.

The alpha cut-off frequency is, by definition, the frequency at which the current gain of a common-base amplifier with a short-circuited output circuit falls to 0·707 of its low-frequency value when the input circuit is supplied with constant current. For uniform-base transistors this frequency is unlikely to exceed 20 Mc/s and a typical value for transistors intended for r.f. applications is 7 Mc/s. Most amplifiers are, however, of the common-emitter type and for this circuit arrangement the cut-off frequency (for short-circuited output and constant-current input) is much lower. The way in which the common-emitter (or α') cut-off frequency depends on the common-base (or α) cut-off frequency can be deduced in the following way.

The current gain α_o' of a common-emitter amplifier at low frequencies is given by

$$\alpha_o' = \frac{\alpha_o}{1 - \alpha_o}$$

where α_o is the current amplification factor at low frequencies. In general the current gain is given by

$$\alpha' = \frac{\alpha}{1 - \alpha}$$

But α falls as frequency is raised according to the expression

$$\alpha = \frac{\alpha_o}{1 + \mathrm{j}f/f_\alpha}$$

where f_α is the alpha cut-off frequency. Substituting for α in the general expression for α' we have

$$\alpha' = \frac{\alpha_o/(1 + \mathrm{j}f/f_\alpha)}{1 - \alpha_o/(1 + \mathrm{j}f/f_\alpha)}$$

$$= \frac{\alpha_o}{1 - \alpha_o + \mathrm{j}f/f_\alpha}$$

$$= \frac{\alpha_o/(1 - \alpha_o)}{1 + \mathrm{j}f/(1 - \alpha_o)f_\alpha}$$

This may be written

$$\alpha' = \frac{\alpha_o'}{1 + jf/f_\alpha'}$$

where f_α' is the α' cut-off frequency. From this it follows that

$$f_\alpha' = (1 - \alpha_o)f_\alpha$$

A typical uniform-base transistor for r.f. applications with $\alpha_o = 0{\cdot}98$ and $f_\alpha = 7$ Mc/s has a value of f_α' given by

$$\begin{aligned} f_\alpha' &= (1 - \alpha_o)f_\alpha \\ &= (1 - 0{\cdot}98) \times 7 \text{ Mc/s} \\ &= 140 \text{ kc/s} \end{aligned}$$

A higher effective value of f_α' could be obtained, at the expense of gain, by the use of negative feedback but the performance of uniform-base alloy junction transistors is in general not good enough for most pulse amplifiers, and drift transistors are preferred.

The performance of a drift transistor as a pulse amplifier is better measured by the frequency f_1. As mentioned on page 35, this is the frequency at which the real part of the current amplification factor has fallen to one half its low-frequency value. This frequency is a most useful criterion of performance because it is also the frequency at which the current gain α' of the common-emitter amplifier has fallen to unity; it therefore gives the gain-bandwidth product for the common-emitter amplifier and a typical value for f_1 is 70 Mc/s. Thus at 35 Mc/s, $(f_1/2)$ α' is 2 and at 7 Mc/s α' is 10. If an α' of 35 is required, the highest working frequency of the transistor is 2 Mc/s.

In practice simple pulse amplifiers such as those implied in the preceding sentences are unlikely to be used. This is because the parameters of transistors of the same type inevitably have a spread and it is difficult to construct two or more amplifiers with a similar performance without careful selection of transistors. To minimise this difficulty negative feedback is used.

Use of Negative Feedback

This has a number of important advantages:

(a) It extends the passband.
(b) It makes the frequency response and the gain of the amplifier less dependent on the transistor parameters. In fact with

a great degree of feedback the gain and the frequency response are to a large extent determined by the constants of a passive network, i.e. the feedback loop. Thus the performance of the amplifier can be predetermined and it is possible to replace the transistors by others with slightly different parameters without significant effect on amplifier performance. It is also possible to manufacture a number of amplifiers all with the same performance within close limits.

(c) By choosing a suitable method of applying feedback the input and output resistances can be made high or low as desired. The precise connections used depend on whether a voltage or a current amplifier is required.

These advantages are not obtained without cost: the application of negative feedback reduces the gain of the amplifier. The way in which negative feedback extends frequency response at the expense of gain is illustrated in Fig. 9.1. The upper frequency

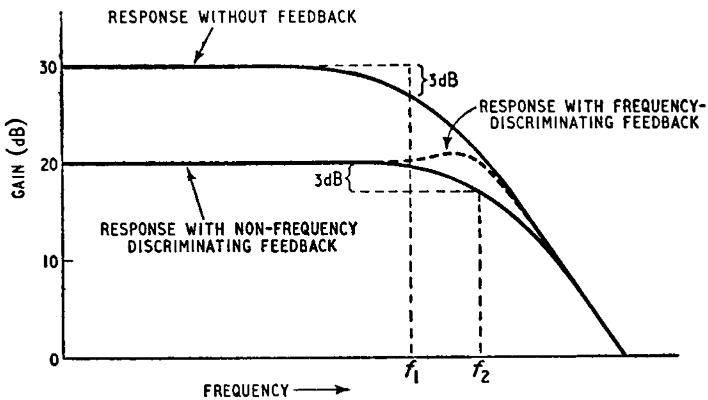


Fig. 9.1. Effect of negative feedback on the frequency response of an amplifier

response curve is that of an amplifier without feedback and is 3 dB down at the frequency f_1, which can be taken as the upper limit of the passband. When feedback is applied, the gain falls at all frequencies but the fall is greater at low that at high frequencies and the new 3-dB loss point is at f_2 representing a considerable extension of the passband. If the negative feedback circuit is so designed that the feedback voltage or current becomes less as frequency rises it is possible to achieve an even greater extension of the frequency response as suggested by the dotted lines. This, however, is usually accompanied by an increase in phase distortion and consequent deterioration in pulse response and it is probably better to adhere to the aperidic feedback represented by the solid curve.

Sometimes it happens that the degree of negative feedback required to give the desired passband is so great that the gain of the amplifier is reduced below the wanted value. It is then necessary to add one or more stages to the amplifier to achieve the required passband and gain. This is an illustration of the general theorem quoted above that the gain-bandwidth product for a particular amplifier is unaffected by feedback. If the effect of feedback is to increase the passband by a factor n, then the low-frequency gain of the amplifier is reduced by the factor n. Thus a gain reduction of 6 dB is accompanied by a doubling of the passband. In designing an amplifier to have a particular bandwidth and gain sufficient stages must be included to give at least this value of gain-bandwidth product.

If a considerable extension of passband is wanted, a considerable degree of feedback must be used. It is not easy to apply a considerable degree of negative feedback to an amplifier because there is a tendency to produce instability at frequencies near the extremes of the passband. This may take the form of oscillation or a sharp peak in the frequency response curve. Great care in design is essential to avoid these effects. One of the methods used to avoid such effects in a multi-stage amplifier is to employ a number of independent feedback loops, each embracing only two transistor stages of amplification. A similar technique is employed in pulse amplifiers using valves.

Some of the methods of applying negative feedback to two-stage voltage and current amplifiers have been described in Chapter 7 but their application to pulse amplification will now be described in more detail.

Use of Direct Coupling

To obtain a good performance from a pulse amplifier at low frequencies it is advantageous to eliminate coupling capacitors as far as possible because these inevitably introduce attenuation and phase shift at low frequencies. Although negative feedback can be used to reduce these effects, the phase shift limits the degree of feedback which can be employed without instability. Direct coupling is thus attractive not only because it provides a zero-frequency response but also because it eliminates phase shift and enables large degrees of feedback to be used without danger of instability. However, because of the phase shifts present near the upper limit of the passband it is usual to arrange that feedback loops embrace only two cascaded stages.

The use of direct coupling in a transistor amplifier introduces another difficulty. We have already seen in Chapter 6 that the leakage current in a common-emitter stage can vary considerably with change in ambient temperature and unless precautions are taken to stabilise d.c. conditions the current can alter by as much as 2 mA for a 30° C change in temperature. By use of stabilising circuits this current change can be reduced by a factor of, say, 10 to 0·2 mA. If such a current is fed directly to another transistor with a current gain of 50, the collector current of this second stage will vary by 10 mA. Such large variations are clearly impossible in a small-signal amplifier in which the second stage may require a mean current of only 1 mA.

If direct coupling is to be used, therefore, the amplifier must include some method of stabilising the no-signal collector current of the second transistor. One method is to limit the zero-frequency gain of the amplifier to a very low value such as 2 by means of direct-coupled feedback. If a larger value of signal frequency gain is required this means that the direct coupling is no longer of use in extending the frequency response to zero frequency although it still enables the coupling capacitor to be omitted and facilitates the application of signal-frequency feedback.

Current Pulse Amplifier

As an illustration of the application of these principles we will consider the design of a current amplifier to have a gain of 100 (40 dB). Two cascaded stages are all that is necessary because without feedback two typical common-emitter amplifiers could provide a current gain of 2,500 (i.e., 50 per transistor). If this is reduced to 100 by negative feedback, the gain reduction factor is 1/25 which represents 28 dB of feedback. The input resistance, output resistance, gain and frequency response of an amplifier with such a large degree of feedback would be almost independent of the transistor parameters except near the cut-off frequency.

The amplifier can take the form of a direct-coupled version of Fig. 7.5 in which feedback is applied via the resistors R_b and R_e. Provided the gain of the transistors is high enough the gain of such an amplifier is given by $(R_e + R_b)/R_e$ with little error. If R_b is large compared with R_e the gain is given by R_b/R_e. To give the required d.c. gain of 2, R_b must be equal to R_e. These resistors cannot be small otherwise the shunting effect of R_b on the input of the first transistor becomes serious. On the other hand R_e cannot be very large otherwise an abnormally-large supply voltage is necessary to give the required 1 mA in R_e. A suitable compromise

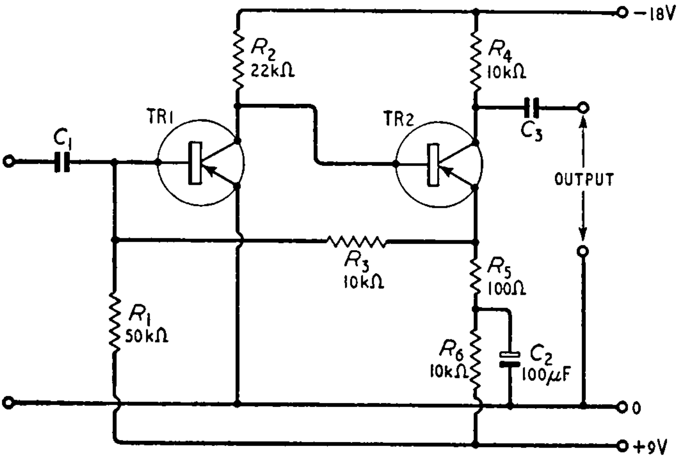


Fig. 9.2. A two-stage current amplifier with negative feedback

value for R_b and R_e is 10 kΩ. Even with this value of R_e a supply voltage exceeding 20 volts is required because the collector resistance for the second transistor (which is effectively the output resistance of the amplifier) can hardly be less than 10 kΩ and this also carries the 1 mA current of the second transistor.

To obtain the required zero-frequency gain R_b and R_e should both be 10 kΩ: to obtain the required signal-frequency gain R_b/R_e must equal 100 and if R_b is 10 kΩ, R_e must be 100 ohms. Both requirements are satisfied if R_e is composed of a 10-kΩ resistor and a 100-ohm resistor in series, the larger resistor being shunted by a large capacitor to eliminate signal-frequency feedback.

The circuit diagram of the current pulse amplifier thus designed is given in Fig. 9.2 in which all component values are shown. A 27-volt supply is used and the emitter of the first transistor is returned directly to a tapping point on this supply. Any resistance included at this point, whether decoupled or not, would reduce the effectiveness of the d.c. stabilising circuit.

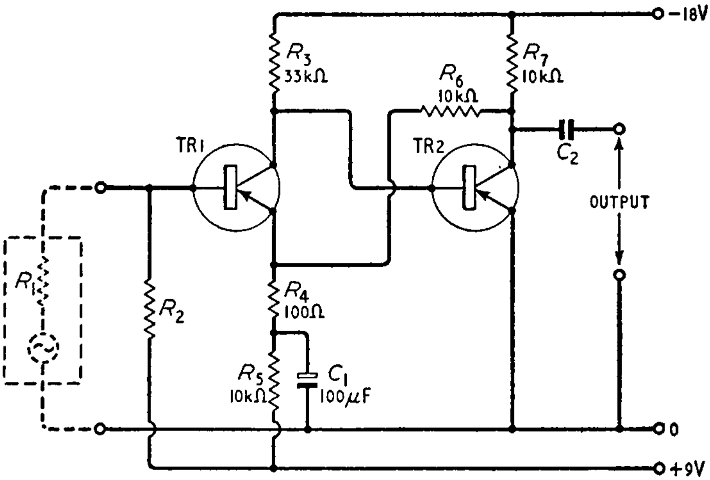


Fig. 9.3. A two-stage voltage amplifier with negative feedback

The amplifier gives the required gain of 100 (40 dB) up to a frequency of say $f_1/20$. By sacrificing gain, however, the response can be maintained up to a higher frequency: for example by making R_2 1 kΩ and R_5 200 ohms the gain is reduced to approximately 10 (20 dB) but is maintained up to nearly $f_1/2$. A further improvement in frequency response is possible by making the feedback frequency-dependent as suggested earlier: this is possible by including an inductor in series with R_3. The inductance should be so chosen that its reactance becomes appreciable compared with R_3 at a frequency of approximately $f_1/2$. The low-frequency response is determined by capacitor C_2 and can be extended as low as desired by making C_2 large enough.

The feedback circuit substantially increases the output resistance of the second transistor and the output impedance of the amplifier is nearly equal to R_4, i.e. 10 kΩ. The input resistance of the first transistor is reduced by the feedback to around 100 ohms. These are the values of input and output resistance over most of the amplifier passband and the ratio of input to output resistance is very small (1/100) as required in a current amplifier. As frequency approaches the upper limit of the passband the feedback becomes less effective causing the input resistance to rise and the output resistance to fall.

Voltage Pulse Amplifier

From the fundamental circuit for a voltage amplifier given in Fig. 7.6 it is possible to develop a pulse amplifier giving a gain of 40 dB over a wide frequency range. Direct coupling can be used between the transistors and this necessitates a d.c.-stabilising feedback circuit analogous to that used in the current amplifier described previously. The circuit diagram for such a voltage amplifier is given in Fig. 9.3.

The feedback components are R_4, R_5, R_6 and C_1 and these give a zero-frequency gain equal to $(R_4 + R_5 + R_6)/(R_4 + R_5)$, i.e. approximately 2 for the component values quoted on the diagram. The signal-frequency gain is equal to R_6/R_4 which is equal to the required value of 100. The high value of R_5 necessitates a 27-volt supply which must be tapped to provide an emitter potential for the second transistor. The tapping point is also used to provide the first transistor with a suitable base potential via a potential divider one arm of which is provided by the resistance of the signal source. To give the first transistor suitable operating conditions the signal source should have a resistance less than 1 kΩ.

The type of negative feedback used in this amplifier gives a high input resistance (approximately 100 kΩ) and a low output resistance (approximately 200 ohms) as required in a voltage amplifier.

The passband obtainable from such an amplifier is slightly better than that of the current amplifier.

10

LARGE-SIGNAL A.F. AMPLIFIERS

Introduction

THE small-signal amplifiers described in Chapter 7 are designed to produce a voltage or a current output. When a voltage output is wanted, the magnitude of the current output is usually of little consequence provided the final transistor can supply it without distortion. Similarly if the amplifier is designed to supply a current output, the magnitude of the voltage output is of secondary importance provided is is not sufficient to cause overloading and consequent distortion. In general the output, whether of voltage or of current, is small compared with the maximum that the transistor could supply and there is little danger of overloading.

In this chapter we shall discuss large-signal amplifiers and here the design principles are quite different. A power output is now required and the voltage swing and the current swing of the amplifier are both of importance. Moreover there is a very real danger of overloading and a major problem is how to obtain the maximum output from a given transistor without distortion.

The final stage of a large signal amplifier may operate in class A or class B and can give an efficiency much higher than is obtainable from a large-signal valve amplifier.

CLASS-A AMPLIFIERS

The circuit of a class-A output stage is given in Fig. 10.1. For maximum efficiency full advantage must be taken of the available collector-current and collector-voltage swing: that is to say, the load resistance and input-signal amplitude must be so chosen that the collector voltage and collector–current swing between zero and twice the quiescent (no-signal) value.

Suppose the battery voltage is V_b and the steady collector current in the absence of an input signal is I_c. Then the power taken from the battery is V_bI_c and most of this is dissipated as heat in the transistor. This static dissipation must not exceed the maximum value quoted by the manufacturers. If a sinusoidal input signal is applied to the transistor, power is supplied to the load but the power taken from the battery remains constant because no change

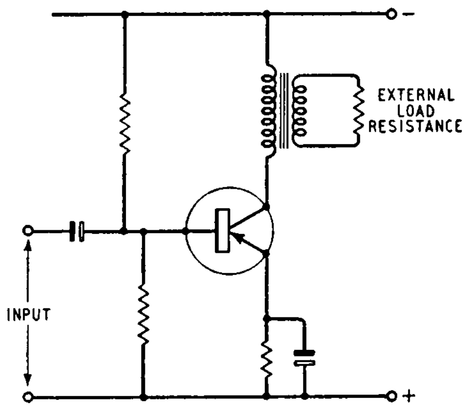


Fig. 10.1. Essential features of a single-ended class-A transistor output stage

has taken place in the average or d.c. component of the collector current.

It follows that the power dissipated in the transistor becomes less when the input signal is applied. If, therefore, the transistor does not become too hot in the absence of an input signal, there will be no danger at all of damaging it by heat when the signal is applied.

When the transistor is delivering its maximum undistorted power, the peak value of the collector voltage is nearly equal to V_b and the peak collector current is I_c. The output power is obtained by multiplying the r.m.s. collector voltage by the r.m.s. collector current: for a sinusoidal output signal these are $V_b/\sqrt{2}$ and $I_c/\sqrt{2}$ and thus

$$\text{maximum power output} = \frac{V_b}{\sqrt{2}} \cdot \frac{I_c}{\sqrt{2}}$$

$$= \frac{V_b I_c}{2}$$

The power taken from the battery is $V_b I_c$ and thus the efficiency is 50 per cent.

This is the theoretical maximum and in practice a transistor class-A stage can approach it very closely. This efficiency applies only for sinusoidal signals and when the transistor is driven to the limit of its output power. For smaller input signals the efficiency is less.

The amplitude of an a.f. signal varies over a range depending on the nature of the signal. For orchestral music the range between the maximum and minimum amplitudes may be as much as 40 dB. For jazz and speech the range is much less.

It can therefore be seen that the efficiency of a class-A amplifier with an a.f. input varies from instant to instant and the average efficiency is in actual practice much less than the theoretical maximum of 50 per cent.

