Mr Millard

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TRANSISTORS

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ACKNOWLEDGMENT

The author is indebted to Messrs. Standard Telephones & Cables Ltd. for permission to use the curves shown in Figures 8 and 9.

CONTENTS

LIST OF SYMBOLS.

A	absolute temperature (zero = − 273°C)
a	current gain (common-base) always less than unity.
a'	current gain (common-emitter) = $\frac{a}{1 - a}$
a'_b	current gain (common-emitter) in the presence of a collector load.
C_c	base-collector capacitance
h_{11}	short-circuit input resistance for common-base connection.
h_{22}	open-circuit output admittance for common-base connection.
h_{12}	reverse voltage feedback ratio for common-base connection.
h_{21}	the current amplification for common-base connection.
h'_{11}, h'_{22}	etc. as above for common-emitter connection
i_e, i_b, i_c	small signal currents (a.c.) e = emitter, b = base, C = collector
I_e,I_b,I_c,	d.c. operating currents.
I_{co}	Base-collector current when I_e is zero (about 10μA).
I'_{co}	emitter-collector current when I_b is zero (about 100μA but very variable)
κ	constant for current-voltage curve of p–n junction, and $\simeq$ 40
n	negative current carrier concentration (electrons)
n_i	intrinsic electron or hole concentration in germanium crystal
r_e,r_b,r_c	components of the active equivalent circuit.
r_m	The resistance which relates emitter current to the open-circuit output voltage $v_2 = r_m i_e$
R_g	internal source resistance.
R_c	collector load resistance.
S	stability factor
$v_i v_2$	signal voltage across input and output terminals respectively.
V_B	Battery voltage.
p	positive current carrier concentration (holes)

TRANSISTORS

THE PHYSICAL OPERATION OF JUNCTION DIODES AND TRANSISTORS

Introduction

The transistor is a device which at present consists of a single crystal of germanium or silicon into which extremely small and carefully controlled amounts of impurity such as arsenic or indium have been introduced, the effect of which is to give the crystal amplifying properties similar to those of a valve.

Progress in transistor development has been so rapid since 1948 that the large-scale application of transistor techniques to electronic engineering has been inhibited until the last few years, because designers of electronic apparatus have been reluctant to embark upon a programme based upon available transistors which, before the production stage is reached, may well be out of date. The situation has become more stable with the advent of the junction transistor although new types giving greater power output or higher frequency operation are in the development stage.

The study of semi-conducting materials which resulted in the development of the transistor has been going on for many years as part of a wider investigation into the nature of solids. *Holes* and *excess electrons*, terms so frequently encountered in present transistor theory, have been quite well-known in the theory of solids since 1931 but the discovery of transistor action in 1948 has given the subject a new technological significance.

From 1940 to 1950 the understanding of semi-conductors greatly increased due to the use of crystal diodes in radar. During an investigation in 1948 into surface conditions on a crystal of germanium, J. Bardeen and W. H. Brattain discovered certain effects which finally led to the invention of the point-contact transistor. The development of transistor theory and the idea of the junction transistor was the work of W. Shockley of the Bell Telephone Laboratories who described its manner of construction and predicted its properties.

The transistor is destined to become increasingly important. Its physical dimensions are small compared with those of a valve; so are its d.c. operating supplies. It is most probable that in electronic apparatus of low power, the transistor will soon be used as commonly as the valve and may eventually replace it. This trend is already apparent in deaf-aid appliances, and in other electronic apparatus where miniaturisation is desirable.

Point-contact transistors, complex in their mode of operation and not predictable in their performance, are being superseded by junction types, so that in this supplement it is proposed to deal only with junction transistors of the *p–n–p* type. The alternative *n–p–n* junction transistor has not so far been developed in this country. As explained later, the term *p* implies positive current carrier, and *n* negative current carrier.

After a brief discussion of the physical nature of *p*- and *n*-type germanium, the rectifying action of the *p–n* junction is described because this leads to a better grasp of the operation of the *p–n–p* junctions which constitute a transistor.

With some physical picture of the internal functioning of the device in mind, the characteristics will be dealt with and a four-terminal equivalent circuit reducing the transistor to a normal piece of a.c. circuitry, will be given. This leads to an investigation of the various methods of operation, contrasting and comparing the transistor's performance, where relevant, with that of the more familiar electronic valve.

The Crystal Structure

There are four outer electrons in the atoms of germanium and silicon, the principal semi-conductors concerned in transistors. In the crystal these atoms are arranged so that by sharing electrons each nucleus becomes associated with eight electrons and forms a very stable arrangement. The link between atoms sharing electrons in this way is known as a *co-valent* bond. (See Fig. 1). (From *co-valency* = the union of two atoms by the sharing of a pair of electrons)*

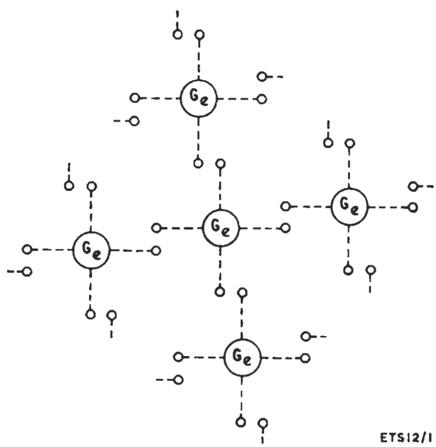


Figure 1.

These bonds may be broken by thermal or other forms of energy thereby releasing electrons from their bonds to become free electrons capable of movement through the lattice. These electrons give the semi-conductor part of its conductivity.

The broken bond, deficient by one electron, may now accept from an adjacent bond an electron which normally has not sufficient energy to become free. The broken bond, or electron deficiency, or hole, may thus move about in the crystal like the bubble in a spirit level. It also contributes to the conductivity, behaving like a 'positive electron'. In the pure material there are as

* Chambers's Technical Dictionary

many holes as electrons and the conductivity they give to the material is called its *intrinsic conductivity*. This increases with temperature.

A state of dynamic equilibrium exists when hole-electron pairs are being generated by thermal or other energy at the same rate as electrons and holes are combining. The rate of recombination is proportional to the electron concentration (n per cm^3) and to the hole concentration (p per cm^3). In the intrinsic material these concentrations are equal, i.e. $n = p = n_i$ so that the rate of recombination $\propto np$

$= rnp$

$= rn_i^2$ for the intrinsic material

where r is the recombination constant and n_i the hole or electron concentration.

The generation of hole-electron pairs is due almost entirely to the thermal energy absorbed by the crystal and is not influenced at all by the concentration of holes and electrons already existing within the crystal. If g is the rate of generation per cubic centimetre, then as the rate of recombination must adjust itself to be equal to the rate of generation for a state of equilibrium, $g = rn_i^2$.

For germanium $n_i = 2{\cdot}25 \times 10^{13}$ per cm^3 at 300°A
whilst for silicon $n_i = 1{\cdot}00 \times 10^{10}$ per cm^3 at 300°A

In the pure germanium crystal there are $4{\cdot}5 \times 10^{22}$ atoms/cm^3. This means there is approximately one broken bond (i.e. two current carriers, electron and hole) for each 2×10^9 atoms.

The addition of impurities to the Crystal

If impurity atoms are added the intrinsic germanium is said to be doped, and extra current carriers are introduced. Thus arsenic, a substance whose atom has five outer electrons, can be fitted into the crystal structure ,but one electron is surplus to the co-valent bond arrangement and is free to move through the lattice and become a current carrier. The arsenic atom is left with a residual positive charge but owing to the high dielectric constant of germanium (16) this implies only a small attraction for the fifth electron. Each atom of arsenic introduced thus adds 1 free electron. In the previous section we said that in the intrinsic state of the crystal there were two current carriers (1 hole + 1 electron) for each 2×10^9 germanium atoms. If one arsenic atom per 2×10^9 germanium atoms is added when the germanium is doped, then to a first approximation (i.e. ignoring a slight increase in the rate of recombination), the number of current carriers rises by 50% and so does the conductivity of the semi-conductor. Such a small trace of impurity element as one part in 2000 million implies germanium of an extremely high purity. The provision of a number of extra electron current carriers in this manner increases the recombination rates of holes and electrons so that when an equilibrium state is achieved, $g = rnp$ as before.

Since g is not influenced by the impurity admixture $np = n_i^2$ and is constant at a given temperature.

An increase of p or of n above the intrinsic concentration results in a similar increase in the number of current carriers present and a consequent increase in conductivity. Thus if the intrinsic concentration is taken as a unit:

$$np = 1$$

and if p is increased to $3p$, n must fall to $\frac{1}{3}n$ but the number of carriers rises from 2 to $3\frac{1}{3}$.

If n is therefore increased by the addition of an impurity above the intrinsic concentration n_i, p is reduced below this level. Electrons become the majority carrier and holes the minority carrier in the doped germanium which is known then as n-type germanium, the n standing for negative current carriers.