As can be seen from Fig. 10.2 the slope of the load line for maximum power output is given by V_b/I_c: this is then the value of the optimum load resistance.

As a numerical example, consider a transistor rated for 100 mW maximum collector dissipation and operating as a class-A output stage from a 6-volt supply.

The maximum undistorted output power is 50 mW and the mean collector current I_c is given by

$$I_c \times 6 = 100 \times 10^{-3}$$

$$\therefore\ I_c = \frac{100 \times 10^{-3}}{6}\text{ A}$$

$$= 17\text{ mA}$$

$$R_l = \frac{V_b}{I_c}$$

$$= \frac{6}{17 \times 10^{-3}}\text{ ohms}$$

$$= 350\text{ ohms}$$

One practical point is that the battery voltage must exceed 6 volts to give a 6-volt swing of collector potential. This is because the steady voltage across the emitter resistor, the primary resistance and the transistor knee voltage must be subtracted from the battery voltage to give the effective collector-emitter voltage.

The output transformer must match the optimum load to the load resistance. Therefore, if the load resistance is 3 ohms,

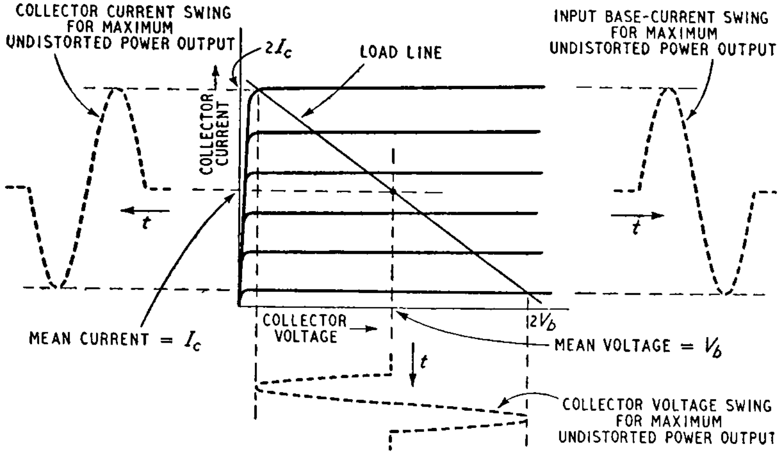


Fig. 10.2. Conditions in a class-A transistor output stage

the transformer must have a ratio of

$$\sqrt{(350/3)} : 1 = \sqrt{117} : 1$$
$$= 11 : 1, \text{ approximately}$$

The primary inductance determines the low-frequency response which is 3 dB down at the frequency for which the inductive reactance is equal to the optimum load. In an a.f. amplifier the 3 dB loss frequency may be 50 c/s and we have

$$2\pi fL = R$$

$$L = \frac{R}{2\pi f}$$

$$= \frac{350}{6{\cdot}28 \times 50}\text{ H}$$

$$= 1{\cdot}1\text{ H}$$

The transformer should have a primary inductance of this value with 17 mA direct current flowing.

Two transistors may be operated in class-A push-pull and a typical circuit is given in Fig. 10.3. The collector-to-collector optimum load is twice that for a single class-A transistor and the output transformer needs twice the primary inductance. Its design is simpler, however, because it is not polarised, the d.c. components of the collector currents flowing in opposite directions in the two halves of the primary winding.

CLASS-B AMPLIFIERS

In a class-B amplifier the base is biased almost to the point of collector current cut-off. In the absence of an input signal, therefore, very little collector current flows but the current increases

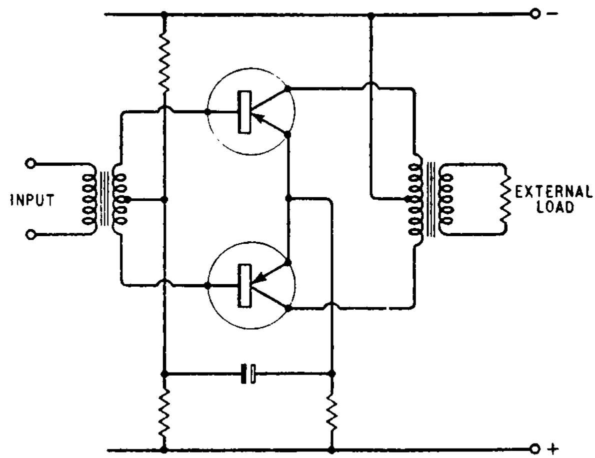


Fig. 10.3. Circuit for a push-pull class-A transistor output stage

as the amplitude of the input signal increases. This leads to economy in running costs because the current taken from the battery tends to be proportional to the signal amplitude and not independent of it as in a class-A amplifier. Positive half-cycles of the input signal are, however, not reproduced in a single class-B amplifier and it is essential to use two transistors in push-pull. Such a pair of transistors can give a power output of 2·5 times the maximum permissible collector dissipation of the two transistors: this is 5 times that available from the same two transistors operating in class-A push-pull. This may be shown in the following way. Let the peak collector current of each transistor be I_p and the battery voltage V_b. Then the peak collector-voltage swing V_p is equal to V_b. The power output from a push-pull stage is given by the product of the r.m.s. collector current $(I_p/\sqrt{2})$ and the r.m.s. collector voltage ($V_p/\sqrt{2}$) and is thus given by $I_pV_p/2$. If the input signal is sinusoidal, the combined collector current is sinusoidal in waveform as pictured in Fig. 10.4. Such a waveform has a mean value, or d.c. component, equal to $2/\pi$ of the peak value (I_p) and the power taken from the battery is thus equal to $2V_bI_p/\pi$, giving the efficiency as

$$\frac{\text{power output from transistors}}{\text{power taken from battery}} = \frac{I_pV_p}{2} \cdot \frac{\pi}{2I_pV_p}$$
$$= \frac{\pi}{4}$$
$$= 78{\cdot}54 \text{ per cent}$$

This is the maximum efficiency of which the stage is capable and to obtain it the optimum load per transistor must be equal to

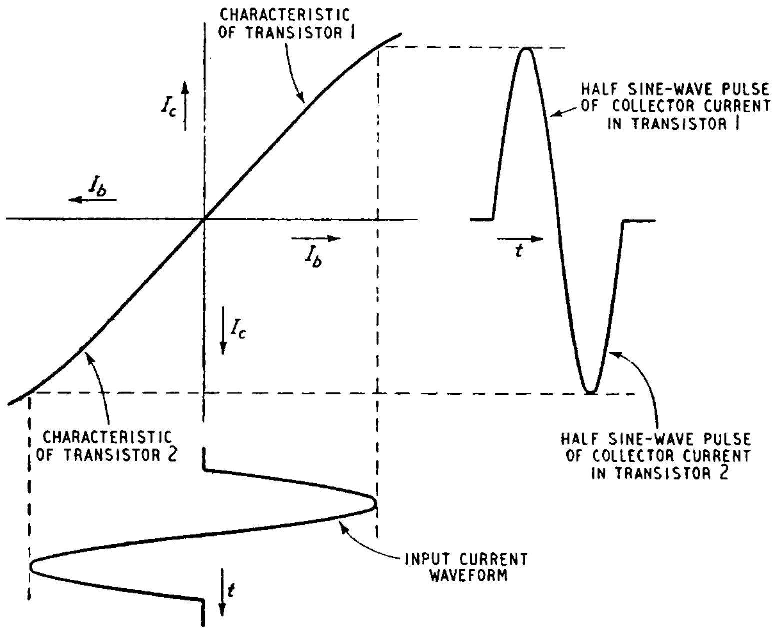


Fig. 10.4. Operation of a push-pull class-B transistor amplifier

(peak voltage)/(peak current). For a class-B push-pull the collector-to-collector load is given by $4V_p/I_p$.

If the load and battery voltage are kept constant and the input signal amplitude is reduced, the efficiency falls linearly due to the increasing failure to make use of the voltage swing available. This is easily shown. Suppose the input voltage is reduced to a times the value which gives maximum output. Then the collector-current swing falls to aI_p and (since the load is constant) the collector-voltage swing falls to aV_p, giving the power output per pair of transistors as $a^2I_pV_p/2$. The mean value of the collector current is now $2aI_p/\pi$ and the power taken from the battery is $2aI_pV_p/\pi$. The efficiency for the reduced input signal is given by

$$\frac{\text{power output from transistors}}{\text{power taken from battery}} = \frac{a^2I_pV_p}{2} \cdot \frac{\pi}{2aI_pV_p}$$

$$= a \cdot \frac{\pi}{4}$$

The efficiency is thus directly proportional to the input signal amplitude.

In transistor amplifiers we are particularly interested in the power P_t dissipated in the transistors themselves, for it is this which causes heating of the junctions and can damage them. Now

$$\begin{array}{lclcl} \text{power dissipated} & = & \text{power taken} & - & \text{power output} \\ \text{in transistors} & & \text{from battery} & & \\ \text{i.e., } P_t & = & \dfrac{2aI_pV_p}{\pi} & - & \dfrac{a^2I_pV_p}{2} \end{array}$$

Differentiating this

$$\frac{dP_t}{da} = \frac{2I_pV_p}{\pi} - aI_pV_p$$

Equating this to zero to find the maximum,

$$a = \frac{2}{\pi}$$

The heat in the transistors is thus a maximum when the signal amplitude is $2/\pi$ (approximately 0·63) times that giving maximum output power. Substituting this particular value of a in the general expression given above we have

$$\text{power output from amplifier} = \frac{a^2I_pV_p}{2}$$

$$= \frac{2I_pV_p}{\pi^2}$$

$$\text{power taken from battery} = \frac{2aI_pV_p}{\pi}$$

$$= \frac{4I_pV_p}{\pi^2}$$

By subtraction,

$$\text{power dissipated in transistors} = \frac{2I_pV_p}{\pi^2}$$

These results show that this particular value of a makes the power output one-half that taken from the battery: in other words it coincides with an efficiency of 50 per cent. The maximum power output from the class-B pair is $I_pV_p/2$. The maximum power dissipated in the transistors is $2I_pV_p/\pi^2$. The ratio of these two quantities is $\pi^2/4$, approximately 2·5 : 1, showing that it is possible to obtain an undistorted output of 2·5 times the rated maximum collector dissipation of the two transistors, i.e., 5 times the maximum collector dissipation of one of them. This applies only for a sinusoidal input.

As a numerical example, consider two transistors each with a maximum collector dissipation of 100 mW. In class-B push-pull it is possible to obtain from these an output power of 500 mW. If the battery supply is 6 volts, this is also the peak value of the collector voltage, and the peak current is given by I_p where

$$P = \frac{1}{2} V_pI_p$$

$$\therefore I_p = \frac{2P}{V_p}$$

$$= \frac{2 \times 500}{6} \text{ mA}$$

$$= 170 \text{ mA approximately}$$

The collector-to-collector load is given by

$$R_l = \frac{4V_p}{I_p}$$

$$= \frac{4 \times 6}{170 \times 10^{-3}} \text{ ohms}$$

$$= 140 \text{ ohms approximately}$$

Driver Stage

The transistor feeding a class-B output stage may be regarded as a large-signal amplifier because it must supply appreciable power to the base circuits of the output stage. The design of the driver stage and the transformer coupling it to the output stage depends on the input resistance of the output stage.

We have so far assumed the input resistance of a common-emitter amplifier to have a value determined by the transistor parameters and the output load resistance. For small input signals this is a reasonable assumption, and the input resistance is commonly taken as 1 kΩ. A class-B stage, however, does not satisfy these conditions: it is biased almost to collector-current cut off and, in the absence of an input signal, has an input resistance higher than that of a class-A amplifier. A class-B stage requires a large input signal to swing the collector current to its maximum limit and for such signals the input resistance is less than that of a class-A amplifier.

Thus the input resistance is no longer a constant but varies with the input signal amplitude, decreasing as the input signal increases. The variation may be from, say, 2 kΩ for small signals to 200 ohms for large signals. This variation of input resistance can be anticipated from inspection of the current-input voltage characteristic for a junction transistor (Fig. 1.12). The slope is small near the origin but increases as the input voltage is increased. The input resistance is measured by the reciprocal of the slope and decreases as signal amplitude increases.

These variations in input resistance can, unless precautions are taken, cause serious distortion. For example, suppose the output stage is driven from a signal source having an output resistance small compared with the minimum input resistance of the output stage. This means that the signal voltages applied to the bases of the output stage are nearly equal to that of the signal source and are therefore substantially undistorted. If the input resistance of the output stage were constant and independent of signal amplitude the input current would also be undistorted. The input resistance does, however, vary with signal amplitude. Thus the input current is not linearly related to the signal voltage and must be distorted.

Transistors are, of course, current-operated devices and if the input current is distorted, the output current must also be distorted. The distortion obtained in this mode of operation (called constant-voltage operation) is most unpleasant: for small signals the input resistance is large and the input current is smaller than required for distortionless amplification. Thus a sine-wave input voltage is reproduced with a waveform similar to that shown in Fig. 10.5. This is known as *cross-over distortion* because it occurs when one transistor is being cut off and the other turned on, i.e. when the

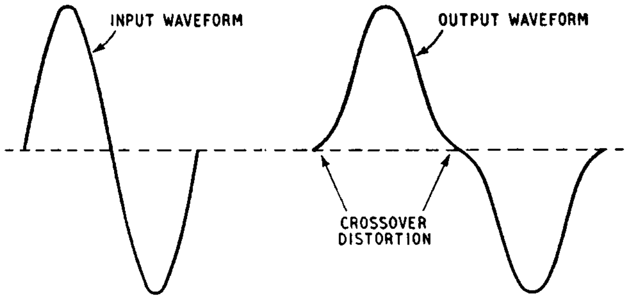


Fig. 10.5. Input and output waveforms for an amplifier giving cross-over distortion

state of conduction is being handed over from one transistor to the other.

To reduce cross-over distortion, the input current for the output stage must be undistorted. The variations in input resistance must therefore have no effect on the input current. To achieve this the output stage must be driven from a high-resistance signal source. Ideally the source resistance should be large compared with the highest value of input resistance likely to occur: we then have constant-current operation. If, however, the driver transformer is designed to satisfy this requirement, the gain of the driver stage is low. This difficulty can be reduced by arranging that the output transistors are not biased to cut off in the absence of a signal but take a small collector current. Thus the transistors operate in class A for small input signals and the variations in input resistance are reduced. It is now possible to use a smaller value of source resistance while still avoiding the worst effects of cross-over distortion and reasonable gain is possible from the driver stage.

For transistors rated for 100 mW maximum collector dissipation, the input resistance falls to approximately 100 ohms at minimum and is probably around 1 kΩ for a standing collector current of 2 mA, a typical value for a class B output stage. If high gain is essential from the driver stage and a little distortion can be tolerated, the driver transformer can be designed to match the output resistance of the driver stage (say 30 kΩ) to the maximum value of the input resistance (1 kΩ). The turns ratio (primary to each secondary winding) required is thus

$$\sqrt{(30{,}000/1{,}000)} : 1 = \sqrt{30} : 1$$
$$= 5{\cdot}5 : 1$$

When better quality is required and a little gain can be sacrificed, this ratio is sometimes decreased to 3 : 1 or even 2 : 1 to give a better approximation to constant-current input for the class-B stage.

Symmetrical Class-B Amplifier

The circuit of the driver stage and class-B output stage of an amplifier is given in full in Fig. 10.6. The driver stage operates in class A and the collector current is stabilised at, say, 3 mA by

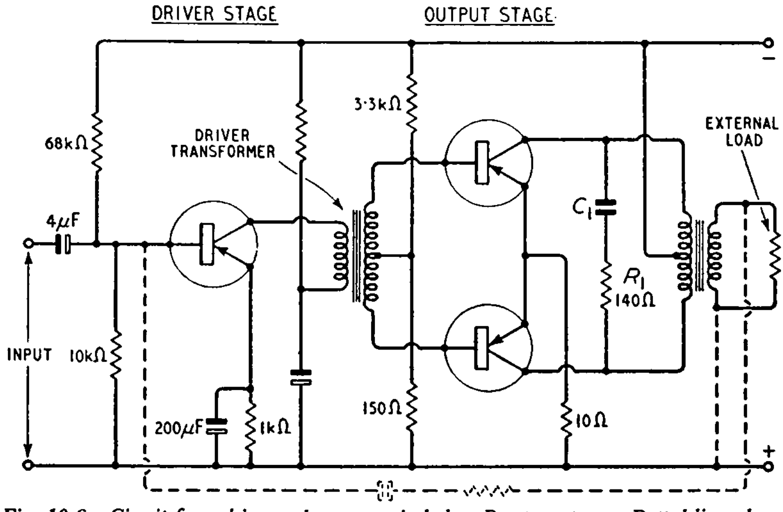


Fig. 10.6. Circuit for a driver and a symmetrical class-B output stage. Dotted lines show how negative feedback can be applied when the driver stage has a current input

the potential-divider method. A similar technique is applied to the output stage which has a common emitter resistance, the potential divider being adjusted to give the required standing collector current. The emitter resistance of a class-B amplifier cannot be decoupled because the mean emitter current is not constant (as in a class-A amplifier) but varies over a wide range depending on the instantaneous amplitude of the input signal. A capacitor in parallel with the emitter resistor would acquire charge on current peaks and discharge in the intervals between peaks, thus introducing unwanted exponential signals into the amplifier. The emitter resistance must therefore be kept small to minimise the negative feedback which would reduce the gain of the output stage.

The emitter resistance is commonly as low as 5 ohms: nevertheless, it gives some measure of stabilisation of the standing collector current and it helps to equalise differences in parameters of the two output transistors. Another resistance which must be small is the lower arm of the potential divider feeding the bases of the output transistors. The input currents of the output stage flow in this resistor and a large value would severely reduce the signal input to the output stage: a value as low as 120 ohms is commonly used. It is not advisable to decouple this resistance by a low-reactance capacitor because on large input signals the capacitor becomes charged by the unidirectional input current flow for the output stage and on small input signals it discharges through the resistance, this also superimposing unwanted exponential signals on the input to the output stage.

Class-B output stages tend to give unwanted ripples on the output signal: these are caused by ringing of the secondary winding of the driver transformer when the associated transistor is cut off. This tendency can be reduced by so constructing the transformer that the two secondary windings are closely coupled, e.g., by winding the secondaries in a bifilar manner. This discourages ringing because the undamped secondary winding is effectively damped by its close coupling to the damped winding. The network R_1C_1 is included to attenuate unwanted high-frequency components such as harmonics of the input signal or oscillations due to ringing. This network together with the leakage inductance L of the output transformer and the resistance R_l of the external load form an LCR combination. Such a network can give an input resistance independent of frequency if the product $R_1 R_l$ is made equal to L/C.

Distortion introduced by the class-B output stage can be reduced by negative feedback. The feedback voltage can be taken from the secondary winding of the output transformer and applied to the driver stage. It should be applied to the base if the driver stage is required to operate with a current input, and to the emitter if the driver stage is required to operate with a voltage input.