If the doping is achieved with a substance such as indium or gallium the atoms of which have only 3 outer electrons then the co-valent bond structure is deficient by one electron and this deficiency, or hole, is free to move about the lattice and contribute to the conductivity. Holes become the majority carriers and the doped germanium is known as p-type germanium, the p standing for positive current carriers. For the successful operation of a transistor, recombination of holes and electrons should not readily occur and the time from generation to recombination of hole-electron pairs should be long. This is often referred to as the 'lifetime'. Recombination occurs most readily in those parts of the crystal structure where strains and imperfections exist so that perfect crystals with treated surfaces are used in the construction of transistors.

The rectifying action of p–n Junctions

A p–n junction is achieved by doping two parts of a single germanium crystal with electron donating impurity (donor impurity) and with electron accepting impurity (acceptor impurity) respectively.

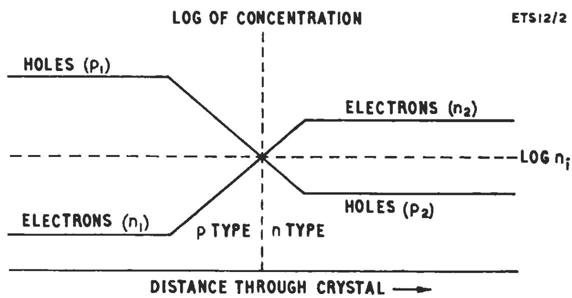


Figure 2.

In the p region the electron concentration is low while in the n region, the hole concentration is low. Diffusion therefore takes place, some holes moving into the n region and some electrons into the p region. This process continues until the overall gain of positive charge by the n region and the overall gain of negative charge by the p region produces an e.m.f. between the two sections sufficient to oppose further migration in either direction. A plot of the logarithm of concentration against distance through the crystal is shown in Fig. 2.

In this equilibrium condition,

$$\log n + \log p = 2 \log n_i = \text{constant.}$$

When an e.m.f. is applied to the crystal making the p region negative and the n region positive, the electrons in the p region are drawn back to the n region and holes in the n region are drawn back to the p region. (Fig. 3a).

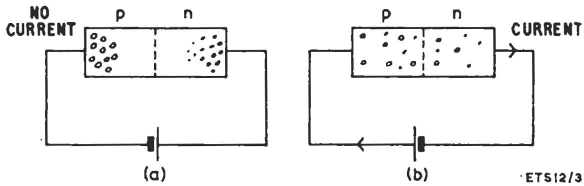


Figure 3.

Once this movement has occurred equilibrium is again established and no further resultant charge movement or current occurs. Of course, minority carriers in each section are able to move across the junction under the action of the external field and a small reverse saturation current flows. If now the applied e.m.f. is reversed, electrons in the n section are driven towards and into the p section, and holes in the p section are driven into the n section. (Fig. 3b). The recombinations of holes with invading electrons in the p section and the recombinations of electrons with invading holes in the n section results in the continued flow of current carriers across the junction, that is a current flows through the 'rectifier'. Emission of holes in the p region starts from the positive electrode and electrons in the n region from the negative electrode.

An electronic idea of the process of rectification may be obtained by imagining a thermionic diode which has the added facility of an anode capable of emitting positrons (i.e. positive electrons).

When a negative voltage is applied to the anode, relative to the cathode, attraction of positrons to the anode and electrons to the cathode occurs and the two clouds might be imagined as contracting about their respective electrodes; no conduction occurs (Fig. 4a).

If now the anode is made positive with respect to the cathode, positrons are repelled from the anode, electrons are repelled from the cathode and recombination occurs as the two types of particles intermingle. Continued emission of positrons and electrons guarantees the continued flow of current through the valve, and the diode conducts as in Fig. 4b. To complete the analogy, the anode should also emit a very few electrons and the cathode a very few positrons so that a small saturated reverse current can flow.

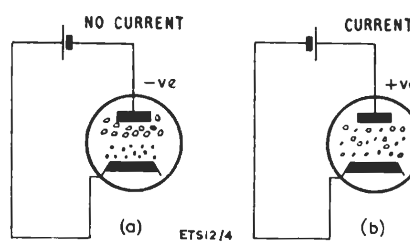


Figure 4.

The current-voltage curve for a typical *p–n* junction is shown in Fig. 5. Although it is normal practice to draw this curve so that the conducting condition occurs in the positive quadrant it is convenient here to reverse the axes as the transistor action to be studied later is concerned principally with the reverse current. The effect of too great a reverse voltage causes a breakdown of the bond structure and an avalanche current to flow. This breakdown is called the *Zener* effect and will be mentioned again later.

The equation for such a curve (excluding breakdown conditions) is

$$i = i_s\,(e^{\kappa V} - 1)$$

where i_s is the small reverse saturation current and V is the applied voltage; κ is a constant and is approximately equal to 40. When V is positive (forward direction), $e^{\kappa V}$ is large compared with unity and i increases exponentially. When V is negative, $e^{\kappa V}$ tends to zero very quickly and i tends to a constant value i_s which is very small.

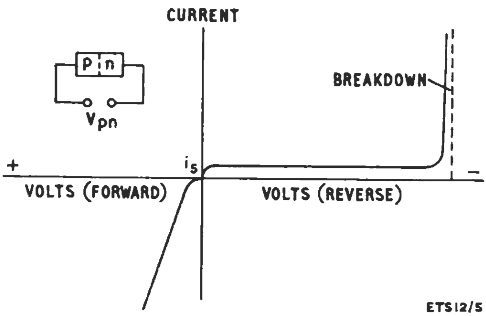


Figure 5.

The *p–n–p* Junction Transistor

The transistor consists of two such *p–n* junctions back-to-back (Fig. 6). The emitter-base junction is biased in the conducting direction so that a hole current is flowing into the base for a relatively small positive emitter voltage. The *n*-type base region is not doped to the same extent as the emitter and its electron concentration is not great although electrons are the majority carriers. This means that most of the current flow across the emitter-base junction is carried by holes from the emitter. The collector-base junction is biased in the reverse or non-conducting direction. In the absence of emitter hole injection into the base only a small saturation current flows. This saturation current is largely independent of the collector voltage up to the Zener breakdown condition as was seen in the case of the single *p–n* junction. If now, emitter and collector are connected to their appropriate potentials, holes which enter the base from the emitter diffuse through it, cross the collector junction and become current carriers, thus increasing the collector current.

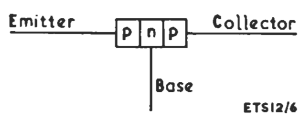


Figure 6.

As the base is very thin the loss of holes by recombination is small and the base current is therefore small. This means that the collector current increase due to hole injection through the base is only a little less than the emitter input current. The enhanced collector current is still independent of the collector voltage.

The device has effectively three terminals so that there are three possible sets of curves showing output current against output voltage for various values of input current. If the input current is made to flow between emitter and base whilst the output is measured between base and collector the transistor is said to be connected in common base (Fig. 7a). It may also be used in common emitter (Fig. 7b) or common collector (Fig. 7c). Let us first consider the common base form of connection, for which typical curves are those shown in Fig. 8.

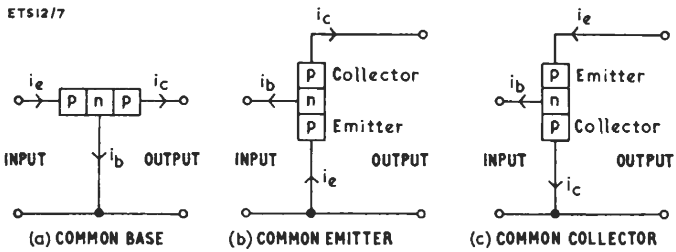


Figure 7.

The ratio of increase in collector current to the increase in emitter current which produces it is the current gain (α) and for junction transistors is always less than unity although it may approach very close to this value. For a typical transistor, $\alpha = 0{\cdot}979$.

When the emitter current I_e is zero, the collector base junction has the normal junction diode characteristic when biased in the reverse direction. This reverse current, usually represented by I_{co}, is very small and of the order of a few mico-amperes.

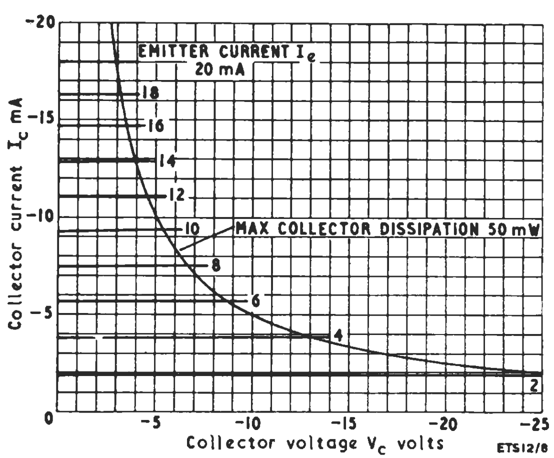


Figure 8.