The latter alternative is preferable if the driver stage directly follows a detector, as may occur in a receiver. The degree of feedback which can be applied is limited by the phase shifts introduced by the driver and output transformers at the extremes of the passband and in practice rarely exceeds 6 dB: even this small degree halves the harmonic distortion of the amplifier.

Asymmetric Class-B Amplifier

An interesting variant of the class-B push-pull circuit is the so-called single-ended or asymmetric circuit illustrated in Fig. 10.7.

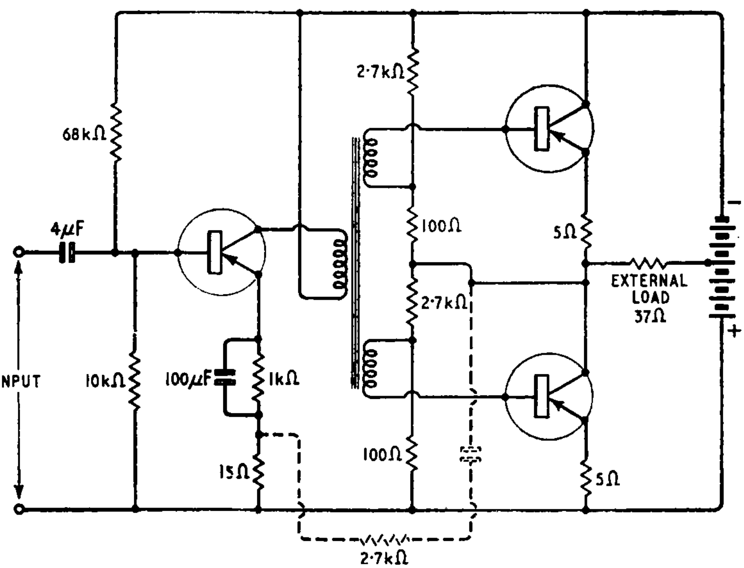


Fig. 10.7. Circuit for a driver and asymmetric class-B output stage. Dotted lines show how negative feedback can be applied when the driver stage has a voltage input

The output transistors are connected in series across the collector supply battery and the load resistance is connected between their junction and a centre tap on the battery. Separate secondary windings are required for the two bases because there is now a difference in the steady voltage of the two. The load resistance equals (peak collector voltage swing)/(peak collector current swing), and for output transistors operating from a 4·5-volt supply and taking a peak collector-current of 120 mA may be as low as 37 ohms. A loudspeaker speech coil can readily be wound to have an impedance of this value and thus there is no need for a loudspeaker matching transformer: this is one of the great advantages of this circuit. It does, however, require a collector supply of higher voltage than a symmetrical circuit and the input transformer must have isolated secondary windings. The elimination of the output transformer is a great asset, however, and makes this circuit particularly suitable for miniature receivers. The use of a centre-tapped battery (as in Fig. 10.7) is not essential. The external load may alternatively be returned to the positive battery terminal (usually at earth potential) provided a capacitor is connected in series with the load to avoid disturbance of d.c. conditions. In an a.f. amplifier employing a 37-ohm load, the capacitance should be not less than 100 μF to avoid loss of bass.

11

A.M. AND F.M. RECEIVERS

Introduction

WE have so far discussed the use of transistors as amplifiers and we shall now consider transistor oscillators, detectors and frequency changers. We shall then have described all the operations necessary in a receiver and can consider the design of complete transistor superheterodyne receivers.

OSCILLATORS

Oscillators have two main sections: a frequency-determining section and a maintaining section. The frequency-determining section commonly consists of an LC circuit or RC network. The maintaining section is commonly a valve or transistor amplifier which must have sufficient gain to make up for losses in the frequency-determining section to give rise to the positive feedback which causes oscillation.

In one type of oscillator the amplifier is used purely as a switch, the valves or transistors being in a state of conduction or of non-conduction (i.e. cut off) for most of the period of oscillation. The changes of state are very rapid and the shape of the input-output characteristic of the amplifier (which is of great importance in class-A amplification) is unimportant in this application. Such oscillators are termed relaxation oscillators and their period of oscillation is usually determined by the time constants of RC or RL networks.

The output waveform of such oscillators is largely made up of linear and exponential sections. Typical of such oscillators are multivibrators and blocking oscillators.

In a second type of oscillator, the maintaining section is used as a linear amplifier and is in operation for the whole or a substantial fraction of each cycle of oscillation. The output waveform is sinusoidal and the frequency-determining system is commonly, though not invariably, an LC circuit. Hartley and Colpitts oscillators are of this type.

In general, transistor oscillators are of the same basic types as thermionic-valve oscillators and their circuits can be deduced by analogy with the valve circuits. It may, however, be necessary to modify the circuit to suit the input and output resistances of the transistors where these differ markedly from those of a valve.

In this chapter we are primarily concerned with the application of junction transistors in sinusoidal oscillators, particularly those

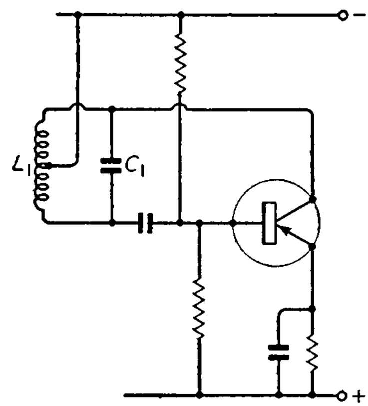


Fig. 11.1. One possible circuit for a transistor Hartley oscillator

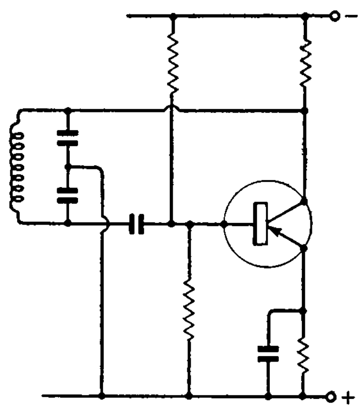


Fig. 11.2. One possible circuit for a transistor Colpitts oscillator

used in receivers. Some information about the use of junction transistors in relaxation oscillators is given in Chapter 12.

When a signal is momentarily induced in an LC circuit, it gives rise to an oscillation which is at the resonance frequency of the circuit and dies away exponentially due to dissipation in the resistance of the circuit. To maintain the oscillation at constant amplitude the loss of power must be made good: this may be achieved by connecting the LC circuit to a source of power such as an amplifier. This source can be regarded as having a negative input resistance which neutralises the positive resistance of the tuned circuit. In the Hartley, Colpitts and Reinartz circuits the negative input resistance is achieved by positive feedback, i.e., by establishing a connection between the input and output circuits of the maintaining amplifier. It is characteristic of these forms of oscillator that at least three connections must be made to the LC circuit to obtain the required positive feedback.

Fig. 11.1 gives the circuit diagram of a Hartley transistor oscillator. The steady component of the transistor collector current is stabilised by the potential divider method and, except for the components associated with the stabilisation, the circuit is similar to that of a

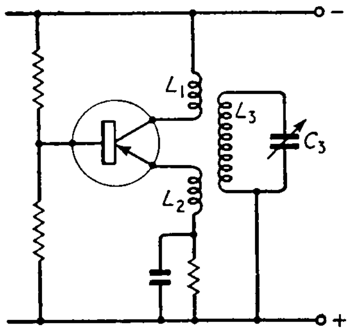


Fig. 11.3. One possible circuit for a transistor Reinartz oscillator

thermionic-valve Hartley oscillator. The frequency-determining section L_1C_1 is connected between collector (anode) and base (grid). The emitter (cathode) is effectively connected to the centre tap of the inductor because both these points are at zero r.f. potential. The circuit illustrated is a series-fed type in which the collector current flows through part of the inductor. There is an alternative shunt-fed circuit in which the transistor has a resistive collector load and the collector is connected to the top end of the inductor by a fixed capacitor.

Fig. 11.2 gives the circuit diagram for a Colpitts oscillator, also d.c.-stabilised by the potential-divider method.

The circuit diagram for a Reinartz-type transistor oscillator is given in Fig. 11.3: this circuit is of interest because it is frequently used in transistor receivers. Regeneration in this circuit is obtained by coupling the collector circuit to the emitter circuit via the mutual inductance between L_1 and L_2. Both inductors are also coupled to the frequency-determining circuit L_3C_3. This circuit is also d.c.-stabilised by the potential-divider method. It could, of course, be simplified by omitting the components L_3 and C_3 and connecting a tuning capacitance across L_1 or L_2, but the form illustrated here enables the moving vanes of the tuning capacitor to be earthed. Moreover, this helps frequency stability by isolating L_3C_3 to some extent from variations in transistor parameters.

A.M. DETECTORS

The purpose of the detector in a receiver is to derive from a modulated carrier a substantially-undistorted copy of the modulation waveform. Frequently an a.m. detector also has to produce an output signal, proportional to the carrier amplitude, which can be used for automatic gain control and in a transistor receiver appreciable power is necessary for a.g.c. Point-contact diodes are often used for detection in transistor receivers: they are satisfactory in providing the a.f. output and the provision of a.g.c. is not difficult. Transistors can be used for a.m. detection also and they have the advantage over point-contact diodes that they give substantial gain (e.g. 20 dB) and can readily supply the power necessary for a.g.c. The circuit of a common-emitter transistor a.m. detector is given in Fig. 11.4. The potential divider method of d.c. stabilisation is again used and the value of R_e must be chosen to give the required value of collector and emitter current. The value of the emitter current

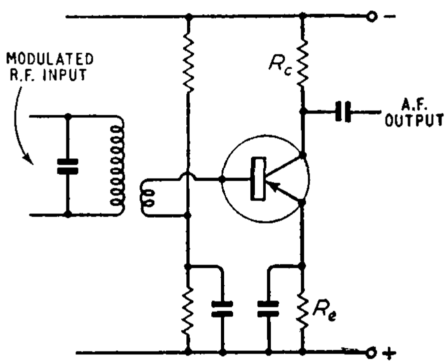


Fig. 11.4. One circuit for a common-emitter transistor a.m. detector

is particularly important in an a.m. detector and its value must be a compromise between the following two conflicting requirements:

(1) The emitter-base junction must be biased to a point on the dynamic characteristic where the curve has a sharp knee. It is essential to use an r.f. type of transistor because the

input is r.f. and the emitter current is usually low when optimum bias is applied.

(2) The transistor also operates as an a.f. amplifier and for maximum gain from such an amplifier the collector current (and hence the emitter current) must be of the order of 1 mA.

Low emitter currents give good demodulation but poor a.f. gain; high emitter currents good a.f. gain but poor demodulation. It is not surprising, therefore, that there is an optimum value of emitter current which gives maximum a.f. output for a given modulated r.f. input, and for germanium transistors this is commonly of the order of 50 μA.

A suitable value for the collector load resistance is 15 kΩ and the voltage drop across this for a collector current of 50 μA is

$$50 \times 10^{-6} \times 15 \times 10^{3} = 750 \times 10^{-3}$$
$$= 0{\cdot}75 \text{ volt}$$

This is the voltage across R_c when there is no input signal. When an input signal is applied the mean collector current rises causing an increased voltage drop across R_c. The collector potential thus becomes less negative, i.e., more positive, the polarity required in one type of a.g.c. circuit. Such a.g.c. is more effective than that obtainable from a simple diode detector because of the amplification available.

The input resistance of a common-emitter amplifier is commonly quoted as approximately 1 kΩ. This applies when the emitter current is approximately 1 mA. In the common-emitter detector, however, the emitter current is much smaller than this and the input resistance is therefore much higher, commonly greater than 10 kΩ. The damping imposed by the detector on the tuned circuit feeding it is thus not as heavy as might be feared. Nevertheless it is usually necessary to feed the detector from a tapping point on the inductor or, as shown in Fig. 11.4, from a small coil closely coupled to the tuning inductor to avoid too great a reduction in the Q value.

AUTOMATIC GAIN CONTROL

A.g.c. is used in radio receivers for two main reasons:

(1) It combats fading. Fading is experienced quite generally in radio reception but it is particularly troublesome during medium-wave reception after nightfall.

(2) It helps to prevent overloading of later stages. In transistor receivers without a.g.c. overloading of the final i.f. stage would occur on strong signals.

A.g.c. is achieved by feeding back to pre-detector stages (usually i.f. stages) a control voltage proportional to the carrier amplitude at the detector. This control voltage has the effect of reducing the gain of the controlled stages. Strong signals give a large control voltage and low i.f. gain; weak signals give a small control voltage and high i.f. gain.

The control voltage must reduce the gain of the i.f. stages and there are two methods by which the gain of a transistor amplifier can be controlled:

(1) By reducing the emitter current, keeping the collector voltage constant. For germanium transistors the gain tends to be proportional to the square root of the emitter current, provided this is below approximately 500 μA.

(2) By reducing the collector voltage, keeping the emitter current constant. For germanium transistors the gain tends to be proportional to the square root of the collector voltage, provided this is below approximately 500 mV.

Both methods of gain control reduce the signal-handling capacity of the amplifier and to avoid distortion due to peak clipping the controlled stages must at all times have small input signals. It is not advisable therefore to control the final i.f. amplifier, for this has to deliver a reasonably large signal to the detector, particularly if this is a diode which has to supply the control voltage. The reduction of signal-handling capacity of the controlled transistors contrasts with the behaviour of automatic-gain-controlled valves: the application of negative bias to the control grid of a variable-mu pentode lengthens the grid base and improves signal-handling capacity.

Gain control by adjustment of emitter current is rarely achieved directly, i.e., by control of the emitter voltage of a common-base amplifier. This is because the input resistance of the emitter circuit is so low (50 ohms is typical) that considerable power input is necessary to produce appreciable change in current and thus in gain.

It is usual to control the emitter current indirectly, e.g., by control of the base potential in a common-emitter amplifier. In this way the current gain of the transistor is utilised for a.g.c.

purposes. Moreover the input resistance of a common-emitter amplifier is higher (1 kΩ is typical) than for a common-base amplifier.

A typical circuit diagram for an amplifier with base injection of a.g.c. voltage is given in Fig. 11.5. The voltage must be positive-going in order to reduce emitter current in a pnp transistor and

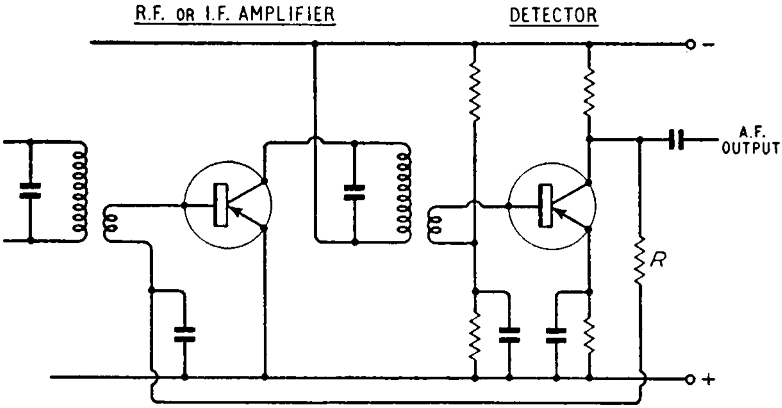


Fig. 11.5. An a.g.c. circuit using emitter-current control

such a polarity can be obtained from a common-emitter detector as shown in this diagram. The resistance R should be so adjusted that, in the absence of an input signal, the controlled stage has a collector current of approximately 500 μA. A larger collector current gives an effective delay to the a.g.c. because little reduction in gain occurs until the collector current has fallen below approximately 500 μA.

For the sake of simplicity, neutralising is omitted from the diagram and no means is indicated of preventing thermal runaway of the r.f. or i.f. amplifier.

A.g.c. by control of collector voltage is also achieved indirectly and a typical circuit is given in Fig. 11.6. This shows a common-emitter stage with the control voltage applied to the base. The collector circuit decoupling resistor R_1 is an essential feature of this circuit. The control voltage must be negative-going in this circuit to increase the collector current on strong signals. The value of the decoupling resistor is so chosen that the collector-emitter voltage is approximately 500 mV in the absence of a signal. Any increase in base current due to a.g.c. action increases emitter current and hence collector current, decreasing collector voltage and gain. A typical value for the decoupling resistance is 15 kΩ. The required polarity of a.g.c. voltage is obtained in this circuit by taking the control voltage from the emitter of a common-emitter detector.

The circuit is again simplified by the omission of neutralising and the means of preventing thermal runaway.

Attention must be paid in the design of a.g.c. circuits to the effects on receiver performance of the inevitable variations in transistor input and output impedances brought about by the changes in d.c. conditions. The input and output impedances of a transistor both have resistive and reactive components. The resistive component causes damping of the input and output tuned circuits and the capacitive component causes mistuning of these circuits. The severity of the damping and the extent of the mistuning both depend on the emitter current and both therefore change with alterations in a.g.c. voltage. The mistuning effect can be minimised by so designing the circuit that the variations

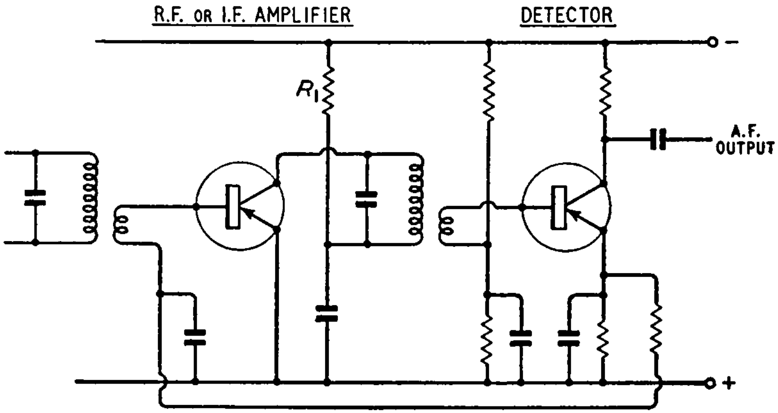


Fig. 11.6. An a.g.c. circuit using collector-voltage control

in transistor capacitance are small compared with that already present in the input and output tuned circuits. The variations in damping are, however, more difficult to minimise, and some decrease in damping is probably inevitable as emitter current is reduced.

However, the variations in damping need not always be undesirable. For example, in an a.g.c. circuit using collector-voltage control the emitter current increases when a strong signal is received. This causes increased damping and hence a wider i.f. passband. This is an advantage in receiving a strong local signal because it permits better quality of reproduction on a signal which is large enough to swamp any interference. When the receiver is tuned to a weak signal, the damping is reduced and the passband narrowed. In the alternative a.g.c. circuit using emitter-current control the emitter current is reduced when a strong signal is tuned in, giving minimum passband on stronger signals. However, this method gives less mistuning than the other and, in general, emitter-current control is preferred.

F.M. DETECTORS

The function of an f.m. detector is to derive from a frequency-modulated signal a substantially-undistorted copy of the modulated waveform impressed on the signal. There are many types of f.m. detector but only two are in common use in f.m. receivers: these are the Foster-Seeley discriminator and the ratio detector. The method of operation of these detectors is complex and only a brief summary of it is given below: for a more complete description the reader is referred to other books*.

Foster-Seeley Discriminator

This discriminator contains two diode detectors so arranged that their outputs are connected in series opposition. The diodes are fed from a double-tuned transformer, the primary and secondary windings of which are resonant at the centre frequency of the passband to be covered. An essential feature of the circuit is that a fraction of the primary voltage is fed to the centre point of the secondary winding. In Fig. 11.7 this is achieved by a connection between the centre point of L_2 and a tapping point on L_1 but the secondary connection could be to the junction of two equal capacitors across L_2 (they could together constitute the tuning capacitance) and the primary connection could be to an inductor closely coupled to L_1 or to a capacitive potential divider across L_1 (formed by two capacitors which may also provide the tuning capacitance).

*For example, B. S. Camies "Principles of Frequency Modulation" Iliffe Books Ltd.

For signals at the centre frequency, diodes D1 and D2 receive equal inputs and the voltages generated across R_1 and R_2 are equal, giving zero resultant voltage across $(R_1 + R_2)$. The effect of the interconnection between primary and secondary windings is that for signals displaced from the centre value one diode receives a bigger input than the other. The voltages across R_1 and R_2 are then no longer equal and there is a net output across $(R_1 + R_2)$,

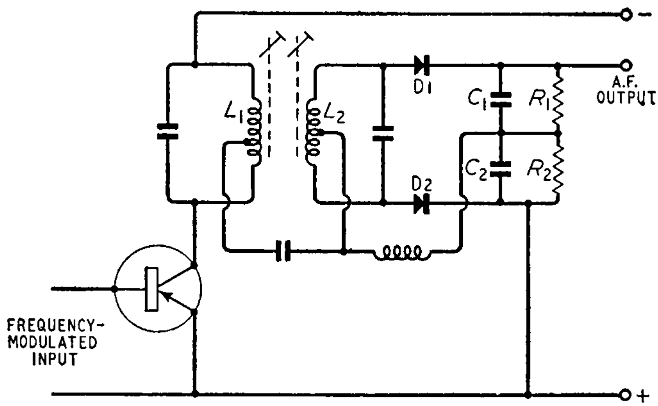


Fig. 11.7. One circuit for a Foster-Seeley discriminator

the polarity depending on the direction of the frequency displacement and the magnitude depending on the extent of the displacement. If, therefore, a frequency-modulated signal is applied to the discriminator, a copy of the modulation waveform is generated across $(R_1 + R_2)$.

The Foster-Seeley discriminator gives zero output at the centre frequency and at this frequency the output is independent of the magnitude of the signal input to the detector. At other frequencies the output of the discriminator is proportional both to frequency displacement and to signal input. Ideally the output of an f.m. detector should be proportional to the frequency displacement but independent of the signal input. If this can be achieved the full advantages of frequency modulation are realised and the receiver is to a large extent immune from interference due to a.m. signals and from the distortion due to multi-path reception. The Foster-Seeley discriminator thus has poor ability to reject a.m. signals and is normally used with a separate limiter stage in order to obtain a satisfactory performance.

Ratio Detector

The ratio detector has much better a.m. rejection than a Foster-Seeley circuit and nearly all commercial receivers employ a ratio detector.

The circuit diagram of one form of ratio detector is given in Fig. 11.8. It has two diodes and a double-tuned transformer with a primary-secondary connection similar to that employed

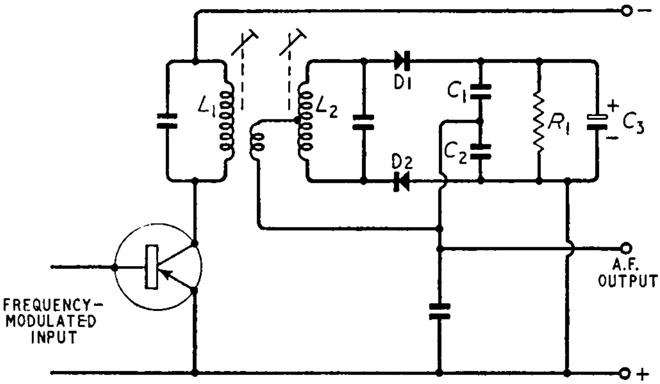


Fig. 11.8. One circuit for a ratio detector

in the Foster-Seeley circuit but the diodes are connected in a series-aiding arrangement and supply a common load resistor. This resistor has a low value to give the heavy damping of the secondary tuned circuit on which the limiting properties of the detector depend. The load resistor is bypassed by a high-value capacitor giving a load time constant of 0·1 second. The diodes conduct continuously when a signal is applied to the detector and give a voltage across the load circuit proportional to the carrier input: the maximum value of this voltage gives an indication of the correct tuning point. The inputs to the two diodes vary with frequency displacement (as in the Foster-Seeley circuit) and the voltages generated across the reservoir capacitors C_1 and C_2 vary also although the total voltage across $(C_1 + C_2)$ is independent of input frequency, being stabilised at a value proportional to carrier input amplitude by the long time constant R_1C_3. One end of the capacitor C_3 is earthed usually and the a.f. output is taken from the junction of C_1 and C_2.

Practical ratio detector circuits commonly employ additional resistors in series with the diodes: by choosing suitable values for these resistors, the a.m. rejection of the detector can be significantly improved.

MIXERS

In a superheterodyne receiver the carrier and sidebands constituting the received signal are in effect translated in frequency to give a new signal with a carrier at the intermediate frequency. This is achieved in a mixer stage in which the received signal is combined with the output of a local oscillator, thereby producing a resultant signal at the sum or difference of the received carrier and oscillator frequencies.

In general, there are two basic principles which can be applied to produce an output with a frequency equal to the difference between the frequencies of two input signals. In one method the two input signals are simply connected in series or in parallel and applied between two electrodes such as the grid and cathode of a valve or the base and emitter of a transistor. If the input-output relationship for the valve or transistor is linear, these two signals are amplified independently and there is no output at any frequency other than frequencies of the two input signals. To produce interaction between the two original signals, thus obtaining an output at the difference frequency, it is essential that the valve or transistor should have a non-linear characteristic, i.e., should behave as a detector.

For this reason the mixers of early superheterodyne receivers were known as first detectors. Stages of this type are known as *additive mixers* and they are usually biased near the point of anode or collector-current cut off to achieve the non-linearity essential for their action.

In another, more modern, type of mixer the received signal and the local oscillator output are in effect multiplied together. This is achieved by applying the two inputs to separate electrodes of a valve or transistor which is so designed that a signal applied to one electrode controls the gain of a signal applied to the other electrode. Two such electrodes are the control grid and suppressor grid of a pentode. In a mixer of this type there is no need for any non-linearity, the output at the difference frequency being produced directly from the effective multiplication of the two input signals. Such mixers are termed *multiplicative*, and most modern mixers are of this type. Clearly it is wrong therefore to refer to such mixers as first detectors because they can be linear devices.

A multiplicative type of mixer requires a valve or transistor with two input electrodes, and mixer valves commonly have a total of at least six electrodes. It is possible that transistors of an analogous type may in time be developed, but at the moment most transistors are triodes with which it is difficult if not impossible to construct a multiplicative type of mixer; transistor mixers must

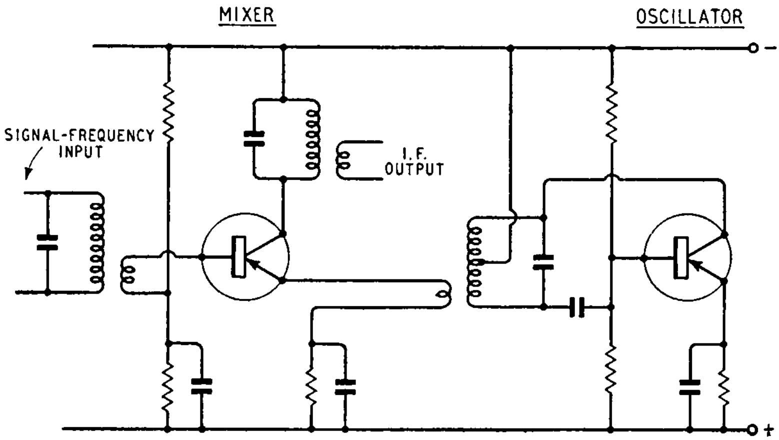


Fig. 11.9. One possible circuit for a two-transistor frequency changer

operate therefore on the additive principle and this can be achieved by applying both input signals between base and emitter. Most transistor receivers employ an additive mixer of this type and one possible circuit is given in Fig. 11.9. This also gives the circuit of the oscillator stage; the combination of the two constitutes a frequency-changer stage.

In this circuit the mixer operates in the common-emitter mode and the two input signals are connected in series and applied to the base-emitter circuit. The operating point is swept over the characteristic by the local oscillation, and the efficiency of the mixing process depends on the amplitude of oscillation injected into the emitter circuit. For small oscillation amplitudes the conversion

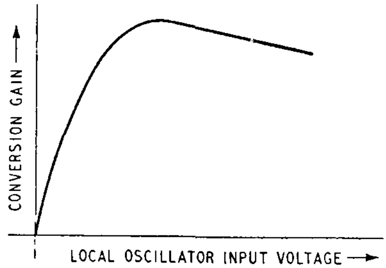


Fig. 11.10. Variation of conversion gain with local-oscillator input voltage

gain, i.e., the ratio of i.f. signal output to r.f. signal input, increases linearly with oscillation amplitude but levels off at a particular amplitude and then falls as shown in Fig. 11.10.

Increase in oscillator amplitude beyond this point causes little change in gain.

In a receiver in which the oscillator has to operate over a range of frequency, some variation in oscillator amplitude is inevitable. In order to prevent such variations causing large changes in conversion gain the oscillator amplitude is usually designed so that it is greater than the minimum critical value which gives maximum gain.

Self-oscillating Mixers

It is quite possible to arrange for a single transistor to perform the functions of mixing and of oscillation, and the conversion gain of such a circuit is little short of that available from circuits in which separate transistors are used for the two purposes. One circuit

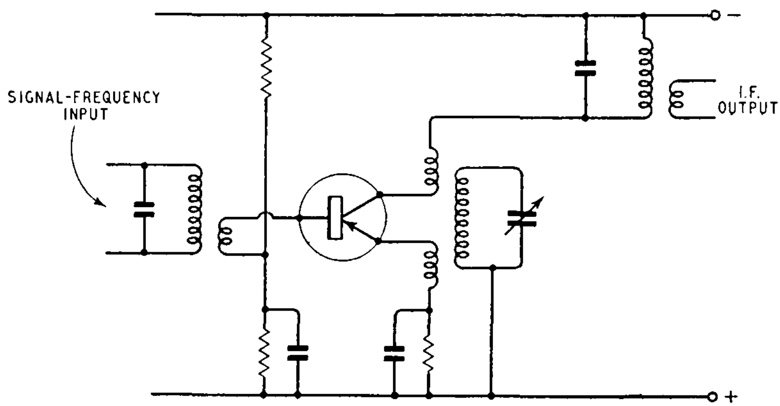


Fig. 11.11. One possible circuit for a self-oscillating mixer

for a self-oscillating mixer is given in Fig. 11.11. This may be regarded as a Reinartz oscillator of the type shown in Fig. 11.3 in which the base circuit is tuned to accept the signal-frequency input and the output circuit is tuned to select the difference-frequency output. The oscillator signal and the signal-frequency input are connected in series between base and emitter to enable the transistor to operate as an additive mixer. Circuits of this type are commonly employed in transistor superheterodyne receivers for a.m. reception.

Fig. 11.12 shows a form of self-oscillating mixer suitable for use in v.h.f. receivers: part of the preceeding r.f. stage is shown on the left of the diagram. The transistor operating conditions are set

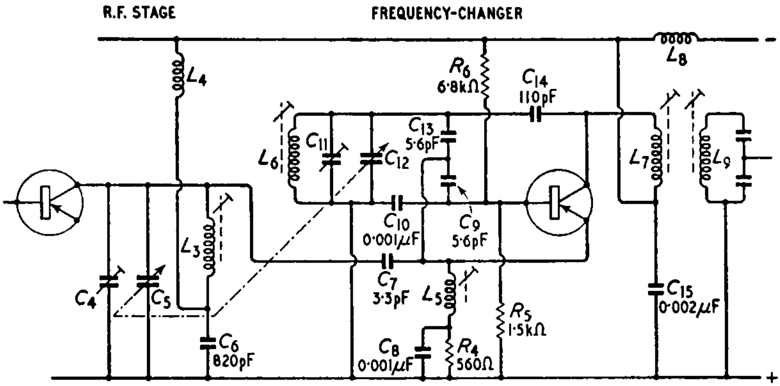


Fig. 11.12. A self-oscillating mixer of the Colpitts' type suitable for use in v.h.f. receivers

at a value suitable for a self-oscillating mixer by the potential divider R_5R_6 which determines the base potential and by the resistor R_4 which determines the emitter current. The transistor circuit is based on that of a Colpitts' oscillator, the two fundamental capacitors C_9 and C_{13} being connected in series between collector and base with the centre point connected to emitter. C_{12} is the oscillator tuning capacitor and C_{11} a trimmer. C_{12} is ganged with C_5 which tunes the signal-frequency inductor L_3 in the collector circuit of the r.f. amplifier. It is desirable to earth the moving vanes of C_5 and C_{12} but if this is done the base of the oscillator is at zero r.f. potential. The emitter cannot also be at zero r.f. potential (as in many Colpitts' oscillators) because C_9 would be short-circuited and oscillation would be impossible. Some reactance is therefore provided in the emitter circuit by the inclusion of inductor L_5. The signal-frequency input to the mixer can now be injected into the emitter circuit via C_7 to enable mixing to take place in the base-emitter diode of the frequency changer. The inclusion of L_5 in the emitter circuit of the frequency changer provides negative feedback which reduces the gain of the stage at the intermediate frequency. The feedback is therefore reduced to a minimum by arranging that L_5 resonates with C_8 at the intermediate frequency. L_5 is made variable to permit this adjustment.

The i.f. output of the frequency changer is selected by the transformer L_7L_9, both primary and secondary windings of which are tuned to the intermediate frequency, usually 10·7 Mc/s. The primary winding is tuned by the 110-pF capacitor C_{14} which is connected in series with L_6 and C_{15} across L_7. Both L_6 and C_{15} have negligible reactance at 10·7 Mc/s and thus C_{14} is effectively in parallel with L_7.

COMPLETE A.M. RECEIVER

Fig. 11.13 gives the circuit diagram of a complete single-waveband superheterodyne receiver using six transistors. The first operates as a self-oscillating mixer, the second and third are i.f. amplifiers, the fourth is an a.f. amplifier and the final two are a push-pull class-B output stage. The detector is a point-contact diode which applies a.g.c. to the first i.f. amplifier. Such a receiver can operate quite satisfactorily from a 6-volt or 9-volt battery and the sensitivity and output power are the equal of (and are sometimes superior to) those of a conventional 4-valve battery-operated receiver. The capacitors can be very small physically because of their low working voltage, and such a receiver can be made very compact. Models using miniature loudspeakers can be made small enough to go into the pocket of a jacket.

To be self-contained, miniature receivers of this type use a Ferrite core aerial consisting of a signal-frequency tuned circuit L_1 wound on a rod or strip of Ferrite. Aerials of this type have now replaced the frame aerials formerly used in portable receivers. Ferrite aerial windings may have a Q value as high as 250 but this is often halved by the proximity of nearby metalwork: such a value is an embarrassment when, as is usual, a 2-gang capacitor is used for tuning the aerial and oscillator circuits. The two circuits can be accurately in step at three frequencies in the band

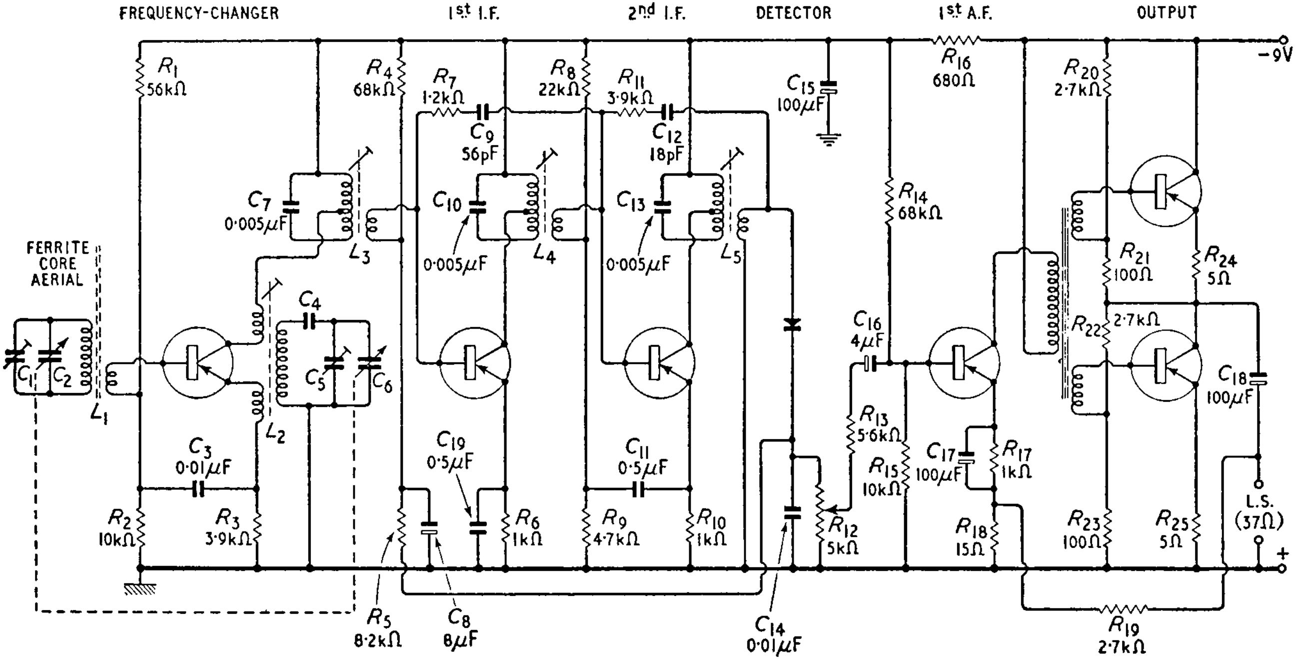


Fig. 11.13. Circuit diagram of complete single-waveband superheterodyne transistor receiver

but at all other frequencies there is a loss in sensitivity due to mistuning of the aerial circuit and this loss is the greater the higher the Q value. On the other hand if the Q is made too low there is a danger of second-channel interference. A working Q of 100 is a reasonable compromise. The aerial coil is coupled to the base circuit of the frequency-changer by means of a secondary winding, and the inductance of this winding can be so chosen that the damping due to the frequency changer gives a working Q of this value.