The collector current, I_c when emitter current flows, consists of this reverse current I_{co} together with a component due to the emitter current: i.e.

$$I_c = I_{co} + \alpha I_e.$$

Turning now to the common-emitter condition we can plot collector current against collector voltage for various values of the base current, as shown in Fig. 9. It may be noted that a change of only 100 microamperes in base current may produce 2 or more milliamperes of collector current so that the current gain α' is large. Its relationship to the current gain α in common-base connection may be determined as follows.

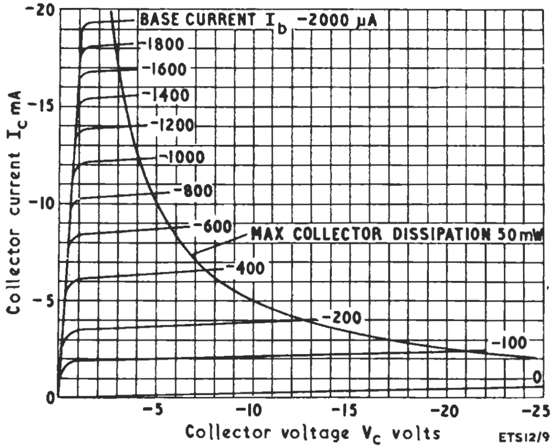


Figure 9.

Current Gain α' (i_c/i_b)

(Throughout this supplement small letters represent small changes in current and voltage. Capital letters represent d.c. conditions). Applying Kirchhoff's first law we have (Fig. 7a).

$$i_e = i_b + i_c \tag{1}$$

and

$$\frac{i_c}{i_e} = \alpha \text{ so that } i_e = \frac{i_c}{\alpha}$$

We may therefore write

$$\frac{i_c}{\alpha} = i_b + i_c$$

i.e.

$$i_c\left(\frac{1}{\alpha} - 1\right) = i_b$$

or

$$\frac{i_c}{i_b} = \frac{\alpha}{1 - \alpha} = \alpha' \tag{2}$$

As α is approximately equal to unity, $\alpha' \simeq \dfrac{1}{1 - \alpha}$

For a typical transistor where $\alpha = \cdot979$ the current gain $\alpha' = \dfrac{0{\cdot}979}{0{\cdot}021} = 46$

The collector current which flows when the base current (I_b) is zero is represented by I_{co}' and may be related to I_{co} in the following way (see Fig. 10). Let I_{co} flow from base to collector, then I_{co} must flow from emitter to base if the base current is to be zero.

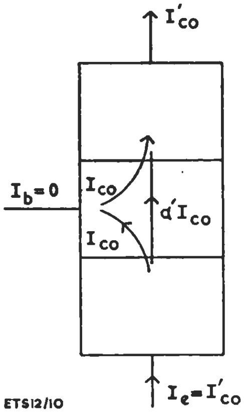


Figure 10.

The emitter to collector current will be $\alpha' I_{co}$ and the total collector current

is $\alpha' I_{co} + I_{co} = I_{co}'$

$$\text{therefore} \qquad I_{co}' = I_{co}(1 + \alpha') \qquad (3)$$

I_{co}' is not as great as might be expected because for low collector currents α' is small. (See graph Fig. 11 showing variation of α' with collector current). A typical value would be about 100μA. Variation in the value of α causes considerable change in the value of $1 - \alpha$ as α is almost unity itself, so as might be expected I_{co}' may vary considerably from one transistor to another.

The base current may be regarded as controlling the flow of holes from the emitter to the collector in an analogous way to that of the grid voltage in an electronic valve controlling the electron flow from cathode to anode. In fact the resemblance of a junction transistor with common or grounded emitter to a valve with common or 'grounded' cathode is quite helpful, especially as this mode of transistor operation is at present the most common one.

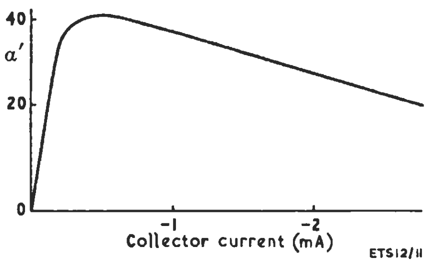


Figure 11.

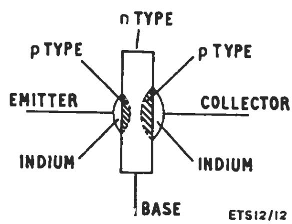


Figure 12.

Physical Construction

Many processes are being used to produce junction transistors. In one method, used for the production of diffused junction types, a thin slice of *n*-type germanium has one blob of indium attached to each side (see Fig.12). Heating causes diffusion of indium into the germanium and the creation of *p*-type zones. Diffusion continues until only a narrow base layer of *n*-type material separates these zones. Leads are attached to the indium blobs and to the *n*-type base and the whole is hermetically sealed in a glass envelope. If the two indium blobs are of equal size either may function as emitter or collector. When such a transistor is intended to provide an appreciable output power the size of the collector blob is increased to allow greater heat dissipation and collector and emitter are not then interchangeable connections.

Circuit Symbol

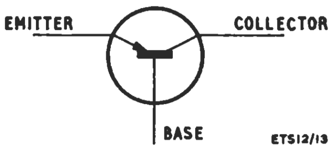


Figure 13.

There is no general agreement yet on the symbol to be used for the junction transistor. The one still most commonly employed is shown in Fig. 13 and is the one recommended by the British Standards Institution in Supplement No. 4 (1956) to B.S. 530: 1948.

COMMON-EMITTER TRANSISTOR CHARACTERISTICS

Collector Characteristics

The characteristics referred to in the first section (Figs. 8 and 9) at first sight resemble those of a pentode valve except that collector current control is more directly proportional to input current than input voltage. There is of course a big difference in the magnitudes of the anode voltage and the collector voltage and this fact tends to disguise the difference in slope resistance which differentiates the transistor in common-emitter connection from the pentode valve.

The transistor output resistance is actually of the order of 20-50 kilohms and resembles a triode in the normal common-cathode connection. The input resistance is 1000 to 2000 ohms, and clearly this low value presents problems of input circuit design not encountered at the same frequencies with thermionic valves. The common-emitter connection permits of maximum power gain and is in frequent use so that most of the analysis of this Supplement is devoted to this mode of operation. Characteristics of the transistor in common-emitter connection are shown in Fig. 9.

An alternating base current produces a corresponding alternating collector current, an alternating grid voltage produces an alternating anode current in a valve. By means of a collector load resistance the alternating collector current produces an alternating voltage which is the amplified version of the input signal voltage at the base. A collector load line may be drawn and an operating point chosen. This point must not lie beyond the permitted maximum collector dissipation nor must it require a collector voltage in excess of some given value usually no greater than 30 volts.

Voltages in excess of the maximum cause a sudden increase in collector-base conductivity and destruction of the transistor unless some limiting resistance is employed.

Apart from the low input impedance already mentioned, the principal feature in circuit design which distinguishes the transistor from the valve is the method of achieving the automatic bias required by the chosen working point. Unlike the valve the transistor characteristics are very sensitive to temperature as might have been expected from the physical operation of the device dealt with in Section 1. Before leaving the general discussion on characteristics it should be mentioned that this sensitivity to temperature changes prevents a point-by-point plot of transistor characteristics. Under normal operating conditions the collector working temperature is constant and the characteristics must be drawn for this temperature. This can be done by using electronic curve tracers where the family of characteristics are portrayed on the end of a cathode-ray tube.

Biasing

Having determined the base biasing current appropriate to the chosen working point, a circuit is required to achieve it. The following expression has been quoted earlier:

$$I_c = I_{co} + a\,I_e$$

where I_{co} is the collector-base current when the emitter current is zero, I_e is the emitter current and a (i_c/i_e) is the current gain, which is always less than unity for a junction transistor of the *p–n–p* variety. This expression is a fundamental relationship applying when the transistor is operating in common-base connection. I_{co} is of the order of 5 to 10μA for germanium transistors and temperature effects would not be very serious with this method of connection. When the transistor is operated in common-emitter connection a similar expression is obtained relating base current to collector current.

Thus

$$I_c = I_{co}' + a' I_b$$

and it will be recalled that I_{co}' is $I_{co}(1 + a')$ (page 13). Since a' is usually much greater than unity we have

$$I_{co}' = a' I_{co}$$

$$\text{so that } I_c = a'\,(I_{co} + I_b)$$

The temperature variation of collector current is due principally to variation in I_{co} although a' also varies. It will be assumed in the following quantitive treatment of biasing circuits that a and therefore a' is constant and I_{co} is the variable factor. Relatively small changes in I_{co} produce large changes in I_{co}'. For example if I_{co} changes from 5 to 20μA due to a rise in temperature, then I_{co}' may change from 75 to 300μA and this change in collector current flowing through the collector load will reduce the collector potential. This change in the working point affects gain and impedances. There is clearly a temperature which so reduces the transistor collector potential that amplification is no longer possible. For these reasons the simple biasing arrangement shown in Fig. 14 is not satisfactory.

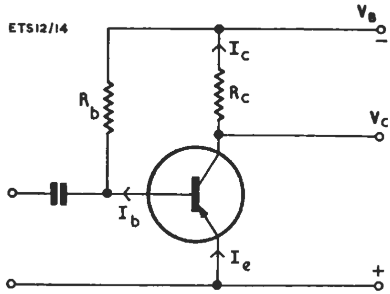


Figure 14.

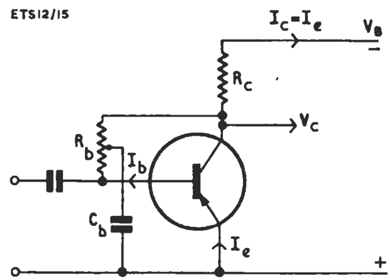


Figure 15.

Assuming the base bias current I_b has been determined and that the base to emitter voltage is negligible (of the order of 0·1v) then $R_b = V_B/I_b$. As I_b is constant any increase in I_{co}' due to increase in the ambient temperature produces an equal rise in I_c and leads to a reduction in the collector voltage.

A measure of stabilisation of the working point is achieved by using the circuit of Fig. 15.

A qualitative appreciation of the action of this circuit is as follows. If I_c tends to rise due to a rise in temperature, then the collector potential will fall and the base biasing current will be reduced. As seen in Fig. 9 any reduction in base current leads to a reduction in the collector current. It is in fact negative feedback and applies equally to the signal the stage is amplifying unless some measure is taken to render it ineffective at signal frequencies. This can be done by earthing a tap on R_b through a suitable capacitor and decoupling the base from collector a.c. voltages (capacitor C_b in Fig. 15).

The quantitative expression for the stabilisation of the working point is arrived at as follows. Collector current for zero signal input, is equal to a constant component, A, together with a component which is proportional to the value of I_{co}., i.e. $I_c = A + S\ I_{co}$ and S is called the Stability factor which should be as small as possible but in fact is always greater than unity. In the common-emitter condition where $I_c = a'\ I_b + I_{co}'$ the Stability factor has its maximum value of $I_{co}'/I_{co} \simeq a'$. The relation between the Stability factor S and the circuit values R_c, R_b of Fig. 20 is derived in Appendix 2(a) and is found to be

$$S = \frac{R_c + R_b}{R_c + (1 - a)R_b}$$

Typical values of R_e and R_b might be 10kΩ and 100kΩ respectively and for $a = \cdot 976$.

$$S = \frac{110}{10 + \cdot 024 \times 100} = \frac{110}{12 \cdot 4} = 8 \cdot 85$$

A very useful stabilising circuit (Fig. 16) consists of a potentiometer R_1 and R_2 which holds the base at a sensibly constant potential together with a resistance in the emitter which then varies the emitter base potential to oppose changes in collector current. It is in fact current negative feedback and if it is not required at working frequencies, the emitter resistance must be by-passed by a large capacitor.

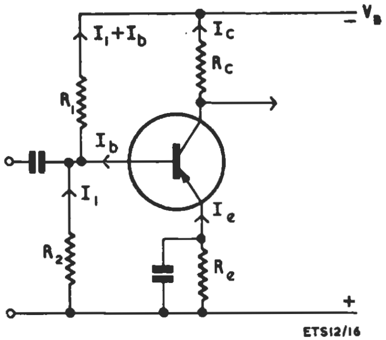


Figure 16.

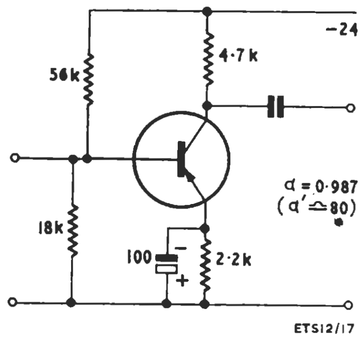


Figure 17.

In this case the stability factor S is given by

$$S = \frac{1 + R_e/R_b}{1 + R_e/R_b - a} \quad \text{(See Appendix 2b)}$$

where $\dfrac{1}{R_b} = \dfrac{1}{R_1} + \dfrac{1}{R_2}$.

For high stability (low value of S) R_b should be small; as it is effectively in shunt with the input resistance of the transistor, its minimum value is determined by the extent to which this shunting is acceptable. There is also a power loss in the potentiometer chain which would also dictate its minimum resistance if overall power efficiency were a criterion. A typical circuit using this method of bias stabilisation might be as shown in Fig. 17.

Here R_b = 56kΩ in parallel with 18kΩ = 13·6kΩ

therefore
$$S = \frac{1 + \dfrac{2 \cdot 5}{13 \cdot 6}}{1 - \cdot 987 + \dfrac{2 \cdot 5}{13 \cdot 6}} = 6 \cdot 1$$

With such stabilisation, a variation of I_{co} of 25μA would cause a collector current variation of 6·1 $\times$ 25μA whereas without stabilisation and a current gain of say 80, the variation would be 80 $\times$ 25μA.

The BBC Designs Department in a transistorised version of the outside broadcast amplifier (OBA/9) has adopted a stability factor of 1·5 in the first stage. This represents a high degree of stabilisation, necessary because of the large collector resistance and low collector voltage used to operate under minimum noise conditions. The second stage has a stability factor of 6·4.

Input characteristics (common emitter)

It was implied earlier that base current rather than base voltage gave a good linear control of the collector current and the non-linear behaviour with voltage control arises from the non-linear input resistance of the transistor.

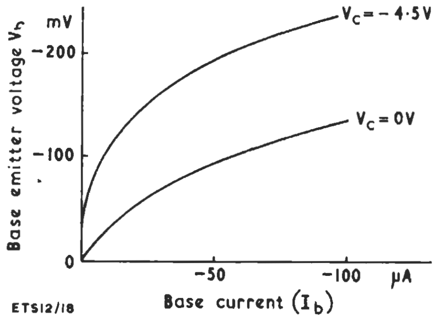


Figure 18.

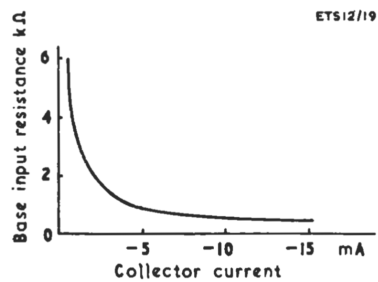


Figure 19.

A graph of I_b against V_b is shown in Fig. 18 for a typical junction transistor for two values of V_c.

The curve of $V_c = 0$ is the ordinary junction diode characteristic in the forward direction at low applied voltages.

The base input resistance may be derived from these graphs and is shown in Fig. 19 as a function of collector current.

At low values of collector current the input resistance can be high but the current gain will be correspondingly low. The output resistance is also high for low collector current.

The input resistance consists of the junction resistance which varies exponentially with applied voltage, in series with the base resistance i.e. the semi-conducting material between the base connection and the junction, the resistance of which varies almost linearly with applied voltage.

An approach can be made to a linear relation between V and I by an increase of external base series resistance which in effect amounts to feeding the base from a current generator rather than from a source of e.m.f. Alternatively an external series emitter resistance R_e may be used which will be shown to be equivalent to the addition of external base resistance of value $\alpha' R_e$ (Equation 14). This emitter feedback also reduces the distortion due to variation of α' with collector current.

Without feedback the base input resistance for small signals is of the order of 1000 ohms and this is so low that it effectively controls the collector load of previous stages when transistors are connected in tandem. This input resistance can be varied in either direction by negative feedback; series applied feedback raises the input impedance whilst shunt applied feedback reduces it.

THE TRANSISTOR EQUIVALENT CIRCUIT

The Transistor Equivalent Circuit

For small signals the transistor may be assumed a linear device and can be represented by a linear network that is active (i.e. contains sources of energy). Several alternative forms of this exist. A frequently encountered form of equivalent circuit drawn for common-emitter operation is shown in Fig. 20 where r_b is the base resistance, r_e the emitter resistance, r_c the collector resistance and r_m a constant relating the generator e.m.f. to the emitter current which is its prime cause. R_c is the load resistance.

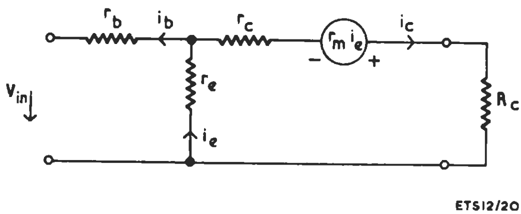


Figure 20.