The frequency-changer circuit is similar to that shown in Fig. 11.3 but the capacitors C_4 and C_5 are included to permit adjustment of the frequency range covered by the receiver and to give the required 3-point tracking. With the tuning capacitors at minimum capacitance C_5 is adjusted to give the required high-frequency tuning limit. With the tuning capacitors at maximum capacitance L_2 is adjusted to give the required low-frequency tuning limit. These adjustments are repeated until no further adjustment is necessary. If the value of C_4 is correctly chosen, the alignment of the receiver will also be perfect at a frequency near the centre of the waveband.

In aligning the receiver the trimmer C_1 is adjusted to give maximum sensitivity near the high-frequency end of the waveband and the inductance L_1 is adjusted to give maximum sensitivity near the low-frequency end. The inductance can be adjusted by sliding the winding along the Ferrite rod or strip.

The remainder of the receiver is conventional, comprising i.f., detector and a.f. stages of types described earlier. One point meriting description is the decoupling network $R_{16}C_{15}$. This is particularly necessary in a receiver such as this which has a class-B output stage. The collector current of a class-B output stage is not steady but has a strong component at the second harmonic of the frequency of audio output. This current, in flowing through the internal resistance of the battery, sets up a voltage which is impressed on the collector voltage for all other stages. This can cause distortion if it affects the gains of these stages. This is particularly likely when the internal resistance of the battery rises due to ageing. The network $R_{16}C_{15}$ reduces distortion due to this cause by attenuating such voltages, the reactance of C_{15} being small compared with R_{16} at audio-frequencies.

Another interesting feature of the receiver is the small forward bias of the detector by the network $R_4R_5R_{12}$. This is deliberately introduced to improve the performance of the detector. The characteristic of a point-contact (and a junction) diode is almost straight where it passes through the origin and detection efficiency is therefore poor for small applied voltages. The bias provided can be chosen to give maximum detector efficiency for small input signals, thus improving receiver sensitivity. Alternatively the bias can be chosen to minimise the peak-clipping which occurs for large signal inputs as a result of the difference in value of the d.c. and a.c. detector loads: this adjustment improves the quality of reproduction from local transmissions.

F.M. RECEIVER

We have now discussed f.m. detectors and the use of transistors in a.f. amplifiers, i.f. amplifiers and frequency changers: we can now consider the design of a transistorised f.m. receiver.

In the first-class service area of an f.m. transmitter the field strength is greater than 1 mV/metre: in the second-class service area (where there may be slight interference from car ignition systems) the signal strength is between 250 μV/metre and 1 mV/metre. The signal delivered to the input terminals of a receiver from a half-wave dipole aerial via a properly-matched feeder is approximately equal to half the field strength, i.e., is 125 μV at the edge of the second-class service area. These signal strengths are however measured using an elevated dipole in the open air. A band-II dipole is 5 feet long and in a portable receiver (such as a transistorised receiver is likely to be) a much smaller aerial must be used: this inevitably gives a smaller signal than an elevated resonant dipole. Moreover portable receivers are likely to be used inside buildings or on the ground where the signal strength can easily be 30 dB below the values measured with an elevated resonant aerial. Thus under unfavourable conditions the signal delivered to the aerial terminals of a portable f.m. receiver can be less than 5 μV.

To obtain good results from an input signal as small as this, a superheterodyne receiver must be used. It is advisable, too, to employ an r.f. stage before the frequency changer: this improves signal-to-noise ratio and helps to minimise oscillator reradiation. The r.f. stage and frequency changer are unlikely to provide a voltage gain of more than 15 and the minimum signal to be expected from the frequency changer is 75 μV. For reasons given earlier a ratio detector is the most likely choice and for adequate a.m. rejection this requires an input to the diodes of not less than 0·5 volt. The i.f. amplifier thus requires a gain of 7,000 and this

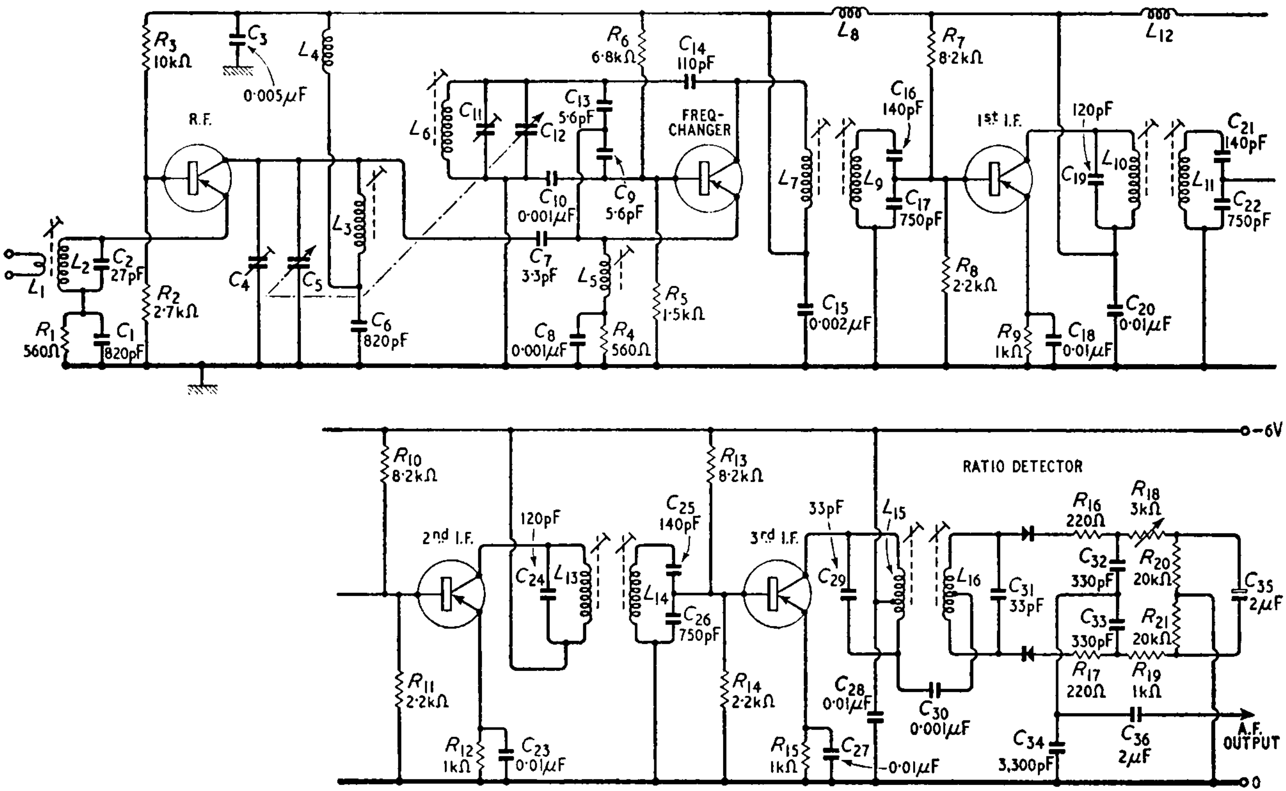


Fig. 11.14. Circuit diagram of a transistorised f.m. receiver up to the a.f. output from the detector

determines the number of i.f. stages required. We have already shown that a 10·7-Mc/s i.f. stage using a non-unilateralised drift transistor gives a voltage gain of about 10 when working into a following transistor and about 65 when working into a ratio detector. A total of three i.f. stages will therefore provide the required gain and the receiver circuit diagram up to the a.f. output from the ratio detector can take the form illustrated in Fig. 11.14. In this the circuit arrangements of frequency changer, i.f. amplifier and ratio detector are all identical with those given earlier. The r.f. stage is a common-base amplifier which at frequencies around 100 Mc/s gives as much gain as a common-emitter type. The input tuned circuit L_2C_1 is heavily damped by the low input resistance of the amplifier and gives so low an effective Q value that there is no advantage in having variable tuning. If the tuning is fixed at the centre of the band to be received the loss due to this input circuit at the ends of the band is negligible.

The a.f. output from the ratio detector is small enough, when the receiver is tuned to a weak signal, to require the use of a three-stage audio amplifier. This may consist of a gain stage followed by a driver stage which in turn feeds a push-pull class-B output stage.

The receiver thus includes a total of nine transistors (five drift types and four uniform-base types) and two point-contact diodes.

Alignment

The receiver should be aligned in the following manner. With the tuning capacitors at minimum capacitance trimmer C_{11} is adjusted to tune in a signal at the high-frequency end of the band—say at 100 Mc/s. Then with the tuning capacitors at maximum capacitance L_6 is adjusted to tune in a signal at the low-frequency end of the band—say at 87·5 Mc/s. The adjustment of C_{11} is now repeated at 100 Mc/s with the tuning capacitors at minimum capacitance. The adjustment of L_6 is next repeated at 87·5 Mc/s with the tuning capacitors at maximum capacitance and these two adjustments are repeated until no further adjustment is required. The receiver now covers the required band. C_4 is adjusted to provide maximum gain at the high-frequency end of the band and L_3 to provide maximum gain at the low-frequency end of the band. These adjustments too, are repeated until no further improvement in gain is obtainable. Finally L_2 is adjusted to give maximum gain at the centre of the band—say at 94 Mc/s.

12

MULTIVIBRATORS

Introduction

TRANSISTORS can be successfully employed in practically all thermionic-valve circuits which use triodes. For example, there are transistor discharger circuits, transistor Miller-integrators, transistor bootstrap circuits and transistor multivibrator circuits to name only a few. We cannot describe all such circuits in one short chapter but we shall describe the transistor multivibrator in some detail.

BI-STABLE MULTIVIBRATOR

The basic circuit of a bi-stable multivibrator using transistors is given in Fig. 12.1. In such a circuit, as in all relaxation oscillator circuits, the transistors are at all times in one or other of two states, that is to say either fully conducting (with collector-emitter potential nearly zero) or completely cut off (with collector current zero except for the inevitable leakage current I_{c0}'). The dissipation in the transistors is low in either state and there is no need for protective circuits to avoid excessive dissipation as in class-A amplifiers. The circuit of Fig. 12.1 is so arranged that when one transistor is conducting, its collector potential ensures that the other transistor is cut off. Thus the circuit has only two possible states:

(1) TR1 conducting and TR2 cut off
(2) TR2 conducting and TR1 cut off

and the circuit is termed bi-stable because without a stimulus from an external source it will remain indefinitely in either state.

If the circuit is initially in state (1), an external signal will cause it to enter state (2) and the change of state is accomplished with great rapidity. The speed of transition is due to the positive feedback inherent in the circuit: this feedback is obvious if the circuit is regarded as that of a two-stage direct-coupled amplifier with the output voltage returned to the input. A second external signal applied to the circuit will cause it to return to state (1) again.

During these changes of state the collector potentials of TR1 and TR2 alternate between a maximum negative value when the transistor is cut off and a minimum when it is fully conducting.

If a recurrent triggering signal is applied to the circuit, the voltage waveform generated at each collector is approximately rectangular in form. Moreover, the frequency of the rectangular signal is half that of the triggering signal, enabling such circuits to be used in frequency dividers or binary counters.

Speed-up Capacitors

In the simple form illustrated in Fig. 12.1 the waveforms generated at the collectors during changes of state have poor rise times. This is because the circuit coupling each collector to the other base consists of a series resistor feeding into the parallel RC combination of the transistor input impedance. Such a network has a response which falls as frequency rises and thus the positive feedback in the circuit is more marked at low than at high frequencies. To achieve steep edges (small rise times) in the output signals the degree

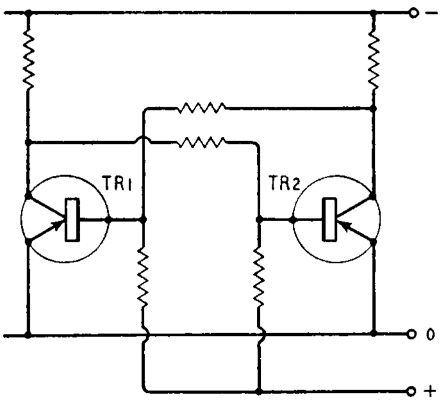


Fig. 12.1. Basic circuit of bistable multivibrator using transistors

of feedback should be independent of frequency: thus the coupling circuits should not have a high-frequency loss. The coupling circuits can be made aperiodic (i.e. non-frequency-discriminating) by shunting the series resistors with capacitors as shown in Fig. 12.2 and the condition for an aperiodic response is that the time constant of the coupling resistor-coupling capacitor combination should equal that of the transistor input impedance. If a smaller value of capacitor is used the rise time of the multivibrator output signals is not improved as much as is possible: if a larger value of capacitor is used the negative-going changes in the collector potentials have a poorer rise time as shown by the dotted curve at the bottom centre of Fig. 12.6 (p. 178).

To avoid the necessity for the separate positive supply for the emitters shown in Fig. 12.1, the two emitters may be bonded and

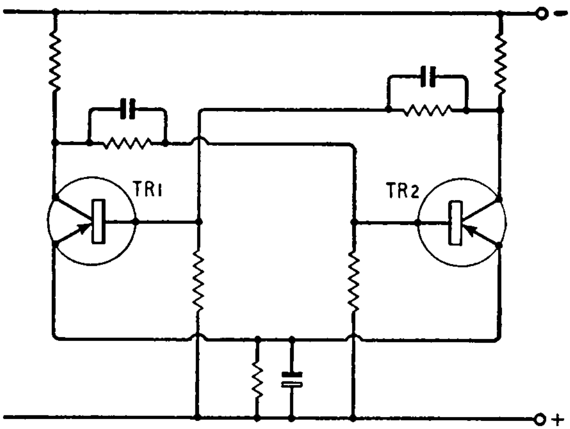


Fig. 12.2. Transistor bi-stable multivibrator with speed-up capacitors

connected to the earth line as shown in Fig. 12.2. In this circuit the emitter current of the conductive transistor ensures that the other is cut off by making its emitter potential more negative than its base potential. It is advisable to decouple the common emitter resistance to preserve steep edges in the output waveform.

Triggering signals applied to a bi-stable transistor circuit to interchange the states of conduction or non-conduction can be applied to the base or the emitter circuit. It is preferable, however, to apply signals to the base because this takes advantage of the α' of the transistors to amplify the triggering signals. If pnp transistors are used, a positive-going signal is required to cut off a conductive transistor or a negative-going signal to make a cut-off transistor conductive. Of these two alternatives the first is in general preferable because it requires a smaller voltage signal, there being no large bias to overcome. Moreover, a triggering signal applied to a conductive transistor is amplified by the transistor causing a much larger triggering signal to be applied to the second transistor.

In counter circuits it is usual to inject a series of unidirectional pulses into a bi-stable circuit and some means is required of directing the pulses alternately to the base circuits of the two transistors. One method of achieving this is by the use of the diode gate circuit illustrated in Fig. 12.3. The input signals are applied to the two capacitors C_1 and C_2 and then to the two base circuits via diodes D1 and D2. The switching action of the diodes can be followed from the following section.

Diode Gate Circuit

Suppose TR1 is conductive. The collector current is a maximum and the collector potential is very low (in practice commonly within 0·1 volt of the emitter potential). The base-emitter voltage is always small and thus there is little difference between the collector and base potentials. The diode D1 is connected between collector and base (via R_1) and there is little potential difference across the diode. The diode is so connected that any positive-going edge applied to capacitor C_1 makes the diode conduct, i.e., low-resistance and the edge is conducted to the base of TR1.

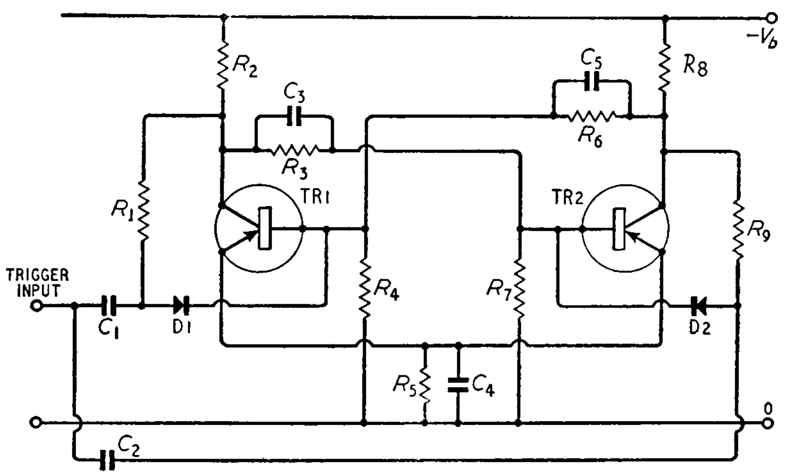


Fig. 12.3. The circuit of Fig. 12.2 with diode triggering gate

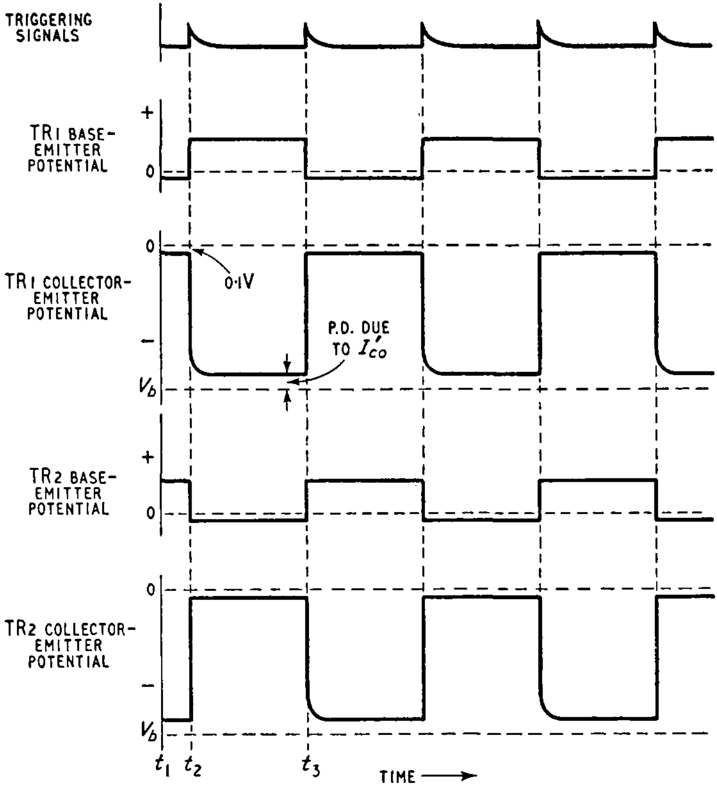


Fig. 12.4. Base and collector waveforms for the bi-stable multivibrator of Fig. 12.3

The edge cannot, however, reach the base of TR2: this transistor is cut off and its collector current is a minimum (consisting of leakage current I_{c0}' only). The collector potential of TR2 is very high and, except for the voltage drop across R_8 due to leakage current would be equal to the supply voltage. There is thus a maximum potential difference across the diode D2 which is connected between collector and base (via R_9). Moreover this potential difference biases D2 in the reverse direction so that it does not conduct when a positive-going edge is applied to C_2 unless this edge has sufficient voltage to offset the static bias (and the amplitude of the edge must be controlled to avoid this).