The current gain for common-emitter operation a' is the ratio of i_b to i_c when R_c is zero.

$$a' = \frac{i_c}{i_b} = \frac{i_c}{i_e - i_c}$$

As r_c is much greater than r_e in normal transistors

$$i_c \simeq \frac{r_m i_e}{r_c} \text{ when } R_c \text{ is zero.}$$

So that
$$\frac{i_c}{i_e} = a = \frac{r_m}{r_c} \tag{4}$$

and
$$\frac{i_c}{i_b} = \frac{a}{1-a} = \frac{\frac{r_m}{r_c}}{1 - \frac{r_m}{r_c}} = \frac{r_m}{r_c - r_m} \tag{5}$$

Input Resistance

Applying Kirchhoff's Laws we have first

$$i_e = i_b + i_c \tag{6}$$

then to the two closed circuits:

$$v_{in} = r_b\, i_b + r_e\, i_e \tag{7}$$

and

$$r_m\, i_e = (r_c + R_c)\, i_c + r_e\, i_e\,. \tag{8}$$

Writing $i_e = i_b + i_c$ in (7) and (8)

$$v_{in} = (r_b + r_e)\, i_b + r_e\, i_c \tag{9}$$

$$r_m\, (i_b + i_c) = (r_c + R_c)\, i_c + r_e\, (i_b + i_c)\,. \tag{10}$$

From (10)

$$i_c = \frac{r_m - r_e}{r_e + r_c + R_c - r_m}\, i_b\,. \tag{11}$$

Substituting in (9)

$$\frac{v_{in}}{i_b} = r_b + r_e + \frac{r_e\,(r_m - r_e)}{r_e + r_c + R_c - r_m}$$

$$= r_b + \frac{r_e(r_c + R_c)}{r_e + r_c + R_c - r_m} = \text{Input resistance } (R_{in}) \tag{12}$$

As might have been expected from an examination of the equivalent circuit this input resistance is a function of the collector load.

Typical values of r_e, r_c etc. are for a junction transistor:—

$$r_e = 20\Omega$$

$$r_b = 800\Omega$$

$$r_c = 1\text{M}\Omega$$

$$r_m = 975\text{k}\Omega$$

and

$$a' = \frac{r_m}{r_c - r_m}$$

$$= \frac{975}{25} = 39$$

Equation 12 may be simplified if r_e is neglected in the denominator, an approximate expression for the input impedance becomes

$$R_{in} = r_b + r_e \frac{(r_c + R_c)}{r_c(1 - a) + R_c} \tag{13}$$

where $$1 - a = \frac{r_c - r_m}{r_c}$$

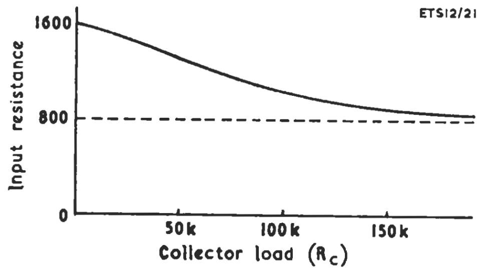


Figure 21.

A graph of input resistance against collector load may be plotted (Fig. 21). The values of r_b, r_e etc. vary with the d.c. working point so that the curve quoted can only give an indication of the order of the input resistance. Where $r_c - r_m$ is much greater than R_c as so often is the case, the input resistance is not affected by variation in the collector load resistance and an expression for the input resistance based on this assumption may be derived from equation (13).

$$R_{in} = r_b + \frac{r_e}{1 - a} \simeq r_b + a' \, r_e \tag{14}$$

From this equation it may be seen that if physical resistance is added in series with r_e the input resistance is increased by a' times the value of the added resistance.

When transistors are connected in tandem and the input resistance of one stage becomes the effective a.c. collector load of the previous stage, R_c is certainly much less than $r_c - r_m$ and the approximate expression above is sufficient. For typical values already quoted,

$$R_{in} = 800 + 40 \times 20 = 1600\Omega \text{ whereas } r_c - r_m = 25\,000\Omega$$

Output Resistance

In a similar way the output resistance of the circuit may be found.

Applying the methods associated with Thevenin's theorem the input generator shown dotted in Fig. 22 is replaced by a short circuit and the resistance between A and B is the equivalent resistance of the Thevenin circuit i.e. the output impedance of the device. The ratio v_2/i_c is the desired value.

Applying Kirchhoff's Laws

$$i_e = i_b + i_c \tag{15}$$

then to the two circuits,

$$0 = (R_g + r_b + r_e)\, i_b + r_e i_c \tag{16}$$

and

$$v_2 + r_m i_e = r_e i_e + r_c i_c \tag{17}$$

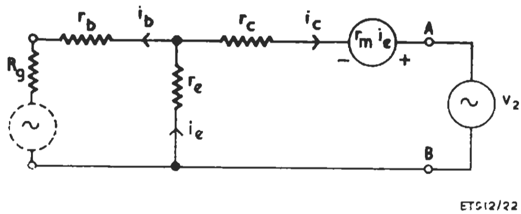


Figure 22.

Substituting for i_e from (15) in (17)

$$v_2 + r_m (i_b + i_c) = r_e i_c + r_e i_b + r_c i_c \tag{18}$$

Substituting for i_b in (18) from (16)

$$v_2 = (r_c + r_e - r_m)\, i_c + (r_e - r_m) \cdot \frac{-r_e}{(R_g + r_b + r_e)} \cdot i_c$$

and

$$\frac{v_2}{i_c} = r_c + r_e - r_m + \frac{r_e (r_m - r_e)}{R_g + r_b + r_e}$$

$$= r_c - r_m + r_e \cdot \frac{R_g + r_b + r_m}{R_g + r_b + r_e} = R_{out} \tag{19}$$

and not surprisingly the expression for output resistance contains the resistance of the generator connected across the base emitter input. Putting in typical values as quoted earlier.

$$R_{out} = 25 \times 10^3 + \frac{20 \times 975 \times 10^3}{R_g + 820} \quad \text{(approx.)}$$

When R_g is zero i.e. the device is supplied by a generator of negligible internal resistance; $R_{out} \simeq 50\text{k}\Omega$

Where R_g is infinity i.e. a high impedance source is used

$$R_{out} = 25 \text{ k}\Omega$$

The expression derived for R_{out} may be simplified by similar approximation to those used for the approximate input resistance.

Writing $$r_c - r_m = r_c(1 - \alpha) \simeq \frac{r_c}{\alpha'}$$

we obtain $$R_{out} = \frac{r_c}{\alpha'} + r_e \cdot \frac{r_m + R_g}{r_e + r_b + R_g}$$

The value of $\frac{r_m + R_g}{r_e + r_b + R_g}$ may vary from $\frac{r_m}{r_e + r_b}$ when R_g is zero to unity when R_g is infinite thus giving a factor multiplying r_e varying from 30 to 1. Even when the factor is 30, 30 r_e is so small compared with r_c/α' that $R_{out} \simeq r_c/\alpha'$

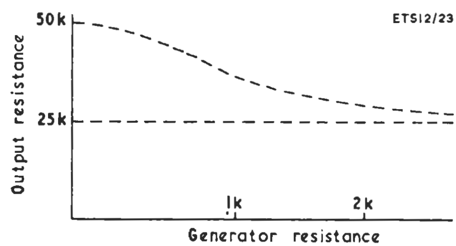


Figure 23.

The variation of output resistance with variation of the generator internal resistance is shown in Fig. 23. The output resistance is seen to depend to some extent on the resistance of the input circuits and the input resistance upon that of the output circuits. There are optimum values of both which give maximum power gain. The transistor working between such impedances is said to be impedance matched. These considerations are very important when dealing with point contact transistors but with junction transistors the variation in input and output impedances with wide variations of generator and load resistance is relatively small, and the transistor can be taken approximately as a current generator of $r_c - r_m$ ohms internal resistance, i.e. $r_c(1 - \alpha)$ or r_c/α' approximately.

For values of r_c of the order of 10^6 ohms and α' of say 40 the output resistance is $25 \cdot 10^3$ ohms as derived earlier.

Current and voltage amplification

The current and voltage amplifications may also be obtained from the equivalent circuit.

Current amplification in the presence of a Collector Load (α'_b)

The equations (6) and (8) derived earlier are the starting point

$$i_e = i_b + i_c \qquad (6)$$

$$r_m i_e = (r_c + R_c)\, i_c + r_e i_e \qquad (8)$$

$\dfrac{i_c}{i_b}$ is the required ratio and obtained from the above relations by eliminating i_e.

$$r_m (i_b + i_c) = (r_c + R_c)\, i_c + r_e (i_b + i_c) \qquad (20)$$

Collecting terms in i_b and i_c

$$i_b (r_m - r_e) = i_c (r_c + R_c + r_e - r_m) \qquad (21)$$

$$\frac{i_c}{i_b} = \frac{r_m - r_e}{r_c + R_c + r_e - r_m} \simeq \frac{r_m}{r_c - r_m + R_c} \qquad (22)$$

i.e.

$$\frac{i_c}{i_b} = \frac{\dfrac{r_m}{r_c - r_m}}{1 + \dfrac{R_c}{r_c - r_m}} = \alpha'_b$$

i.e.

$$\alpha'_b = \frac{\alpha'}{1 + \dfrac{R_c}{r_c - r_m}}$$

which tends to α' as R_c tends to zero.