A positive-going edge applied to C_1 and C_2 thus reaches the base of TR1 but not that of TR2. This edge initiates the regenerative change of state characteristic of multivibrators which ends with TR1 cut off and TR2 conducting. A second positive-going edge applied to the circuit is now conducted to the base of TR2 but not to that of TR1. This triggers off a second change of state, hastened by regeneration, which ends with TR2 cut off and TR1 conductive. This is the state originally postulated.

Waveforms

The waveforms for a triggered bi-stable multivibrator are given in Fig. 12.4. At the time t_1 TR1 is conductive, the base-emitter

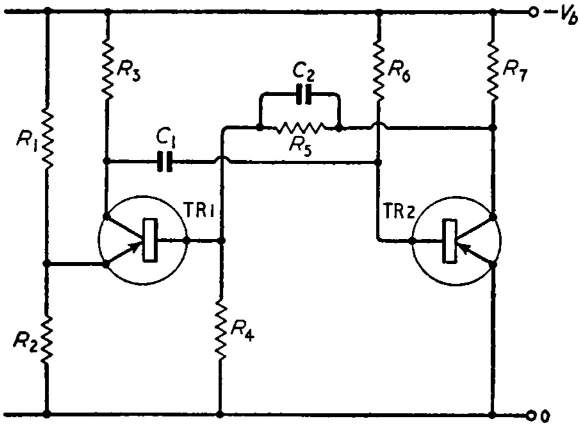


Fig. 12.5. Transistor monostable multivibrator

potential being zero and the collector-emitter potential a minimum. At the same instant TR2 is cut off, the base-emitter potential being positive and the collector-emitter potential a maximum. When $t = t_2$ a triggering pulse is received by TR1 and interchanges the states. At $t = t_3$ a second triggering edge is received (by TR2 this time) and interchanges the states again. Two triggering signals are thus required to produce one complete rectangular wave from TR1 or TR2 collector circuit.

MONOSTABLE MULTIVIBRATOR

If a coupling resistor is removed from a bi-stable multivibrator, the resulting circuit has one direct and one capacitance inter-transistor coupling. Such a circuit is illustrated in Fig. 12.5.

TR1 is biased by the potential divider R_1R_2 but TR2 has no bias components other than R_6 which is returned to the negative supply line from the base.

The mode of operation of a circuit of this type is quite different from that of a bi-stable circuit and will be illustrated by reference to the waveforms given in Fig. 12.6.

At a time t_1 TR2 is assumed conductive: the d.c. conditions in the circuit are such that TR1 is cut off. The circuit will remain in this state indefinitely unless it is compelled to leave it by an externally-applied signal.

Thus this state (TR2 conductive and TR1 cut off) is a stable one and is similar in nature to either of the states of a bi-stable multivibrator.

At t_2 TR1 is made conductive by applying a negative-going edge to its base or preferably (as shown in Fig. 12.6) by applying a positive-going edge to TR2 base. The collector potential of TR1 makes a positive excursion (*AB* in Fig. 12.6). The voltage across a capacitor cannot be changed instantaneously and this edge is transferred without loss in amplitude by C_1 to the base of TR2 (*ab*), cutting TR2 off. The circuit is now in the second of its two possible states (TR1 conductive and TR2 cut off) but it cannot remain in it permanently because C_1 begins immediately to discharge through R_6 and the output circuit of TR1. As the discharge proceeds the potential at the base of TR2 becomes less positive (*bc*) until a point (*c*) is reached at which TR2 begins to conduct. As soon as TR2 is conducting strongly enough for the overall gain of the two transistor stages to exceed unity, regeneration occurs and the circuit changes state rapidly leaving TR2 conducting and TR1 cut off. The circuit is now back in its stable state again.

The state in which TR1 is conducting and TR2 cut off is thus an unstable state because the circuit cannot remain indefinitely in it. It will automatically revert to the stable state after an interval without the aid of external triggering signals. External signals are needed to bring about the change from the stable to the unstable state but not to accomplish the reverse change. The duration of the unstable state can be given any desired value within certain limits by appropriate choice of circuit constants.

Circuits such as this which have one stable and one unstable state are termed monostable. Their chief application is in producing accurately timed delays.

The delay is equal to the period of the unstable state and we thus need to be able to calculate the component values needed to

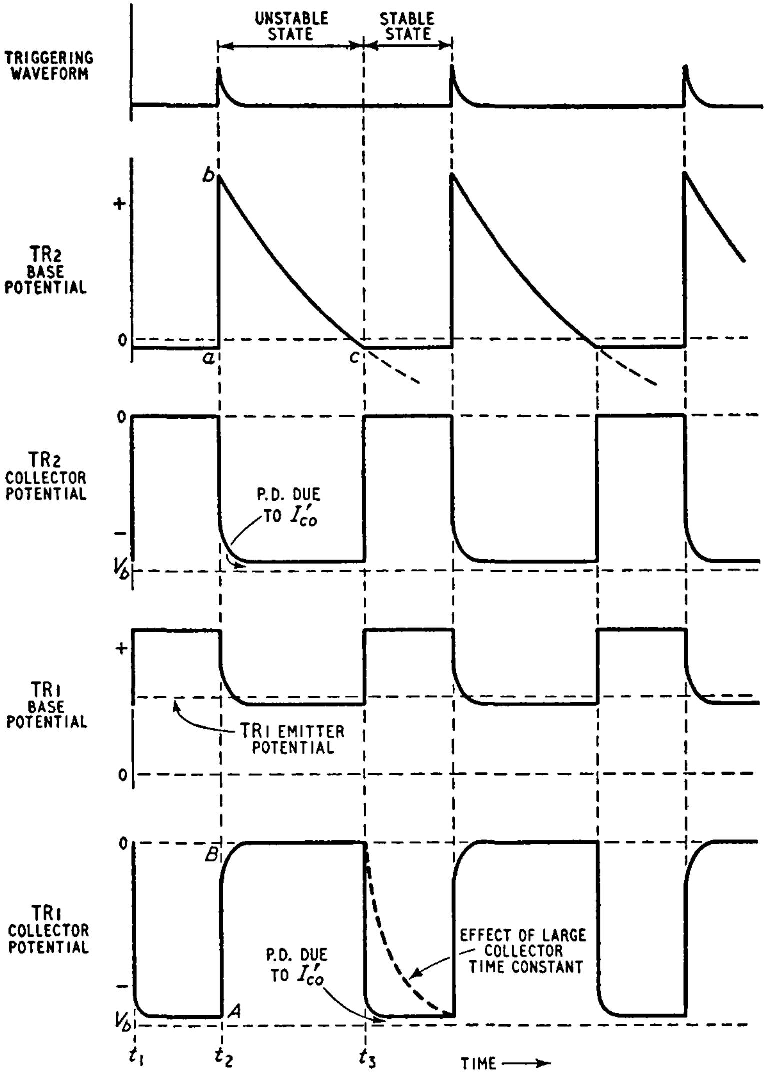


Fig. 12.6. Base and collector waveforms for the monostable multivibrator of Fig. 12.5

produce a wanted delay. This can be achieved in the following manner.

Duration of Unstable State

The duration of the unstable state is equal to the time taken for the potential at the base of the non-conductive transistor to fall almost to zero.

First consider the conditions in the circuit in the stable state: TR1 is cut off and TR2 conductive. If TR1 were an ideal transistor its collector potential would be equal to that of the supply $-V_b$. In practice there is a voltage drop across R_3 due to the leakage current I_{c0}' in TR1. However, provided we are prepared to accept an approximate answer we can neglect the voltage drop due to I_{c0}' and assume that the collector potential is $-V_b$. TR2 is conductive and its base potential is slightly negative with respect to earth potential: again we can take this potential as being equal to earth potential to obtain an approximate answer. Thus the voltage across C_1 is equal to V_b, the right-hand plate being positive with respect to the other.

The unstable period begins when TR1 is suddenly switched on. The collector current rises at this instant to a value which is sufficient with proper design to bring the collector potential almost to zero. This excursion is handed on to TR2 by C_1 and the base potential immediately becomes $+V_b$. C_1 now begins to discharge through R_6 (and the output circuit of TR1 but the resistance of this circuit is normally small compared with R_6 and will be neglected) and if the discharge were completed the base potential of TR2 would fall exponentially with time from $+V_b$ to $-V_b$. However, as soon as the base potential reaches zero (the halfway point in the discharge) TR2 begins to conduct and the unstable period is abruptly terminated.

Thus the duration of the unstable state is given approximately by the time taken for the voltage in a circuit of time constant R_6C_1 to fall to half its initial value. The fall in voltage is given by the equation

$$V_t = V_0 e^{-t/R_6C_1}$$

where V_t is the voltage after an interval t and V_0 is the initial voltage. This may be written in the form,

$$\log_e \frac{V_0}{V_t} = \frac{t}{R_6C_1}$$

from which

$$t = R_6C_1 \log_e V_o/V_t$$

The time taken for the voltage to fall to half its initial value is obtained by putting V_o/V_t equal to 2 in the above expression. We then have

$$t = R_6C_1 \log_e 2$$
$$= 0{\cdot}6931R_6C_1$$

Thus to obtain a duration of say 1 millisecond

$$R_6C_1 = \frac{t}{0{\cdot}6931}$$
$$= \frac{1}{0{\cdot}6931} \text{ millisec}$$
$$= 1{\cdot}44 \text{ millisec}$$

Any values of R_6 and C_1 would thus appear to be suitable provided their product (time constant) is 1·44 milliseconds. In fact, if a good approximation to rectangular output signals is required, there are limitations to the values of R_6 and C_1 which can be used.

When TR1 is cut off at the end of the unstable period, its collector potential tends to go negative. This tends to drive the base of TR2 negative but as soon as TR2 starts to conduct its input resistance becomes very low and the base potential is effectively stabilised near zero. Thus the collector potential of TR1 can go negative only by charging C_1 through the load resistor R_3. The time constant R_3C_1 in fact determines the rise time of the voltage edge generated at TR1 collector and this edge may be used for triggering other circuits when the monostable circuit is used as a delay generator. This edge must thus be steep and this requires a small value of time constant R_3C_1. (The effect of a large value of R_3C_1 is shown in dotted lines in Fig. 12.6.) R_3 cannot be made less than a certain value, otherwise an abnormally-high value of collector current will be required to reduce the collector potential to near zero when TR1 is conducting. If the supply is -6 volts R_3 can be 3 kΩ and a collector current of 2 mA is required—a reasonable value. This is the minimum value of R_3 and a short rise time can be achieved only by using small values of C_1. If, however, C_1 is made small, R_6 must be large to give the desired value of time constant R_6C_1 on which the duration of the unstable period depends. R_6 supplies the base current to TR2 when TR2 is conductive and this sets an upper limit to the value of R_6 which can be used. For example if TR2 is to take 2 mA collector current and if α' if 50, the base current

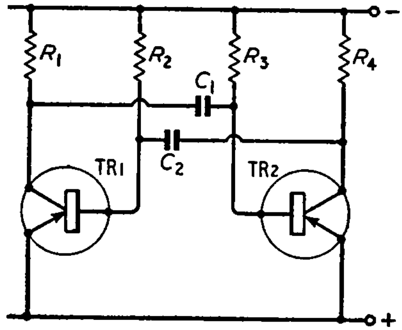


Fig. 12.7. Transistor astable multivibrator

is given by 2/50 mA, i.e. 40 μA. If the supply voltage is 6 volts R_6 is given by $6/(40 \times 10^{-6})$, i.e., 150 kΩ. This is the maximum permissible value and it is advisable to use a smaller value, say, 100 kΩ to ensure that the collector potential is near zero when TR2 conducts. Thus the minimum value of C_1 which can be used is given by

$$C_1 = \frac{1{\cdot}44 \times 10^{-3}}{R_6}$$
$$= \frac{1{\cdot}44 \times 10^{-3}}{100 \times 10^3} \text{ F}$$
$$= 0{\cdot}014 \ \mu\text{F}$$

The time constant governing the positive edge developed at the collector of TR1 is given by

$$R_3C_1 = 3 \times 10^3 \times 0{\cdot}014 \times 10^{-6} \text{ second}$$
$$= 0{\cdot}042 \text{ millisec}$$

The rise time is approximately twice the time constant and is thus less than 0·1 millisecond.

ASTABLE MULTIVIBRATORS

If both the coupling resistors are removed from a bi-stable multivibrator, the resulting circuit has two capacitance couplings. Such a circuit has two unstable states and automatically switches from one to the other continuously without need for external triggering signals. Circuits of this type are termed astable and are free-running relaxation oscillators. They can, however, readily be synchronised at the frequency of a recurrent signal applied to them.

Fig. 12.7 gives the circuit diagram of an astable multivibrator. No emitter bias is used and both bases are returned via resistors to the negative supply line. The behaviour of the circuit will be illustrated by the waveforms shown in Fig. 12.8. At t_1 TR1 is cut off by a positive signal at the base and the collector potential is a negative maximum. At the same time TR2 is conducting, having a slightly negative base potential, the collector potential being nearly zero. TR2 is in a stable state but TR1 is not, because the potential at its base is moving negatively as C_2 discharges through R_2. At t_2 the potential at TR1 base is sufficiently negative for TR1

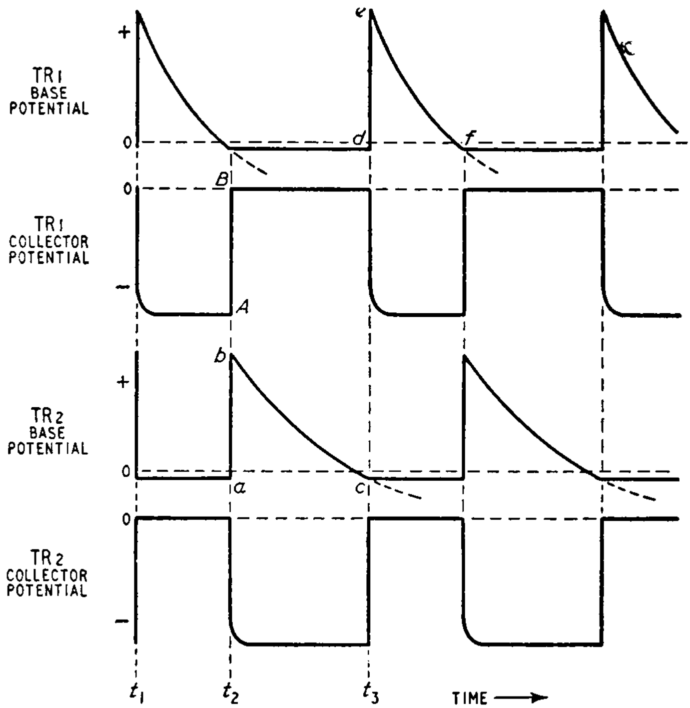


Fig. 12.8. Base and collector waveforms for the astable multivibrator of Fig. 12.7

to start conducting. As explained for the monostable multivibrator, this starts a regenerative action which causes TR1 to become abruptly conductive. The resultant steep positive voltage (*AB*) at TR1 collector is transferred by C_1 to the base of TR2 (*ab*), cutting TR2 off. There now follows the exponential change in base potential (*bc*) previously described. It keeps TR2 cut off for a period given approximately by $0{\cdot}69R_3C_1$ seconds which terminates at time t_3 when TR2 becomes abruptly conductive. TR2 is now in a stable state but TR1 is not, being cut off by a positive edge (*de*) which decays exponentially (*ef*). This keeps TR1 cut off for a period given approximately by $0{\cdot}69R_2C_2$. This is the circuit condition assumed initially. Thus the cycle continues, the period of oscillation being equal to the sum of the durations of the two unstable periods

$$T = 0{\cdot}69(R_3C_1 + R_2C_2)$$

If $R_2 = R_3 = R$ and $C_1 = C_2 = C$ the multivibrator is symmetrical and generates square waves (equal mark space ratio) at both collectors. The period of oscillation is given by

$$T = 1{\cdot}38RC$$

The free-running frequency of such a multivibrator is given by

$$f = \frac{1}{T}$$
$$= \frac{1}{1{\cdot}38RC}$$

A multivibrator required to be synchronised at the line frequency (approximately 10 kc/s) of the British Television System would be designed to have a natural frequency somewhat lower than this—say 7 kc/s. We have already seen that suitable values for the load resistors and base resistors are 3 kΩ and 100 kΩ: we have now to determine a suitable value for the coupling capacitors. From the above expression we have.

$$C = \frac{1}{1{\cdot}38Rf}$$
$$= \frac{1}{1{\cdot}38 \times 10^5 \times 7 \times 10^3}\ \mathrm{F}$$
$$= 0{\cdot}001\ \mu\mathrm{F}\ \text{approximately}$$

SYNCHRONISING OF MULTIVIBRATORS

Bi-stable and monostable multivibrators are triggered by signals which cause the circuit to leave a stable state. Such a technique cannot be applied to astable circuits because these have no stable states. The synchronising signals injected into an astable circuit are designed to terminate the unstable periods earlier than would occur naturally. It follows that the natural frequency of the circuit must be lower than the frequency of the synchronising signals.

To terminate the unstable periods unnaturally early the synchronising signals can take the form of negative-going pulses applied to one or both of the bases of the transistors. The amplitude of the edges is important. If the natural frequency of the multivibrator is small compared with the frequency of the synchronising signals, a small-amplitude signal may cause the multivibrator to run at a simple fraction, say 1/5th of the sync frequency. Increase in the amplitude of the synchronising signal may cause the multivibrator frequency to jump to $\frac{1}{4}$ of the synchronising frequency. Further increase may cause the frequency to jump to $\frac{1}{3}$ the synchronising frequency, etc. Thus a synchronised multivibrator can be used as a frequency divider but close control of the synchronising signal amplitude is necessary to obtain a consistent division ratio.

HOLE STORAGE

The voltage step generated at the collector of a transistor is steeper when the transistor is abruptly turned on than when it is turned off. One reason for this has already been suggested: when a transistor is cut off the collector potential can only change at the rate determined by the time constant of the collector load resistance and the following coupling capacitor. When the transistor is turned on, however, the state of charge of the coupling capacitor is not altered and the step can thus be very steep.

There is, however, another effect which tends still further to degrade the rise time when the transistor is cut off. This is an effect, internal to the transistor, which causes the collector current to persist for a few microseconds after the emitter current has been cut off. This occurs in transistors which have been driven hard into conduction so that the collector-emitter potential is nearly zero. In these circumstances the emitter injects more current carriers (holes in a pnp transistor) into the base region than are required to give the collector current and the excess carriers are stored ready to be swept into the collector region to prolong the collector current for a short period after the emitter current has been cut off. This hole-storage effect can be avoided by so designing the circuits that the collector-emitter potential does not approach zero but is limited to a value such as 1 volt: this can be achieved by use of diodes.