The short circuit current amplification is the value of α_b when R_c the collector load is zero and is clearly equal to α'.

Inserting typical values for a junction transistor (values of resistances in kilohms)

$$\alpha_b = \frac{975}{1000 - 975 + R_c} = \frac{975}{25 + R_c}$$

When R_c is zero $\alpha'_b = \alpha' = 39$

,, R_c is 25k $\alpha'_b = 19$

,, R_c is 5k, a more typical value, $\alpha'_b = 32{\cdot}5$.

Assuming the input impedance values derived earlier for given load resistances, the voltage amplification may be calculated from the derived value of current amplification. For example with this transistor when $R_c = 5\text{k}\Omega$ the input resistance is found from equation (12) to be 1470 ohms.

$$\text{Input current} = \frac{\text{input voltage}}{1470} = i_b.$$

$$i_c = \alpha'_b\, i_b = 32{\cdot}5\, \frac{\text{input voltage}}{1470}$$

$$\text{and output voltage} = R_c i_c = \text{input voltage} \times \frac{32{\cdot}5 \times 5000}{1470}$$

$$\text{Hence the voltage amplification} = \frac{\text{Output voltage}}{\text{Input voltage}} \times \frac{32{\cdot}5 \times 5000}{1470}$$

$$= 110 \text{ or a gain of } 40{\cdot}8\text{dB(V)}$$

If the transistor were in fact driving a further stage with an input impedance similar to its own, then R_c would effectively be about 1500 ohms and the voltage gain would be equal to the current gain and given by approximately

$$20 \log_{10} \alpha'$$

which in this case would be 31·6dB.

The h parameters

The subject of equivalent transistor circuits and parameters cannot be left without reference to an alternative and frequently used system based on the h, or hybrid parameters. The transistor is treated as a four-terminal network with input and output voltages and currents (Fig. 24).

v_2 and i_2 are both written as a function of i_1 and v_2, i.e.

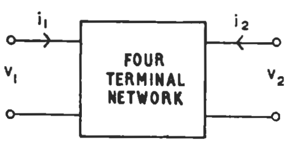


Figure 24.

$$v_2 = h_{11}i_1 + h_{12}v_2 \qquad (23)$$

$$i_2 = h_{21}i_1 + h_{22}v_2 \qquad (24)$$

It will be noticed that h_{11} is a resistance and h_{22} is an admittance whereas h_{12} and h_{21} are pure numbers. These parameters are useful in that they are easily obtained by direct measurement and for those whose mathematical skill permits them to use four-terminal networks as such without enquiring into their internal make-up, they make a most convenient starting point.

We have seen that the parameters are based on small amplitude signals and a fixed d.c. operating condition; they are therefore determined as follows:

To find the value of h_{11} the output of the transistor is shorted to a.c. making v_2 zero, and from (23) $h_{11} = v_1/i_1$ i.e. h_{11} is the input resistance when the output is shorted. From (24), with i_1 equal to zero the second equation gives $h_{22} = i_2/v_2$; this is the output admittance with the input open circuit. When a voltage is applied at the output terminals with the input open circuit ($i_1 = 0$), h_{12} is equal to v_1/v_2. This is often called the reverse voltage feedback ratio.

When v_2 is zero the second equation gives h_{21} as equal to i_2/i_1 and this is a pure number, the current amplification.

The symbols h_{11}, h_{12}, etc. are reserved for the common-base connection and for the common-emitter connection the parameters are designated by a dash, i.e. h'_{11}, h'_{12} etc. In order to illustrate the calculation of these parameters let us consider the junction transistor previously used and determine for it the common-emitter parameters. The input resistance for zero collector load was found to be 1600Ω so that

$$h'_{11} = 1600\Omega$$

and the output resistance when the input was open circuited ($Rg = \infty$) was 25kΩ so that

$$1/h'_{22} = 25\text{k}\Omega \text{ or } h'_{22} = 40\mu \text{ mhos.}$$

The current gain when the collector load was zero was $\dfrac{r_m}{r_c - r_m}$ which gives

$$h'_{21} = \frac{r_m}{r_c - r_m} = 39$$

Finally h'_{12} is found with the aid of equation 17 which is modified to

$$v_2 + r_m i_e = r_e i_c + r_c i_c$$

because by definition $i_1 = 0 = i_b$ and therefore $i_e = i_c$.

The voltage v_1 appearing across the input terminals is $r_e i_c$ so that

$$h'_{12} = \frac{v_1}{v_2} \simeq \frac{i_c r_e}{i_c(r_c + r_e - r_m)} \simeq \frac{r_e}{r_c - r_m} = \frac{20}{25\,000} = \frac{1}{1250}$$

The h' parameters are of practical use in that their values give directly the approximate input and output impedances (h'_{11} and $1/h'_{22}$), current amplification and feedback ratio (h'_{21} and h'_{12}) and allow a rapid estimate of transistor performance to be made.

Other Applications

Common-Base and Common-Collector Connections (Fig. 25)

The most frequently encountered application of the junction transistor in a.f. amplifiers is in the common-emitter connection as this gives greatest power and voltage amplification. Hence the analysis of the previous chapter has been concerned with common-emitter operation, which, like the valve with common cathode gives phase reversal from input to output. Two other

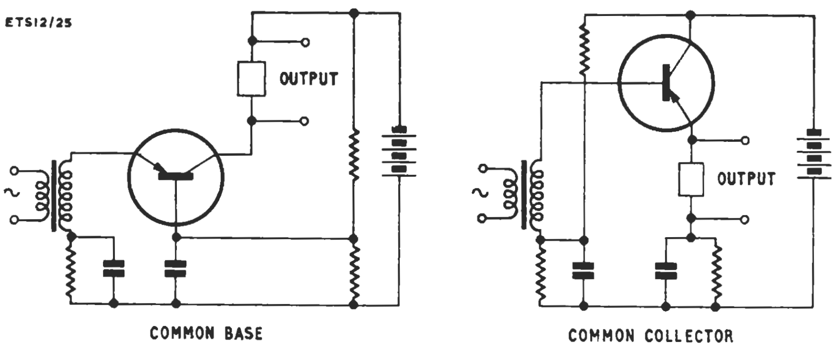


Figure 25.

modes of connection are possible. The common-base connection, which has properties similar to those of a valve with earthed grid, and the common collector, which compares with the earthed anode more usually referred to as the cathode follower. Tables shown below give the approximate expression for input and output impedances, current and voltage gains for the three circuit arrangements.

It will be assumed that the emitter resistance r_e is much smaller than $r_c - r_m$ which is itself much smaller than r_c

i.e. $$r_e \leqslant r_c - r_m \leqslant r_c$$

and also $$r_e \leqslant R_c \leqslant r_c - r_m$$

In the cases considered in this Supplement, the values of the equivalent circuit constants have been as follows:

$$r_e = 20\Omega,\ r_c - r_m = 25\text{k}\Omega,\ r_c = 1\text{M}\Omega$$

The collector load resistance R_c has had values in the region 1kΩ to 10kΩ so that the approximations made above are applicable.

Common Emitter (the case dealt with in this supplement)

	Approx. formula	*Typical Values*
Input resistance	$r_b + \alpha' r_e$	1600 ohms
Output resistance	r_c/α'	25kΩ
Current Amplification	α'	40
Voltage Amplification	$\alpha' \dfrac{R_c}{\text{Input resistance}}$	125 (R_c=5kΩ)

Common Base

	Approx. formula	*Typical Values*
Input Resistance	$r_e + r_b(1 - \alpha)$	40 ohms
Output Resistance	r_c	10^6 ohms
Current Amplification	α	·979
Voltage Amplification	$\dfrac{\alpha R_c}{\text{Input resistance}}$	125 (R_c = 5kΩ)

Common Collector

	Approx. formula	*Typical Values*
Input Resistance	$R_c/1 - \alpha$	200k (R_c = 5k)
Output Resistance	$r_e+(r_b+R_g)(1-\alpha)$	5k (R_g = 200k)
Current Amplification	$\dfrac{1}{1-\alpha} \simeq \alpha'$	40
Voltage Amplification	1	1

Circuits showing the applications of the transistor in common-base and common-collector connection are given in Fig. 25.

Push-Pull Operation

The non-linear base input resistance of the transistor leads to the presence of harmonic distortion in the output which is principally second harmonic. This may be reduced by push-pull operation in Class A.

For a given collector dissipation, greater power outputs than are possible with Class A operation may be achieved with Class B. Further the low no-signal power consumption is attractive where batteries are concerned. The maximum theoretical efficiency of Class B operation is 78% and efficiencies of 75% have been achieved with transistors. In Class A the maximum output of transistors each having a maximum collector dissipation of 150mW might be about 125mW, whereas in Class B the same pair could provide 1 watt.

The distortion is greater and if low distortion is an essential consideration Class A push-pull must be employed.

For typical Class A push-pull and Class B push-pull circuits, see Fig. 26.

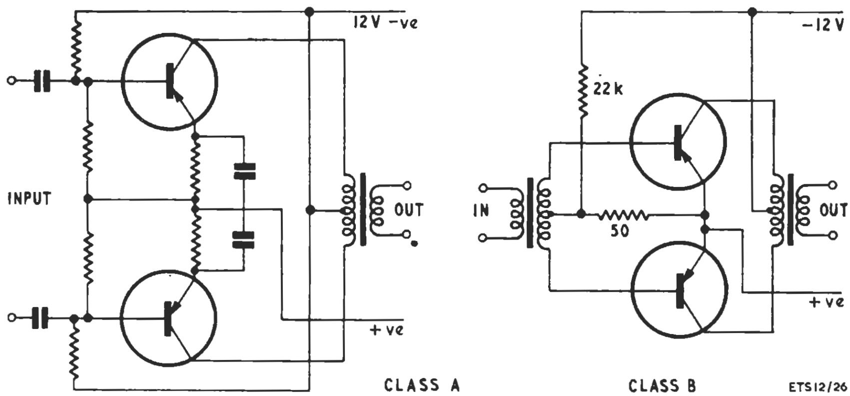


Figure 26.

A.C. Negative Feedback (n.f.b.)

The use of d.c. feedback for the purpose of stabilisation of the working point has already been dealt with in some detail. A.C. feedback is also used to reduce distortion and gain variation, the principle of operation being similar to that for valve circuits. The need for feedback is in general greater with transistors because of the non-linear relation between the base-emitter voltage and current. Feedback is also used, as with valve circuits, to control the input and output resistance of the device.

The application of n.f.b. is complicated by the variation of phase shift with frequency which occurs in the transistor itself (see section on r.f. application) so that instead of applying all the required feedback over a number of stages it is usual to supply some to each stage and a reduced amount overall.

Examples of feedback arrangements are shown in Figs. 27, 28, 29 and 30.

Input Noise

When it is desired to amplify low-level signals such as those derived from a microphone, sound track or television camera tube, the limit of useful amplification is determined by the noise level occurring at the amplifier input If the signal to be amplified is of the same order of magnitude as the noise there is little to be gained by amplifying it. The lower limit to the input noise as far as the amplifier itself is concerned would be the noise generated by the random electron movement in a pure resistance equal in value to the internal resistance of the source connected to the amplifier. In fact the first and sometimes the first two stages of the amplifier add a good deal to the noise level.

When point-contact transistors were first introduced their high noise level made them unsuitable for the amplification of signals below 1 milliwatt (zero level). With the advent of the junction transistor the situation has greatly improved and selected specimens used in the best operating conditions produce noise levels which approach within a few dB of those produced by valves.

Transistor noise differs from the noise produced in a pure resistance; in the latter the noise power (per unit range of frequency) is independent of frequency. In the transistor the expression for noise power at audio frequencies contains a term which varies inversely as the frequency thus giving maximum noise at the lowest frequency. This tends to make the design of d.c. or very-low-frequency amplifiers for small signals rather difficult.

Transistor noise is at a minimum when the collector current is low and the resistance of the signal source connected to the transistor has a value in the range 500 – 1,000 ohms.

Radio-frequency Amplification

At present (1958) transistors available in this country are capable of amplifying signals up to 15 to 20Mc/s and no doubt the frequency limit will eventually be raised still higher. This limitation is dependent on the drift time of holes across the base and results in a variation of α with frequency given approximately by the expression $\alpha = \dfrac{\alpha_1}{1 + jf/f_c}$ or $|\alpha| = \dfrac{\alpha_1}{\sqrt{[1+(f/f_c)^2]}}$ where α is the current gain at f c/s, α_1 is the low-frequency current gain and f_c is the frequency at which the current gain has fallen to 3dB of its low frequency value. f_c is known as the alpha cut-off frequency and is the one quoted by manufacturers for transistors capable of h.f. operation. The values of the circuit elements in the equivalent circuit are dependent on frequency and their impedances are complex. These factors conspire to make the high-frequency transistor equivalent circuit at best only a rough approximation to the actual device over a given frequency band. Such a simplified equivalent circuit is shown in Fig. 31.

C_c implies a close coupling between input and output circuits which is undesirable and may lead to instability. Neutralisation is often employed to overcome this effect and may be achieved by a circuit of the type shown in Fig. 32.

The mode of operation of the neutralisation is better appreciated if the transistor equivalent circuit is employed and the circuit redrawn as in Fig. 33.

When the bridge is balanced i.e. $r_b/R_b = z_c/Z_c$, then variation in impedance of the tuned circuit does not influence the input conditions and vice versa.

In the case of the junction transistor the value of r_c and consequently R_c are so high as to have little shunting action on c_c and C_n, and R_c is usually omitted. The neutralised junction r.f. amplifier is then as shown in Fig. 34. For satisfactory operation R_b should be approximately equal to r_b.

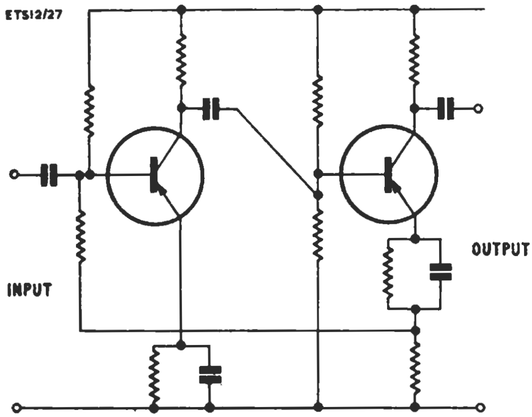
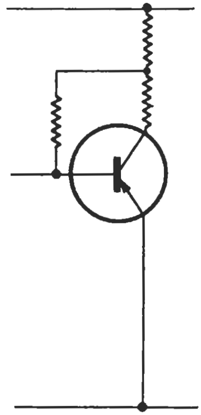


Figure 27.



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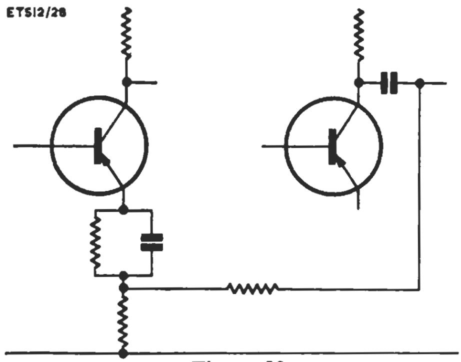

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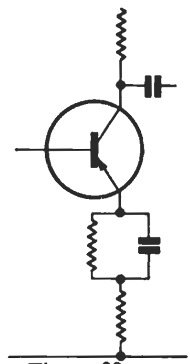

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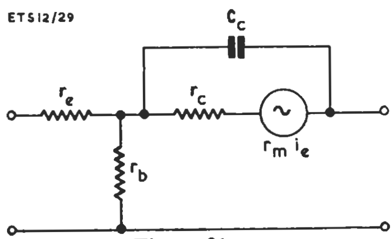

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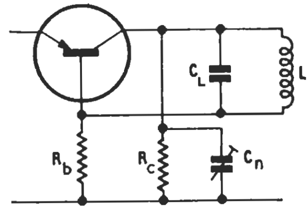

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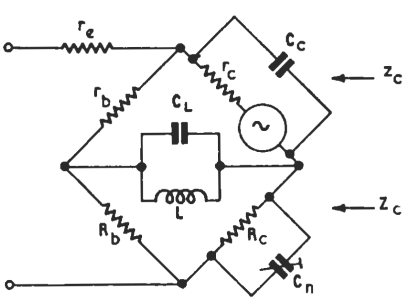

Figure 33.

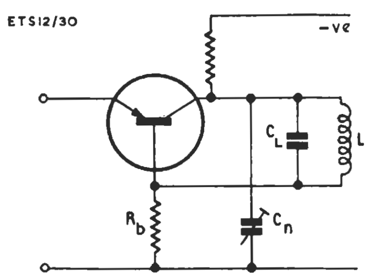

Figure 34.

Another difficulty arising from the use of transistors in r.f. amplifiers is the relatively low values of output and input resistances. This means that the tuned circuit must be tapped or loosely coupled to the transistor to preserve selectivity. A typical coupling arrangement is shown in Fig. 35.

A tap on the coil ensures the correct match to the 20 000–30 000 ohm collector output resistance whilst a tap between two chosen values of capacitance matches the base input of the next stage (1000–2000 ohms).