13

FURTHER APPLICATIONS OF JUNCTION DIODES AND TRANSISTORS

Introduction

IN this chapter we shall describe a number of miscellaneous applications of junction diodes and transistors which do not properly belong in earlier chapters.

Simple Voltage-stabilising Circuit

It was pointed out in Chapter 1 that junction diodes could be used as a source of stable voltage and to begin this chapter we shall describe a simple circuit suitable for use where only a small current is required from the stable voltage source.

An example of such a requirement may occur in a transistorised car radio where the supply for the oscillator transistor may require stabilising against changes in car battery voltage to secure good frequency stability and hence stable tuning. A suitable circuit is illustrated in Fig. 13.1. The junction diode must have a breakdown voltage equal to the value of the stabilised voltage required

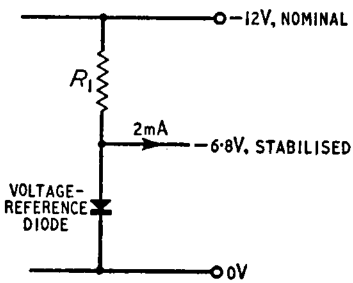


Fig. 13.1. Simple junction-diode voltage stabilising circuit suitable for small currents

and the value of R_1 must be chosen to give an operating point on the nearly-vertical part of the diode characteristic (Fig. 13.2). The diagram represents conditions in the circuit. The load line AB meets the axis at A at a voltage equal to the supply voltage, say 12 volts. AB meets the diode characteristic at B and this point corresponds with the stabilised output voltage. OC represents the voltage drop across the diode and AC the voltage drop across R_1. In the chosen example the stabilised voltage is 6·8 volts.

The slope of AB corresponds to the resistance R_1 and this clearly may vary within limits (as suggested by the dotted lines AD and AE) without much effect on the value of the stabilised voltage but

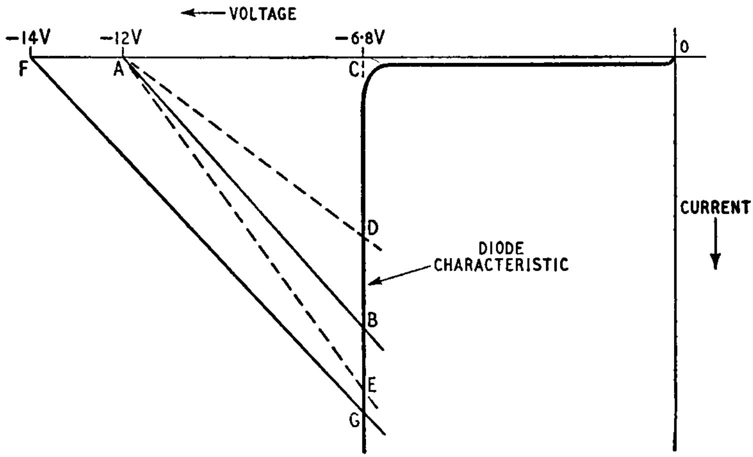


Fig. 13.2. Illustrating the operation of a simple voltage-stabilising circuit

it is preferable to choose a value for R_1 which keeps the dissipation in the diode well within the maximum value prescribed by the makers. For example if the maximum dissipation is 50 mW we can choose to dissipate half of this, 25 mW, at a battery voltage of 12 volts. The voltage across the diode is 6·8 volts and the diode current must be 25/6·8, i.e., 3·7 mA. This current is supplied via R_1 together with the current for the load (the oscillator transistor). If the load current is 2 mA, the total current in R_1 is 5·7 mA. The voltage across R_1 is 5·2 volts and the required value of R_1 is given by

$$R_1 = \frac{5{\cdot}2}{5{\cdot}7 \times 10^{-3}}$$
$$= 900\ \Omega \text{ approximately}$$

The supply voltage may easily rise to 14 volts when the car dynamo is running. The effect such a voltage rise has on the stabilised voltage is illustrated by the load line FG which is parallel to AB (thus representing the same value of resistance R_1) but meets the axis at F corresponding to 14 volts. FG meets the diode characteristic at G representing a greater diode current than before (point B). The new stabilised voltage corresponds to point G which, because of the extreme steepness of the diode breakdown characteristic is at almost the same voltage as before (point B). The dissipation in the diode is now greater than before the increase in supply voltage and care must be taken to see that the maximum safe dissipation is not exceeded when the supply voltage is at its maximum.

The effectiveness of the circuit depends on the steepness of the diode characteristic which is usually expressed as a slope resistance. This may be as low as 5 ohms, showing that a change of diode current of 10 mA gives an alteration in breakdown voltage of only $5 \times 10 = 50$ mV.

Voltage-stabilising Circuit including Transistor and Voltage-reference Diode

The maximum current which can be drawn from a simple voltage-stabilising circuit of the type described above is limited but larger currents can be obtained by use of a current amplifier in conjunction with a voltage-reference diode. One possible circuit is illustrated in Fig. 13.3. This can be regarded as an emitter-follower (analogous to a cathode follower), the base (grid) of which is held at a constant voltage by a junction-diode circuit of the type described above. Such a circuit can provide a current magnification of, say, 50, and, provided a suitable transistor is used, currents of the

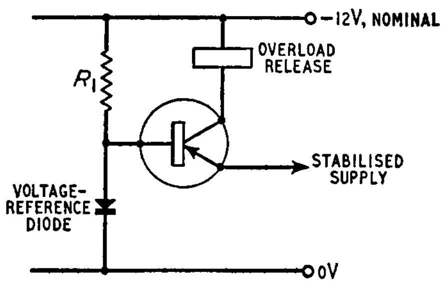


Fig. 13.3. Voltage-stabilising circuit suitable for supplying large currents

order of 1 A can be supplied, the input current to the transistor being of the order of 20 mA.

The regulation of a circuit of this type can be determined in the following way. Transistors capable of supplying 1 A may require a base-emitter voltage of 500 mV to give this value of emitter current. It follows that the output voltage must change by 500 mV when the current drawn from the circuit changes by 1 A. The effective source resistance is thus 0·5 ohm.

There is a danger that the transistor could be damaged if the output of the regulated supply were short-circuited for this would cause a large current to flow through the transistor causing excessive dissipation in it. The transistor can be protected against such damage by including in the collector circuit a quick-acting overload release device or a resistor which limits the collector current to a safe value.

Junction-diode a.f.c. Circuit

As mentioned in Chapter 1 the capacitance of a reverse-biased junction diode varies with the bias voltage. Such a diode can therefore be used for automatic frequency control and Fig. 13.4 gives a circuit diagram which can be used in an f.m. receiver for this purpose. Not all junction diodes are suitable for this application: for some types the damping due to the resistive component of the diode impedance may be sufficient to reduce the oscillation amplitude to a low value or even to prevent oscillation altogether. Diodes with very low damping have been developed for use in a.f.c. circuits.

L_1C_1 is the oscillator tuned circuit and the junction diode is connected across the circuit via the isolating capacitors C_2 and C_3.

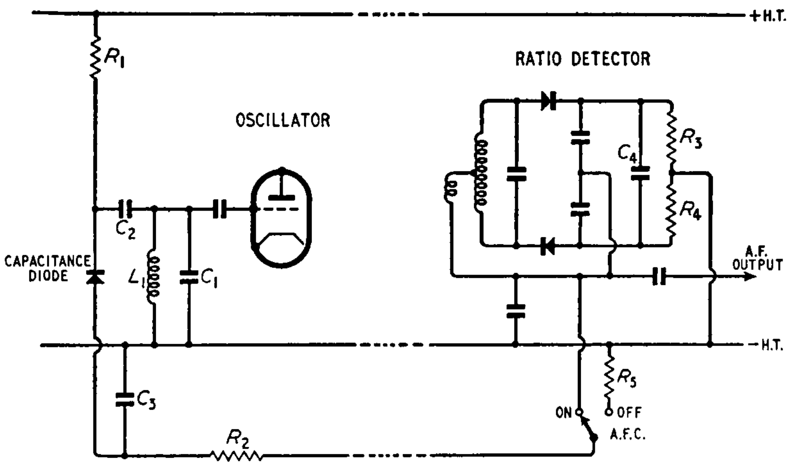


Fig. 13.4. Circuit illustrating the use of a junction diode to give a.f.c.

The diode is reverse-biased from the supply by the potential divider R_1R_2 and provided that R_1 is reasonably high in value, say, more than 40 kΩ, the damping of the oscillator circuit by this resistor should not seriously reduce the oscillation amplitude. The capacitance of the diode is effectively in parallel with C_1 and alteration in diode bias causes an alteration in oscillator frequency. To obtain a.f.c. the diode bias must be controlled automatically by the degree of mistuning and this can be achieved by returning R_2 to the d.c. output of the discriminator. If the discriminator is a Foster-Seeley type, R_2 may be connected directly to the detector output provided the connection is made on the detector side of the output coupling capacitor. If a ratio detector is used the connection of R_2 is not so straightforward.

The d.c. output of a ratio detector circuit of the type illustrated in Fig. 11.8 has two components: one component varies with tuning and reverses in polarity at the correct tuning point, the relationship between voltage and frequency being substantially linear over a limited frequency range. This is the component responsible for the a.f. output of the detector and is the component required for a.f.c. purposes. The second component is a positive voltage obtained from the long time constant circuit $R_3R_4C_4$: this also varies with tuning but has a maximum at the correct tuning point. To obtain good a.f.c. performance it is advisable to eliminate the second component from the d.c. output of the detector. This can readily be done by earthing the mid-point of the resistor as shown in Fig. 11.4. This gives what might be termed a " balanced " form of ratio detector which gives zero output voltage at the correct tuning point: the d.c. output of such a detector varies with tuning in the same manner as that of a Foster-Seeley discriminator and R_2 can then be returned to the detector as shown in Fig. 13.4. The junction diode bias must not be affected by a.f. signals in the detector output and these are therefore prevented from reaching the diode by the capacitor C_3 which forms with R_2 a potential divider which considerably attenuates all audio frequency signals.

A.f.c. circuits of this type can be extremely effective, reducing mistuning effects by a factor of as much as 10 : 1. Manual tuning can be very difficult with a.f.c. and it is desirable to have some means of switching a.f.c. off whilst tuning is being carried out. As soon as the wanted signal is tuned in, a.f.c. is switched on to minimise subsequent tuning drift. Fig. 13.4 indicates one method of switching a.f.c. off. The resistor R_5 is approximately equal to the d.c. resistance of the ratio detector. Such a resistor is necessary to enable the a.f.c. to be switched off and on when the receiver is accurately in tune, without alteration of the bias across the capacitance diode.

Use of Transistor to increase Relay Sensitivity

Electromagnetic relays enable one or more circuits to be switched on and off by a controlling current much smaller than the controlled current. For example a relay requiring 5 mA of input current can control a circuit carrying 5 amperes.

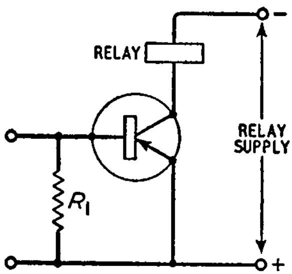


Fig. 13.5. Use of a transistor to increase relay sensitivity

By use of a transistor as a current amplifier, the sensitivity of a relay can be greatly increased. As we have seen, a common-emitter amplifier can easily give a current gain of 50 and by using such an amplifier with the relay mentioned above only 100 μA of input current is needed to control the 5-ampere circuit.

A suitable circuit diagram is given in Fig. 13.5. The base of the transistor is returned to the emitter via R_1 causing a collector current too small to operate the relay. A small input current causes the transistor to conduct and the relay to operate.

A desirable practical precaution is to connect a diode across the relay winding to prevent generation of large collector voltages across the relay coil when the transistor input is removed and the collector current is cut off. Such voltages can exceed the collector breakdown voltage and can damage the transistor.

The use of a transistor with a relay is a convenient circuit arrangement because the transistor requires no supplies other than that required for the relay and takes up very little space.

D.C. Converters

Transistors are particularly useful in the construction of d.c. converters, units which can be made remarkably compact and which can convert power from a low voltage source (e.g., 6 volts) to a higher voltage (e.g., 120 volts) with an efficiency which can approach 85 per cent and is seldom less than 60 per cent.

The transistor in such a converter is used as a switch which interrupts the d.c. supply from the low-voltage source to produce alternating current. This is stepped up in voltage by a transformer or resonant circuit to give a high-voltage supply which is rectified and smoothed to obtain the high-voltage output. For high efficiency the power dissipated in the transistor itself must be small. The power dissipated in a transistor is of course given by the product of the collector current and the collector-emitter voltage. The power is therefore small when the collector current is nearly zero, that is to say when the transistor is cut off and also when the collector-emitter voltage is nearly zero, that is to say, when the transistor is fully conducting. Thus the design of the d.c. convertor must be such that the transistor is at all times either fully conducting or cut off. This is achieved by using the transistor as an astable relaxation oscillator which generates rectangular waves.

The basic principles of one type of d.c. converter circuit can be illustrated as shown in Fig. 13.6 in which the transistor is shown for simplicity as a switch. When the switch is closed a magnetic field begins to grow in the transformer core. This takes a little time to reach its maximum value and whilst it is still growing a voltage equal to nV_b is generated in the secondary winding where $1 : n$ is the step-up turns ratio of the transformer. The polarity of the rectifier is such that this secondary e.m.f. drives a current through the load. If the switch were left closed the magnetic

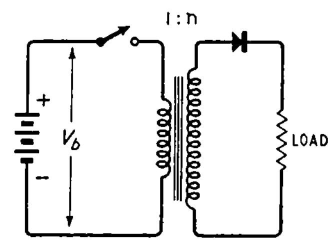


Fig. 13.6. Basic circuit of one type of transistor d.c. converter

field would in time reach its maximum value and become static: there would then be no secondary voltage. However, before this instant is reached, the switch is opened and the magnetic field collapses suddenly, setting up a secondary e.m.f. with a polarity opposite to that set up on closing the switch. This e.m.f. does not cause current to be delivered to the load because the polarity is in the opposite direction to that necessary to drive current through the rectifier.

A simple circuit such as this will behave as a d.c. converter and has the merit of good regulation but the efficiency is low primarily because of the loss of power associated with the e.m.f. generated when the switch is opened. To achieve high efficiency this power

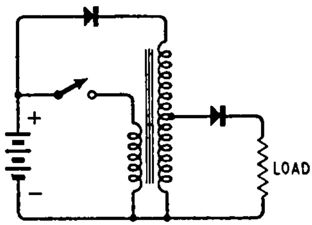


Fig. 13.7. Basic converter circuit with provision for returning reverse secondary e.m.f. to the low-tension battery

must be put to use and one of the methods used is to return the power to the low-tension source as indicated in Fig. 13.7. By replacing the switch by a transistor we obtain the complete circuit of a d.c. converter shown in Fig. 13.8. The collector and base

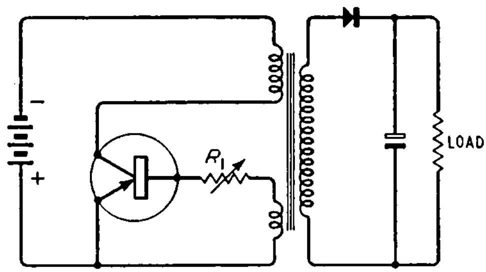


Fig. 13.8. Complete circuit of transistor d.c. converter

circuits of the transistor are coupled to give positive feedback and consequent oscillation. Considerable feedback is necessary to drive the transistor hard into conduction and cut off. The ratio of the periods of conduction and non-conduction can be controlled by adjustment of the value of the resistor R_1, which determines the base bias of the transistor. Frequently d.c. converters of this type (termed the transformer type) employ two transistors operating in push-pull.

In an alternative (ringing-choke) form of d.c. converter, power is delivered to the load during the period of cut off of the switching transistor. As before, the transistor switches the low-tension source across an inductor, thus establishing a magnetic field as before, and then breaks the primary circuit. The e.m.f. generated across the inductor by the collapse of current on breaking the battery circuit causes the rectifier to conduct and thus supplies power to the load. The circuit diagram is thus similar to that of a transformer converter but employs an inductor instead of a transformer. The ringing-choke circuit can be as efficient as the transformer type but the regulation is inferior.

Oscillation frequencies in d.c. converters may lie between 500 c/s and 10 kc/s. If a converter is required to be particularly compact the transformer and smoothing capacitor must be small. This is practicable, provided the working frequency is high, and the tendency is therefore to have high working frequencies in compact converters.

Photo-diode

It is shown in Chapter 1 that the current which flows across a reverse-biased pn junction is carried by minority carriers, i.e., by

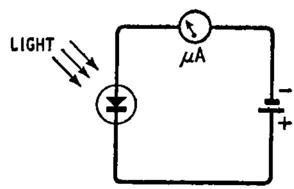


Fig. 13.9. Simple circuit using a photo-diode

the electrons and holes liberated by breakdown of the covalent bonds of the intrinsic semiconducting material. This current is substantially independent of the reverse bias voltage, provided this exceeds approximately 1 volt, but can be increased by heating the material or by allowing light to fall on it: both give the semiconductor atoms more energy and cause more covalent bonds to break. Where sensitivity to light is undesirable junction diodes and transistors are sealed in opaque containers: if sensitivity to light is required a transparent container is employed.

A junction diode in a transparent container is known as a photo-diode and can be used to indicate the presence of light. An obvious form of circuit is that illustrated in Fig. 13.9. The current which flows in such a circuit when the diode is in darkness is due entirely to thermal dissociation of covalent bonds and increases rapidly as temperature rises. It was known as the reverse current in Chapter 1 but in photo-diodes is usually known as the dark current. The ratio of light to dark current thus decreases as temperature rises.

The output power from a photo-diode is limited and amplification is essential if greater power is required, e.g., to operate a milli-ammeter or a relay. Amplification can be provided by a transistor direct-coupled to the photo-diode as shown in Fig. 13.10. In this

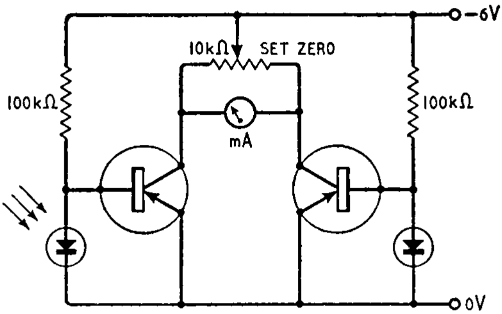


Fig. 13.10. Light-meter circuit employing two photo-diodes and two transistors

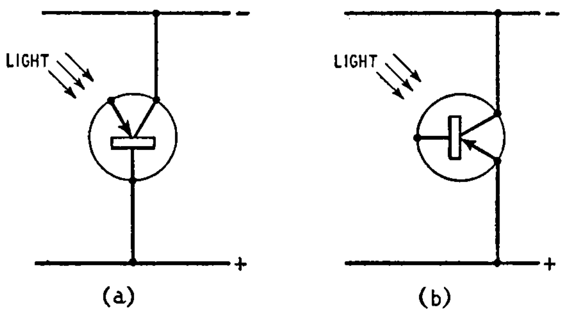


Fig. 13.11. Photo-sensitivity of a transistor arranged as at (b) is considerably greater than when arranged as at (a)

circuit two photo-diodes and two transistors are used in a balanced circuit which largely eliminates the effects of temperature changes and gives a meter reading which depends only on the illumination falling on one of the photo-diodes.