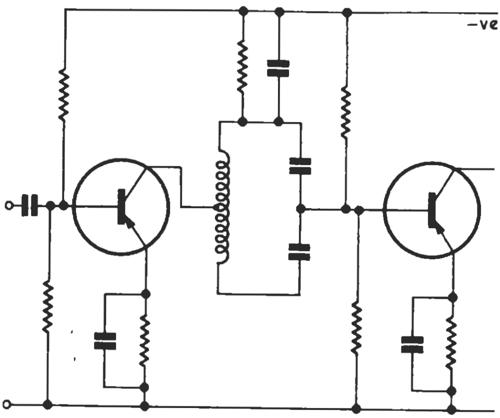

Figure 35.

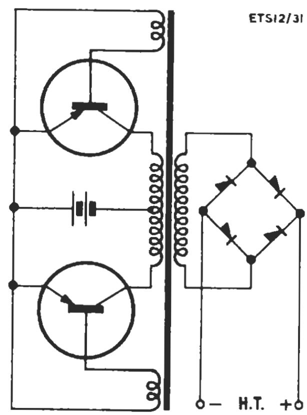

Figure 36.

Transistor Oscillators

The circuits employed with junction transistor oscillators are very similar to those employing valves. Two significant differences lie in the greater efficiency of the transistor as a d.c. to a.c. convertor and the operation from low voltage supplies. By rectification of the a.c. produced, d.c. voltages up to several kilovolts may be obtained. The h.t. battery in portable devices employing valves may be replaced by a small transistor d.c. converter operating from the l.t. battery.

A simplified circuit of such a device employed in a portable tape recorder is shown in Fig. 36.

Efficiencies of 90% can be achieved at a power output level of 0·5 watt.

Future developments

Various developments are in progress for increasing the useful operating frequency by a reduction in base thickness, and by the production of an electric field across the base to hasten the diffusion process.

Transistors are available now in the audio-frequency range delivering output powers of several watts with collector currents of several amperes. No doubt these powers will be considerably increased in future years. In the foreseeable future valves will certainly hold their own in really high-power radio transmitters and in the field of v.h.f. and u.h.f. techniques.

Appendix 1

GLOSSARY OF TERMS

Acceptor	A doping impurity containing 3 electrons in its outer orbits and leading to *p*-type material.
alpha (α)	The current gain from emitter to collector—always less than 1.
α'	The current gain from base to collector.
alpha cut-off frequency	The frequency at which the value of α has fallen 3dB below its low-frequency value.
avalanche	The build up of collector current due to carrier bombardment producing further carriers and leading to breakdown.
bottoming	A failure of the transistor to amplify due to very low collector voltage. (This is usually caused by a large collector current flowing through the collector load resistance.)
breakdown	The sudden rise of collector current which occurs when the collector voltage exceeds a critical value and leads to the destruction of the transistor unless limited.
common emitter, base, collector	The particular electrode which is common to input and output circuits. (Sometimes referred to as grounded or earthed.)
donor	A doping impurity with 5 electrons in its outer orbit and leading to *n*-type material.
doping	The addition of impurities to germanium, or silicon, to achieve an enhanced conductivity either by holes or excess electrons.
forward direction	Applying to a *p–n* junction. When the *p* region is positive relative to the *n* region, current flows for low applied voltages.
gallium	A doping element containing three outer orbital electrons; an acceptor impurity.
germanium	An element with four outer orbital electrons. It has co-valent bonds in its crystal structure and is the raw material for the majority of transistors.
hole	An electron deficiency; a co-valent bond which is short of one electron.
hole injection	The flow of holes from a *p*-type region into an *n*-type region.
hole storage	Retention of holes in an *n*-type region during hole flow which prolongs transit time and limits frequency response.

indium	An element used as an acceptor impurity. (See also gallium.)
intrinsic material	Pure material containing equal numbers of holes and excess electrons.
junction	The region of transition between *n*- and *p*-type regions of a doped semi-conducting crystal.
lattice	The symmetrical arrangement of atoms in a crystal.
recombination	The coincidence of excess electron and hole resulting in the removal of two current carriers; the acceptance by a deficient bond of an electron, with a release of energy.
semiconductor	A material depending for its slight conductivity on the presence of hole-electron pairs resulting from crystal bonds disrupted by thermal agitation.
stability factor	A factor which indicates the effect of temperature variation on collector current. (It should be as small as possible.)
short-circuit current gain.	The current gain for zero collector load resistance.
turnover voltage	See Zener voltage.
Zener voltage	The collector voltage which causes breakdown.

Appendix 2

BIAS STABILISATION

(a) Consider the circuit of Fig. 15, repeated here for convenience, and assume the base-emitter voltage to be negligible, usually of the order of 0·1 volt.

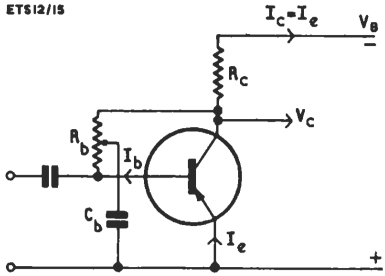

Figure 15. Bias circuit achieving stabilisation (1)

We have
$$I_e = I_b + I_c \qquad \text{(A1)}$$
$$I_c = I_{co} + aI_e \qquad \text{(A2)}$$
$$R_b I_b + R_c I_e = V_B \qquad \text{(A3)}$$

from these circuit equations an equation is required relating I_c to I_{co}.

from (A2)
$$I_e = \frac{I_c - I_{co}}{a} \qquad \text{(A4)}$$

from (A1)
$$I_b = I_e - I_c = \frac{I_c - I_{co}}{a} - I_c \qquad \text{(A5)}$$

from (A3)
$$R_b\left(\frac{I_c - I_{co}}{a} - I_c\right) + R_c\left(\frac{I_c - I_{co}}{a}\right) = V_B \qquad \text{(A6)}$$

Rearranging
$$I_c\,(R_b - a\,R_b + R_c) = I_{co}\,(R_c + R_b) + a\,V_B$$

i.e.
$$I_c = I_{co}\,\frac{R_c + R_b}{R_c + (1 - a)\,R_b} + \frac{V_B}{R_c + (1 - a)\,R_b}$$

The Stability factor, the slope of the graph of I_c against I_{co} is clearly:—

$$S = \frac{R_c + R_b}{R_c + (1 - a)\, R_b} \tag{A7}$$

(b) The circuit shown in Fig. 16 is repeated below.

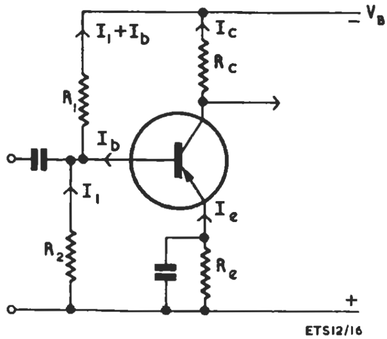

Figure16. Bias circuit achieving stabilisation (2)

With the same assumptions as before, the circuit equations are

$$I_e = I_b + I_c \tag{A8}$$

$$I_c = I_{co} + aI_e \tag{A9}$$

$$R_eI_e = R_2I_1 = V_B - R_1\,(I_1 + I_b) \qquad \text{(A10 and A11)}$$

Again an equation is required relating I_e to I_{co} .

From (A9) $$I_e = \frac{I_c - I_{co}}{a} \tag{A12}$$

therefore $$I_1 = \frac{R_e}{R_2} \cdot \frac{I_c - I_{co}}{a} \tag{A13}$$

$$I_b = I_e - I_c = \frac{I_c - I_{co}}{a} - I_c \tag{A14}$$

substituting in (A11)

$$R_e\left(\frac{I_c - I_{co}}{a}\right) = V_B - R_1\left(\frac{R_e}{R_2} \cdot \frac{(I_c - I_{co})}{a} + \frac{I_c - I_{co}}{a} - I_c\right)$$

rearranging

$$I_c\left(\frac{R_e}{R_1} + \frac{R_e}{R_2} + 1 - a\right) = I_{co}\left(\frac{R_e}{R_1} + \frac{R_e}{R_2} + 1\right) + a\,\frac{V_B}{R_1}$$

so that

$$I_c = I_{co} \frac{1 + R_e\left(\frac{1}{R_1} + \frac{1}{R_2}\right)}{1 - \alpha + R_e\left(\frac{1}{R_1} + \frac{1}{R_2}\right)} + \frac{\alpha V_B}{R_1\left[1 - \alpha + R_e\left(\frac{1}{R_1} + \frac{1}{R_2}\right)\right]}$$

The Stability factor S is the coefficient of I_{co} in the above expression

i.e.

$$S = \frac{1 + R_e\left(\frac{1}{R_1} + \frac{1}{R_2}\right)}{1 - \alpha + R_e\left(\frac{1}{R_1} + \frac{1}{R_2}\right)} \qquad \text{(A15)}$$

Writing

$$\frac{1}{R_1} + \frac{1}{R_2} = \frac{1}{R_b},$$

(A15) becomes

$$S = \frac{1 + R_e/R_b}{1 + R_e/R_b - \alpha} \qquad \text{(A16)}$$