To set up the circuit the two photo-diodes are screened from light and the potentiometer is adjusted to give zero meter reading. When one of the photo-diodes is now exposed to light the meter gives an indication proportional to the illumination and the meter can, in fact, be calibrated in terms of illumination.

Photo-transistors

A photo-transistor may alternatively be used to produce an output greater than is possible from a photo-diode. The mechanism of the amplification inherent in a photo-transistor can be explained in the following way.

First consider a transistor connected to a supply as indicated in Fig. 13.11 (a). It was pointed out at the beginning of Chapter 6 that the small leakage current I_{c0} which flows in such a circuit arises from dissociation of covalent bonds and increases rapidly as tem-

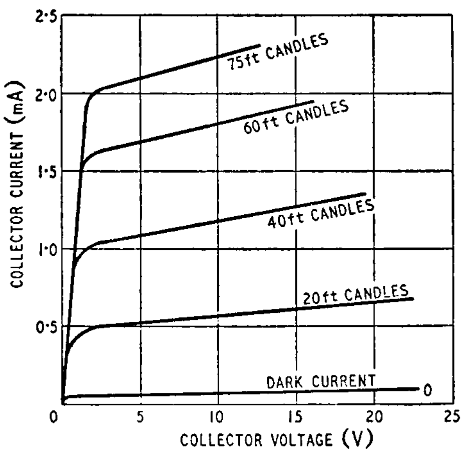


Fig. 13.12. Characteristics of a photo-transistor

perature rises. This current also increases if light falls on the transistor because this also breaks up covalent bonds. The leakage current in this circuit is that of the reverse-biased collector-base junction and is of the same order as that of a photo-diode. Such a transistor circuit therefore provides no greater output than is available from a photo-diode.

Now consider the transistor connected to the supply as shown in Fig. 13.11 (b). Chapter 6 shows that the leakage current I_{c0}' for such a circuit is $(1 + \alpha')$ times I_{c0}, i.e., is much greater than that of the transistor connected as in Fig. 13.11 (a). It is in fact the magnitude of this leakage current which necessitates the use of protective circuits for bias stabilisation in common-emitter amplifiers. The leakage current may be due to hole-electron pairs released by heat or released by light and thus a transistor used in a circuit arrangement such as that of Fig. 13.11 (b) can produce a considerable increase in collector current when light falls on the base region. This is illustrated in Fig. 13.12 which gives the collector current-collector voltage characteristics for a photo-transistor plotted with

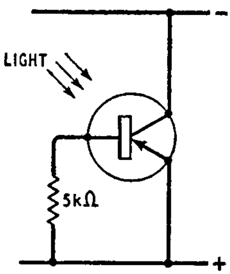


Fig. 13.13. A resistor connected between base and emitter of a photo-transistor can be used to improve the ratio of light current to dark current

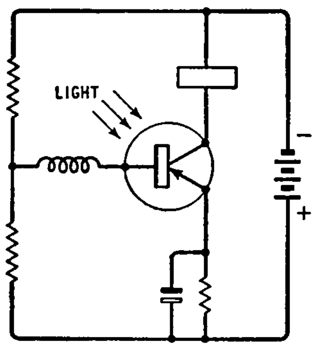


Fig. 13.14. Circuit using a photo-transistor with an interrupted light input

incident light as the parameter. The characteristics are similar in shape to those of a common-emitter amplifier. The curves show that a change of collector current of 0·5 mA can be produced by a change in light input of 20 ft-candles.

If the base circuit of the photo-transistor is open-circuited as shown in Fig. 13.11 (b), the variations in collector current due to temperature changes are considerable. Thus the ratio of light current to dark current is limited and may vary in practice from 100 at 25° C to 10 at 45° C. Provided the photo-transistor is not required to work in surroundings where wide variations in temperature are likely to occur the very simple circuit of Fig. 13.11 (b) may be satisfactory.

Where wide variations in temperature are inevitable it is preferable to have a larger ratio of light current to dark current. This can be achieved by use of a resistor connected between base and emitter as shown in Fig. 13.13. With a resistor of approximately 5 kΩ the ratio of light current to dark current is now increased to 400 at 25° C falling to 20 at 45° C.

There are some applications of photo-electric devices where the light input is " chopped "; this occurs, for example, where the device is used for counting articles moving along a conveyor belt. A photo-transistor circuit suitable for such an application is illustrated in Fig. 13.14. This employs the potential-divider method (see Chapter 6) of dark-current stabilisation but to avoid reduction in light current an inductor is included in series with the base lead to the photo-transistor. To give maximum output this inductor should be parallel-resonant at the frequency of the light variation.

14

FURTHER SEMICONDUCTOR DEVICES

Introduction

THE preceding text has shown the usefulness and the versatility of transistors and has also mentioned their chief disadvantages, namely the temperature-dependence of their characteristics and the restricted upper limit to their working frequencies. Of these limitations the last-mentioned is probably the most serious and efforts are constantly being made to improve the high-frequency performance. There are a number of new techniques under development which seem to offer great promise and there is no doubt that the next few years will see increases in the working frequencies of transistors. In conclusion we shall list what appear

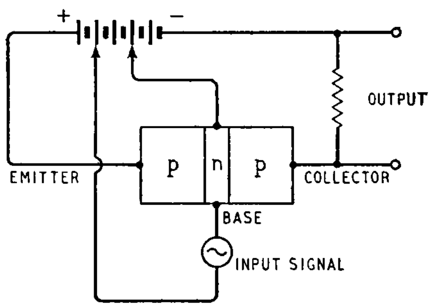


Fig. 14.1. Basic features of tetrode transistor

to be the most important new types of transistor. In some of these, improvements in the high-frequency performance are obtained by reducing the base resistance $r_{bb'}$, the collector capacitance $c_{b'c}$ or both.

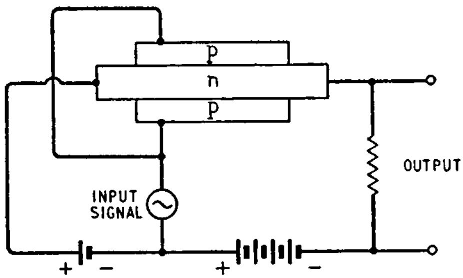
Tetrode Transistor

The tetrode transistor is similar to a conventional pnp or npn triode transistor but has two base connections arranged as shown in Fig. 14.1. One connection is treated as the input terminal as in a conventional transistor and the other is connected to a source of steady potential, negative with respect to the emitter potential in a pnp tetrode transistor. This steady base bias reduces the effective area of the base-emitter junction by forcing the current carriers towards the input connection, thus reducing the effective area of the emitter-base junction and hence the base resistance r_{bb}' and the collector capacitance $c_{b'c}$. The transverse bias of the base region increases the length of path the injected carriers have to travel to reach the collector, increases their chances of neutralisation and thus reduces the value of α and α'.

Surface-barrier Transistor

This type of transistor employs a thin slab of semi-conducting material to the opposite faces of which metal electrodes are attached to form the emitter and collector connections. Two circular recesses are etched in the semi-conducting material until the wall

Fig. 14.2. Basic features of field-effect transistor



is thin enough to give the required high-frequency performance. Electrodes are then deposited in the recesses by electrolytic action, this being essential to obtain the necessary perfection of contact between the electrodes and semi-conducting material. By this means it is possible to produce transistors with cut-off frequencies of the order of 50 Mc/s and with collector capacitances as small as 3 pF.

Field-effect Transistor

This is a form of transistor operating on principles entirely different from those of point-contact and junction transistors. Provided the manufacturing difficulties can be overcome, this type of transistor offers great possibilities as a v.h.f. amplifier. It consists of a thin slab of high-resistance germanium (say n-type) through which a longitudinal current is passed as indicated in

Fig. 14.2. On either face of the slab are two thin layers of low-resistance (p-type) germanium. The pn junction formed between the p-type and n-type material is biased in the reverse direction to produce areas adjacent to the junction where there are very few current carriers.

Thus the longitudinal current flowing in the centre slab is confined to a region near the centre of the slab where current

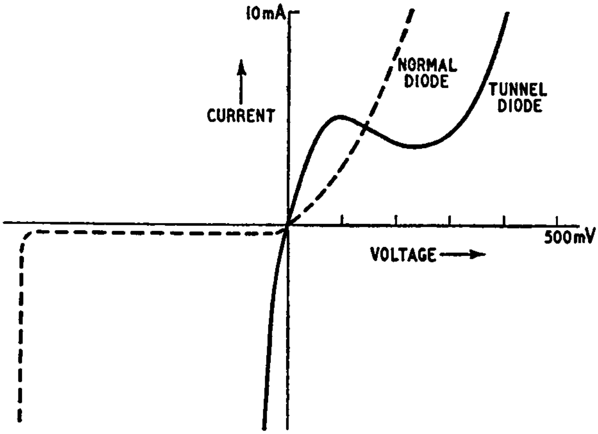


Fig. 14.3. Characteristic of a tunnel diode (solid) compared with that of a normal junction diode (dashed)

carriers still exist. By increasing the reverse bias, the longitudinal current can be reduced and if a signal to be amplified is connected in series with the bias, the longitudinal current can be modulated and an amplified output can be obtained by passing this current through a suitable impedance. The input resistance of such a transistor is that of a reverse-biased diode and is therefore much higher than that of a conventional transistor the input resistance of which is that of a forward-biased diode.

Tecnetron

This is a semi-conducting device of French design which may be regarded as a development of the field-effect transistor. It consists of a crystal of, say, n-type germanium in the form of a cylindrical rod with a narrow neck near the centre surrounded by an indium ring. A current is sent along the rod via electrodes at the ends and a negative control voltage is applied between the ring and one end of the rod. This control voltage produces an electric field within the rod which deflects current carriers towards the centre of the rod. The conducting cross-section and hence the resistance of the rod thus depend on the applied signal. The device is claimed to have a mutual conductance of over 0·5 mA/V at 200 Mc/s with an input resistance of the order of 1 MΩ shunted by 0·2 pF.

Tunnel Diode

A pn junction with a very thin junction region has a characteristic shaped as shown in the solid line in Fig. 14.3: for comparison the characteristic for a diode with a thicker junction region is given in dotted lines. As would be expected breakdown for the thin-junction diode occurs at a very low value of reverse bias and indeed there is in effect no region of high reverse resistance. An unexpected feature of the characteristic is, however, the region of negative slope which occurs at a forward bias: this was first reported by Esaki

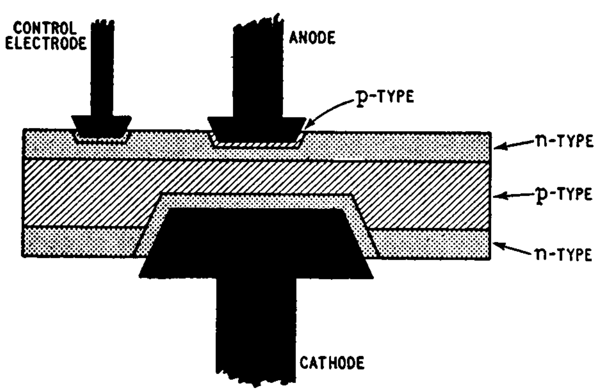


Fig. 14.4. Construction of a controlled rectifier

in 1958. Such a pn junction can be manufactured with very low capacitance and oscillators making use of the negative-resistance kink can function at frequencies as high as thousands of Mc/s. These diodes are termed tunnel diodes and offer great promise as very high-frequency oscillators although the power output is very limited. They can also be used as high-frequency amplifiers but because the tunnel diode has only two terminals there is no isolation between the input and output circuits and it is very difficult to construct a cascaded amplifier.

PNPN Transistor (Trinistor or Controlled Rectifier)

This is a four-layer semiconductor device with a structure of the form shown in Fig. 14.4, which can be represented more simply as in Fig. 14.5. It has two stable states, one in which the resistance is very low (the conductive state) and the other in which the resistance is very high (the non-conductive state). The device can be switched very rapidly from non-conduction to conduction and very little power is needed to bring about this change of state. Thus the

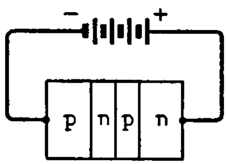


Fig. 14.5. Reverse-biased controlled rectifier

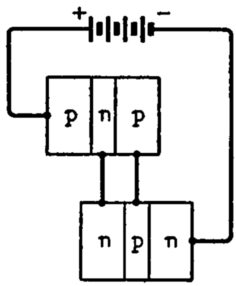


Fig. 14.6. Forward-biased rectifier represented as two direct-coupled transistors

pnpn transistor has properties similar to those of a thyratron (gas-filled valve) but it is far more efficient; the device is used mainly for switching and power control purposes, e.g., as a controlled rectifier.

To understand the mode of action of the pnpn transistor, the device can be regarded as made up of two triode transistors, one of pnp type and the other of npn type, direct-coupled as shown in Fig. 14.6. Suppose a voltage is applied to the outermost regions of the device, as shown in Fig. 14.5. The polarity of this voltage is such as to reverse bias the two outer junctions of the device: these are the base-emitter junctions of the two triode transistors and both triodes are thus cut off. Very little current can flow through the device and with this polarity of applied voltage the pnpn transistor is in its non-conductive state.

Now suppose the polarity of the applied voltage is reversed: the positive terminal of the supply is now connected to the outermost p-region and the negative terminal to the outermost n-region. The applied voltage now biases the two outer junctions in the forward direction and the outermost p- and n-regions can now act as emitters in the two triode transistors. The centre junction (which is the base-collector junction for both triodes) is reverse-biased and most of the applied voltage acts across this junction.

Let the current which flows through the device be I. This current is carried in the pnp transistor by holes which originate in the emitter region, cross the base region and enter the collector region. The fraction of the total number of holes emitted which reach the collector region is α_1, the current amplification factor of the pnp transistor. The fraction $(1 - \alpha_1)$ represents the number of holes which combine with electrons in the base region and so fail to reach the collector. The holes which are lost by combination with electrons constitute a current of $(1 - \alpha_1)I$.

Now the base region of the pnp transistor is also the collector region of the npn transistor and the electrons which combine with the holes in this region are obtained from the emitter of the npn transistor. Electrons are, of course, the current carriers in an npn transistor and the fraction of those leaving the emitter which arrive at the collector is α_2, the current amplification factor of the npn transistor. Electrons reaching the collector thus carry a current of $\alpha_2 I$. If this current equals that carried by the holes from the pnp transistor, we have

$$(1 - \alpha_1)\, I = \alpha_2\, I$$

giving

$$\alpha_1 + \alpha_2 = 1$$

This relationship could also be deduced by considering the inner p-region of the pnpn device. This can be regarded as the collector of the pnp transistor or the base of the npn transistor. By equating the number of holes entering the collector region with the number of electrons lost by combination in the base region, we can again deduce the same result.

This result expresses a limiting condition: provided the sum of α_1 and α_2 is less than unity, the device is stable even though it is forward-biased. If, however, the alphas total unity (each equal to 0·5, say) the transistor is on the verge of instability. Any current amplification occurring in the reverse-biased centre junction (due to ionisation by collision, for example) increases both α_1 and α_2, causing their sum to exceed unity. Equilibrium is then impossible and the current I rises rapidly to a very high value. The process is regenerative because any increase in I causes enhanced ionisation, which in turn increases I. Such breakdown occurs naturally if the applied voltage is raised to a high value (causing a sufficiently intense electric field across the centre junction) and in a controlled rectifier the design might be such that breakdown occurs at an applied voltage of 350 volts.

So far we have assumed that there is no external connection to the inner p- or n-region. Suppose that such a connection is used to apply a negative voltage to the inner n-region (relative to the

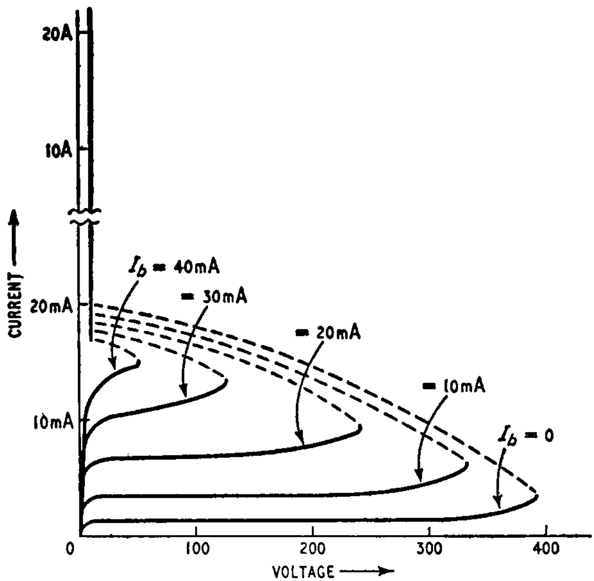


Fig. 14.7. Characteristics of a controlled rectifier

outermost p-region) that this supplies a current of I_b electrons to the device. It is a property of transistors, particularly those employing silicon, that the current gain α depends on the emitter current, being small for small emitter currents and increasing as the emitter is increased. The application of a forward bias current to the inner n-region increases the emitter current from the outer p-region, so increasing the current gain α_1 of the pnp transistor and the sum of the two alphas. Breakdown now occurs more easily, i.e., at a lower applied voltage than in the absence of external bias. Moreover, by increasing I_b, the breakdown voltage can be reduced as low as desired: in practice, the breakdown voltage can be reduced to a few volts only. It is, of course, alternatively possible to control the breakdown voltage by a bias applied to the inner p-region and this has to be made positive with respect to the outermost n-region to reduce the breakdown voltage.

The characteristics of the controlled rectifier are given in Fig. 14.7: they are similar to those of a gas-filled triode valve (thyratron). The greatest control power required is only 5 volts at 100 mA: this can control a current of 50 A at 250 volts. With no control signal the rectifier presents a very high resistance and an applied

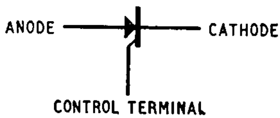


Fig. 14.8. Symbol for controlled rectifier

voltage of 350 volts is required to cause breakdown. The resistance then falls to a very low value and currents of up to 50 A can flow through the rectifier with less than 2 volts drop across it. Once breakdown has occurred, the control signal may be removed but the low resistance will remain until the forward current in the rectifier has fallen to a low value. Thus the device can be fired by very short-duration pulses applied to the control terminal. When firing occurs the build-up of current in the rectifier can be very rapid: rise times of 1 microsecond can be achieved.

The controlled rectifier can be used in a.c. power control circuits, in d.c. converters and in voltage-regulated d.c. supply circuits.

The symbol for a controlled rectifier is given in Fig. 14.8.

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